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Internal Assessment Test 2 – Dec. 2025

Subject/Code: Microwave Engineering and Antenna Theory / BEC701				
Date:01/12/2025	Duration: 90 mins	Max. Marks: 50	Semester: 07	Branch: ECE

Sl.	Answer any FIVE FULL Questions	Marks	CO	RBT
1	Define the following parameters of antenna: (a) Radiation Pattern, (b) Directivity, (c) Gain, (d) Beam area, (e) Radiation Intensity.	10	CO3	L2
2	a. Derive an expression for directivity in terms of maximum effective aperture. b. A radio link has 150 watts transmitter connected to an antenna of 2 m^2 aperture at 2 GHz. The receiving antenna has an aperture of 1.5m^2 and is located at 10 km. Find the power delivered to the receiver.	5 5	CO3 CO3	L2 L3
3	Obtain the field pattern for a linear uniform array of 6 isotropic sources spaced $\lambda/2$ distance apart. The power is applied with equal amplitude and in phase. Also find HPBW and FNBW.	10	CO4	L3
4	a. Define and explain the principle of pattern multiplication. b. Find the directivity of a source with a sine squared pattern (doughnut power pattern)	5 5	CO4 CO4	L2 L3
5	Derive the expressions for far field components of a short dipole.	10	CO4	L2
6	Obtain the expression for radiation resistance of small loop antennas.	10	CO5	L2
7	Draw the structure of a pyramidal horn antenna. Use the principle of equality of path length and bring out the optimum horn dimension.	10	CO5	L2
8	Explain the Yagi Uda Array antenna with necessary equations and diagrams.	10	CO5	L2

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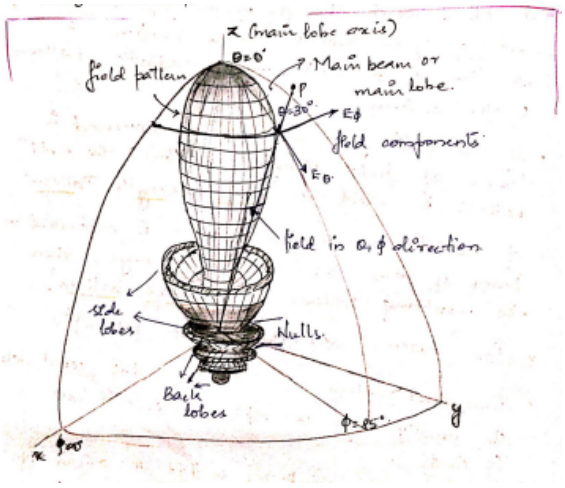
Scheme of evaluation:

Q. No	Contents	Marks
1	a. Radiation Pattern	2
1	b. Directivity	2
1	c. Gain	2
1	d. Beam Area	2
1	e. Radiation Intensity	2
2	a. Definition + Derivation	2+3=5
2	b. Problem (Approach + final Answer)	3+2=5
3	Problem: Formula Approach Answer Diagram	3+3+2+2=10
4	a. Pattern Multiplication (Diagram + explanation)	2+3=5
4	b. Diagram+ Directivity Derivation	2+3=5
5	Derivation (Short Dipole Antenna): Diagram Expressions Far field components	1+4+5=10
6	Derivation (Loop Antenna): Diagram Expressions Radiation Resistance	1+4+5=10
7	Pyramidal Horn: Diagram Principle of equality of path length Optimum horn dimension	2+5+3=10
8	Yagi Uda array: Diagram Explanation Pattern	2+5+3=10

Solution:

1. a) Radiation pattern:



* Radiation pattern of an antenna is nothing but a graph which shows the variation in actual field strength of electromagnetic field at all points which are at equal distance from the antenna. Obviously, the graph of radiation pattern will be 3-dimensional & hence can't be completely represented on a plain paper.
* In other words, the graphical representation of radiation of an antenna as a function of direction is given the name radiation pattern of the antenna.

b) Directivity

Directivity of an antenna = ratio of maximum power density $P(\theta,\phi)_{max}$ (watts/m²) to its average value over a sphere as observed in far field of an antenna.

Thus, $D = \dfrac{P(\theta,\phi)_{max}}{P(\theta,\phi)_{av}}$ (Directivity from pattern) ——(1.25)

The average power density over a sphere is given by:

$$P(\theta,\phi)_{av} = \frac{1}{4\pi}\int_{\phi=0}^{\phi=2\pi}\int_{\theta=0}^{\theta=\pi} P(\theta,\phi)\sin\theta\, d\theta\, d\phi$$

$$= \frac{1}{4\pi}\iint_{4\pi} P(\theta,\phi)\, d\Omega \quad (W/s_r) \qquad ——(1.26)$$

$$\therefore \text{Directivity, } D = \frac{P(\theta,\phi)_{max}}{\frac{1}{4\pi}\iint_{4\pi} P(\theta,\phi)\, d\Omega} = \frac{1}{\frac{1}{4\pi}\iint_{4\pi}\left[P(\theta,\phi)/P(\theta,\phi)_{max}\right] d\Omega} \qquad ——(1.27)$$

and

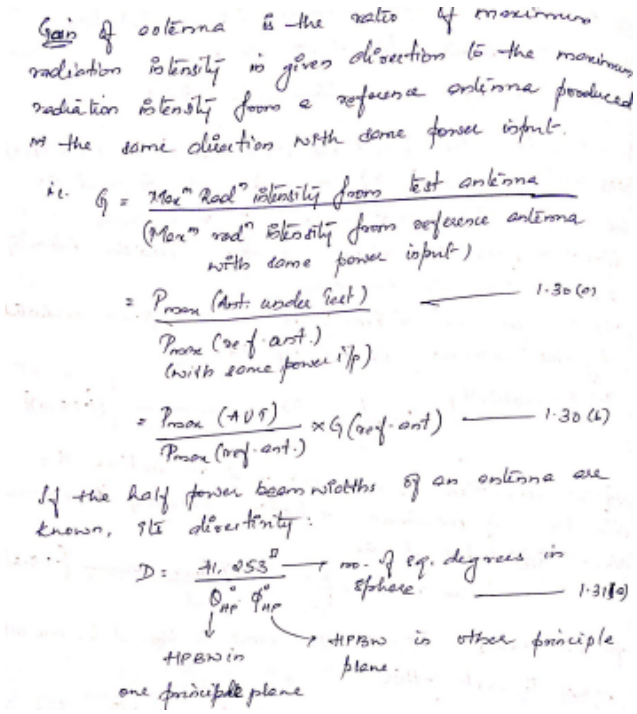
$$D = \frac{4\pi}{\iint_{4\pi} P_n(\theta,\phi)\, d\Omega} = \frac{4\pi}{\Omega_A}$$ (Directivity from beam area Ω_A) ——(1.28)

↳ Normalized power pattern

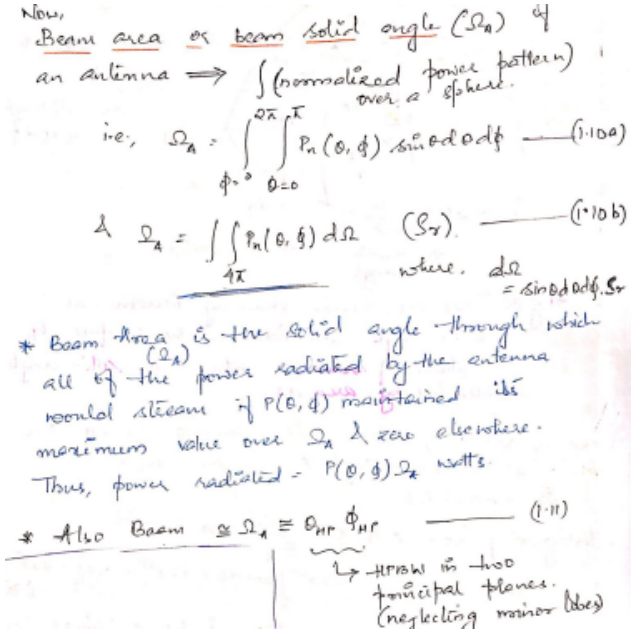
Thus, the directivity is the ratio of the area of a sphere (4π sr) to the beam area Ω_A of the antenna.

Smaller the beam area, larger the directivity D.

c) Gain:



d) Beam Area:



e) Radiation Intensity:

η_0.

If magnitude of Poynting vector P_r is multiplied by the square of radius r at which radiation intensity is measured, then we get the power per unit solid angle i.e. radiation intensity.

$\therefore$ Radiation intensity $\equiv U(\theta, \phi) = P_r \cdot r^2 \equiv U$ ——(1.17a)

or $U = P_r \cdot r^2$ ——(1.17b)

or $U = \frac{1}{2} \frac{E^2(\theta, \phi) r^2}{\eta_0}$ ——(1.17c)

and $U = \frac{\text{Differential power } (dW/r)}{\text{Differential element of solid angle } (d\Omega)}$ ——(1.18)

2. a) Directivity in terms of effective aperture:

Let the Poynting vector (or power density) of the plane wave be S watts per sq. meter and the area (physical aperture) of the horn be A_p m^2.

If horn extracts all the power from the wave over its entire physical aperture, then the total power (P) absorbed from the wave is:

$$P = \frac{E^2}{2} A_p = S A_p \quad \text{——(1.32)}$$

Thus the horn is regarded as having an aperture, the total power it extracts from a passing wave being proportional to the aperture or area of its mouth.

The field response of the horn is not uniform across the aperture A_p since $\vec{E}$ at the sidewalls should be ideally zero (PPP).

True effective aperture (A_e) of the horn is < physical aperture (A_p) of

$\therefore$ Aperture efficiency, $\eta_{ap} = \frac{A_e}{A_p}$ ——(1.33)

(dimensionless)

Consider an antenna with effective aperture (A_e) which radiates all of its power in a conical pattern of beam area Ω_A

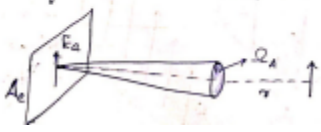


Fig (7) Radiation over beam area Ω_A from A_e

Assuming uniform field E_a over the aperture, the power radiated is:

$$P = \frac{E_a^2}{Z_0} A_e \quad \text{(Watts)} \qquad \text{——— (1.34(a))}$$

Z_0 → intrinsic impedance of air (or vacuum) (377Ω)

Consider a uniform field E_r in the far field at a distance r, the power radiated is given by:

$$P = \frac{E_r^2}{Z_0} r^2 \Omega_A \quad \text{(Watts)} \qquad \text{——— (1.34(b))}$$

~~Also gain $G = \frac{4\pi A_e}{\lambda^2}$~~

Equating 1.34(a) & (b) & noting that

$$E_r = \frac{E_a A_e}{r\lambda} \qquad \text{——— (1.35)}$$

we get:

$$E_r^2 = \frac{E_a^2 A_e}{r^2 \Omega_A} = \frac{E_a^2 A_e^2}{r^2 \lambda^2}$$

$$\Rightarrow \frac{A_e}{\lambda^2} = \frac{1}{\Omega_A} \Rightarrow \boxed{\lambda^2 = A_e \Omega_A} \quad (m^2) \qquad \text{——— 1.36}$$

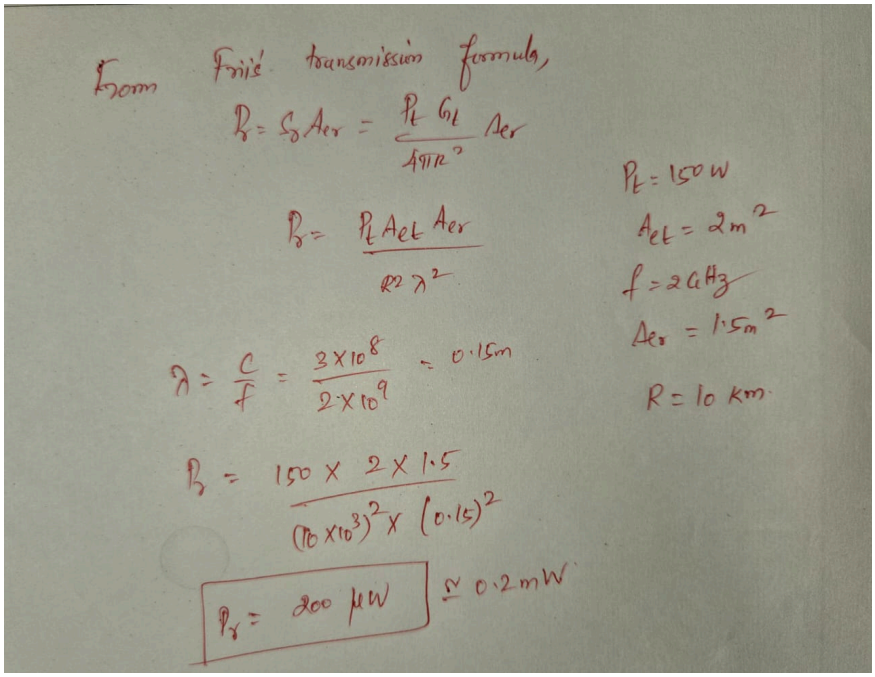
→ Antenna-beam area relation

Thus, from (1.34) & (1.28),

$$\boxed{D = \frac{4\pi}{\Omega_A} = \frac{4\pi A_e}{\lambda^2}} \qquad \text{——— 1.37}$$

→ Directivity from aperture

b) Problem:



3) Linear Uniform Array Problem:

Obtain the field pattern for a linear uniform array of 6 isotropic sources spaced $\lambda/2$ distance apart. The power is applied with equal amplitude and in phase. Also find HPBW and FNBW.

Array factor (field pattern)

Let θ be the angle from the array axis (broadside at $\theta = 90°$).

Define $\psi = kd\cos\theta$ with $k = 2\pi/\lambda$. Since $d = \lambda/2$,

$$kd = \pi \quad \Rightarrow \quad \psi = \pi\cos\theta.$$

The array factor is

$$AF(\theta) = \sum_{n=0}^{5} e^{jn\psi} = e^{j\frac{5\psi}{2}}\,\frac{\sin\left(\frac{6\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)} = e^{j\frac{5\psi}{2}}\,\frac{\sin(3\psi)}{\sin(\psi/2)}.$$

Magnitude (normalized by the maximum $AF_{\max} = N = 6$) is

$$E_{norm}(\theta) = \frac{\left|\sin(3\psi)\right|}{6\left|\sin(\psi/2)\right|}, \qquad \psi = \pi\cos\theta.$$

(The phase factor $e^{j5\psi/2}$ only shifts phase; it does not affect the magnitude.)

The broadside maximum occurs at $\psi = 0$ (i.e. $\cos\theta = 0 \Rightarrow \theta = 90°$), where the limiting value of the ratio is 6.

First-null beamwidth (FNBW)

Nulls occur where the numerator $\sin(3\psi) = 0 \Rightarrow 3\psi = m\pi$. The first nulls adjacent to the main lobe are for $m = \pm 1$:

$$\psi = \pm\frac{\pi}{3} \quad \Rightarrow \quad \pi\cos\theta = \pm\frac{\pi}{3}$$

so

$$\cos\theta = \pm\frac{1}{3}.$$

Hence the first-null angles are

$$\theta_1 = \cos^{-1}\left(\tfrac{1}{3}\right) \approx 70.5288°, \quad \theta_2 = \cos^{-1}\left(-\tfrac{1}{3}\right) \approx 109.4712°.$$

Therefore the **FNBW**

$$\text{FNBW} = \theta_2 - \theta_1 \approx 109.4712° - 70.5288° \approx 38.94°.$$

Half-power beamwidth (HPBW)

Solve for angles where $|AF(\theta)| = 6/\sqrt{2}$ (half-power points) near the broadside main-lobe.

Numerically solving that equation gives the two half-power angles:

$$\theta_{HP1} \approx 81.4049°, \quad \theta_{HP2} \approx 98.5951°$$

so the **HPBW**

$$\text{HPBW} = \theta_{HP2} - \theta_{HP1} \approx 17.19°.$$

4) a) Pattern Multiplication:

The field pattern of an array of nonisotropic but similar point sources is the product of the pattern of the individual source and the pattern of an array of isotropic point sources having the same locations, relative amplitudes, and phase as the nonisotropic point sources.

This principle may be applied to arrays of any number of sources provided only that they are similar. The individual nonisotropic source or antenna may be of finite size but can be considered as a point source situated at the point in the antenna to which phase is referred. This point is said to be the "phase center."

The above discussion of pattern multiplication has been concerned only with the field pattern or magnitude of the field. If the field of the nonisotropic source and the array of isotropic sources vary in phase with space angle, i.e., have a phase pattern which is not a constant, the statement of the principle of pattern multiplication may be extended to include this more general case as follows:

> The total field pattern of an array of nonisotropic but similar sources is the product of the individual source pattern and the pattern of an array of isotropic point sources each located at the phase center of the individual source and having the same relative amplitude and phase, while the total phase pattern is the sum of the phase patterns of the individual source and the array of isotropic point sources.

The total phase pattern is referred to the phase center of the array. In symbols, the total field E is then

$$E = \underbrace{f(\theta, \phi)F(\theta, \phi)}_{\text{Field pattern}} \underbrace{/f_p(\theta, \phi) + F_p(\theta, \phi)}_{\text{Phase pattern}} \tag{4}$$

where

$f(\theta, \phi)$ = field pattern of individual source
$f_p(\theta, \phi)$ = phase pattern of individual source
$F(\theta, \phi)$ = field pattern of array of isotropic sources
$F_p(\theta, \phi)$ = phase pattern of array of isotropic sources

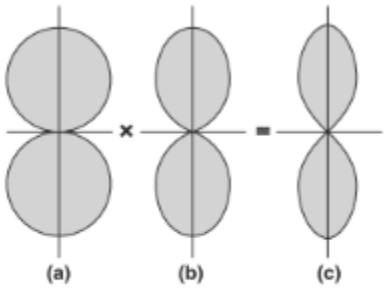


Figure 5–22 Example of pattern multiplication. Two nonisotropic but identical point sources of the same amplitude and phase, spaced $\lambda/2$ apart and arranged as in Fig. 5–21, produce the pattern shown at (*c*). The individual source has the pattern shown at (*a*), which, when multiplied by the pattern of an array of two isotropic point sources (of the same amplitude and phase) as shown at (*b*), yields the total array pattern of (*c*).

b) Directivity of sine squared pattern:

EXAMPLE 5–6.4 Source with Sine-Squared (Doughnut) Power Pattern

A source has a sine-squared radiation-intensity power pattern. The radiation-intensity pattern is given by

$$U = U_m \sin^2\theta \tag{8}$$

The power pattern is shown in Fig. 5–7*a*. This type of pattern is of considerable interest because it is the pattern produced by a short dipole coincident with the polar ($\theta = 0$) axis in Fig. 5–7*a*. Applying (5–4–3), the total power radiated is

$$P = U_m \int_0^{2\pi} \int_0^{\pi} \sin^3\theta \, d\theta \, d\phi = \frac{8}{3}\pi U_m \tag{9}$$

If P is the same as for the isotropic source,

$$\frac{8}{3}\pi U_m = 4\pi U_0$$

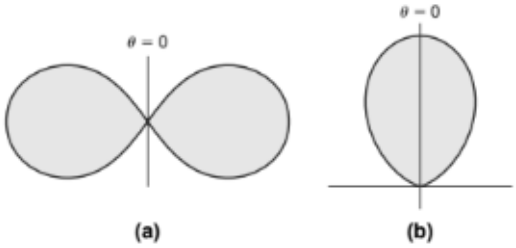


Figure 5–7 (*a*) Sine-squared power pattern and (*b*) unidirectional cosine-squared power pattern.

and

$$\text{Directivity} = \frac{U_m}{U_0} = \frac{3}{2} = 1.5 = D \qquad \textit{Ans.} \tag{10}$$

5) Far field components of a short dipole:

Let us now proceed to find the fields everywhere around a short dipole. Let the dipole of length L be placed coincident with the z axis and with its center at the origin as in Fig. 6–2. The relation of the electric field components, E_r, E_θ and E_ϕ, is then as shown. It is assumed that the medium surrounding the dipole is air or vacuum.

In dealing with antennas or radiating systems, the propagation time is a matter of great importance. Thus, if a current is flowing in the short dipole of Fig. 6–3, the effect of the current is not felt instantaneously at the point P, but only after an interval equal to the time required for the disturbance to propagate over the distance r. We have already recognized this in Chap. 5 in connection with the pattern of arrays of point sources, but here we are more explicit and describe it as a *retardation* effect.

Accordingly, instead of writing the current I as[1]

$$I = I_0 e^{j\omega t} \tag{1}$$

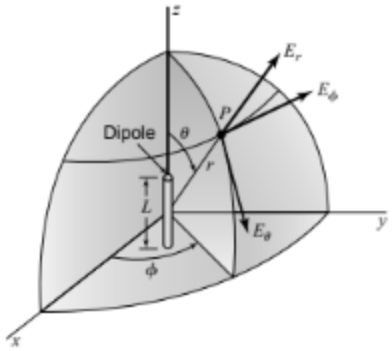


Figure 6–2 Relation of dipole to coordinates.

$$[I] = I_0 e^{j\omega[t-(r/c)]} \tag{2}$$

where $[I]$ is called the *retarded current*. Specifically, the retardation time r/c results in a *phase retardation* $\omega r/c = 2\pi f r/c$ radians $= 360° \; fr/c = 360° \; t/T$, where $T = 1/f =$ time of one period or cycle (seconds) and $f =$ frequency (hertz, Hz = cycles per second). The brackets may be added as in (2) to indicate explicitly that the effect of the current is retarded.

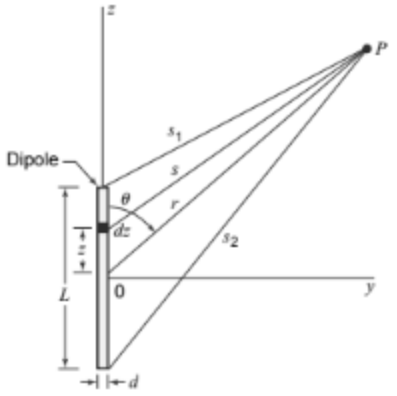


Figure 6–3a Geometry for short dipole.

$$A_z = \frac{\mu_0}{4\pi} \int_{-L/2}^{L/2} \frac{[I]}{s} \, dz \tag{2}$$

where $[I]$ is the retarded current given by

$$[I] = I_0 e^{j\omega[t-(s/c)]} \tag{3a}$$

In (3) and (3a),

$z =$ distance to a point on the conductor
$I_0 =$ peak value in time of current (uniform along dipole)
$\mu_0 =$ permeability of free space $= 4\pi \times 10^{-7}$ H m^{-1}

If the distance from the dipole is large compared to its length ($r \gg L$) and if the wavelength is large compared to the length ($\lambda \gg L$), we can put $s = r$ and neglect the phase differences of the field contributions from different parts of the wire. The integrand in (3) can then be regarded as a constant, so that (3) becomes

$$A_z = \frac{\mu_0 L I_0 e^{j\omega[t-(r/c)]}}{4\pi r} \tag{4}$$

The retarded scalar potential V of a charge distribution is

$$V = \frac{1}{4\pi\varepsilon_0} \int_V \frac{[\rho]}{s} \, d\tau \tag{5}$$

where $[\rho]$ is the retarded charge density given by

$$[\rho] = \rho_0 e^{j\omega[t-(s/c)]} \tag{6}$$

and $d\tau$ = infinitesimal volume element

ε_0 = permittivity or dielectric constant of free space = 8.85×10^{-12} F m^{-1}

Since the region of charge in the case of the dipole being considered is confined to the points at the ends as in Fig. 6–1*b*, (5) reduces to

$$V = \frac{1}{4\pi\varepsilon_0}\left\{\frac{[q]}{s_1} - \frac{[q]}{s_2}\right\} \tag{7}$$

From (6–1–1) and (3a),

$$[q] = \int [I]\,dt = I_0 \int e^{j\omega[t-(s/c)]}\,dt = \frac{[I]}{j\omega} \tag{8}$$

Substituting (8) into (7),

$$V = \frac{I_0}{4\pi\varepsilon_0 j\omega}\left[\frac{e^{j\omega[t-(s_1/c)]}}{s_1} - \frac{e^{j\omega[t-(s_2/c)]}}{s_2}\right] \tag{9}$$

Referring to Fig. 6–3*b*, when $r \gg L$, the lines connecting the ends of the dipole and the point P may be considered as parallel so that

$$s_1 = r - \frac{L}{2}\cos\theta \tag{10}$$

and

$$s_2 = r + \frac{L}{2}\cos\theta \tag{11}$$

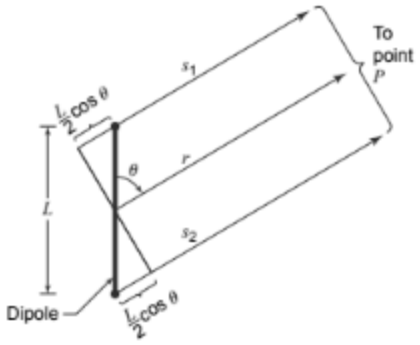


Figure 6–3b Relations for short dipole when $r \gg L$.

Substituting (10) and (11) into (9), it may be shown that the fields of a short electric dipole are:

Electric fields of short dipole — ***General case***

$$E_r = \frac{I_0 L\cos\theta\, e^{j\omega[t-(r/c)]}}{2\pi\varepsilon_0}\left(\frac{1}{cr^2} + \frac{1}{j\omega r^3}\right) \tag{12}$$

$$E_\theta = \frac{I_0 L\sin\theta\, e^{j\omega[t-(r/c)]}}{4\pi\varepsilon_0}\left(\frac{j\omega}{c^2 r} + \frac{1}{cr^2} + \frac{1}{j\omega r^3}\right) \tag{13}$$

Electric and magnetic fields of short dipole — ***Far-field case***

$$E_\theta = \frac{j\omega I_0 L\sin\theta\, e^{j\omega[t-(r/c)]}}{4\pi\varepsilon_0 c^2 r} = j\frac{I_0\beta L}{4\pi\varepsilon_0 cr}\sin\theta\, e^{j\omega[t-(r/c)]} \tag{17}$$

$$H_\phi = \frac{j\omega I_0 L\sin\theta\, e^{j\omega[t-(r/c)]}}{4\pi cr} = j\frac{I_0\beta L}{4\pi r}\sin\theta\, e^{j\omega[t-(r/c)]} \tag{18}$$

Taking the ratio of E_θ to H_ϕ as given by (17) and (18), we obtain

$$\frac{E_\theta}{H_\phi} = \frac{1}{\varepsilon_0 c} = \sqrt{\frac{\mu_0}{\varepsilon_0}} = 376.7\ \Omega \qquad \textit{Impedance of space} \tag{19}$$

6) Radiation Resistance of small loop antenna:

To find the radiation resistance of a loop antenna, the Poynting vector is integrated over a large sphere yielding the total power P radiated. This power is then equated to the square of the effective current on the loop times the radiation resistance R_r (Foster-1).

$$P = \frac{I_0^2}{2} R_r \tag{1}$$

where I_0 = peak current in time on the loop. The radiation resistance so obtained is the value which would appear at the loop terminals connected to the twin-line, as in Fig. 7–9*a* and coaxial line as in Fig. 7–9*b*. It is assumed that the current is uniform and in phase for any radius a, this condition being obtained by means of phase shifters, multiple feeds or other devices (see Fig. 7–8). The average Poynting vector of a far field is given by

$$S_r = \tfrac{1}{2}|H|^2 \operatorname{Re} Z \tag{2}$$

where $|H|$ is the absolute value of the magnetic field and Z is the intrinsic impedance of the medium, which in this case is free space. Substituting the absolute value of H_θ from (7–4–10) for $|H|$ in (2) yields

$$S_r = \frac{15\pi(\beta a I_0)^2}{r^2} J_1^2(\beta a \sin\theta) \tag{3}$$

The total power radiated P is the integral of S_r over a large sphere; that is,

$$P = \iint S_r \, ds = 15\pi(\beta a I_0)^2 \int_0^{2\pi} \int_0^{\pi} J_1^2(\beta a \sin\theta) \sin\theta \, d\theta \, d\phi \tag{4}$$

$$P = 30\pi^2(\beta a I_0)^2 \int_0^{\pi} J_1^2(\beta a \sin\theta) \sin\theta \, d\theta \tag{5}$$

In the case of a loop that is small in terms of wavelengths, the approximation of (7–6–1) can be applied. Thus (5) reduces to

$$P = \frac{15}{2}\pi^2(\beta a)^4 I_0^2 \int_0^{\pi} \sin^3\theta \, d\theta = 10\pi^2\beta^4 a^4 I_0^2 \tag{6}$$

Since the area $A = \pi a^2$, (6) becomes

$$P = 10\beta^4 A^2 I_0^2 \tag{7}$$

Assuming no antenna losses, this power equals the power delivered to the loop terminals as given by (1). Therefore,

$$R_r \frac{I_0^2}{2} = 10\beta^4 A^2 I_0^2 \tag{8}$$

and

Small loop $\quad R_r = 31{,}171 \left(\frac{A}{\lambda^2}\right)^2 = 197 C_\lambda^4 \quad (\Omega)$ ***Radiation resistance*** (9)

or $\quad R_r \simeq 31{,}200 \left(\frac{A}{\lambda^2}\right)^2 \quad (\Omega)$ (10)

This is the radiation resistance of a small single-turn loop antenna, circular or square, with uniform in-phase current. The relation is about 2 percent in error when the loop perimeter is $\lambda/3$. A circular loop of this perimeter has a diameter of about $\lambda/10$. Its radiation resistance by (10) is nearly 2.5 Ω.

The radiation resistance of a small loop consisting of one or more turns is given by (Alford-1),

Small loop $\quad R_r = 31{,}200 \left(n\frac{A}{\lambda^2}\right)^2 \quad (\Omega)$ ***Radiation resistance*** (10*a*)

where n = number of turns

7) Pyramidal Horn Antenna:

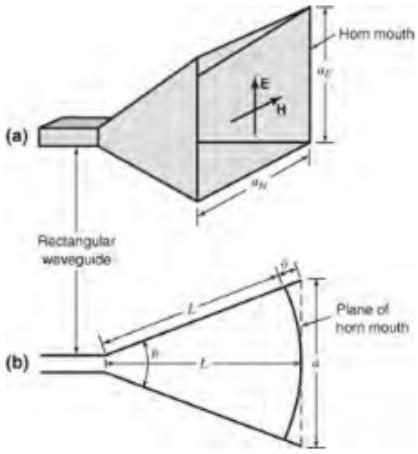


Figure 7–41 (*a*) Pyramidal horn antenna. (*b*) Cross section with dimensions used in analysis. The diagram can be for either E-plane or H-plane cross sections. For the E plane the flare angle is θ_E and aperture a_E. For the *H* plane the flare angle is θ_H and the aperture a_H. See Fig. 7–42.

The *principle of equality of path length (Fermat's principle)* is applicable to the horn design but with a different emphasis. Instead of requiring a constant phase across the horn mouth, the requirement is relaxed to one where the phase may deviate, but by less than a specified amount δ, equal to the path length difference between a ray traveling along the side and along the axis of the horn.

From Fig. 7–41,

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

$$\sin\frac{\theta}{2} = \frac{a}{2(L+\delta)} \tag{2}$$

$$\tan\frac{\theta}{2} = \frac{a}{2L} \tag{3}$$

where

θ = flare angle (θ_E for E plane, θ_H for H plane), deg
a = aperture (a_E for E plane, a_H for H plane), m
L = horn length, m
δ = path length difference, m

From the geometry we have also that

$$L = \frac{a^2}{8\delta} \qquad (\delta \ll L) \tag{4}$$

and

$$\theta = 2\tan^{-1}\frac{a}{2L} = 2\cos^{-1}\frac{L}{L+\delta} \tag{5}$$

Optimum horn dimensions

$$\delta_0 = \frac{L}{\cos(\theta/2)} - L = \text{optimum } \delta \tag{6}$$

$$L = \frac{\delta_0\cos(\theta/2)}{1-\cos(\theta/2)} = \text{optimum length} \tag{7}$$

8) Yagi Uda Array:

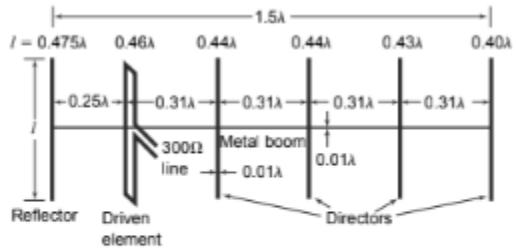


Figure 8–31 Modern-version 6-element Yagi-Uda antenna with dimensions. It has a maximum directivity of about 12 dBi at the center of a bandwidth of 10 percent at half-power.

A typical modern-version 6-elementYagi-Uda antenna is shown in Fig. 8–31. It consists of a driven element (folded λ/2 dipole) fed by a 300- 2-wire transmission line (twin line), a reflector and 4 directors. Dimensions (lengths and spacings) are indicated on the figure. The antenna provides a gain of about 10 dBi (maximum) with a bandwidth at half-power of 10 percent. By adjusting lengths and spacings appropriately (tweeking), the dimensions can be optimized, producing an increase in gain of another decibel (Chen-1, 2; Viezbicke-1).
However, the dimensions are critical.
The inherently narrow bandwidth of the Yagi-Uda antenna can be broadened to 1.5 to 1 by lengthening the reflector to improve operation at low frequencies and shortening the directors to improve high-frequency operation. However, this is accomplished at a sacrifice in gain of as much as 5 dB.

For the 1.5 λ array of Fig. 8–31, what is (*a*) HPBW, (*b*) axial ratio, (*c*) gain and (*d*) the pattern?

■ **Solution**

(*a*) HPBW = 44° in plane of elements, HPBW = 64° in plane perpendicular to elements (from pattern).
(*b*) AR = infinite (pure linear polarization).
(*c*) Gain = 9.4 dBi by pattern integration.
(*d*) Pattern is shown in Fig. 8–32.

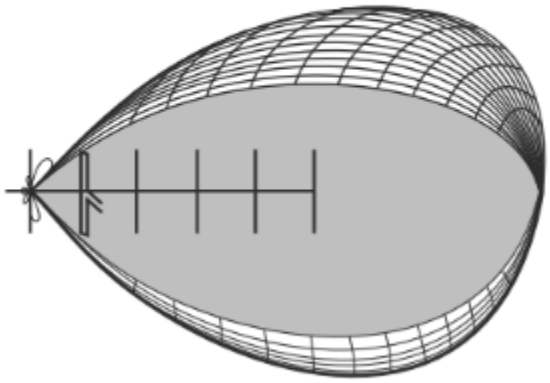


Figure 8–32 Power pattern of the Yagi-Uda array of Fig. 8–31. The narrower pattern is in the plane of the elements.