From: O. P. Gandhi Antenna Engineering - Class Notes
Supplementary material on Slot Antennas (p. 250 of the Text)

6.6 Slot Antennas

For fast moving vehicles, it is usually necessary to design antennas that do not protrude appreciably from the surface of the vehicle. Slot antennas where the slot aperture may in fact be filled by an insulating dielectric fit the above requirement of a smooth surface perfectly. A slot antenna may be fed by a transmission line connected across it at lower frequencies or is fed by a waveguide at microwave frequencies. In order to derive the radiation pattern of a slot antenna, we shall first show that the *radiation pattern of a thin* rectangular slot is identical to that of a *complementary* metallic (dipole) strip which would just fit the slot opening. The only difference is that the orientations of $\vec{E}$ and $\vec{H}$ are interchanged in the two cases.

A slot antenna and its complementary (flat strip) antenna are shown in Fig. 6.10.

In order to derive the property that the radiaton pattern of a slot antenna resembles that of a complementary metallic strip dipole (except for the interchange of the orientations of $\vec{E}$ and $\vec{H}$ between the two), we use the *Babinet's principle*[5] of optics.

Babinet's principle states that the field at any point behind a plane having a screen, if added to the field at the same point wh[illegible] the complementary screen of case 2 (Fig. 6.11) is substituted, is equal to

[5] F. A. Jenkins and H. A. White, *Fundamentals of Optics*, McGraw-Hill Book Company, New York, 1957.

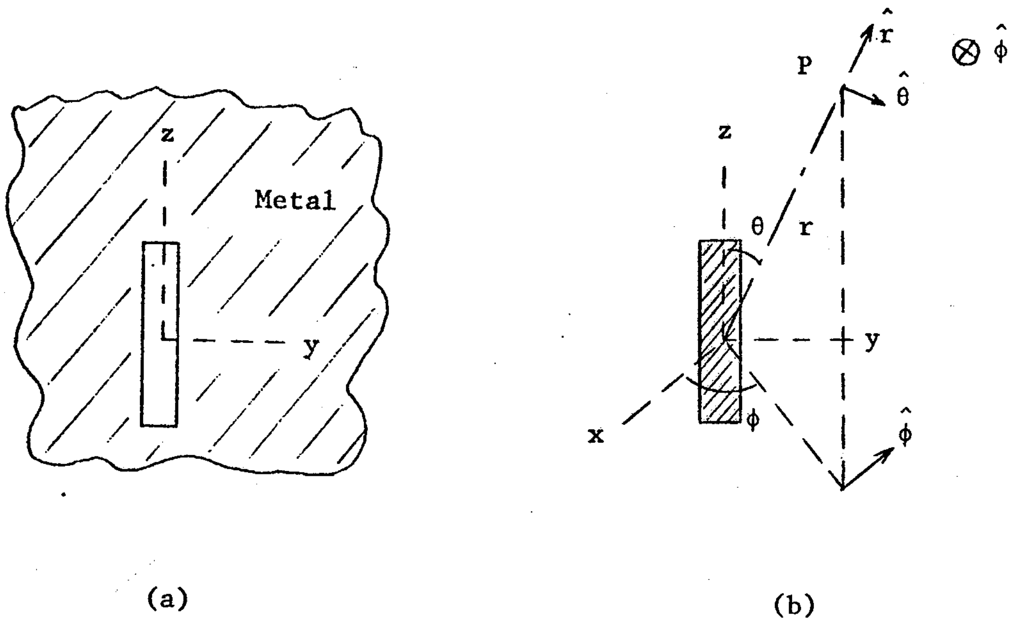


Fig. 6.10. (a) A slot antenna in a conducting screen and its
(b) complementary (flat strip) metallic antenna.

Note: In drawing the complementary antenna, the metal portions of the original antenna are replaced by free space and vice versa. The name complementary is used because of this property.

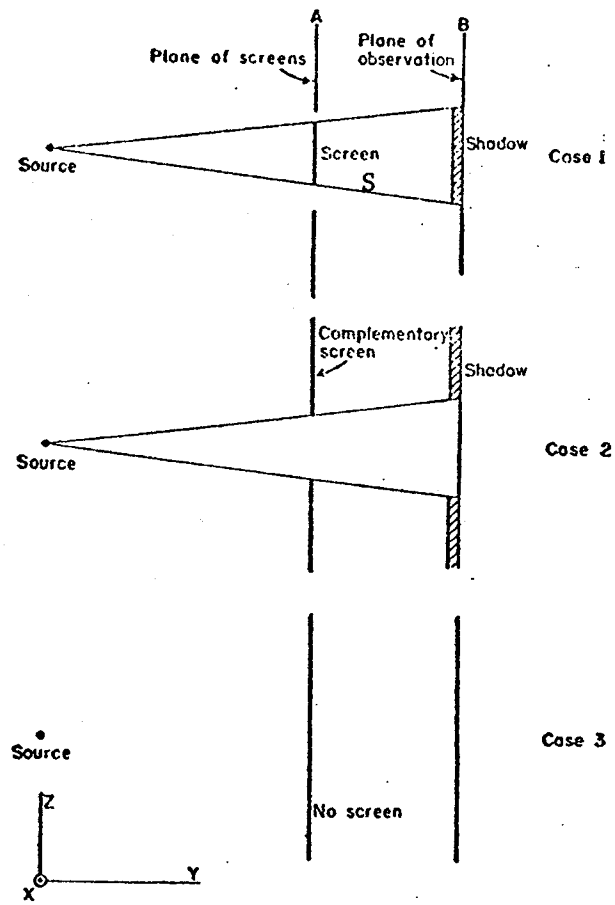


Fig. 6.11. Optical illustration of Babinet's principle.

the field at the point when no screen is present.

The optical illustration of Babinet's principle is shown in Fig. 6.11. In case 1, let a perfectly absorbing screen be placed at plane A. Then at plane B of the screen there is a region of shadow as indicated. Let the field at plane B be some function $f_1(x, y, z)$.

For case 2, the screen S is replaced by the complementary screen such that the metal portion is replaced by a free space and vice versa, and let the field at plane B be given by $f_2\ (x, y, z)$. In case 3 with no screen present at plane A, let the field at plane B be given by $f_1\ (x, y, z)$.

From the Babinet's principle, therefore,

$$f_1\ (x, y, z) + f_2\ (x, y, z) = f_o\ (x, y, z) \tag{6.84}$$

From Fig. 6.11 it is easily seen that the shadows created at plane B are complementary. The source may be a point source as in the example above or a distribution of sources. The principle applies also when the points of observation are not necessarily on a plane as in Fig. 6.11.

Babinet's principle has been extended and generalized by Booker[6] to define the vector nature of the electromagnetic fields.

By looking at Fig. 6.12 illustrating a thin slot in a metallic

---

[6] H. G. Booker, "Slot Aerials and Their Relation to Complementary Wire Antennas," *JIEE*, London, Vol. 93, Part IIIA, No. 4, 1946.

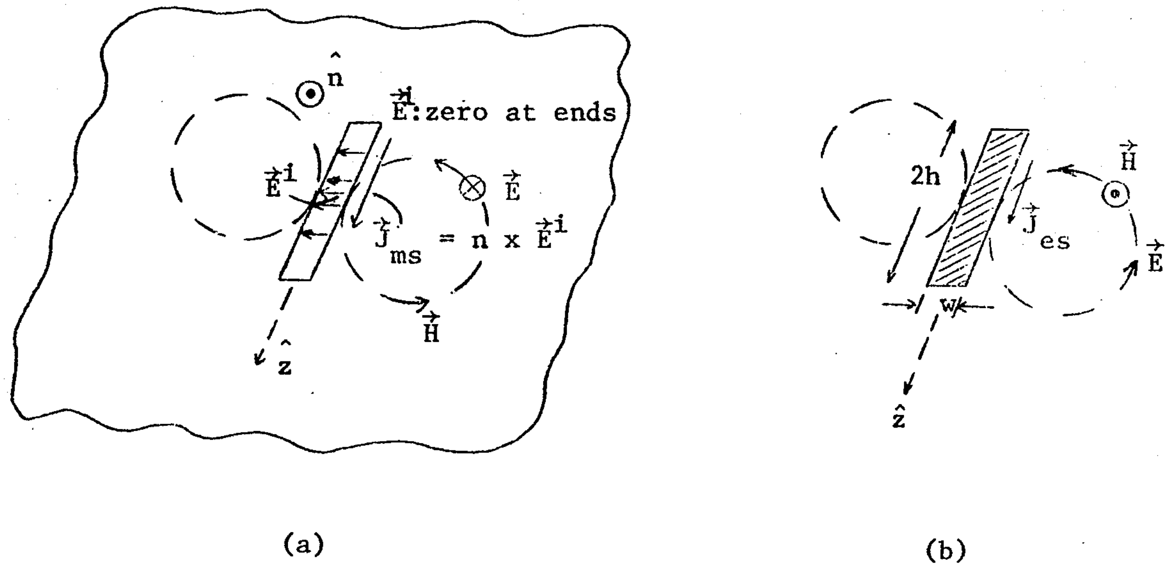


Fig. 6.12. (a) Thin slot in a metallic screen.
(b) Complementary (metallic sheet) dipole.

screen and its complementary (metallic sheet) dipole, it is obvious that the source of radiation in (a) is the distribution of the electric field and its equivalent Schelkunoff magnetic sheet current $\vec{J}_{ms}$ (parallel to the orientation of the slot which is the S axis) which is zero at the two ends of the slot (since $\vec{E}^i$, being a tangential field to the slot edges, is zero at the edges and $\vec{J}_{ms} = \hat{n} \times \vec{E}^i$ where $\hat{n}$ is the unit vector normal to the plane of the slot). The source of radiation in Fig. 6.12, on the other hand, is an electric current (which may be expressed in terms of the surface of sheet current by dividing by w the width of the strip) which, too, is zero at the two edges. The radiation pattern due to the linear antenna of

Fig. 6.12.b has been studied at length in Chapter 2. In section 2.9, the electric and magnetic fields due to a current distribution were derived as being $\hat{\theta}$ and $\hat{\phi}$ directed, respectively.

The following comparison of the radiation from the complementary dipole and that due to the slot antenna may be drawn.

Fields Due to a Linear Antenna (c)

(complementary metallic strip antenna)

Source of radiation: electric current distribution $\vec{J}_{es}$.

From Eq. (2.76)

$$\vec{A}_c = \frac{1}{4\pi}\int_{S'} \frac{\vec{J}_{es}\, e^{j(\omega t - kR)}}{R}\, dS'$$

$$= \frac{I_m}{2\pi r_o k}\, e^{j\left(\omega t - kr_o\right)}$$

$$\cdot \left\{\frac{\cos(kh\cos\theta) - \cos(kh)}{\sin^2\theta}\right\}\hat{z}$$

$$\equiv A_{cz}\,\hat{z} \tag{6.86}$$

Fields Due to a Slot Antenna (s)

Schelkunoff surface current equivalent to the electric field distribution,

$$\vec{J}_{ms} = \hat{n} \times \vec{E}^i \tag{6.85}$$

From Eq. (6.6),

$$\vec{A}_s = \frac{1}{4\pi}\int_{S'} \frac{\vec{J}_{ms}\, e^{j(\omega t - kR)}}{R}\, dS'$$

$$= \frac{E_m}{2\pi r_o k}\, e^{j\left(\omega t - kr_o\right)}$$

$$\cdot \left\{\frac{\cos(kh\cos\theta) - \cos kh}{\sin^2\theta}\right\}\hat{z}$$

$$\equiv A_{sz}\,\hat{z} \tag{6.87}$$

where h is the half length of the slot or its complementary dipole.

I_m is the peak value of the current at the location of the current maximum on the radiator. From Eq. 2.78

$$\vec{H} = \nabla \times \vec{A}_c = jkA_{cz} \sin\theta\, \hat{\phi} \quad (6.88)$$

From the Maxwell's equation, Eq. 1.40, since $\vec{J} \equiv 0$ in free space,

$$\vec{E} = \frac{\nabla \times \vec{H}}{j\omega\varepsilon_o} = \sqrt{\frac{\mu_o}{\varepsilon_o}}\, jkA_{cz} \sin\theta\, \hat{\theta} \quad (6.90)$$

E_m is the peak value of the electric field at the location of the maximum in the slot. From Eq. 6.8

$$\vec{E} = \nabla \times \vec{A}_s = jkA_{sz} \sin\theta\, \hat{\phi} \quad (6.89)$$

From Eq. 1.39

$$\vec{H} = \frac{\nabla \times \vec{E}}{-j\omega\mu_o} = -\sqrt{\frac{\varepsilon_o}{\mu_o}}\, jkA_{sz} \sin\theta\, \hat{\theta} \quad (6.91)$$

The coordinates (r, θ, ϕ) and the respective unit vectors are marked in Fig. 6.10 (b).

It can therefore be seen that the field produced by a thin slot is the same as that of a complementary metallic sheet antenna except that the directions of $\vec{E}$ and $\vec{H}$ are interchanged. Whereas for linear sheet antennas the $\vec{E}$ is $\hat{\theta}$ directed and $\vec{H}$ is $\hat{\phi}$ directed, for slot antennas the electric field is $\hat{\phi}$ directed and $\vec{H}$ is $\hat{\theta}$ directed.

The Poynting vector or the radiation intensity distribution for the two antennas, being the product of $\vec{E}$ and $\vec{H}$, is identical for the slot antenna to that of the complementary metallic sheet dipole. As with linear antennas, there is no radiation for $\theta = 0$, so also for slot radiators there is no radiation along the length of the slot.

The same doughnut shaped variation of the radiation pattern is obtained from a slot of length 2h as with a linear antenna of length 2h.

6.6.1 Typical Slot Antennas and Their Impedances

A typical slot antenna for relatively broadband operation and its complementary dipole are shown in Fig. 6.13 (a) and (b). The

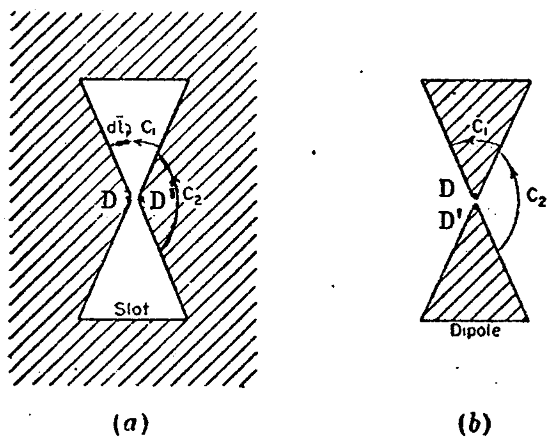


Fig. 6.13. Slot antenna and complementary dipole antenna.

driving terminals for each antenna are given by DD' and it is assumed that they are separated by an infinitesimal distance. For the complementary linear antenna, the current is continuous across DD' while voltage is not, and the converse is true for the terminals of the slot.

Let us develop a relationship between the feed point impedances of a slot and equivalent dipole antennas.

Feed point impedances of a slot and complementary dipole antennas:

For the complementary dipole

The potential difference $V_{DD'}$ between the feed points can be expressed in terms of the line integral of the electric field.

$$V_{DD'} = \lim_{c_2 \to 0} \int_{c_2} \vec{E}_{dipole} \cdot \vec{d\ell} \qquad (6.92)$$

where the integration is taken over the path c_2 indicated in Fig. 6.13 (b).

The feed point currents can similarly be written as follows:

$$I_{DD'} = 2 \lim_{c_1 \to 0} \int_{c_1} \vec{H}_{dipole} \cdot \vec{d\ell} \qquad (6.94)$$

The factor 2 enters because one-half the enclosed integral is taken, the line integral over the other side being equal by symmetry.

Driving point impedance Z_{dipole} of the complementary antenna therefore is

$$Z_{dipole} = \frac{V_{DD'}}{I_{DD'}}$$

For the slot antenna

$$V_{DD'} = \lim_{c_1 \to 0} \int_{c_1} \vec{E}_{slot} \cdot \vec{d\ell} \qquad (6.93)$$

The integration in this case is over the path of the electrical field c_1 marked in Fig. 6.13 (a).

$$I_{DD'} = 2 \lim_{c_2 \to 0} \int_{c_2} \vec{H}_{slot} \cdot \vec{d\ell} \qquad (6.95)$$

Driving point impedance Z_s of the slot is

$$Z_s = \frac{V_{DD'}}{I_{DD'}}$$

$$= \frac{\lim_{c_2 \to 0} \int_{c_2} \vec{E}_{dipole} \cdot d\vec{\ell}}{2\lim_{c_1 \to 0} \int_{c_1} \vec{H}_{dipole} \cdot d\vec{\ell}} \qquad (6.96)$$

$$= \frac{\lim_{c_1 \to 0} \int_{c_1} \vec{E}_{slot} \cdot d\vec{\ell}}{2\lim_{c_2 \to 0} \int_{c_2} \vec{H}_{slot} \cdot d\vec{\ell}} \qquad (6.97)$$

But from Eqs. 6.90, 6.91, and 6.88, 6.89,

$$\vec{E}_{dipole} = \frac{\mu_o}{\varepsilon_o} \vec{H}_{slot} \frac{I_m}{E_m} \qquad (6.98)$$

and

$$\vec{H}_{dipole} = \frac{I_m}{E_m} \vec{E}_{slot} \qquad (6.99)$$

From Eqs. 6.96 to 6.99, therefore,

$$Z_{dipole} = \frac{\frac{I_m}{E_m} \frac{\mu_o}{\varepsilon_o} \lim_{c_2 \to 0} \int_{c_2} \vec{H}_{slot} \cdot d\vec{\ell}}{\frac{I_m}{E_m} 2\lim_{c_1 \to 0} \int_{c_1} \vec{E}_{slot} \cdot d\vec{\ell}} = \frac{\eta^2}{4Z_s} \qquad (6.100)$$

where $\eta = \sqrt{\mu_o/\varepsilon_o} = 377$ ohms is the intrinsic impedance of free space.

From Eq. 6.100 the driving point impedance of a slot antenna is given by $(377)^2/4Z_{dipole}$ where Z_{dipole} is the driving point impedance of the metallic sheet antenna complementary to the slot.

Some typical electric dipoles,[7] their complementary slot antennas

---

[7] J. D. Kraus, *Antennas*, McGraw-Hill Book Company, New York, 1950, Chapter 13.

and the methods of feeding these are illustrated in Fig. 6.14. In each of the cases of Fig. 6.14, the slot width w is taken to be twice the diameter D of the circular conductor of electric dipole.

In Fig. 6.14 (a) a half wave dipole antenna and its complementary half wavelength slot antenna are shown. For the antenna length to diameter ratio $2h/d \geq 20{,}000$; i.e., for infinitesimally thin dipole antenna, the antenna impedance $Z_{dipole} = 73 + j42\ \Omega$ (from graphs in Fig. 2.33 (a) and (b)). The input impedance of the $\lambda/2$ slot antenna therefore is $(377)^2/4(73 + j42) = 363 - j211\ \Omega$. Power may be fed to the slot antenna from a 363 Ω characteristic impedance transmission line provided the capacitive $-j211\ \Omega$ part of the antenna input impedance is compensated by an inductor of value $j211\ \Omega$. The slot antenna of a somewhat reduced length ($\sim 0.475\ \lambda$) would, on the other hand, present a purely resistive impedance of $(377)^2/4 \times 70 = 530\ \Omega$, simplifying the feeder arrangement on account of the lack of need for a compensating reactance.

It should be noted, however, that the characteristic impedances required for the feeder lines for the slot antennas of Fig. 6.14 (b) and (d) are too large for coaxial lines that are preferred at the VHF/UHF frequencies where the slot antennas are advantageously used. A reduced "full" wavelength slot antenna of $L \simeq 0.925\ \lambda$ would, on the other hand, present an input impedance of $(377)^2/4 \times 710 = 50\ \Omega$, facilitating the power feeding to this antenna by a coaxial line of $Z_o = 50\ \Omega$.

From the above discussion, it is obvious that a $0.925\ \lambda$ slot

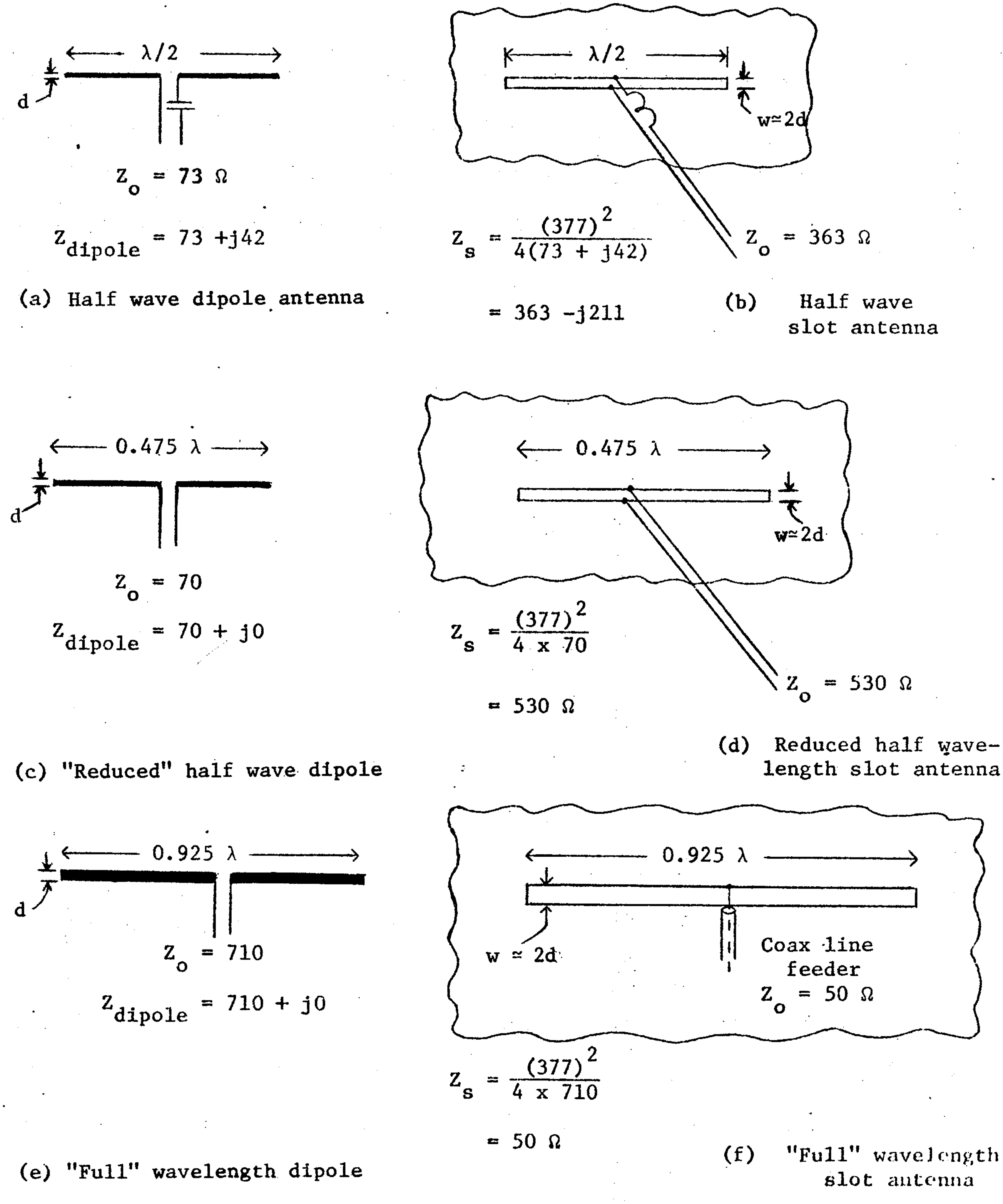


Fig. 6.14. Electric dipoles and their complementary slot antennas.

is ideal from the point of view of feeding from a standard 50 ohm coaxial line. If, on the other hand, a $\lambda/2$ length slot of Fig. 6.14 (b) is preferred, say on account of its radiation pattern, and a coaxial line (rather than open wire) type feeder is desired, an off-center feed illustrated in Fig. 6.15 may be used.

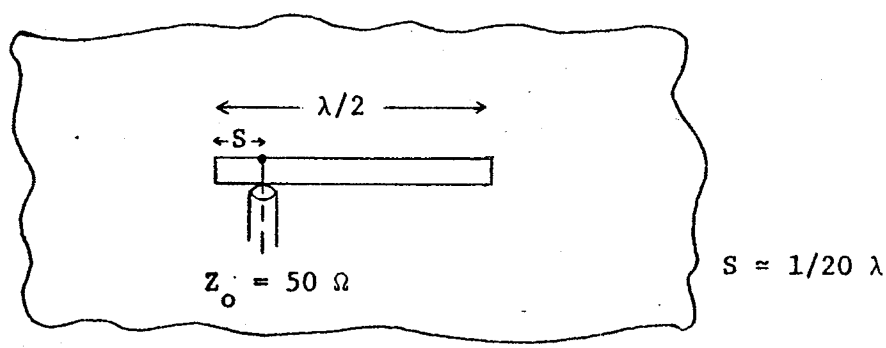


Fig. 6.15. A half wavelength slot antenna fed from a coaxial line of 50 ohm characteristic impedance.

6.6.2 Fishtail or Batwing Antenna

From the foregoing discussion on slot antennas, it is seen that a perfectly conducting sheet is needed. In practice, however, the fishtail or batwing construction of tubular conductors illustrated in Fig. 6.16 has been found to do a reasonable job of replacing a solid infinite conducting sheet. The batwing construction has a much lower weight and wind resistance and is quite popular for television transmitting antennas.

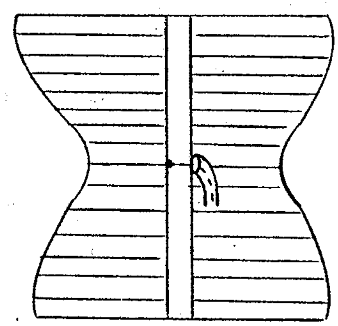


Fig. 6.16. Batwing antenna.

6.6.3 Folded Slot Antennas

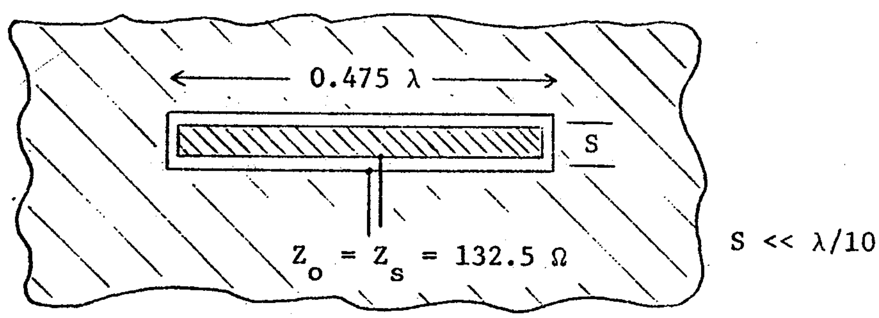


Fig. 6.17. A two-folded slot antenna.

The complementary antenna of a two-folded slot is a two-folded strip dipole which was discussed at length in Section 5.2. Since the impedance of the complementary strip antenna for arms of identical widths (from Section 5.2) is quadrupled, the impedance of a two-folded slot is reduced by a factor of 4 [since $Z_s = (377)^2/4Z_{dipole}$].

For a two-folded 0.475 λ length slot antenna, the input impedance, therefore, is

$$Z_s = \frac{530}{4} = 132.5 \text{ ohms} \tag{6.101}$$

which is not as high as the 530-ohm impedance of a single slot of 0.475 λ length.

For a three-folded slot antenna (each arm of length 0.475 λ), similarly

$$Z_s = \frac{530}{9} = 59 \text{ ohms} \tag{6.102}$$

and a coaxial line of 59-ohm characteristic impedance can be used to feed such a slot antenna.

As in folded dipoles, it is necessary to keep the separation S between the various arms much smaller than $\lambda/10$.

6.6.4 Boxed-in Slot Antenna

Since the radiation pattern of a slot antenna is identical to that of the complementary linear antenna, a slot antenna radiates therefore equally on both sides of the screen. In order to make the radiation pattern unidirectional, a boxed-in slot antenna of Fig. 6.18 may be used.

For the depth d of the box of $\lambda/4$, the radiation sent in the back direction undergoes a 180° phase shift upon reflection at the back metallic wall of the box and a 180° phase shift going and returning; i.e., a total distance $2d = \lambda/2$. This results in a phase coherent addition on the front side, meaning a four-times as much

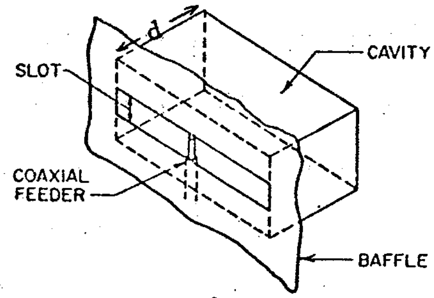


Fig. 6.18. Boxed-in rectangular slot antenna.

radiation intensity on the front side (half volume) as that of a simple slot antenna radiating bidirectionally. A four-times improvement in the radiation intensity for half the volume means a two-fold improvement in the radiated power or a two-fold increase in the driving point (radiation) resistance.

For a 0.475 λ long slot this corresponds to a slot impedance of 2 x 530 = 1060 + j0 ohms.

Slots provide a particularly desirable form of high frequency antennas for high speed vehicles. By closing the aperture with insulating material, such an antenna does not affect the streamlining of the vehicle. Slots are, however, limited to use at UHF frequencies and above because to radiate efficiently, the slot length must be on the order of a half wavelength or more and this implies reasonable slot dimensions only at UHF or higher frequencies.

6.6.5 Annular Slot Antennas[8]

In aircraft applications particularly, there is frequently a requirement for an antenna with a radiation pattern similar to that of a monopole above ground; i.e., with a radiation independent of the angle in the plane perpendicular to that of the monopole but which is flush mounted with the aircraft surface. Annular antennas shown in Fig. 6.19 (a), (b), and (c) have the radiation patterns desirable for such applications. The annular antennas are excited from a coaxial line whose conductor diameters are increased to the required diameters of the annular slot antenna. Because the electric field of the TEM mode in the coaxial line bringing power to the radiator is purely radial (see Section 6.3), the equivalent magnetic sheet current given by $\hat{n} \times \vec{E}$ is directed along the ring which is a situation similar to that of a loop antenna (Section 5.7). The radiation pattern of an annular slot antenna is therefore similar to that of the complementary loop antenna of the same dimensions as that of the slot except that the $\vec{E}$ and $\vec{H}$ field orientations for the two antennas are interchanged. The radiation is zero in the direction normal to the plane of the annular slot and is symmetrically distributed with respect to the azimuthal angle. The maximum of the radiation pattern is in the plane of the slot ($\theta = 90°$) and the polarization of the radiated electric field is $\hat{\theta}$-directed.

The arrangement of Fig. 6.19 (c) has a cavity backing which

---

[8] W. L. Weeks, *Antenna Engineering*, McGraw-Hill Book Company, New York, 1968, Chapter 6.

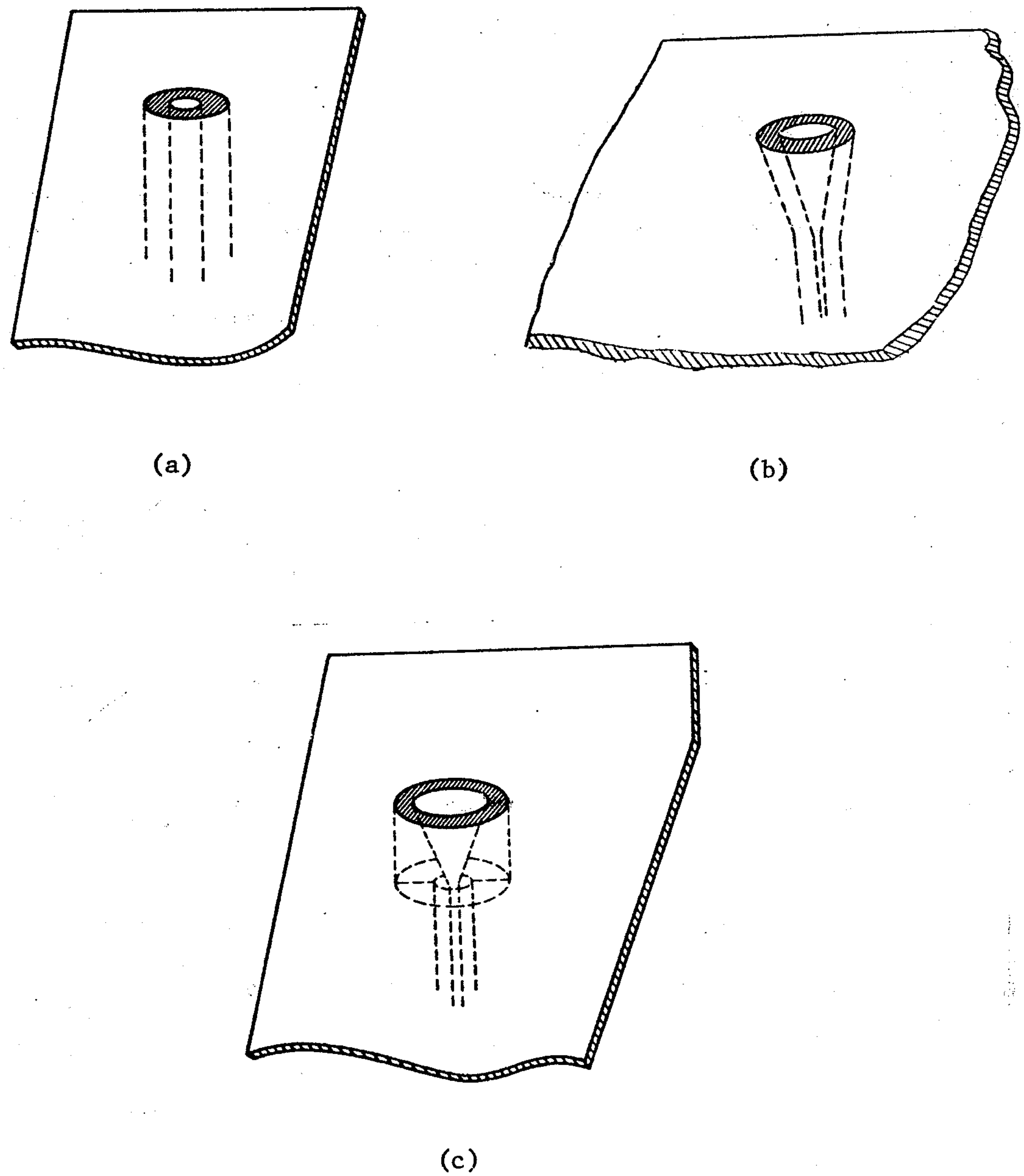


Fig. 6.19. Annular slot antennas: (a) Coaxial line terminated in a plane; (b) coaxial input line tapered to the required final dimensions; (c) annular slot with cavity backing.

would tend to reduce the bandwidth of the antenna compared to the arrangement of Fig. 6.19 (b).

6.7 Turnstile Antennas

A turnstile antenna[9] consists of two identical dipoles arranged in a crossed fashion as shown in Fig. 6.20 (a). For a 90° out-of-phase excitation $\left(I_1 = -jI_2 = I_m\right)$, the resulting electric field is given by (from Eq. 3.3)

$$\vec{E} = \vec{E}_1 + \vec{E}_2 = -\sqrt{\frac{\mu_o}{\varepsilon_o}}\,\frac{I_m\hat{\theta}}{2\pi r_o}\left\{F(\theta)\ \sin\left(\omega t - kr_o\right) + F\left(\frac{\pi}{2} + \theta\right)\ \cos\left(\omega t - kr_o\right)\right\} \tag{6.103}$$

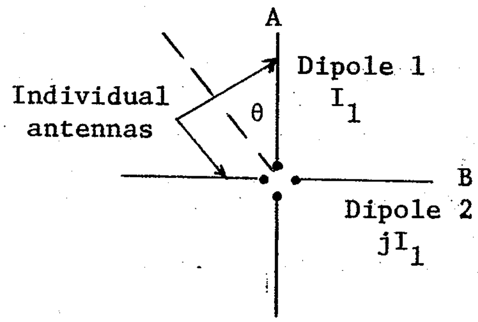
where I_m is the peak value of the current in antenna 1 at the point of current maximum. For the dipole length less than or equal to $\lambda/2$, the maximum value of the current is at the driving (center) point of the dipole.

In writing Eq. 6.103, the real parts of the two electric fields due to the dipoles are vectorially added (from Eq. 3.3). The fact that the angle to the field point is $\pi/2 + \theta$ for the second dipole carrying a current $I_2 = jI_1$ (90° out of phase with I_1) is also used in Eq. 6.103.

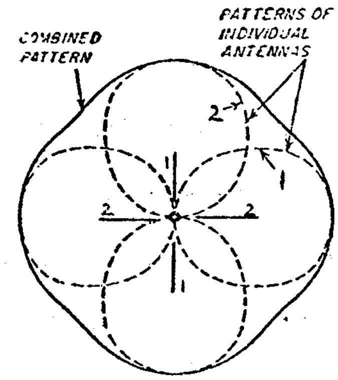
The directional patterns of the individual antennas and of the combination, in the plane of the turnstile, is illustrated in Fig. 6.20 (b). Since the individual radiations are in time quadrature,

---

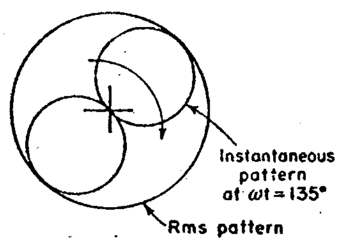
[9] G. H. Brown, "The Turnstile Antenna," *Electronics*, Vol. 9, 1936, p. 15.



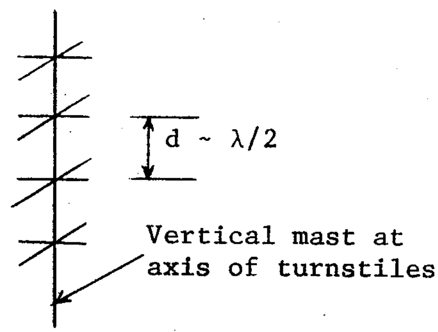
(a) A turnstile antenna for two crossed dipoles.



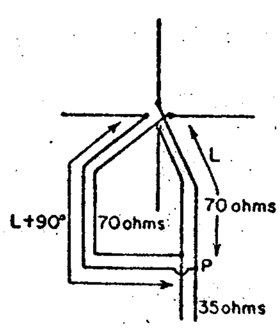
(b) Directional pattern of a turnstile antenna in the plane of turnstile.



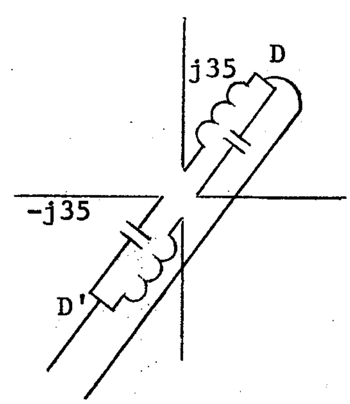
(c) The directional pattern of a turnstile of short dipoles.



(d) A four-bay turnstile antenna.



(e) Feeding by transmission lines of different lengths.



(f) Lumped element method.

Fig. 6.20. Some turnstile antenna feeder arrangements.

the total radiation in any direction from the turnstile is the square root of the sum of the square of the radiations from the individual antennas in that particular direction. As a result the pattern is very nearly circular in the plane of the turnstile as shown in Fig. 6.20 (b).

If the individual antennas are <u>short dipoles</u> (rather than $\lambda/2$ length dipoles), $F(\theta) = kh/2 \sin\theta$ from Eq. 3.2 and Eq. 6.103 can be written as:

$$\vec{E} = -60 \frac{I_1 \hat{\theta}}{r_o} \frac{kh}{2} \left\{ \sin\theta \sin\left(\omega t - kr_o\right) + \cos\theta \cos\left(\omega t - kr_o\right) \right\}$$

$$= -60 \frac{I_1 \hat{\theta}}{r_o} \frac{kh}{2} \cos\left(\omega t - kr_o - \theta\right) \tag{6.104}$$

At a given distance r_o, at any value of θ, the maximum value of E is the same and is given by $30\left(I_1 kh/r_o\right)\hat{\theta}$. Hence the field distribution is exactly circular as shown in Fig. 6.20 (c).

A turnstile antenna may be conveniently mounted on a vertical mast as shown in Fig. 6.20 (d). To increase the vertical plane directivity, several turnstile units can be stacked at about $\lambda/2$ intervals and excited in phase with one another to obtain the in-phase addition of the radiations of the various turnstile elements in the plane perpendicular to the mast. The arrangement of Fig. 6.20 (d) is called a "four-bay" turnstile.

In order that the currents on the $\lambda/2$ dipoles of a turnstile be in phase quadrature, the dipoles may be connected to separate nonresonant lines (of characteristic impedance equal to the load impedance) of unequal length. For the terminal impedance of each dipole in a single-bay turnstile being 70 + j0 (0.475 λ rather than 0.5 λ length dipoles), 70 Ω coaxial lines are used as connecting lines to P as in the schematic diagram of Fig. 6.20.e, with the length of one line 90 electrical degrees ($\lambda/4$) longer than the other. In such a case, the dipoles will be driven with currents of equal magnitude but 90° out of phase. By connecting a 35 Ω characteristic impedance line between the junction P and the transmitter, the entire transmission line system is matched.

An easier method of obtaining the required phase quadrature is by the use of lumped series reactances as shown in Fig. 6.20.f. By introducing a series inductive reactance of +j70 for one dipole and a series capacitive reactance of -j70 for the other, the currents in the two dipoles can be written in terms of the feed points DD' voltage $V_{DD'}$:

$$I_1 = \frac{V_{DD'}}{70 + j70} = \frac{V}{99} \underline{/-45^\circ} \tag{6.105}$$

$$I_2 = \frac{V_{DD'}}{70 - j70} = \frac{V}{99} \underline{/+45^\circ} \tag{6.106}$$

so that I_1 and I_2 are equal in magnitude but 90° out of phase. The net impedance at DD' due to the two parallel impedances 70 + j70 and 70 - j70 is given by

$$Z_{DD'} = \frac{1}{\{1/(70 + j70)\} + \{1/(70 - j70)\}} = 70 + j0 \text{ ohms} \qquad (6.107)$$

This may be conveniently matched to a 70 Ω characteristic impedance coaxial by using a bazooka balun arrangement of Fig. 4.27.d. Note that in the feeder arrangement of Fig. 6.20.f, the required series reactances of ± j70 Ω for the two dipoles are connected in two halves of ± j35 Ω for each of the arms for balanced feeding.

In several applications it is desirable to radiate or receive signals that are *circularly polarized*. For the 90° out-of-phase crossed dipoles forming the turnstile antenna, it can be readily shown that the radiation in the directions perpendicular to the plane of the dipole is composed of two equal magnitude right angle polarized and 90° out-of-phase components resulting, thereby, in circular polarization of the total fields.

The reason for the $\lambda/2$ separation between the various elements of the mast-mounted turnstile antenna of Fig. 6.20.d is now clear. This is to reduce by cancellation the radiation along the mast from the various elements that are excited in phase but arrive at the field points 180° out of phase from one another.

6.8 The Superturnstile Antenna

In order to obtain a low VSWR over a considerable bandwidth and to obtain zero radiation in the directions perpendicular to the plane of the turnstile, the turnstile described above has been modified to the form shown in Fig. 6.21. In this arrangement called a "super-

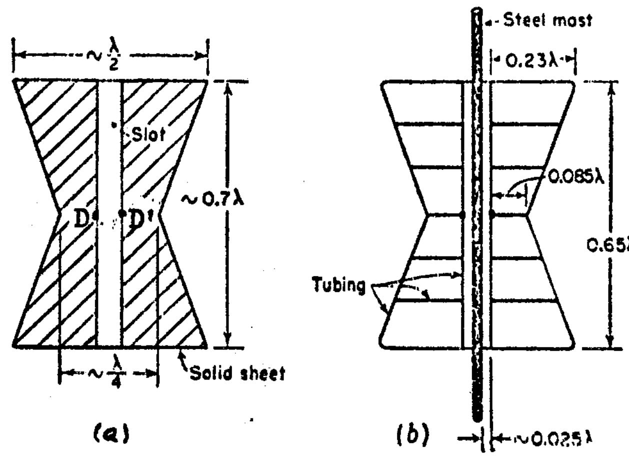


Fig. 6.21. Each element of the superturnstile antenna. (a) Solid sheet batwing construction; (b) gridded batwing construction showing also method of mounting on mast.

turnstile," the linear dipole elements are replaced by slots in flat sheets at right angles to each other or their electrical equivalent, the gridded batwing type construction shown in Fig. 6.21.b. The terminals are at DD'. The arrangement gives a VSWR of about 1.1 or less over about a 30 percent bandwidth which makes it convenient as a *mast-mounted TV transmitting antenna* (Fig. 6.22) for frequencies as low as 50 MHz.

Unlike the simple turnstile of Fig. 6.20.d, there is relatively

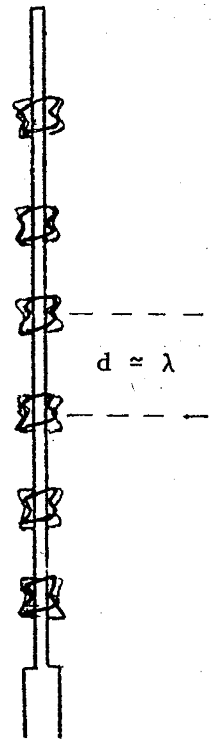


Fig. 6.22. Six-bay superturnstile antenna.

little radiation in the axial direction (along the mast). For a narrower beam width in the vertical plane, the superturnstile bays are stacked at intervals of about λ between centers and excited in phase with one another. The maximum allowed spacing d of nearly a wavelength is used to obtain the minimum beam width while keeping the principal radiation only in the horizontal plane (see section 3.2.1).