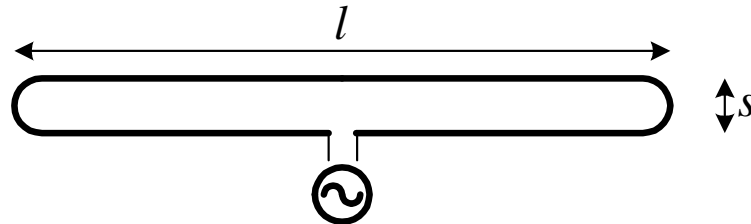


LECTURE 11: Practical Dipole/Monopole Geometries. Matching Techniques for Dipole/Monopole Feeds

(The folded dipole antenna. Conical skirt monopoles. Sleeve antennas. Turnstile antenna. Impedance matching techniques. Dipoles with traps.)

1. Folded Dipoles

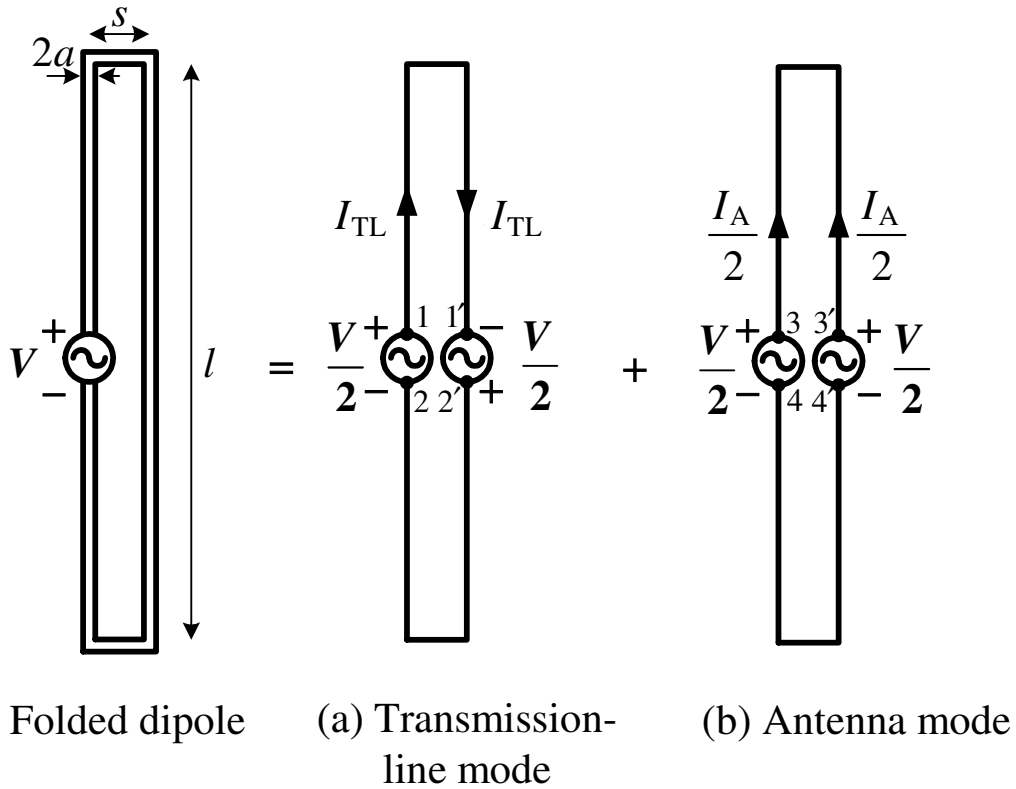


The folded dipole is a popular antenna for reception of TV broadcast signals. It has essentially the same radiation pattern as the dipole of the same length l but it provides four times greater input impedance when $l \approx \lambda / 2$. The input resistance of the conventional half-wavelength dipole is $R_{in} \approx 73 \Omega$ while that of the half-wavelength folded dipole is about 292Ω . Wire antennas do not fit well with coaxial feed lines because of the different field mode; thus, balanced-to-unbalanced transition is required. However, they are ideally suited for twin-lead (two-wire) feed lines. These lines (two parallel thin wires separated by a distance of about 8 to 10 mm) have characteristic impedance $Z_0 \approx 300 \Omega$. Therefore, an input antenna impedance of $(4 \times 73) \Omega$ matches well the 2-wire feed lines. The separation distance between the two wires of the folded dipole s should not exceed 0.05λ .

The folded dipole can be analyzed by decomposing its current into two modes: the transmission-line (TL) mode (or odd mode) and the antenna mode (or even mode). This analysis, although approximate¹, illustrates the four-fold impedance transformation.

In the TL mode (see Figure (a) below), the source terminals $1-2'$ and $2-1'$ are at the same potential and can be connected by a short without changing the mode of operation. Each of the two 2-wire transmission lines is fed by a voltage source $V / 2$ and is terminated by a short.

¹ G.A. Thiele, E.P. Ekelman, Jr., L.W. Henderson, "On the accuracy of the transmission line model for the folded dipole," *IEEE Trans. on Antennas and Propagation*, vol. AP-28, No. 5, pp. 700-703, Sep. 1980.



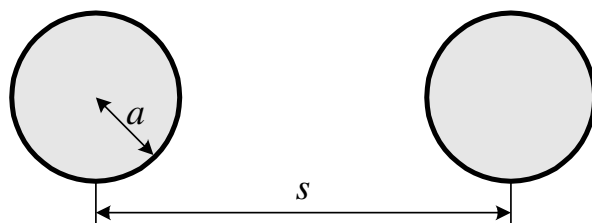
The input impedance of each shorted transmission line of length $l/2$ is

$$Z_{\text{TL}} = \left[Z_0 \left(\frac{Z_L + jZ_0 \tan(\beta l / 2)}{Z_0 + jZ_L \tan(\beta l / 2)} \right) \right]_{Z_L=0}, \quad (11.1)$$

$$\Rightarrow Z_{\text{TL}} = jZ_0 \tan\left(\frac{\beta l}{2}\right). \quad (11.2)$$

Here, Z_0 is the characteristic impedance of the 2-wire transmission line formed by the two segments of the folded wire. It can be calculated as

$$Z_0 = \frac{\eta}{\pi} \operatorname{arccosh}\left(\frac{s}{2a}\right) = \frac{\eta}{\pi} \ln \left[\frac{s/2 + \sqrt{(s/2)^2 - a^2}}{a} \right]. \quad (11.3)$$



Usually, the folded dipole has a length of $l \approx \lambda / 2$. Then,

$$Z_{\text{TL}(l=\lambda/2)} = jZ_0 \tan(\pi / 2) \rightarrow \infty. \quad (11.4)$$

If $l \neq \lambda / 2$, the more general expression (11.2) should be used. On each dipole side, the current in the TL mode is

$$I_{\text{TL}} = \frac{V}{2Z_{\text{TL}}}. \quad (11.5)$$

We now consider the antenna mode. The generators' terminals 3–3' (and 4–4') are with identical potentials. Therefore, they can be connected electrically without changing the conditions of operation. The following assumption is made: an equivalent dipole of effective radius

$$a_e = \sqrt{as} \quad (11.6)$$

is radiating excited by $V / 2$ voltage. Since usually $a \ll \lambda$ and $s \ll \lambda$, the input impedance in the antenna mode Z_A is assumed equal to the input impedance of an infinitesimally thin dipole of the respective length l . If $l = \lambda / 2$, then $R_A = 73 \ \Omega$. The current in the antenna mode is

$$I_A = \frac{V}{2Z_A}. \quad (11.7)$$

The current on each arm of the equivalent dipole is

$$\frac{I_A}{2} = \frac{V}{4Z_A}. \quad (11.8)$$

The total current of a folded dipole is obtained by superimposing both modes. At the antenna input (left arms)

$$I_{\text{in}} = I_{\text{TL}} + \frac{I_A}{2} = V \left(\frac{1}{2Z_{\text{TL}}} + \frac{1}{4Z_A} \right), \quad (11.9)$$

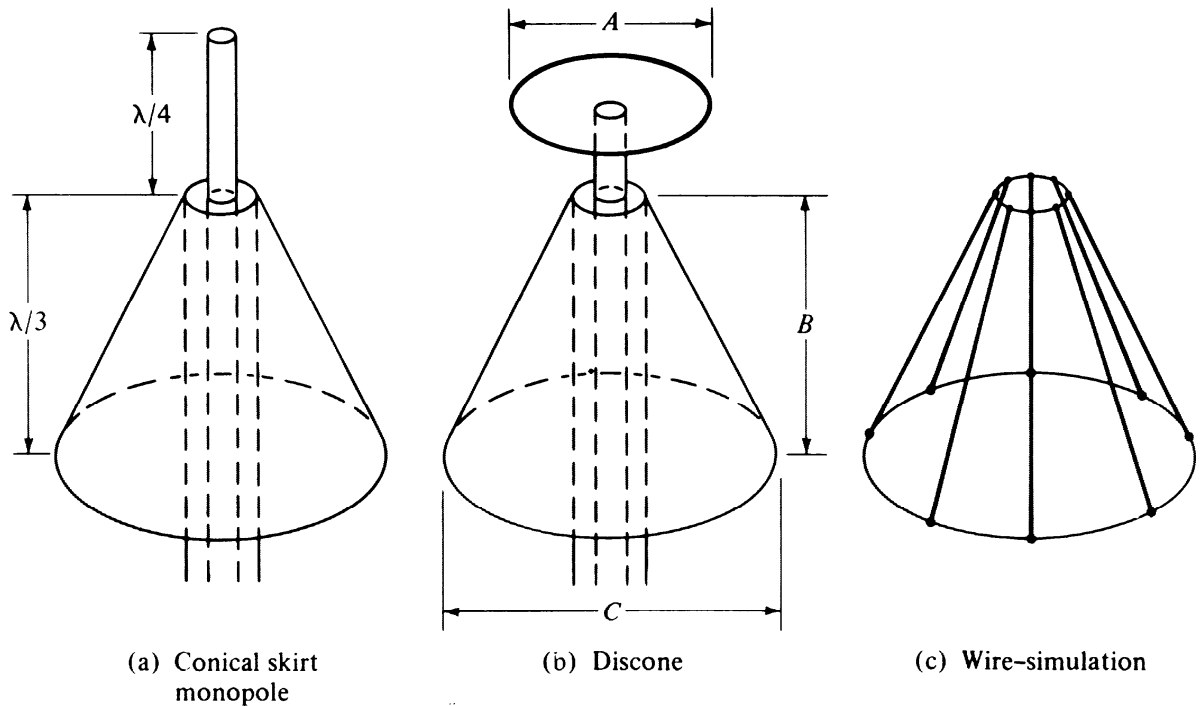
$$\Rightarrow Z_{\text{in}} = \frac{V}{I_{\text{in}}} = \frac{4Z_{\text{TL}}Z_A}{2Z_A + Z_{\text{TL}}}. \quad (11.10)$$

When $l = \lambda / 2$ (half-wavelength folded dipole), then $Z_{\text{TL}} \rightarrow \infty$, and

$$\text{Re } Z_{\text{in}} = 4 \text{Re } Z_A |_{l=\lambda/2} \approx 292 \ \Omega. \quad (11.11)$$

Thus, the half-wavelength folded dipole is well suited for direct connection to a twin-lead line of $Z_0 \approx 300 \Omega$. It is often made in a simple way: a suitable portion (the end part of the twin-lead cable of length $l = \lambda/2$) is cut along its mid-line so that it is split into two wires, which are bent and soldered into a folded dipole.

2. Conical (Skirt) Monopoles and Discones

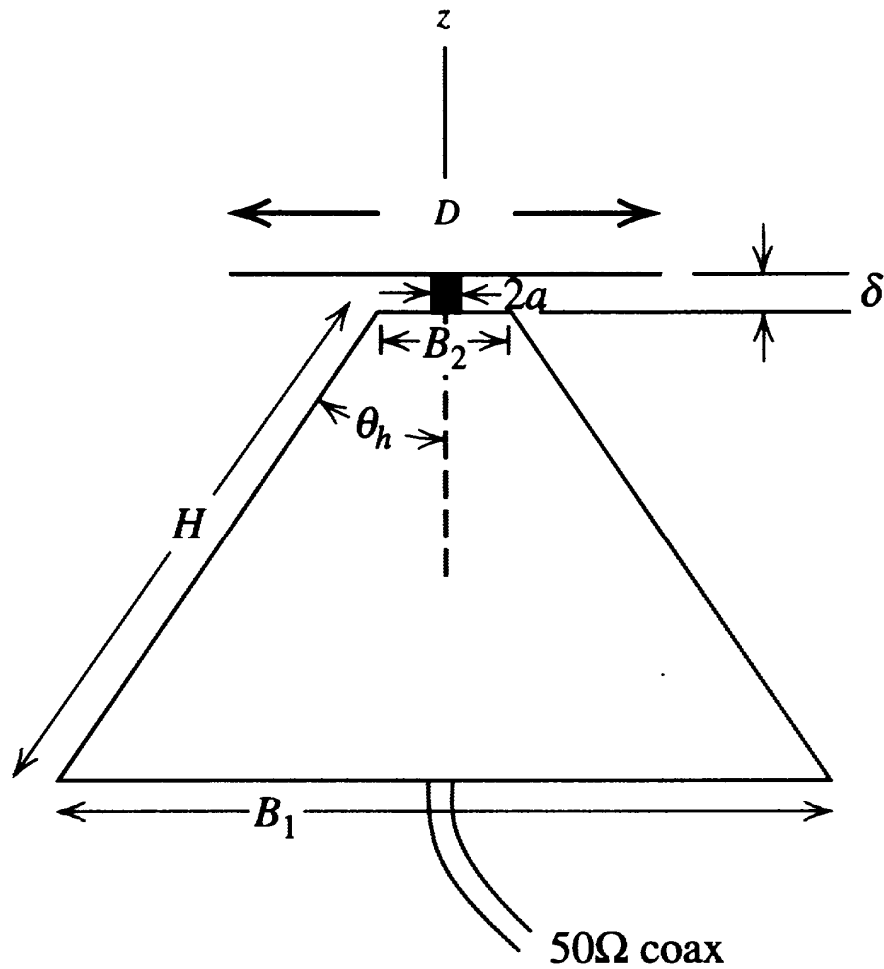


These monopoles have much broader impedance frequency band (a couple of octaves) than the ordinary quarter-wavelength monopoles. Just like other monopoles, they are conveniently fed through coaxial cables. They are a combination of the two basic antennas: the monopole/dipole antenna and the biconical antenna. The discone and conical skirt monopoles find wide application in the VHF (30 to 300 MHz) and the UHF (300 MHz to 3 GHz) spectrum for FM broadcast, television and mobile communications.

There are numerous variations of the dipole/monopole/cone geometries, which aim at broader bandwidth rather than shaping the radiation pattern. All these antennas provide omnidirectional radiation.

The discone (disk-cone) is the most broadband among these types of

antennas. This antenna was first designed by Kandoian² in 1945. The performance of the discone in frequency is similar to that of a high-pass filter. Below certain effective cutoff frequency, it has a considerable reactance and produces severe standing waves in the feed line. This happens approximately at wavelength such that the slant height of the cone H is $\approx \lambda / 4$.

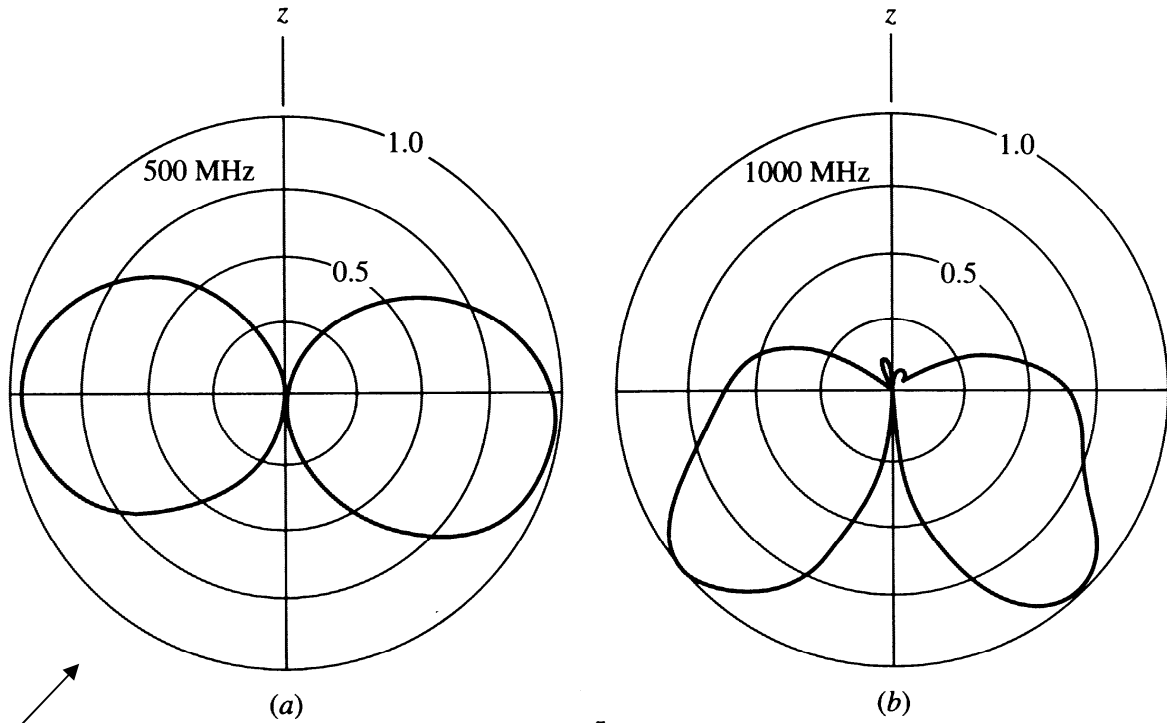


Typical dimensions of a discone antenna at the central frequency are: $D \approx 0.4\lambda$, $B_1 \approx 0.6\lambda$, $H = 0.7\lambda$, $45^\circ \leq 2\theta_h \leq 75^\circ$ and $\delta \ll \lambda$. The typical input impedance is designed to be 50Ω , for which optimum design formulas are given by Nail³: $B_2 \approx \lambda_u / 75$ at the highest operating frequency, $\delta \approx (0.3 \div 0.5)B_2$, and $D \approx 0.7B_1$.

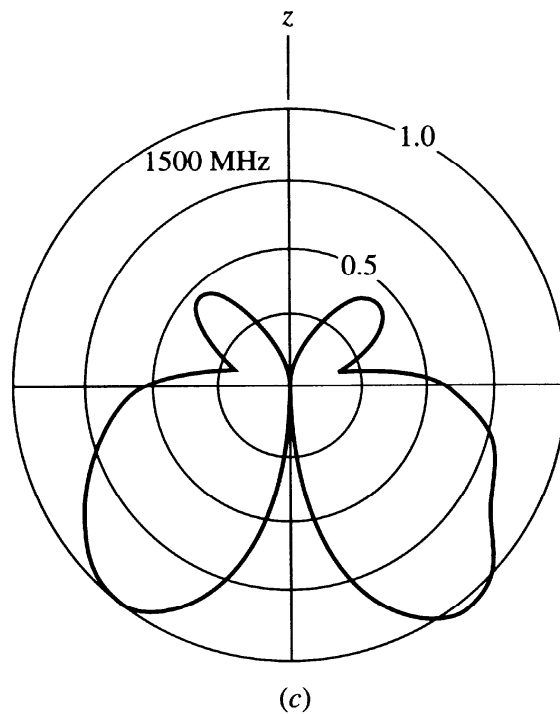
² A.G. Kandoian, "Three new antenna types and their application," *Proc. IRE*, vol. 34, pp. 70W-75W, Feb. 1946.

³ J.J. Nail, "Designing discone antennas," *Electronics*, vol. 26, pp. 167-169, Aug. 1953.

Measured patterns for a discone: $H = 21.3$ cm, $B = 19.3$ cm, $\theta_h = 25^\circ$:



Similar to a short dipole



Similar to an infinite conical monopole

[Balanis]

3. Sleeve (Coaxial) Dipoles and Monopoles

The impedance of dipole/monopole antennas is very frequency sensitive. The addition of a sleeve to a dipole or a monopole can increase the bandwidth up to more than an octave and fine-tune the input impedance.

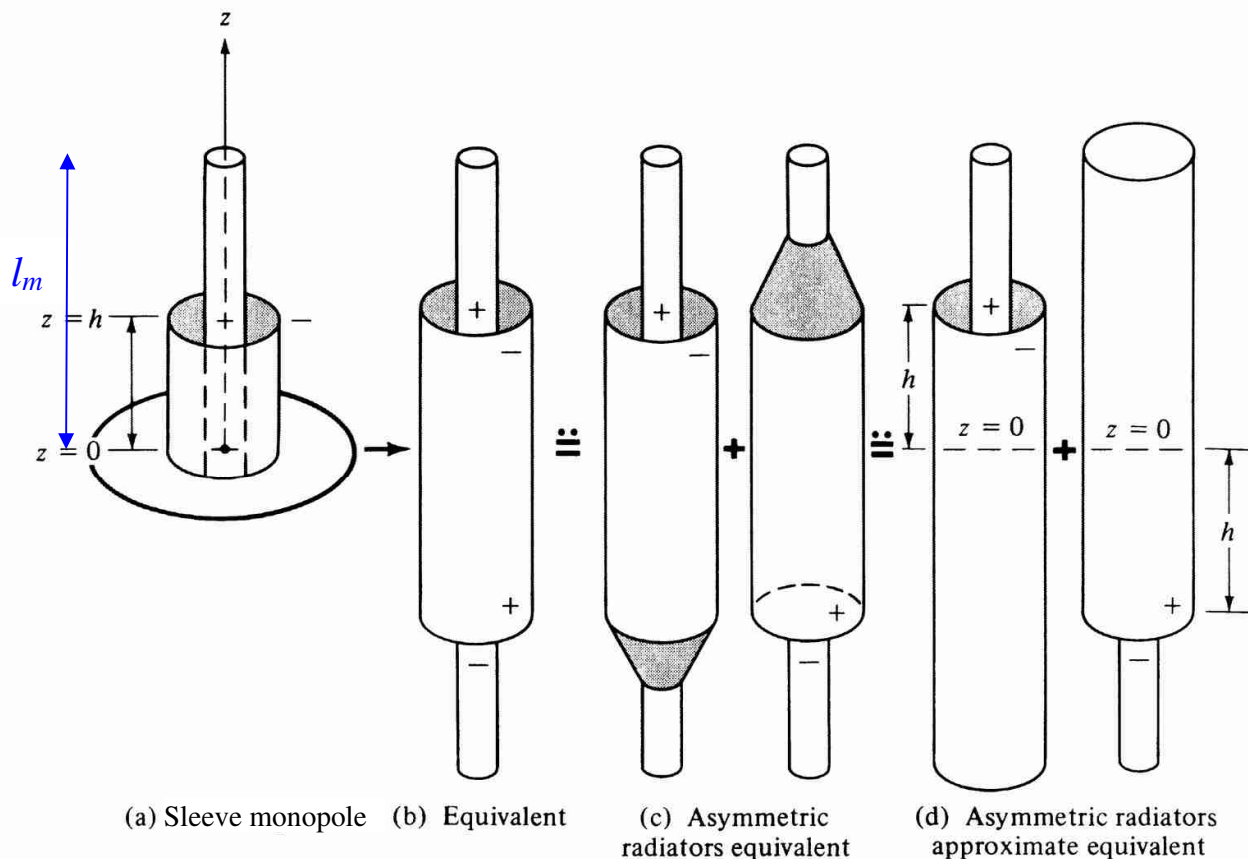


Figure 9.17 Sleeve dipole and its equivalents. (SOURCE: W. L. Weeks, *Antenna Engineering*, McGraw-Hill, New York, 1968) [Balanis]

This type of antenna closely resembles two asymmetrically fed dipoles and can be analyzed using the approximation in (d). The outer shield of the coaxial line is connected to the ground plane, but it also extends above it a distance h [see (a)] in order to provide mechanical strength, impedance tuning and impedance broadband characteristics. The equivalent in (d) consists of two dipoles, which are asymmetrically driven at $z' = +h$ or $z' = -h$ (offset feed). When analyzing the field of the two asymmetrically driven dipoles, we can ignore the change in diameter occurring at the feed point.

The input impedance Z_{as} of an asymmetric dipole can be related to the impedance Z_s of a center-fed (symmetric) dipole of the same length l using the assumption for sinusoidal current distribution [see Lecture 9],

$$I(z') = \begin{cases} I_0 \sin[\beta(0.5l - z')], & 0 \leq z' \leq l/2 \\ I_0 \sin[\beta(0.5l + z')], & -l/2 \leq z' \leq 0. \end{cases} \quad (11.12)$$

The impedances Z_{as} and Z_s relate to the radiated power Π through their respective current magnitudes, $I(z' = h)$ and $I(z' = 0)$. Imposing the condition that for a given current distribution the power delivered to the antenna remains the same regardless of the feed position, leads to

$$Z_s I_s^2 = Z_{as} I_{as}^2. \quad (11.13)$$

Thus,

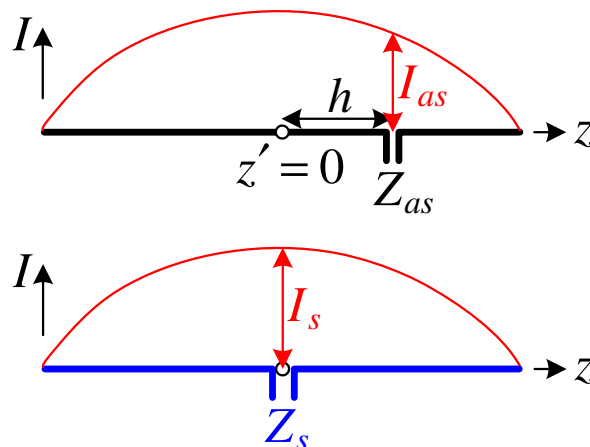
$$Z_{as}(h) = Z_s \left[\frac{I(z' = 0)}{I(z' = h)} \right]^2 = Z_s \frac{\sin^2(\beta l / 2)}{\sin^2 \left[\beta \left(\frac{l}{2} - h \right) \right]}. \quad (11.14)$$

For a half-wavelength dipole, $\beta l / 2 = \pi / 2$ and $\sin(0.5\pi - \beta h) = \cos(\beta h)$. Thus, the relation (11.14) reduces to

$$Z_{as}(h) = \frac{Z_s}{\cos^2(\beta h)}, \text{ for } l = \frac{\lambda}{2}. \quad (11.15)$$

The center-fed dipole impedance Z_s can be calculated with the *emf* method; see Eqs. (10.40)-(10.41) in Lecture 10.

The relation between the input impedances of the symmetric and asymmetric dipole feeds is illustrated in the figure below.



It is now obvious that we can tune the input impedance of a dipole by moving the feed point off-center. In the case of a sleeve monopole, this is

achieved by changing h , i.e., shortening or extending the sleeve along the stub.

Let us examine the equivalent antenna structure in Figure (d). It consists of two asymmetrically driven dipoles in parallel. The total input current is

$$I_{in} = I_{as}(z' = +h) + I_{as}(z' = -h). \quad (11.16)$$

The input admittance is

$$Y_{in} = \frac{I_{in}}{V_{in}} = \frac{I_{as}(z' = +h) + I_{as}(z' = -h)}{V_{in}} = \frac{I_{as}(z' = h)}{V_{in}} \left[1 + \frac{I_{as}(z' = -h)}{I_{as}(z' = +h)} \right] \quad (11.17)$$

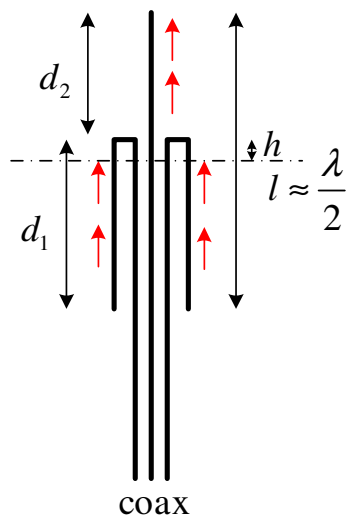
$$\Rightarrow Y_{in} = Y_{as} \left[1 + \frac{I_{as}(z' = -h)}{I_{as}(z' = +h)} \right]. \quad (11.18)$$

Since the two dipoles in (d) are geometrically identical and their currents are equal according to image theory, $I_{as}(z' = -h) = I_{as}(z' = +h)$, it follows from (11.18) that

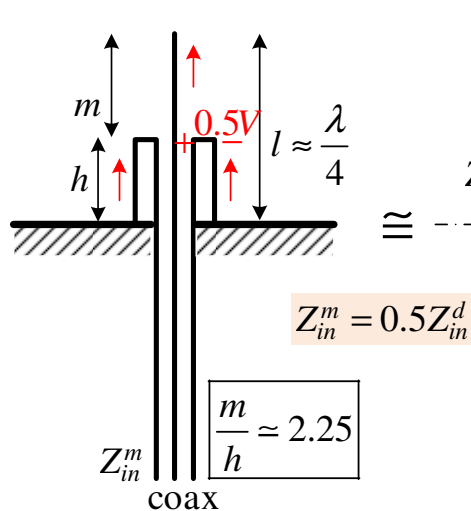
$$Y_{in} \approx 2Y_{as}. \quad (11.19)$$

Thus, the impedance of the sleeve monopole in (a) is twice smaller than the impedance of the respective asymmetrically driven dipole [one of the dipoles in (d)]. This conclusion is in agreement with the general relation between the impedance of a monopole above a ground plane and its respective dipole (of doubled length) radiating in open space (see Lecture 9).

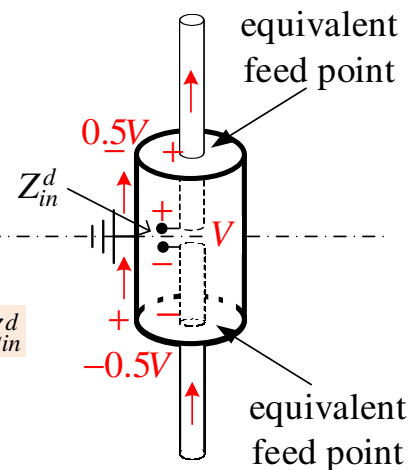
The first sleeve-dipole resonance occurs at a length $l_m \approx \lambda / 4$. The other important design variable is the monopole-to-sleeve ratio $\eta = (l_m - h) / h$. It has been experimentally established that $\eta = 2.25$ yields optimum (nearly constant with frequency) radiation patterns over a 4:1 band. The value of η has little effect on the radiation pattern if $l_m \leq \lambda / 2$, since the current on the outside of the sleeve has approximately the same phase as that on the top portion of the monopole. However, for longer lengths, the ratio η has notable effect on the pattern since the current on the outside of the sleeve is not necessarily in phase with that on the top portion of the monopole. Some practical sleeve dipoles and monopoles are shown below.



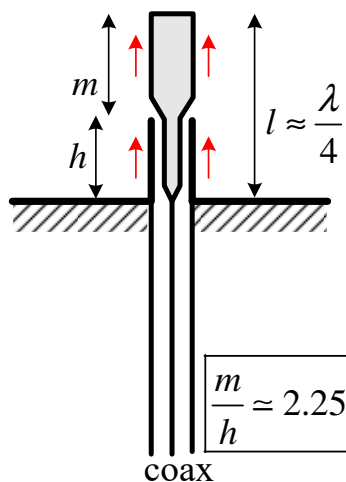
(a) asymmetrically-fed sleeve dipole



(b) sleeve monopole



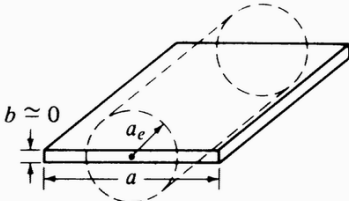
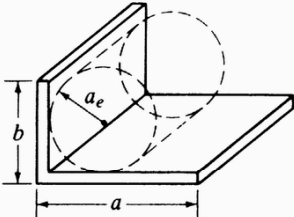
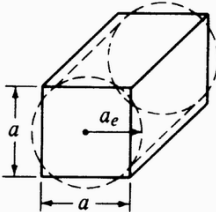
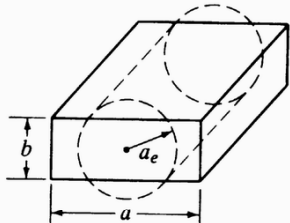
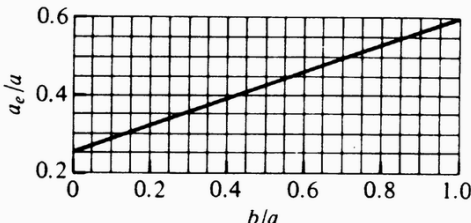
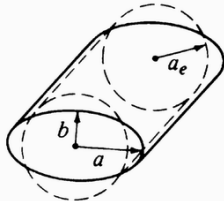
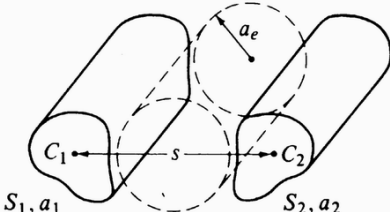
(c) sleeve dipole equivalent to monopole in (b)



(d) another sleeve monopole

So far, we have assumed that the cross-section of the wire is circular of radius a . An electrically equivalent radius can be obtained for some uniform wires of non-circular cross-sections. This is helpful when calculating the impedance of dipoles made of non-circular wires. The equivalent radii for certain wires are given below.

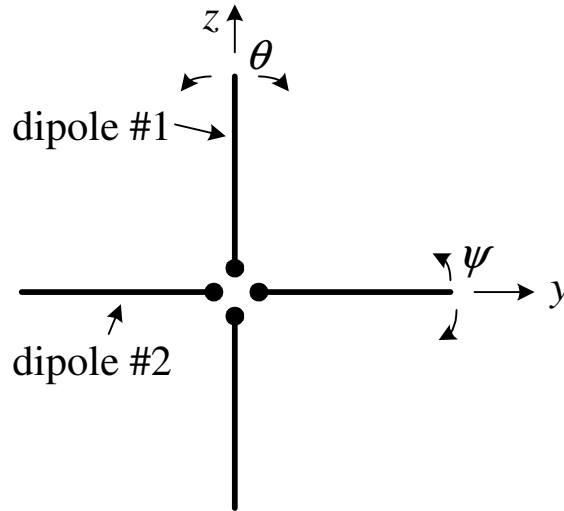
Table 9.3 CONDUCTOR GEOMETRICAL SHAPES AND THEIR EQUIVALENT CIRCULAR CYLINDER RADII

Geometrical Shape	Electrical Equivalent Radius
	$a_e = 0.25a$
	$a_e \approx 0.2(a + b)$
	$a_e = 0.59a$
	
	$a_e = \frac{1}{2}(a + b)$
	$\ln a_e \approx \frac{1}{(S_1 + S_2)^2} \times [S_1^2 \ln a_1 + S_2^2 \ln a_2 + 2S_1 S_2 \ln s]$ <p>S_1, S_2 = peripheries of conductors C_1, C_2 a_1, a_2 = equivalent radii of conductors C_1, C_2</p>

[Balanis]

4. Turnstile Antenna

The *turnstile* antenna is a combination of two orthogonal dipoles *fed in phase-quadrature*. This antenna is capable of producing circularly polarized field in the direction, which is normal to the dipoles' plane. It produces an isotropic pattern in the dipoles' plane (the θ -plane) of linearly (along $\hat{\theta}$) polarized wave. In all other directions, the wave is elliptically polarized.



In the $\varphi = 90^\circ, 270^\circ$ plane (the yz plane in which the dipoles lie), the field is a superposition of the individual dipole fields, the patterns of which as a function of time are

$$\bar{E}_\theta^{(1)}(t) = \sin \theta \cos \omega t, \quad (11.20)$$

$$\bar{E}_\psi^{(2)}(t) = \sin \psi \cdot \cos(\omega t \pm \pi / 2) = \pm \sin(\pi / 2 - \theta) \cdot \sin \omega t = \pm \cos \theta \cdot \sin \omega t. \quad (11.21)$$

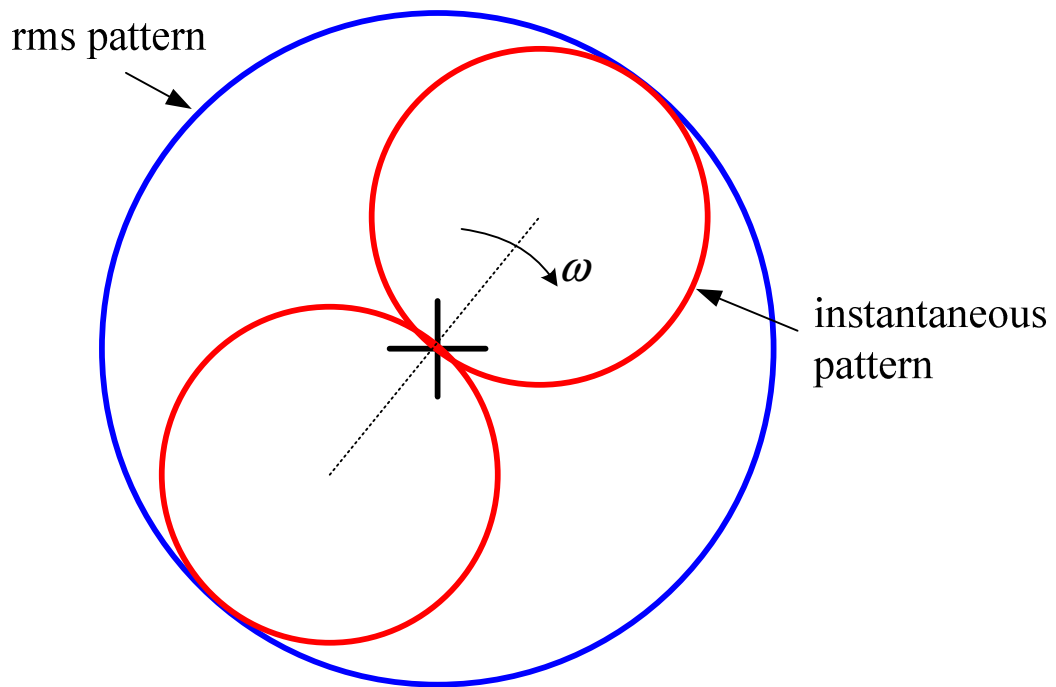
In the yz plane, the ψ -component of a vector is actually a θ -component. Equations (11.20) and (11.21) define the total normalized field as

$$\bar{E}_\theta(t, \theta) = \sin \theta \cos \omega t \pm \cos \theta \sin \omega t, \quad (11.22)$$

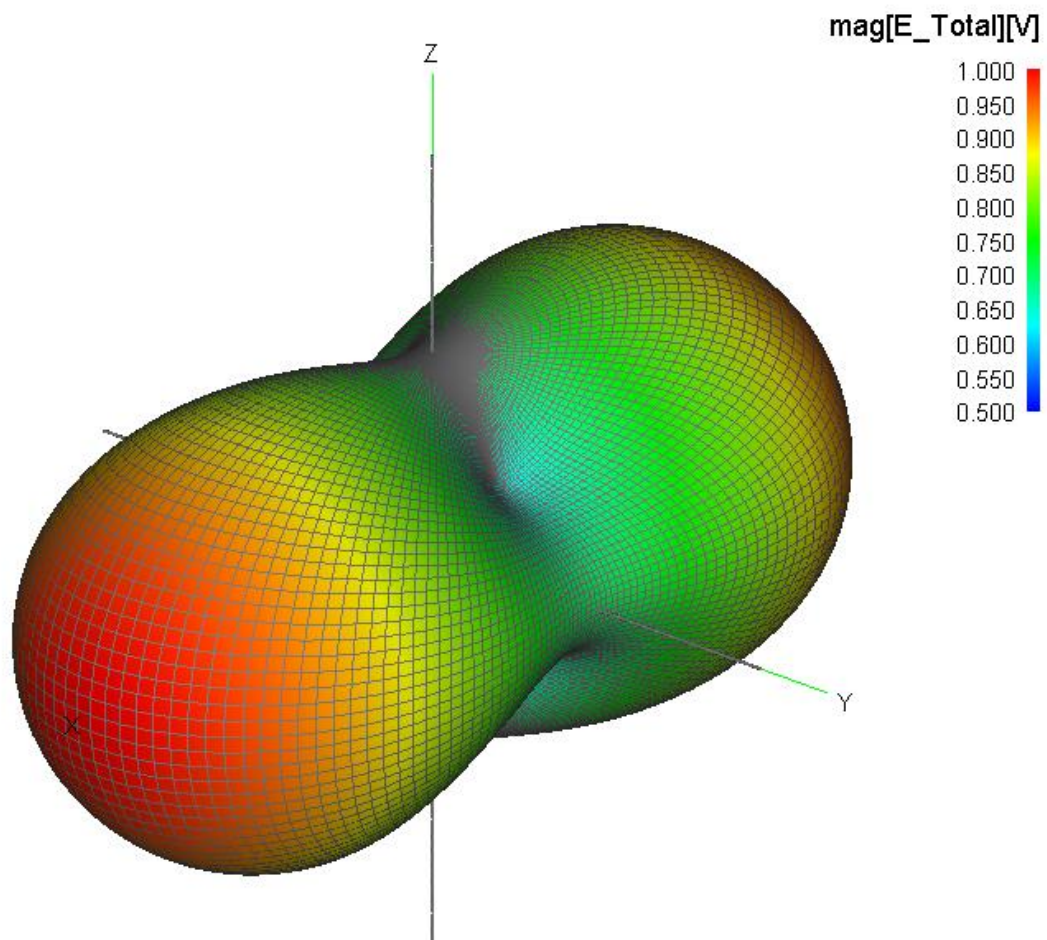
which reduces to

$$\bar{E}_\theta(t, \theta) = \sin(\theta \pm \omega t). \quad (11.23)$$

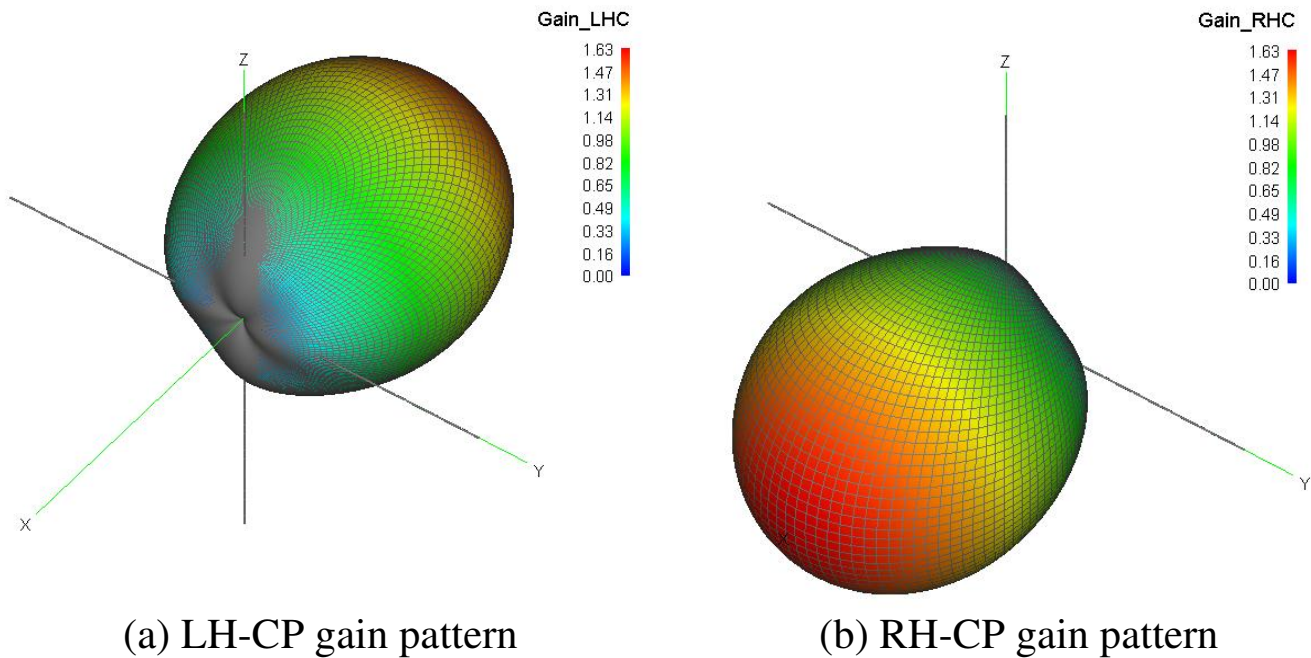
The *rms* normalized pattern is circular, $\bar{E}_{\theta, \text{rms}}(\theta) = 1$, although the instantaneous pattern rotates with an angular frequency ω .



3-D Turnstile Total Field Magnitude Pattern:

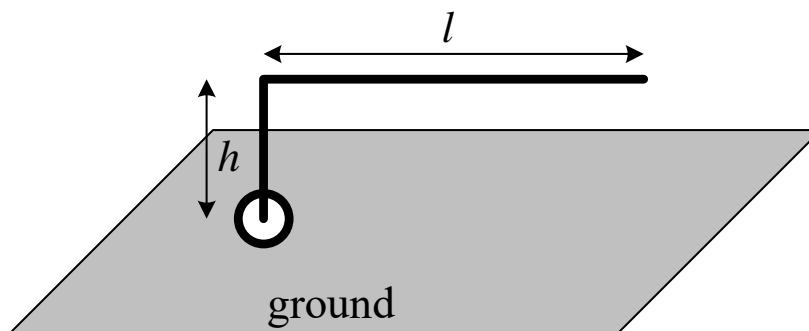


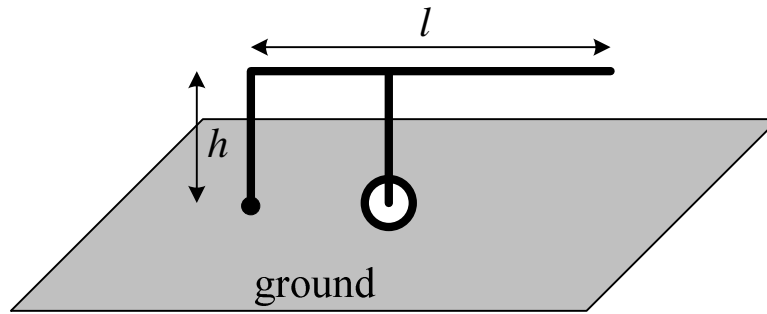
3D Gain Patterns of Trunstile ($V_z = 1$, $V_y = 1e^{j\pi/2}$)



5. Inverted-L and Inverted-F Antennas

The L-antenna (or inverted-L antenna) can be viewed as a bent monopole. The bending of the monopole results in a reduced size and low profile. The first designs were made for missile applications (R. King *et al.*, “Transmission-line missile antennas,” *IRE Trans. Antennas Propagat.*, Jan. 1960, pp. 88-90). The reduction of the monopole’s height results in reduced radiation resistance and bandwidth. Besides, the main angle of radiation is depressed as there is substantial radiation not only from the monopole h but also from the arm l . A popular amateur-radio antenna is the $\lambda/4$ inverted-L where $h = l = \lambda/8$.





The inverted-F antenna uses a shifted feed point (a “tap”) along the bent arm l , to obtain better impedance match (offset feed).

Both the inverted-L and the inverted-F antennas can be analyzed using equivalent transmission line models. Their patterns in both principal planes are not very different from those of a monopole.

A popular antenna for mobile handsets is the planar inverted-F (PIFA) or its variations:

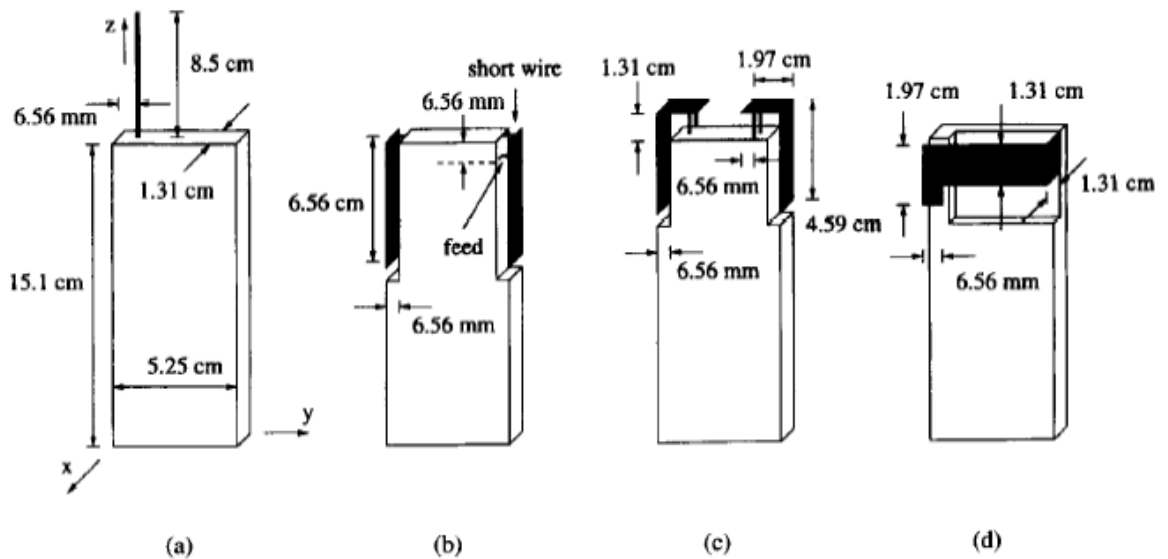


Fig. 1. Antenna geometries and dimensions for hand-held transceiver units: (a) monopole; (b) side-mounted dual PIFA; (c) top-mounted dual BIFA; (d) back-mounted PIFA. The chassis dimensions (105 cm^3) shown in (a) apply to all four configurations.

Figure from M.A. Jensen *et al.*, “EM interaction of handset antennas and a human in personal communications,” *Proceedings of the IEEE*, vol. 83 , No. 1, Jan. 1995, pp. 7-17. (Note: BIFA stands for *bent inverted F antenna*)

6. Matching Techniques for Wire Antennas

There are two major issues when constructing the feed circuit: impedance matching and balanced-unbalanced matching.

6.1. Impedance matching

$$\text{VSWR} = \frac{1+|\Gamma|}{1-|\Gamma|}. \quad (11.24)$$

The reflected-to-incident power ratio is given by $|\Gamma|^2$ (where Γ is the reflection coefficient). In terms of the VSWR, this is

$$|\Gamma|^2 = \left(\frac{\text{VSWR} - 1}{\text{VSWR} + 1} \right)^2. \quad (11.25)$$

The transmitted-to-incident power ratio is given by

$$|T|^2 = 1 - |\Gamma|^2. \quad (11.26)$$

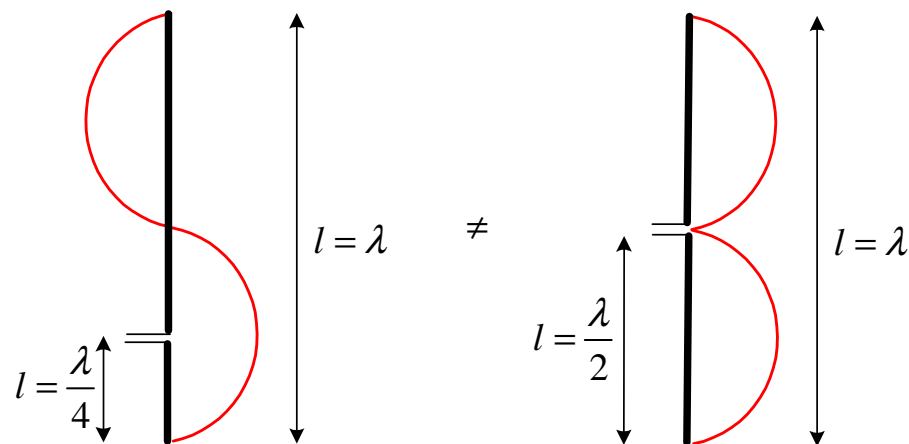
Impedance mismatch is undesirable not only because of the inefficient power transfer. In high-power transmitting systems, high VSWR leads to maxima of the standing wave which can cause arcing. Sometimes, the frequency of the transmitter can be affected by severe impedance mismatch (“frequency pulling”). Excessive reflections can damage the amplifying stages in the transmitter.

TABLE: VSWR AND TRANSMITTED POWER

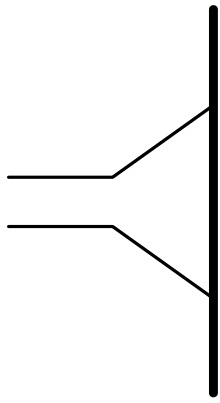
VSWR	$ \Gamma ^2 \times 100\%$	$ T ^2 \times 100\%$
1.0	0.0	100.0
1.1	0.2	99.8
1.2	0.8	99.2
1.5	4.0	96.0
2.0	11.1	88.9
3.0	25.0	75.0
4.0	36.0	64.0
5.0	44.4	55.6
5.83	50.0	50.0
10.0	66.9	33.1

A common way to find the proper feed location along a dipole or monopole is to feed off-center, which provides increase of the input impedance with respect to the center-feed impedance according to (11.14) and (11.15). For example, the input resistance of a center-fed half-wavelength dipole is approximately 73Ω , which is well suited for a $75\text{-}\Omega$ coaxial line if proper care is taken of the balanced-to-unbalanced transition. However, it is not well matched to a $300\text{-}\Omega$ antenna cable where we do not have to worry about a balanced-to-unbalanced transition since both the antenna and the cable are balanced. Greater values of the dipole input impedance (close to 300Ω) are achieved by moving the feed off center. Similarly, the quarter-wavelength monopole has an input resistance of approximately 37Ω and usually the sleeve-type of feed is used to achieve greater values of the antenna input impedance such as 50Ω to achieve better impedance match to a $50\text{-}\Omega$ coaxial cable.

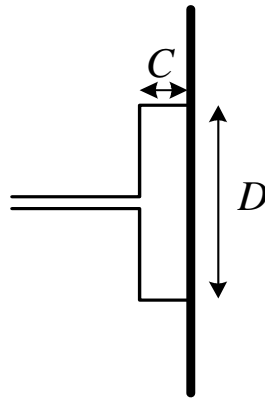
As a word of caution, the off-center feed is not symmetrical and can lead to undesirable phase reversal in the antenna if $l > \lambda/2$. This may profoundly change the radiation pattern.



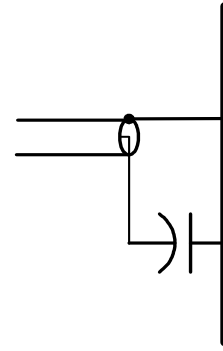
To avoid current phase reversal, symmetrical feeds for increased impedance are used. A few forms of *shunt matching (or shunt feed)* are shown below:



(a) Delta match



(b) Tee match



(c) Gamma match

We explain the principles of operation of the T-match only, which is the simplest of all to design and which gives the basic idea for all shunt feeds. Similarly to the folded-dipole analysis, the T-match interconnection together with the antenna can be viewed as two shorted transmission lines (in TL mode of operation) and a dipole (in an antenna mode of operation), which is longer than the two shorted TLs. The shorted TLs are less than quarter-wavelength long, and, therefore, they have an inductive reactance. This reactance is usually greater than the capacitive reactance of the dipole and an additional tuning lumped capacitor might be necessary to achieve better match. As the distance D increases, the input impedance increases (current magnitude drops). It has a maximum at about $D = l / 2$ (half the dipole's length). Then, it starts decreasing again, and when $D = l$, it equals the folded-dipole input impedance. In practice, sliding contacts are made between the shunt arms and the dipole for impedance adjustments and tuning. Note that shunt matches radiate and may alter the patterns and the directivity. To reduce the parasitic radiation, the wire separation C must be small.

The Gamma-match is essentially the same as the T-match, only that it is designed for unbalanced-balanced connection.

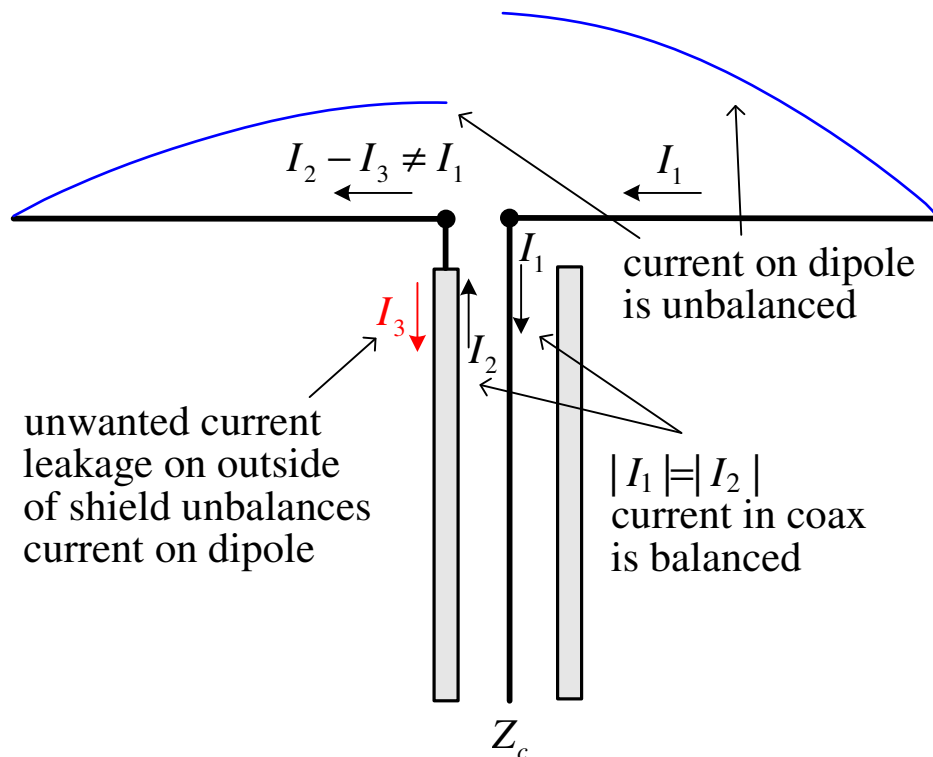
Additional matching devices are sometimes used such as quarter-wavelength impedance transformers, reactive stubs for compensating antenna reactance, etc. These devices are studied in Microwave Engineering courses.

6.2. Balanced-to-unbalanced feed

Sometimes, when high-frequency devices are connected, their impedances

may be well matched and still we may observe significant reflections. This is sometimes referred to as “field mismatch.”

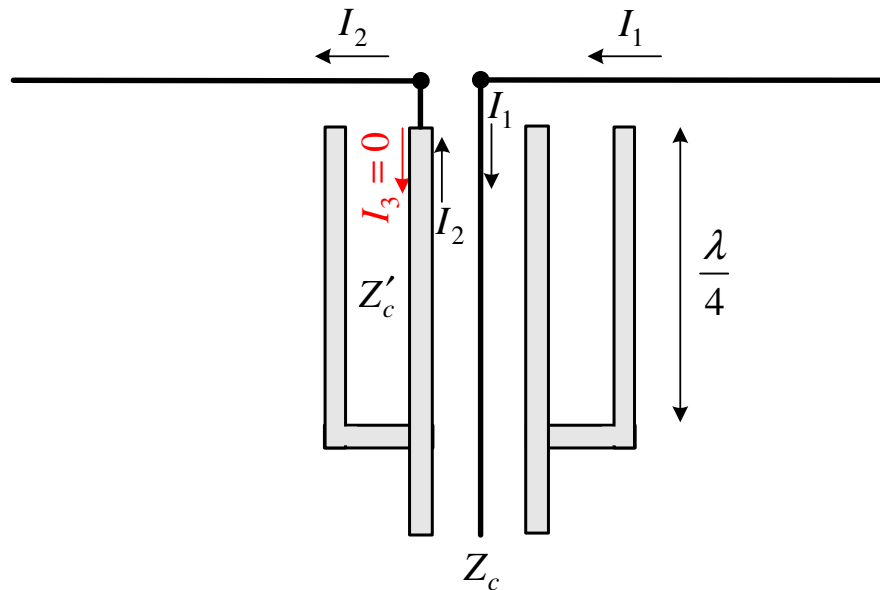
A typical example in antennas is the interconnect between a coaxial line of $Z_c = 75 \ \Omega$ and a half-wavelength dipole of $Z_{in} = 73 \ \Omega$. The reflections are much more severe than one would predict using equation (11.25). This is because the field and the current distributions in the coaxial line and at the input of the wire dipole are very different [see figure below]. The unequal currents on the dipole’s arms unbalance the antenna and the coaxial feed and induce currents on the outside of the coax shield which are the reason for parasitic radiation. To balance the currents, various devices are used, called *baluns* (balanced-to-unbalanced transformer).



A. Sleeve (bazooka) balun 1:1

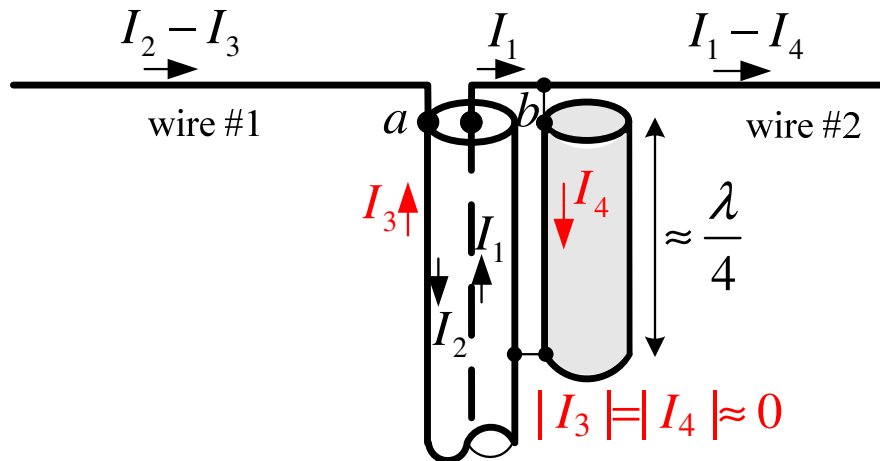
The sleeve and the outer conductor of the coaxial feed form another coaxial line, which has a characteristic impedance of Z'_c . This line is shorted quarter-wavelength away from the antenna input terminals. Thus, its input impedance is very large and results in: (i) suppression of the currents on the outer shield (I_3), and (ii) no interference with the antenna input impedance, which is in parallel

with respect to the coaxial feed. This is a narrowband balun, which does not transform the impedance (1:1 balun). It is not very easy to construct.



B. Folded balun 1:1 (split-coax balun, $\lambda / 4$ -coax balun)

This 1:1 balun is easier to make. It is also narrowband.



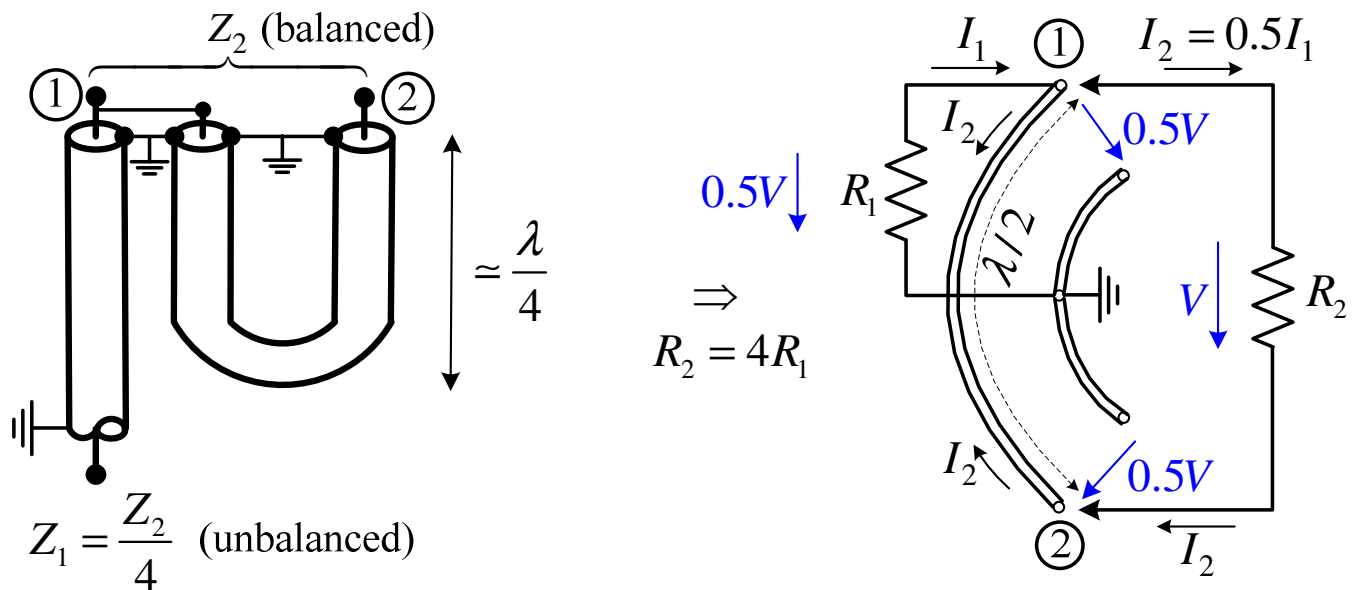
The outer shields of the feeding coaxial line and the additional coax-line section form a twin-lead transmission line, shorted a distance $\approx \lambda / 4$ away from the antenna input. This line's input is in parallel with the antenna's terminals but does not affect the overall impedance because it has infinite impedance. The additional piece of coaxial line re-directs a portion of the I_1 current, which induces the twin-lead current I_4 . The currents I_3 and I_4 are well balanced ($I_3 = I_4$) because the current of wire #1 ($I_2 - I_3$) would induce as much current

at the outer coaxial shield I_3 , as the current of wire #2 ($I_1 - I_4$) would induce in the outer shield of the auxiliary coaxial piece I_4 . This is due to the structural similarity of the two interconnects; see nodes (a) and (b) in the Figure. Thus,

$$\frac{I_3}{I_2 - I_3} = \frac{I_4}{I_1 - I_4}.$$

Since $I_1 = I_2$ in the feeding coaxial line, it is also true that $I_3 = I_4$. Thus, the current on the outer coaxial shield is effectively canceled from a certain point on ($\approx \lambda / 4$).

C. Half-wavelength coaxial balun 1:4

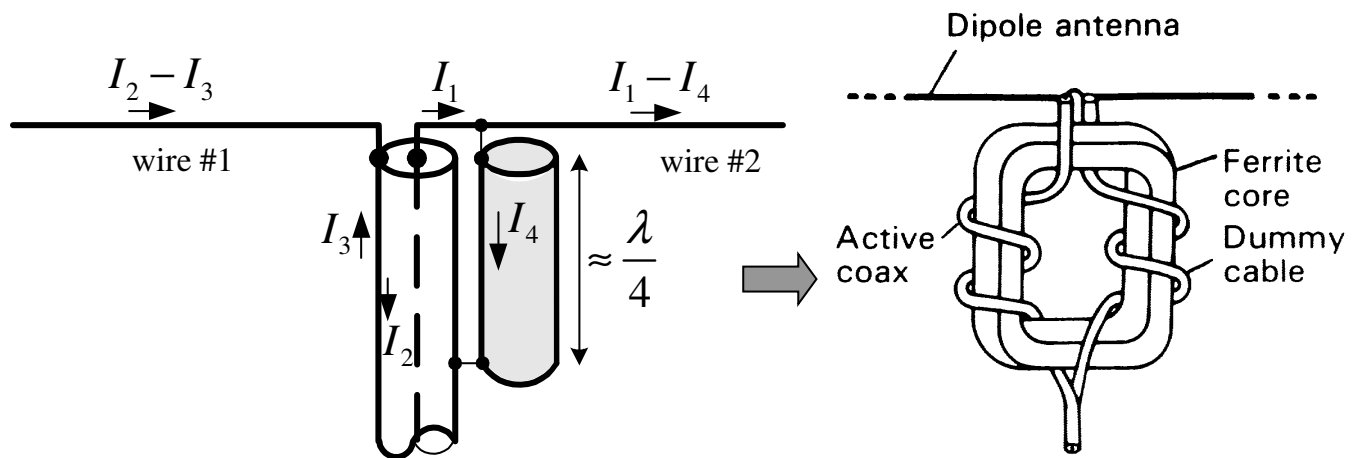


Typically, a coaxial feed of $Z_c = 75 \Omega$ would be connected with such a balun to a folded dipole of $Z_A \approx 292 \Omega$ [see equation (11.11)]. The principle of operation is explained by the equivalent circuit on the right. The auxiliary piece of coaxial line ($\lambda/2$ long) transforms the input voltage at terminal 1 to a voltage of the same magnitude but opposite polarity at terminal 2. It also splits the input current into two equal parts (why?). Thus, the load “sees” twice larger voltage and twice smaller current compared to the case without balun. Hence the 4-fold increase in impedance.

D. Broadband baluns

All baluns described above are narrowband because of the dependence on the wavelength of the auxiliary transmission-line sections. Broadband baluns for high-frequency applications can be constructed by tapering a balanced transmission line to an unbalanced one gradually over a distance of several wavelengths (microstrip-to-twin-lead, coax-to-twin-lead).

At lower frequencies (below UHF), tapered baluns are impractical, and transformers are used for impedance adjustment and balancing the feed. Often, ferrite-core bifilar wound-wire baluns are preferred for their small dimensions and broadband characteristics (bandwidths of 10:1 are achievable). A ferrite-core transformer 1:1, which is equivalent to the folded balun 1:1, but is much more broadband, is shown below.

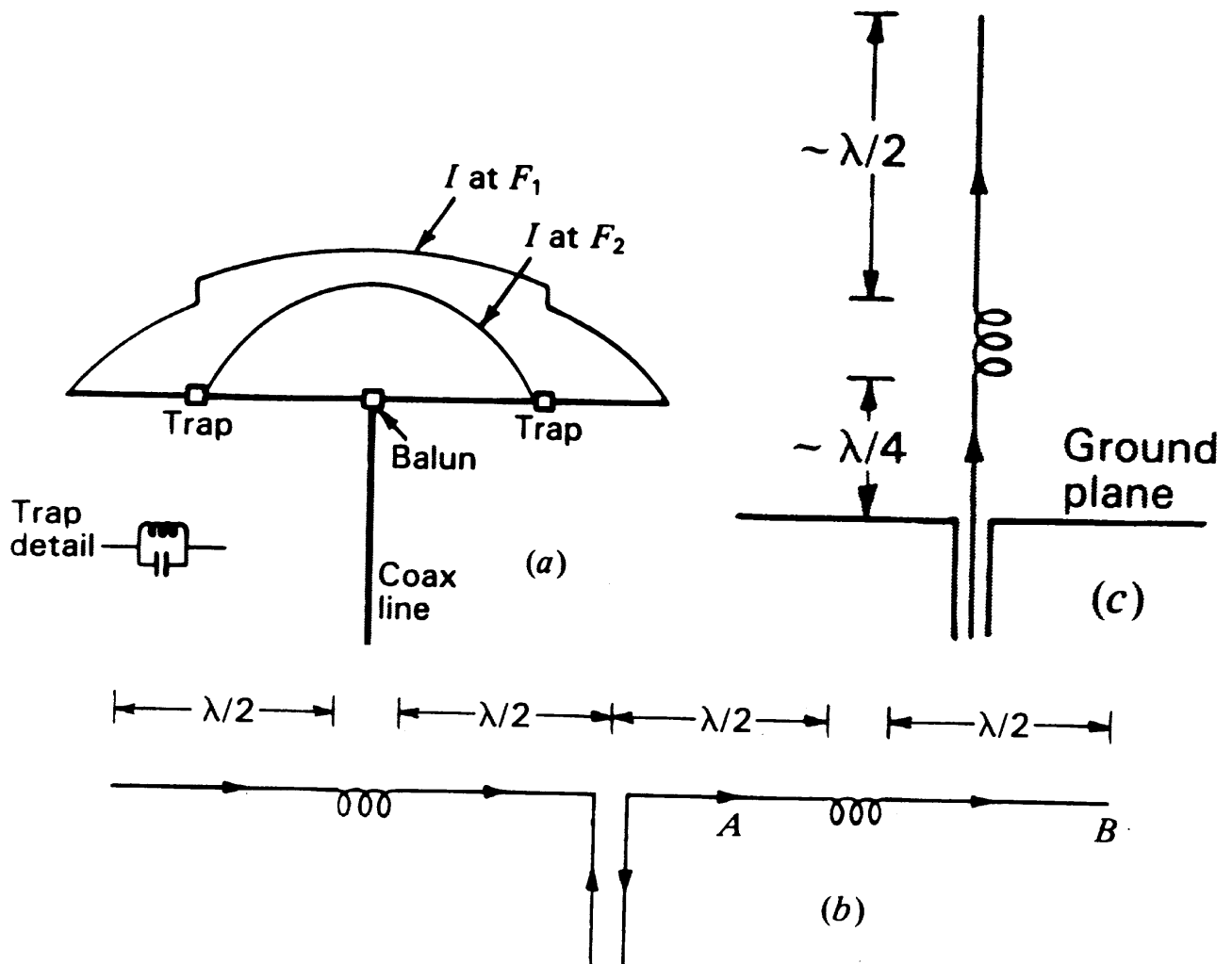


The transmission line formed by the outer shields of the two coaxial cables is now a very high-impedance line because of the high relative permeability of the ferrite core. Thus, it does not disturb significantly the antenna input impedance, which is in parallel with it. At the same time, its electrical length is not critical. The ferrite core is also lossy, effectively damping all currents excited on the outer shields of the coaxial (active and dummy) cables.

7. Dipoles with Traps

In many wideband applications, it is not necessary to have frequency-independent antennas (which are often expensive), but rather antennas that can operate at two (or more) different narrow bands. Typical example is the type of multi-band antennas in cellular communication systems. A dual-band antenna

can be constructed from a single center-fed dipole (or its respective monopole) by means of tuned traps [see figure (a) below]. The same principle applies to monopoles as well. Each trap is a tuned parallel LC circuit (although it does have parasitic resistance as well, typically associated with the coil). At frequency f_1 , for which the whole dipole is $\approx \lambda/2$ long, the trap behaves like an inductor because the coil's susceptance is larger than that of the capacitor, $B_L = 1/(\omega_1 L) \gg B_C = \omega_1 C$. This reduces slightly the resonant length of the dipole or monopole [see figure (a) below] and has to be taken into account. At another frequency $f_2 > f_1$, the traps become resonant ($B_L + B_C = 0$) and effectively cut off the outer portions of the dipole, making the dipole shorter and resonant at this higher frequency. If the traps, for example, are in the middle of the dipole arms, then $f_2 \approx 2f_1$ is the resonant frequency for half the dipole and the antenna can operate equally well at two frequencies separated by an octave.



[Figures from Kraus, *Antennas*, 2nd ed., p. 744]

It should be noted that the isolation of the outer portion of the dipole depends not only on the high impedance of the traps but also on the impedance of this outer portion. When the outer arm portions are about $\lambda/4$ long, they have very low impedance compared to the trap's impedance and are effectively mismatched, i.e., their currents are negligible. This is not the case if the outer portions are $\lambda/2$ long, when they represent very high impedance themselves and become impedance matched to the inner portion of the dipole “terminated” by the traps. They are then no longer isolated and support significant current. In fact, resonant traps cause current phase reversal and can thus be used to construct electrically long dipoles (longer than a wavelength) that support in-phase current along the whole dipole [see figure (b) above].

A coil alone can form a trap at certain (very high) frequency because of its own distributed capacitance. It can also act as a 180° phase shifter (the coil may be viewed as a coiled-up $\lambda/2$ wire element). The use of this property is illustrated in figure (b) which shows how one can construct an array of 4 in-phase $\lambda/2$ -elements with a single feed and achieve a gain of 6.4 dBi. Figure (c) shows the $3\lambda/4$ monopole, which is obtained from the dipole in (b) by cutting the dipole at point A, and mounting it above a ground plane. This is a common antenna for cordless phone handsets. Its gain can be as high as 8.3 dBi and it has an input resistance of $\approx 150 \ \Omega$.