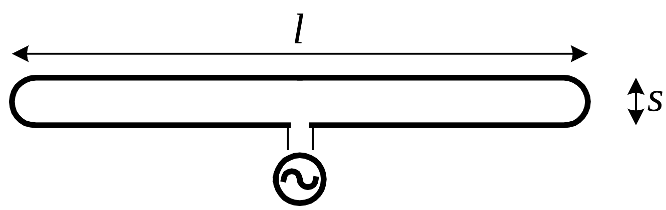
LECTURE 10: Other Practical Dipole/Monopole Geometries. Matching Techniques for Dipole/Monopole Feeds.

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

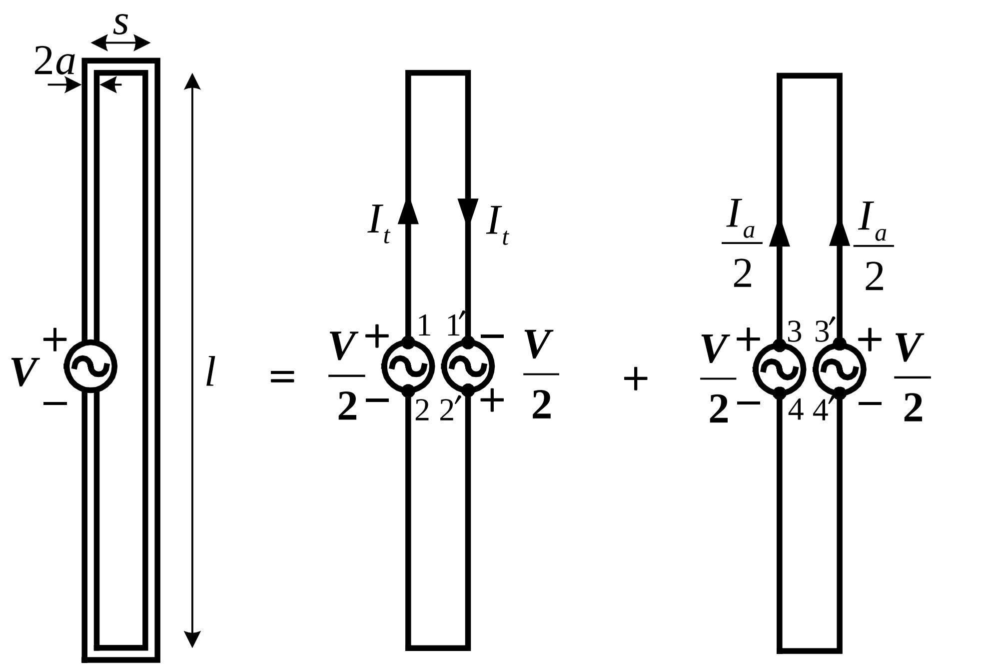
1. Folded dipoles



The folded dipole is a very popular antenna for reception of TV broadcast signals. It has essentially the same pattern as the dipole of the same length l but it provides four times greater input impedance when $l \simeq \lambda/2$. The length of a single-wire dipole is usually $\lambda/4 \le l \le \lambda$ for best directivity with no side lobes. Most often, $l \simeq \lambda/2$. The input resistance then is $R_{in} \simeq 73\ \Omega$. Wire antennas are not fit to coaxial feed lines because of the different field structure. However, they are ideally suited for twin-lead (two-wire) feed lines. These lines (two parallel thin wire lines separated by a distance of about 8-10 mm) have $Z_c \approx 300\ \Omega$. Therefore, an input antenna impedance of $(4\times 73)\ \Omega$ is perfect for matching to 2-wire feed lines. The separation distance between the two wires of the folded dipole should not exceed 0.05λ.

The folded dipole can be analyzed by decomposing its current into two modes: the transmission-line mode and the antenna mode. This analysis, albeit approximate[1], illustrates the four-fold impedance transformation.

[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, Sept. 1980.



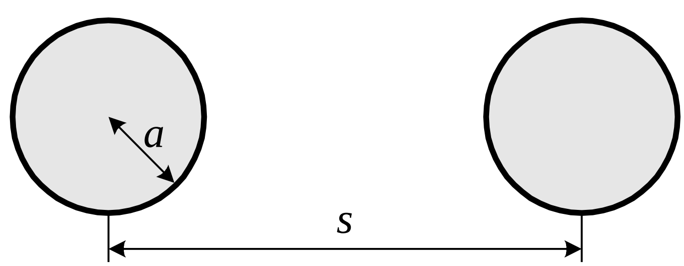
The input impedance at the terminals $1-1'$ and $2-2'$ can be determined as the input impedance of a shorted transmission line of length $l/2$:

$$Z_t = \left[Z_0 \left(\frac{Z_L + jZ_0 \tan(\beta l/2)}{Z_0 + jZ_L \tan(\beta l/2)} \right) \right]_{Z_L=0} \tag{10.1}$$

$$\Rightarrow Z_t = jZ_0 \tan\left(\frac{\beta l}{2}\right) \tag{10.2}$$

Here, Z_0 is the characteristic impedance of a 2-wire transmission line:

$$Z_0 = \frac{\eta}{\pi} \operatorname{arccosh}\left(\frac{s}{2a}\right) = \frac{\eta}{\pi} \ln\left[\frac{s/2 + \sqrt{(s/2)^2 - a^2}}{a}\right] \tag{10.3}$$



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

$$Z_t(\lambda/2) = jZ_0 \tan(\pi/2) \to \infty \quad (10.4)$$

If $l \neq \lambda/2$, the more general expression (10.2) should be used. The current in the transmission-line mode is:

$$I_t = \frac{V}{2Z_t} \quad (10.5)$$

Let us consider now the antenna mode. The generators' terminals $3-3'$ (and $4-4'$) are with identical potentials. Therefore, they can be connected without loss of generality. The following assumption is made: an equivalent dipole of effective radius

$$a_e = \sqrt{as} \quad (10.6)$$

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

$$I_a = \frac{V}{2Z_a} \quad (10.7)$$

The current at each leg of the equivalent dipole is obviously

$$\frac{I_a}{2} = \frac{V}{4Z_a} \quad (10.8)$$

The total current of a folded dipole is obtained by combining both modes. At the input

$$I_{in} = I_t + \frac{I_a}{2} = V\left(\frac{1}{2Z_t} + \frac{1}{4Z_a}\right) \quad (10.9)$$

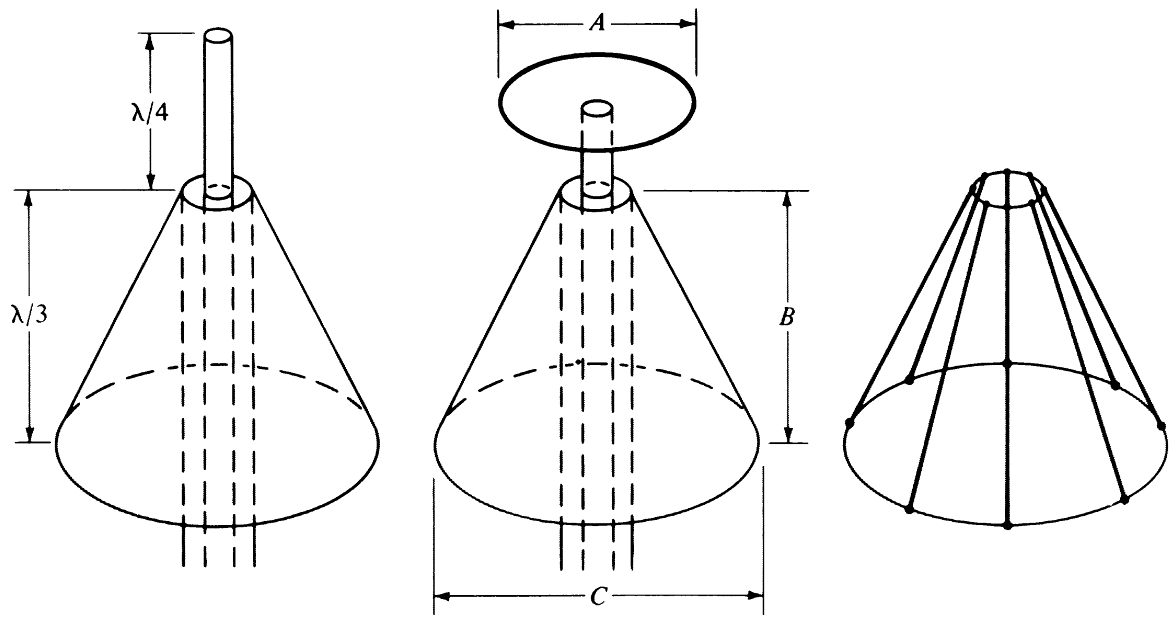
$$\Rightarrow Z_{in} = \frac{4Z_t Z_a}{2Z_a + Z_t} \quad (10.10)$$

When $l = \lambda/2$ (half-wavelength folded dipole), then $Z_t \to \infty$, and

$$\Rightarrow Z_{in} = 4Z_a \Big|_{l=\frac{\lambda}{2}} \simeq 292\ \ \Omega \quad (10.11)$$

Thus, the half-wavelength folded dipole is very well suited for direct connection to a twin-lead line ($Z_c \approx 300\ \ \Omega$). It is often made in a very simple way: a suitable portion (the end part of the twin-lead cable of length $l = \lambda/2$) is separated into two single wire leads, which are bent to form the folded dipole.

2. Conical (skirt) monopoles and discones



(a) Conical skirt monopole (b) Discone (c) Wire-simulation

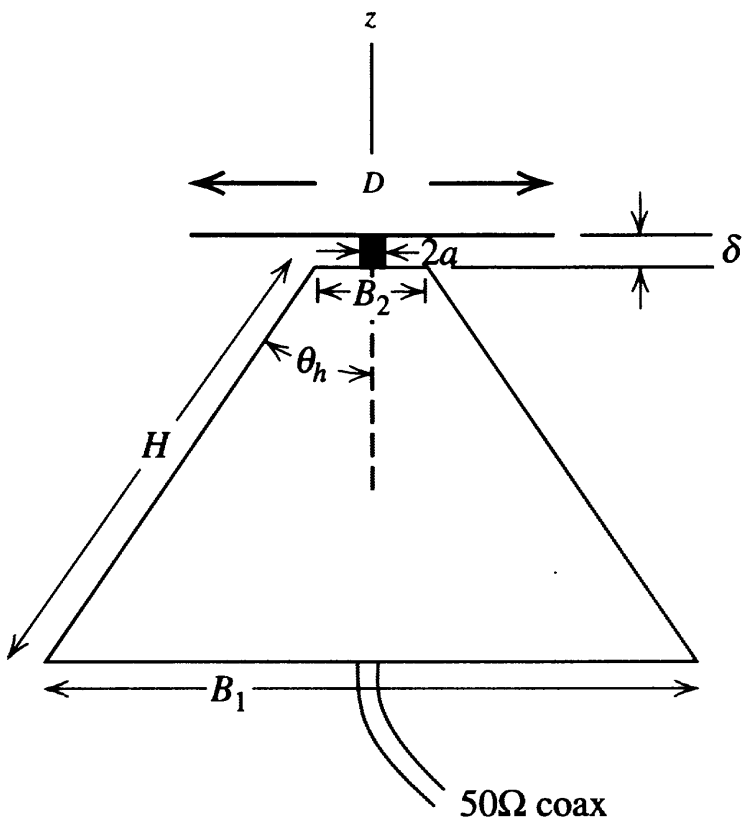
These monopoles have much broader frequency band for their impedance variations (a couple of octaves) than the ordinary quarter-wavelength monopoles. They are a combination of the two basic antennas: the monopole/dipole antenna and the biconical antenna. The discone and conical skirt monopoles find wide application in the VHF (30-300 MHz) and the UHF (300 MHz - 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 Kandoian[2] 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

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

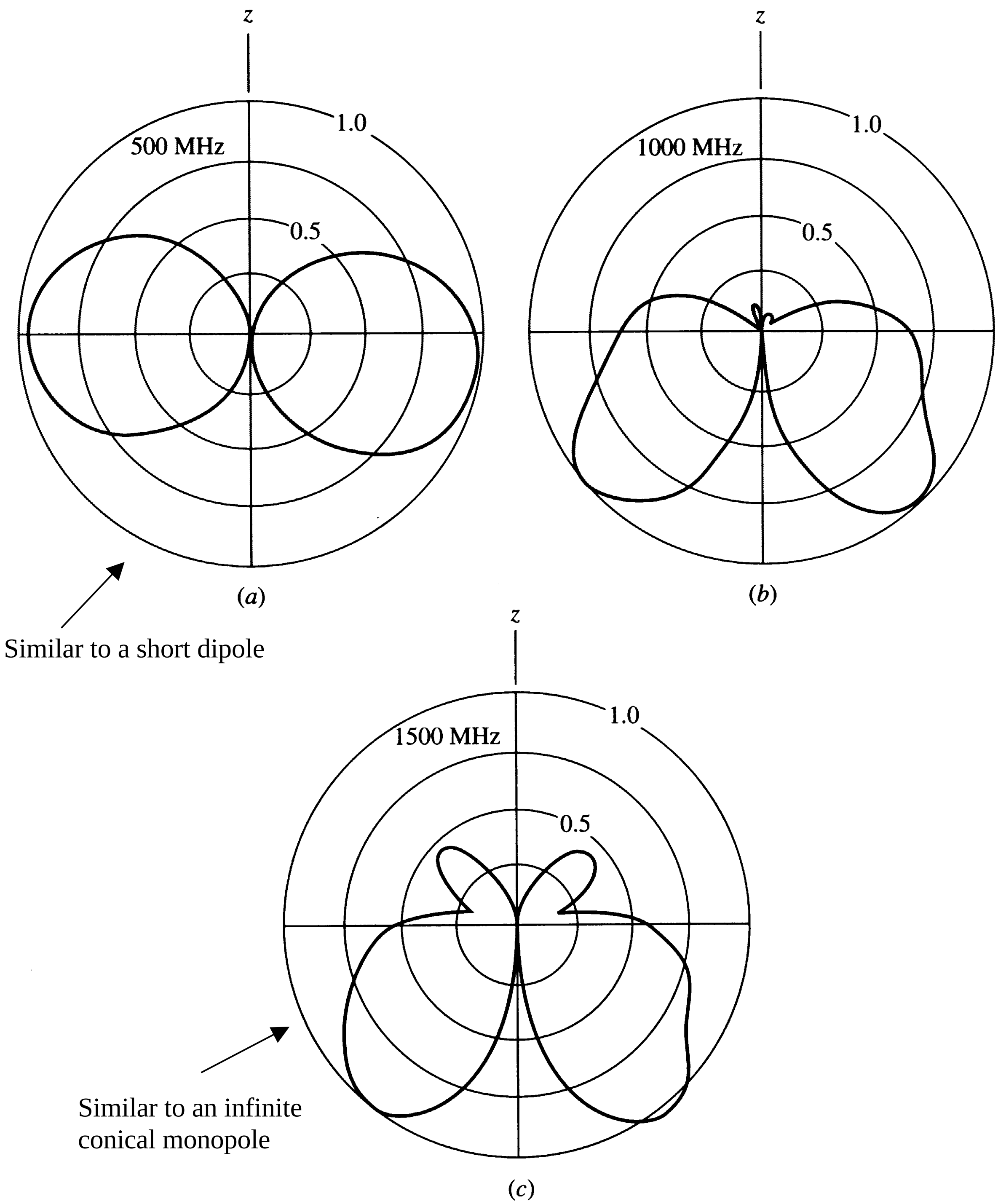
happens approximately at wavelength such that the slant height of the cone is $\approx \lambda/4$.



Typical dimensions of a discone antenna at central frequency are: $D \approx 0.4\lambda$, $B_1 \approx 0.6\lambda$, $H = 0.7\lambda$, $45^\circ \le 2\theta_h \le 75^\circ$ and $\delta \ll \lambda$. The typical input impedance is designed to be 50 Ω. Optimum design formulas are given by Nail[3]: $B_2 \approx \lambda_u / 75$ at the highest operating frequency, $\delta \approx (0.3 \div 0.5)B_2$, and $D \approx 0.7B_1$.

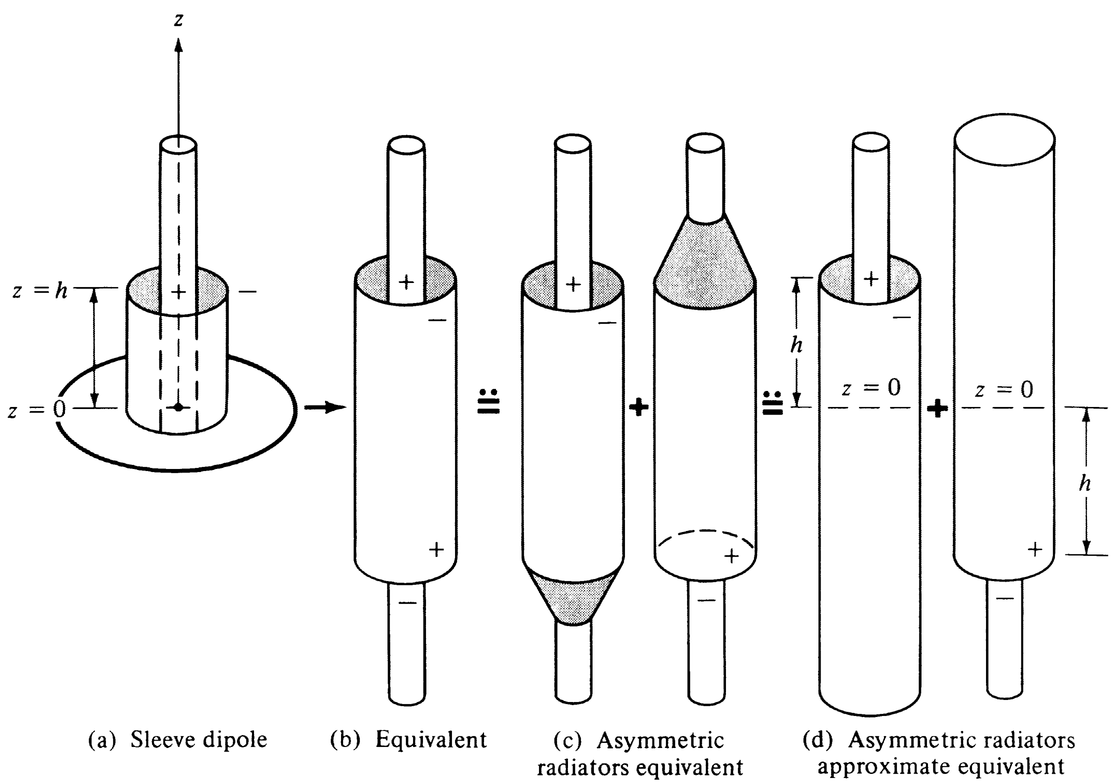
[3] J.J. Nail, "Designing discone antennas," *Electronics*, vol. 26, pp. 167-169, Aug. 1953.

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



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.



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

This type of antenna closely resembles an asymmetric dipole, 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, 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 $z' - h$. When analyzing the field of the two asymmetrically driven dipoles, one can ignore the change in diameter occurring at the feed point.

The input impedance of an asymmetric dipole can be related to its self-impedance approximately as:

$$Z_{as} = \frac{Z_m}{\sin^2\left[\beta\left(\frac{l}{2} - h\right)\right]} \tag{10.12}$$

where h is the off-center displacement. The relation (10.12) was already derived when the centered-feed impedance of a dipole of arbitrary length was analyzed in Lecture 7. A symmetrically driven dipole would have an input impedance of:

$$Z_s = \frac{Z_m}{\sin^2\left(\beta\frac{l}{2}\right)} \tag{10.13}$$

For a half-wavelength dipole $(l = \lambda/2)$, it is easy to show that the relation between the input impedance of the asymmetric dipole Z_{as} and the center-fed symmetric dipole Z_s is:

$$Z_{as} \simeq \frac{Z_s}{\cos^2(\beta h)} \tag{10.14}$$

The general expression is:

$$Z_{as}(h) \simeq Z_s \frac{\sin^2\left(\beta\frac{l}{2}\right)}{\sin^2\left[\beta\left(\frac{l}{2} - h\right)\right]} \tag{10.15}$$

From (10.15) or from (10.12) it is obvious that one can control the input impedance by shortening or extending the sleeve along the stub.

The equivalent antenna structure in (d) actually consists of two asymmetrically driven dipoles. The total input current is:

$$I_{in} = I_{as}(z' = +h) + I_{as}(z' = -h) \tag{10.16}$$

The input admittance is:

$$Y_{in} = \frac{I_{in}}{V_{in}} = \frac{I_{as}(z' = +h) + I_{as}(z' = -h)}{V_{in}} = \frac{I_{as}(z = h)}{V_{in}}\left[1 + \frac{I_{as}(z = -h)}{I_{as}(z = +h)}\right] \tag{10.17}$$

$$\Rightarrow Y_{in} = Y_{as}\left[1 + \frac{I_{as}(z' = -h)}{I_{as}(z' = +h)}\right] \tag{10.18}$$

Here, $Y_{as} = 1/Z_{as}$ can be calculated from (10.15). The assumption for a sinusoidal current distribution dictates that the currents in (10.18) should be calculated from the formula:

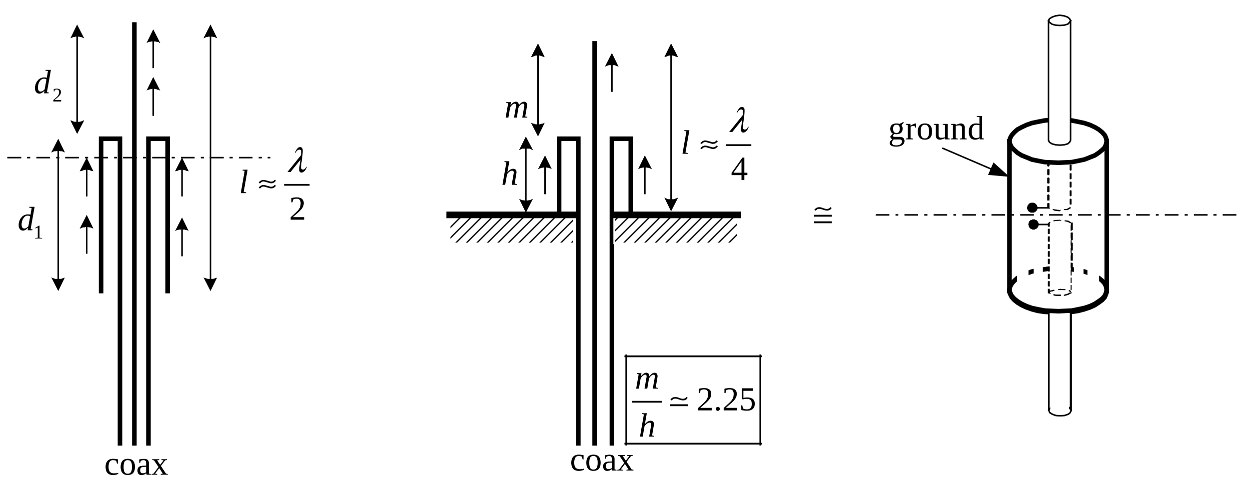
$$I(z') = I_m \sin\left[\beta\left(\frac{l}{2} - |z'|\right)\right] \qquad (10.19)$$

Since the two dipoles in (d) are geometrically identical, it follows from (10.18) that:

$$Y_{in} \simeq 2Y_{as} \qquad (10.20)$$

The first sleeve monopole resonance will occur at a length of approximately $l \simeq \lambda/4$. The other important design variable is the monopole-to-sleeve ratio $\eta = (l-h)/h$. 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 $\eta = (l-h)/h$ has little effect on the radiation pattern if $l \le \lambda/2$, since the current on the outside of the sleeve will have approximately the same phase as that on the top potion of the monopole itself. However, for longer lengths, the ration η has marked effect on the radiation patterns since the current on the outside of the sleeve will not necessarily be in-phase with that on the top portion of the monopole.

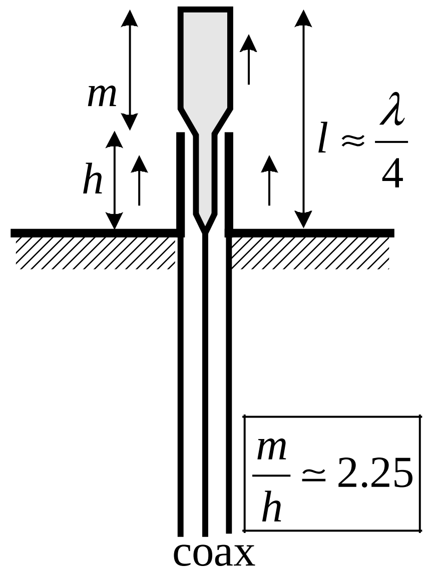
Some practical geometries:



(a) assymertically-fed sleeve dipole

(b) sleeve monopole

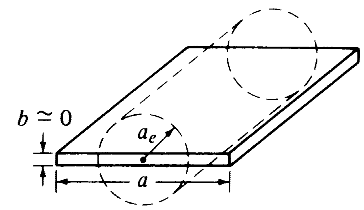
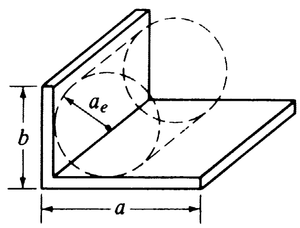
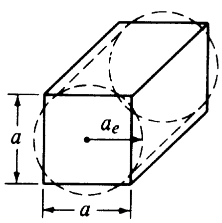
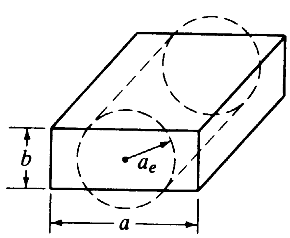
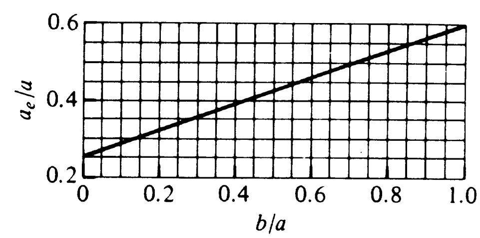
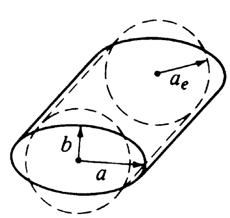
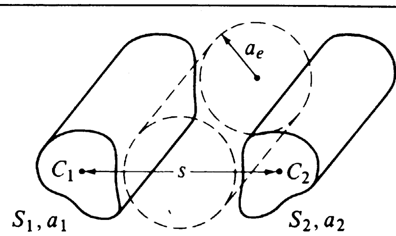
(c) sleeve dipole



(d) another sleeve monopole

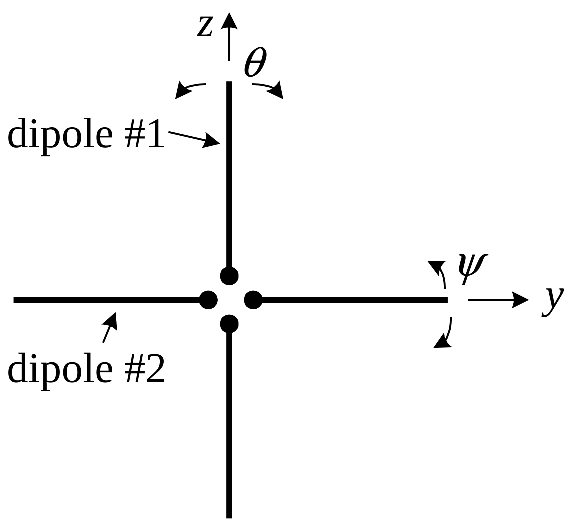
Up to now, it was always assumed that the cross-section of the wire is circular of radius a, when deriving the expressions for the input impedance. An electrical equivalent radius can be obtained for some uniform wires of non-circular cross-section. This is very helpful when calculating the impedance of dipoles made of non-circular cross-section 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
	$a_e = 0.25a$
	$a_e \simeq 0.2(a + b)$
	$a_e = 0.59a$
	
	$a_e = \frac{1}{2}(a + b)$
	$\ln a_e \simeq \frac{1}{(S_1 + S_2)^2}$ $\times [S_1^2 \ln a_1 + S_2^2 \ln a_2 + 2S_1S_2 \ln s]$ S_1, S_2 = peripheries of conductors C_1, C_2 a_1, a_2 = equivalent radii of conductors C_1, C_2

4. Turnstile antenna

The turnstile antenna is a combination of two orthogonal in space 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 $\varphi = 90^\circ$- plane (the $y-z$ plane in which the dipoles lie), the field is a superposition of the fields whose patterns are:

$$\overline{E}_\theta^{(1)}(t) = \sin\theta\cos\omega t \tag{10.21}$$

$$\overline{E}_\psi^{(2)}(t) = \sin\psi\cos(\omega t \pm \pi/2) = \pm\sin\omega t \cdot \sqrt{1-\sin^2\theta\sin^2\varphi} = \\ = \pm\cos\theta\cdot\sin\omega t \tag{10.22}$$

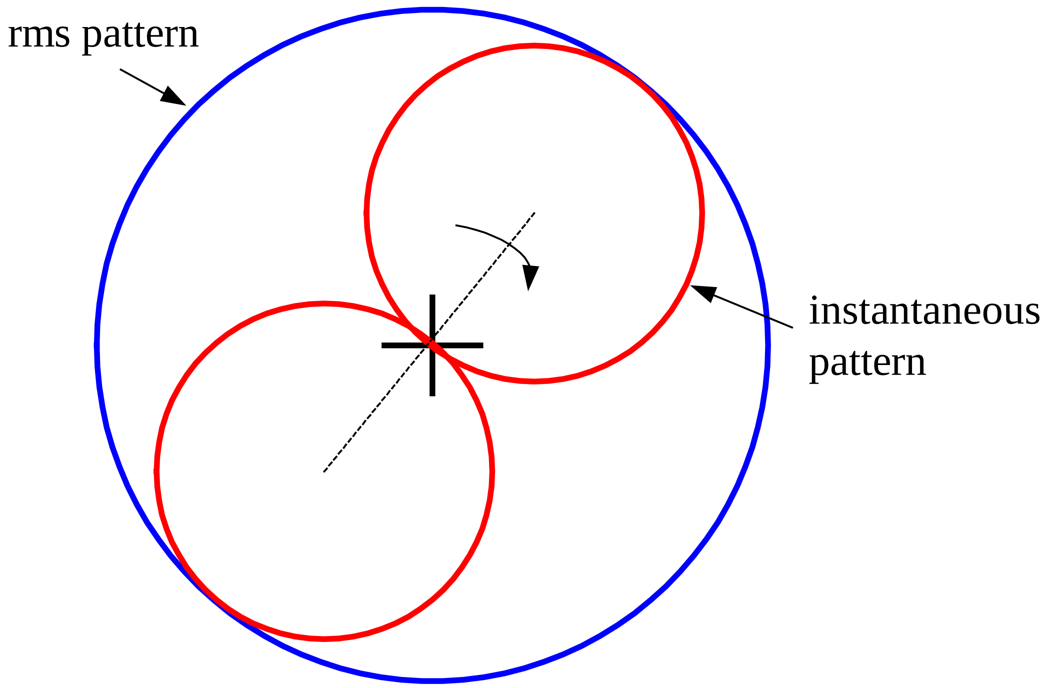
In the $\varphi = 90^\circ$- plane, the ψ-component of a vector is actually a θ-component. Equations (10.21) and (10.22) define a total field of:

$$\overline{E}_\theta(t,\theta) = \sin\theta\cos\omega t \pm \cos\theta\sin\omega t \tag{10.23}$$

which reduces to

$$\overline{E}_\theta(t,\theta) = \sin(\theta \pm \omega t) \tag{10.24}$$

The rms pattern is circular, although the instantaneous pattern rotates.



5. Matching techniques for wire antennas

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

5.1. Impedance matching

$$\text{VSWR} = \frac{1+|\Gamma|}{1-|\Gamma|} \tag{10.25}$$

Reflected power in terms of VSWR:

$$|\Gamma|^2 = \left(\frac{\text{VSWR}-1}{\text{VSWR}+1}\right)^2 \tag{10.26}$$

Transmitted power:

$$|T|^2 = 1-|\Gamma|^2 \tag{10.27}$$

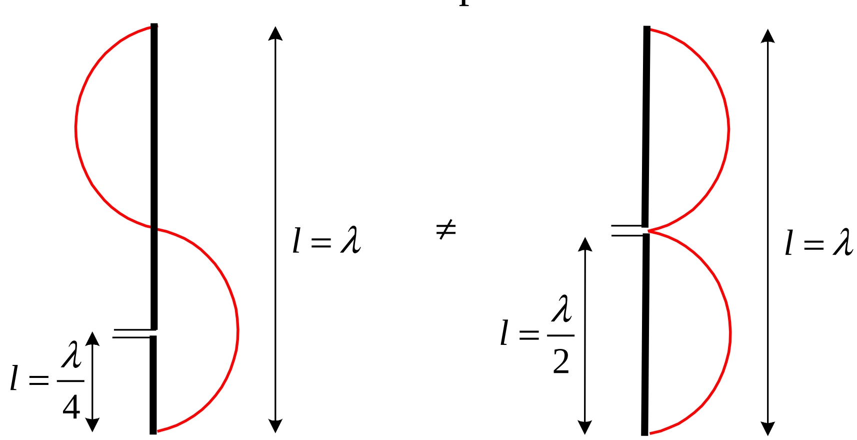
Impedance mismatch is undesirable not only because of the inefficient power transfer. In high-power transmitting systems, high VSWR leads to maxima of the standing wave, which can cause arcing. Sometimes, the frequency of the transmitter can be affected by severe impedance mismatch (“frequency pulling”).

TABLE: VSWR and Transmitted Power

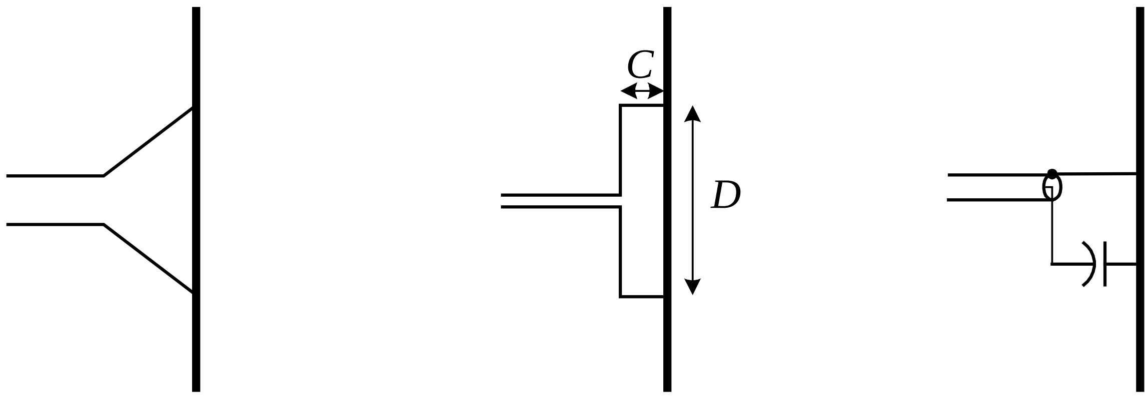
VSWR	$\lvert \Gamma \rvert^2 \times 100\%$	$\lvert T \rvert^2 \times 100\%$
1.0	0.0	100.0
1.1	0.2	99.8
1.2	0.8	99.2
1.5	4.0	96.0
2.0	11.1	88.9
3.0	25.0	75.0
4.0	36.0	64.0
5.0	44.4	55.6
5.83	50.0	50.0
10.0	66.9	33.1

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 equation (10.14). The input resistance of a half-wavelength dipole is approximately $73\ \Omega$, which is well suited for standard coaxial lines. The quarter-wavelength monopole has an input resistance of approx. $37\ \Omega$, and, usually the sleeve-type of feed is used to achieve greater values of the antenna input impedance. The folded dipole is excellent to feed with $300\ \Omega$ twin-lead line.

However, the off-center feed is unsymmetrical and can lead to undesirable phase reversal in the antenna if $l > \lambda/2$. This will profoundly change the radiation pattern. To avoid such cases, 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

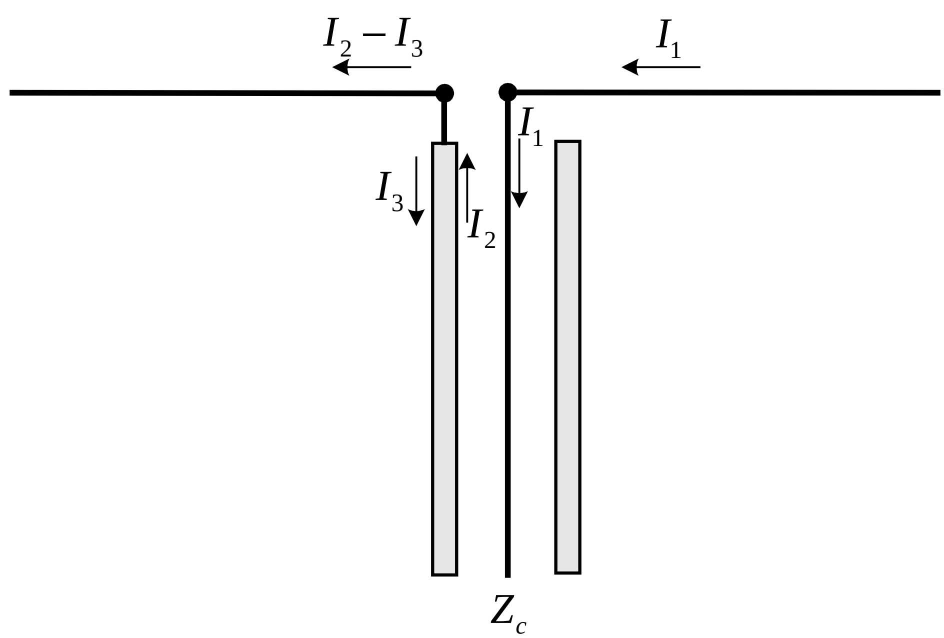
We shall explain the principles of operation of the T-match only, which is the simplest of all to design and which gives the basic idea for all shunt feeds. The T-match interconnection can be viewed as two shorted transmission lines and a very-wide-gap dipole in parallel with respect to the twin-lead cable. The shorted transmission lines are less than quarter-wavelength long, and, therefore they have an inductive reactance. This reactance is usually greater than the capacitive reactance of the wide-gap dipole, and an additional tuning lumped capacitor might be used to achieve better match. As the distance D increases, the input impedance increases, too. 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 may radiate, which is very undesirable at the operating frequency band.

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 well studied and described in courses in Microwave Engineering.

5.2. Balanced-to-unbalanced feed

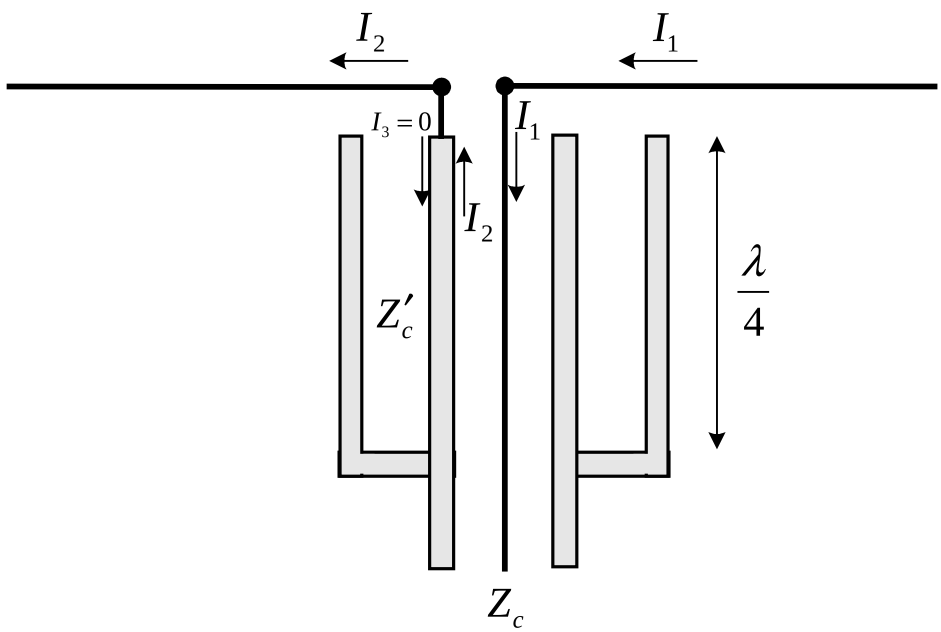
Sometimes when high-frequency devices are connected, their impedances (in a quasi-static sense) might be well matched, and still one might observe significant reflections. This is sometimes referred to as “field mismatch”. A typical example in antenna technology is the interconnect between a coaxial 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 (10.26). This is because the field and the current distributions in the coaxial line and at the input of the wire dipole are very different.



The unequal currents at the dipole’s legs unbalance the antenna and the coax. To balance the currents, various devices are used, called baluns (balanced-to-unbalanced transformer).

1) <u>Sleeve (bazooka) balun 1:1</u>

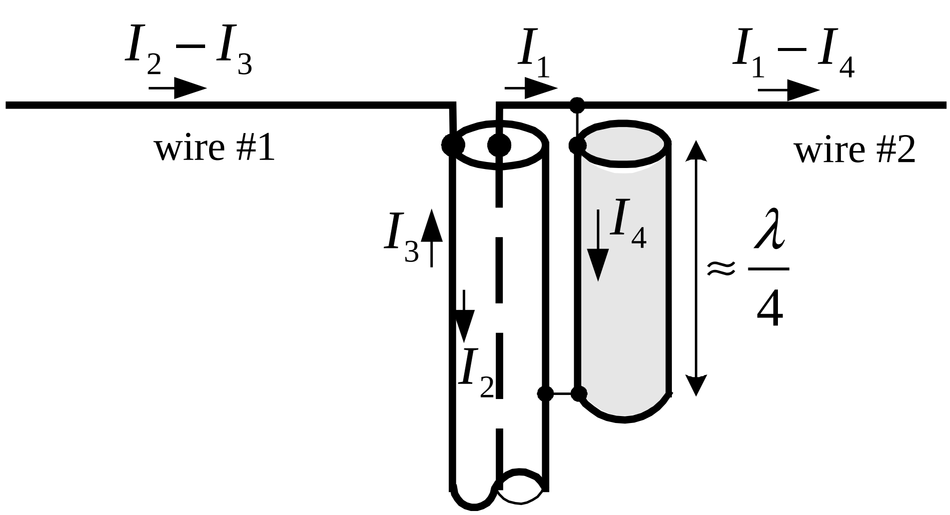
The sleeve and the outer conductor of the coax form another coax line, which has a characteristic impedance of Z_c'. This line is shorted quarter-wavelength away from the antenna input terminals.



This is a narrowband balun, which does not have impedance-transformer capability (1:1 balun). It is not very easy to construct.

2) <u>Folded balun 1:1</u> (split-coax balun, $\lambda/4$-coax balun)

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

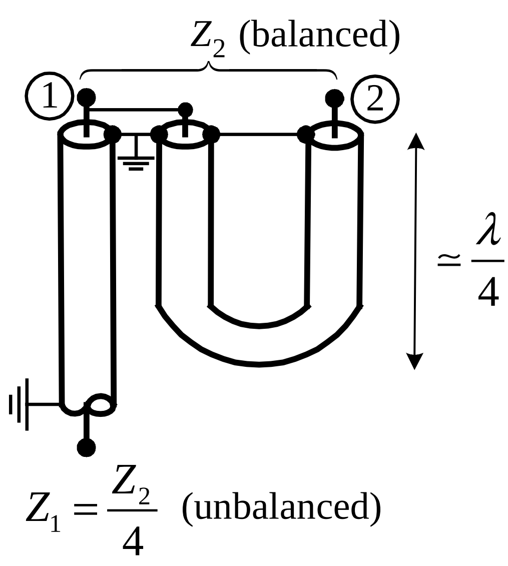


The outer shield of the feeding coax line and the additional coax-line section form a two-lead transmission line, shorted a distance $\approx \lambda/4$ away from the antenna input. This line is in parallel with the antenna but does not affect the overall impedance because it has infinite impedance at the antenna terminals. The additional coax line redirects a portion of the I_1 current, which induces the two-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 coax shield I_3, as the current of wire #2 $(I_1 - I_4)$ would induce in the outer shield of the auxiliary coax I_4 (note the geometry similarity of the interconnects), i.e.

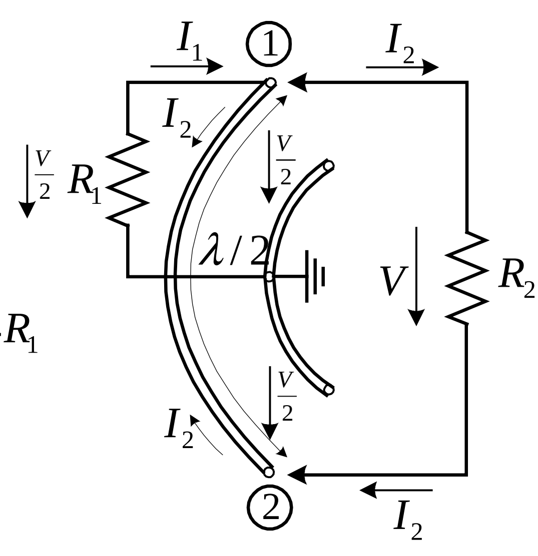
$$\frac{I_3}{I_2 - I_3} = \frac{I_4}{I_1 - I_4}$$

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

3) Half-wavelength coaxial balun 1:4



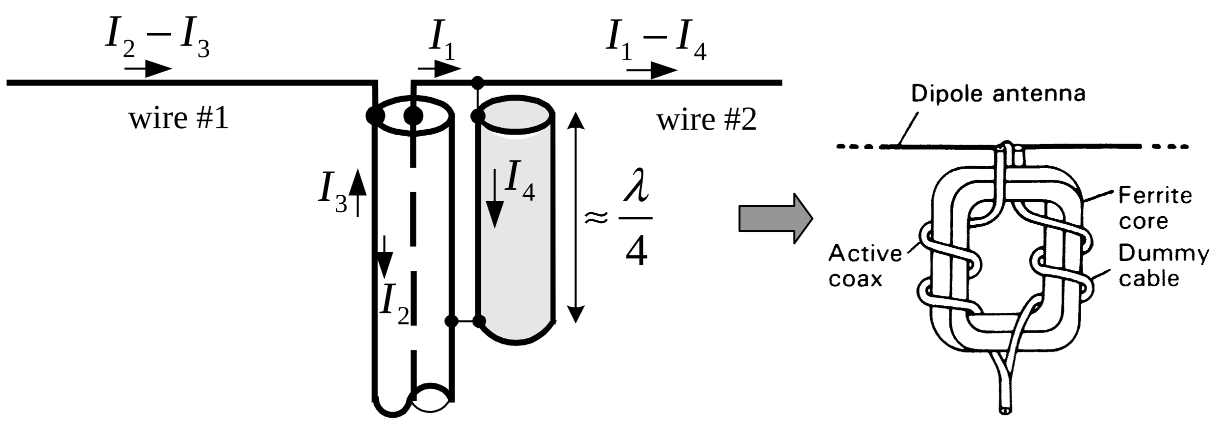
$\Rightarrow$
$R_2 = 4R_1$



Typically, a coax feed of $Z_c = 75\ \Omega$ would be connected with such a balun to a folded dipole of $Z_A \simeq 292\ \Omega$ (see equation (10.11)).

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

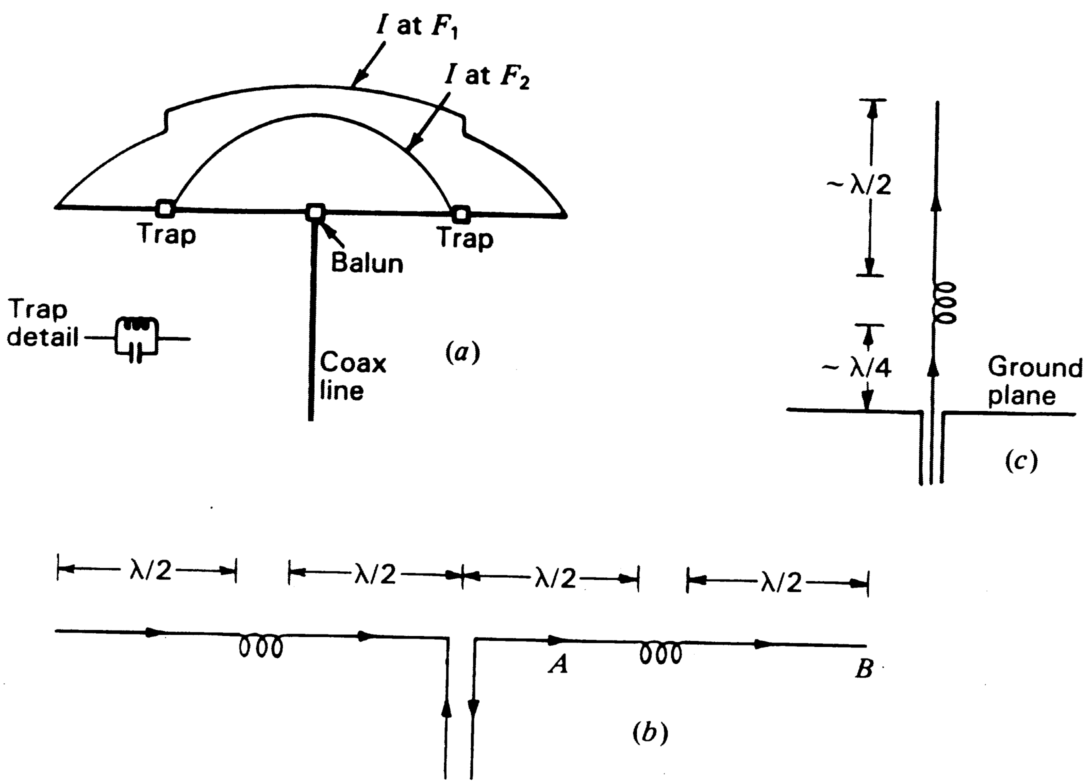
At lower frequencies (below UHF) tapered baluns are impractical, and transformers are used for impedance adjustment and balancing the feed. Often ferrite core bifilar wound wire baluns are preferred for their small dimensions and broadband characteristics (bandwidths of 10:1 are achievable). A ferrite-core 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 lines is now a very high-impedance line because of the high relative permeability of the ferrite core. Thus, its length does not depend critically on λ, in order not to disturb the antenna input impedance.

6. Dipoles with traps

In many wideband applications, it is not necessary to have frequency-independent antennas (which are more expensive and difficult to manufacture) but rather an antenna that can operate at two (or more) different bands. Typical example is the dual-band antennas in PCS and 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. Each trap represents a tuned parallel *LC* circuit. At frequency f_1, for which the whole dipole is $\simeq \lambda/2$ long, the trap is typically an inductor. This reduces slightly the resonant length of the dipole, and has to be taken into account in the antenna design. At another frequency $f_2 > f_1$, the traps become resonant and effectively cut the outer portions of the dipole, making the dipole much shorter and resonant at this new frequency. If the traps, for example, are in the middle of the dipole's legs, then $f_2 = 2f_1$ and the antenna can operate equally well at two frequencies separated by an octave. It should be noted that the isolation of the outer portions of the dipole depends not only on the high impedance of the trap but also on the impedance of this outer portion. When the outer 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. However, this is not the case if the outer portions were $\lambda/2$ each.



When the outer portions of the dipole are about $\lambda/2$ each they represent very high impedance themselves in series with the trap. They are no longer isolated. A coil only can form a trap at certain (very high) frequencies because of its own distributed capacitance. This trap would now act as a 180° phase shifter. Figure (b) 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 cellular telephony and PCS handsets. Its gain in 8.3 dBi and it has an input resistance of $\approx 150\ \Omega$.