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# Absorption Efficiency of Receiving Antennas

J. Bach Andersen, *Life Fellow, IEEE*, and Aksel Frandsen, *Member, IEEE*

**Abstract**—A receiving antenna with a matched load will always scatter some power. This paper sets an upper and a lower bound on the absorption efficiency (absorbed power over sum of absorbed and scattered powers), which lies between 0 and 100% depending on the directivities of the antenna and scatter patterns. It can approach 100% as closely as desired, although in practice this may not be an attractive solution. An example with a small endfire array of dipoles shows an efficiency of 93%. Several examples of small conical horn antennas are also given, and they all have absorption efficiencies less than 50%.

**Index Terms**—Antennas, receiving, scattering.

## I. INTRODUCTION

USUALLY, the scattering properties of receiving antennas are not considered, since most important antenna properties, such as pattern, gain, and impedance, are the same as those for transmitting antennas. However, there are a few situations where the scattering is of importance, and recently there have been discussions in the literature [1], [2] concerning the absorption efficiency, which we here define as the ratio between the absorbed power and the sum of the absorbed and scattered powers. It is generally believed that this quantity is less than or equal to 50%, although it has been shown [1], [3] that this need not be the case. The scattering depends on the matching and can be divided into two parts: one that is due to mismatch at the load, called the antenna scattering or reradiation, since it has the same pattern as the transmitting antenna; and one that is what is left when the antenna is matched, called the residual or structural scattering. Only the residual scattering is considered in this paper.

Several known cases are worth mentioning. The first is the classical minimum scattering antenna, which is a one-mode antenna (like a thin dipole), which has the same scattering pattern as transmitting pattern. It follows from a simple analysis that the absorption efficiency is 50%, i.e., an equal amount of power is absorbed and scattered. The second example is an electrically large flat antenna (or array), where the forward scattering pattern again equals the transmitting pattern, since the sources have the same distribution. The sources of the scattering pattern must equal the incident with a minus sign to create the shadow (zero field) area behind the absorber. Again the absorption efficiency equals 50%. This type of argumentation has led many to believe that the absorption efficiency always equals 50%. In this paper, we demonstrate that there can be a significant variety of absorption efficiencies.

Recently, Munk [4] published an extensive analysis of the bistatic scattering from arrays in the back half-space, showing

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that the back scattering may be reduced to zero. This is usually what is of practical interest; our approach here is different, seeking to find basic limitations on the absorption efficiency, which is related to the total scattered power, not just in a certain region of space. Although it is known by examples that the absorption efficiency may be larger than 50%, it is not known how close to 100% it can be. It is one of the results of this paper that theoretically it may approach 100% as closely as desired, although of course not without a price. It is worth emphasizing that the absorption efficiency as defined here is not related to the aperture efficiency, which is a quantity related to the absorption area relative to the area of a uniformly illuminated aperture [2]. The aperture efficiency is a receiving-transmitting property of an antenna, and as such not necessarily related to the scattering from the antenna.

This paper is organized such that first the fundamental bounds are given as dictated by the universally valid *forward scattering theorem*, also known as the *optical theorem*. In Section III, some examples are presented, including a case of an antenna with high absorption efficiency. A discussion concludes this paper.

## II. FUNDAMENTAL RELATIONSHIPS FOR THE SCATTERING FROM ANTENNAS

As shown in Fig. 1, a receiving antenna can be thought of as an absorbing scatterer, and general relationships for scatterers can thus be applied.

As is customary and useful, we define the *total* field as the sum of the *incident* and *scattered* fields, i.e.,

$$\begin{aligned} \mathbf{E}_t &= \mathbf{E}_i + \mathbf{E}_s \\ \mathbf{E}_s &= \frac{e^{-jkr}}{kr} \mathbf{F}(\hat{\mathbf{r}}) \end{aligned} \quad (1)$$

where  $\mathbf{F}(\hat{\mathbf{r}})$  is called the far-field pattern function. The corresponding scattered power flow is, from Poynting's vector

$$\mathbf{S}_s = \frac{\hat{\mathbf{r}}}{2\zeta_0(kr)^2} |\mathbf{F}(\hat{\mathbf{r}})|^2 \quad (2)$$

with  $\zeta_0$  being the free-space impedance.

Assume a plane wave incident from a direction  $(\theta, \phi) = (\theta_i, \phi_i)$

$$\mathbf{E}_i(\mathbf{r}) = \mathbf{E}_0 \exp(-jk\hat{\mathbf{k}}_i \cdot \mathbf{r}) \quad (3)$$

where  $\mathbf{E}_0$  is a complex vector, and  $\hat{\mathbf{k}}_i$  is the propagation vector

$$\hat{\mathbf{k}}_i = -(\sin\theta_i \cos\phi_i \hat{\mathbf{x}} + \sin\theta_i \sin\phi_i \hat{\mathbf{y}} + \cos\theta_i \hat{\mathbf{z}}). \quad (4)$$

The power density of the incident field is determined from Poynting's vector for  $\mathbf{E}_i$

$$\mathbf{S}_i = \frac{\hat{\mathbf{k}}_i}{2\zeta_0} |\mathbf{E}_0|^2. \quad (5)$$

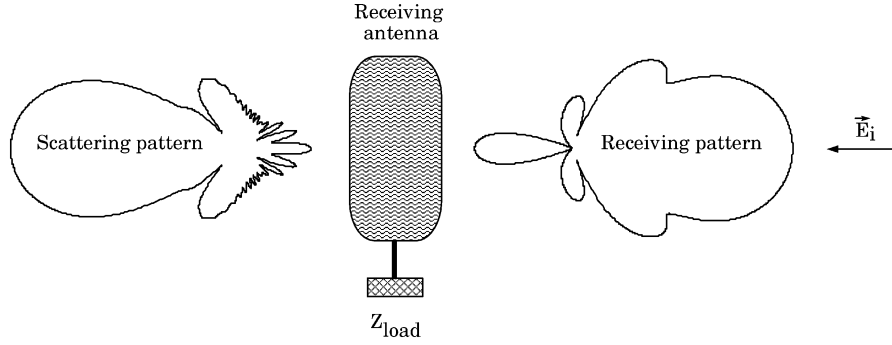


Fig. 1. Matched receiving antenna has a receiving as well as a scattering pattern. The antenna receives the incident field from the direction  $(\theta, \phi) = (\theta_i, \phi_i)$  with a directivity  $D_a$ , given by its receiving properties, and scatters in the forward direction  $(\theta, \phi) = (\pi - \theta_i, \pi + \phi_i)$  with a scattering directivity  $D_s$ . The antenna receiving and scattering patterns are in general quite different.

We can now define the (bistatic) scattering cross section (or radar cross-section) as

$$\sigma_b = 4\pi r^2 \frac{\mathbf{S}_s \cdot \hat{\mathbf{r}}}{\mathbf{S}_i \cdot \hat{\mathbf{k}}_i} = 4\pi \frac{|\mathbf{F}(\hat{\mathbf{r}})|^2}{k^2 |\mathbf{E}_0|^2}. \quad (6)$$

It has the dimension of area and signifies an equivalent source of power, equal to the area times the incident power density, giving the same power density as the scattered field in the direction  $\hat{\mathbf{r}}$ . It is thus a normalized measure of the radiation pattern of the scattered energy.

Consider now the total scattered power obtained by integrating over the far-field sphere

$$P_s = \iint \mathbf{S}_s \cdot \hat{\mathbf{r}} dS = \frac{1}{2\zeta_0 k^2} \iint |\mathbf{F}(\hat{\mathbf{r}})|^2 d\Omega \quad (7)$$

and define in turn the total scattering cross-section as the area that extracts the scattered power from the incident power density

$$\begin{aligned} \sigma_s &= \frac{P_s}{\mathbf{S}_i \cdot \hat{\mathbf{k}}_i} = \frac{1}{k^2 |\mathbf{E}_0|^2} \iint |\mathbf{F}(\hat{\mathbf{r}})|^2 d\Omega \\ &= \frac{1}{4\pi} \iint \sigma_b d\Omega. \end{aligned} \quad (8)$$

It is natural to make an analogy with a transmitting antenna, where the total radiated power now corresponds to the scattered power from the scatterer, and we easily find that the directivity  $D_s$  of the scattering pattern in the direction  $\hat{\mathbf{r}}$  equals

$$D_s(\hat{\mathbf{r}}) = \frac{4\pi |\mathbf{F}(\hat{\mathbf{r}})|^2}{\iint |\mathbf{F}(\hat{\mathbf{r}})|^2 d\Omega} = \frac{\sigma_b(\hat{\mathbf{r}})}{\sigma_s}. \quad (9)$$

An important general relationship for scatterers, which we will find useful for receiving antennas as well, is the so-called *optical theorem* or *forward scattering theorem*, which relates the total scattered and absorbed powers to the forward scattered field  $\mathbf{F}(\hat{\mathbf{k}}_i)$ . In the notation used previously, the optical theorem [5], [6] reads

$$\begin{aligned} \sigma_t &= \sigma_a + \sigma_s = \frac{4\pi}{k^2} \text{Im} \left\{ \frac{\hat{\mathbf{p}}_0 \cdot \mathbf{F}(\hat{\mathbf{k}}_i)}{\mathbf{E}_0} \right\} \\ &= \frac{4\pi}{k^2} \text{Im} \{ \hat{\mathbf{p}}_0 \cdot \mathbf{F}(\hat{\mathbf{k}}_i) \} \end{aligned} \quad (10)$$

in the case of a unit incident field linearly polarized along  $\hat{\mathbf{p}}_0$ . Here  $\sigma_a$  is the absorption cross-section. Normalizing the cross-

sections to  $\lambda^2$ , and assuming that  $F_l$  is the component along the incident polarization  $F_l = \hat{\mathbf{p}}_0 \cdot \mathbf{F}(\hat{\mathbf{k}}_i)$ , the equation is reduced to

$$\sigma_t = \sigma_a + \sigma_s = \frac{1}{\pi} \text{Im} \{ F_l(\hat{\mathbf{k}}_i) \}. \quad (11)$$

The optical theorem states that the total cross-section of an obstacle, i.e., the sum of the absorption and scattering cross-sections, is simply related to the  $\hat{\mathbf{p}}_0$ -component of the forward scattered field in the direction of the incident wave, i.e., directly in the shadow of the scatterer.  $\sigma_t$  is also sometimes denoted the *extinction cross-section*, since it represents the total power loss from the incident field due to scattering and absorption by the obstacle.

By utilizing that the imaginary part is less than or equal to the absolute value we arrive at the following fundamental inequality

$$\sigma_a + \sigma_s \leq \sqrt{\frac{\sigma_b(\hat{\mathbf{k}}_i)}{\pi}} \quad (12)$$

or, from (9)

$$\sigma_a + \sigma_s \leq \sqrt{\frac{D_s(\hat{\mathbf{k}}_i)}{\pi}} \sigma_s. \quad (13)$$

Equation (13) provides an interesting possibility for studying the bounds of the scattering cross-section as a function of the absorption cross-section and the directivity of the scattered field. Note that the scattered power may tend to zero and still maintain the absorbed power if the directivity of the scattered power tends to infinity.

Now let  $(D_s(\hat{\mathbf{k}}_i)/\pi) = \alpha$  in (13). We then get the following second-degree inequality for  $\sigma_s$ :

$$\sigma_s^2 + (2\sigma_a - \alpha)\sigma_s + \sigma_a^2 \leq 0 \quad (14)$$

which leads to the following lower and upper bounds on  $\sigma_s$ :

$$\alpha/2 - \sigma_a - \sqrt{\alpha(\alpha/4 - \sigma_a)} \leq \sigma_s \leq \alpha/2 - \sigma_a + \sqrt{\alpha(\alpha/4 - \sigma_a)}. \quad (15)$$

The two bounds meet when  $\sigma_a = \alpha/4$ , equivalent to

$$D_s = 4\pi\sigma_a = D_a \quad (16)$$

since the cross-sections are normalized to  $\lambda^2$ .

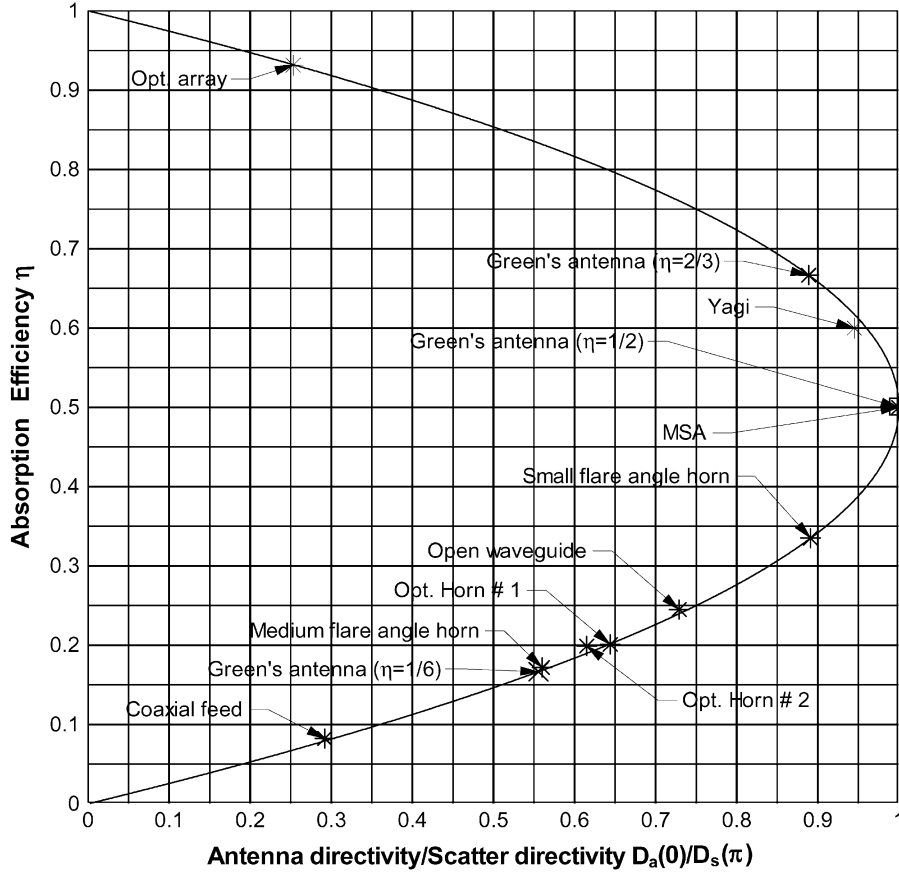


Fig. 2. Upper and lower bounds on absorption efficiency as a function of antenna directivity and scatter directivity. The points refer to the examples in Section III.

The right side of (16) is the fundamental relationship between absorption area and directivity for any antenna. In all cases considered here, we have assumed a lossless and matched antenna, so there are no reflections from the load. It should also be remembered that  $D_a$  is the directivity of the antenna in the direction looking toward the source ( $-\hat{\mathbf{k}}_i$ ), while  $D_s$  is the directivity of the scattered field in the forward direction ( $\hat{\mathbf{k}}_i$ ), opposite to the direction to the source (Fig. 1). It also follows that at the point where the two bounds coalesce

$$\sigma_s = \alpha/2 - \sigma_a = 2\sigma_a - \sigma_a = \sigma_a \quad (17)$$

i.e., the two directivities as well as the two cross-sections are the same. This situation is valid for many antennas, from a simple dipole to a large flat aperture antenna.

Let us define the absorption efficiency  $\eta$  as

$$\eta = \frac{\sigma_a}{\sigma_a + \sigma_s}. \quad (18)$$

An upper and lower bound on  $\eta$  may now be found from (15).

Consider the right-hand-side inequality of (15)

$$\sigma_s \leq \alpha/2 - \sigma_a + \sqrt{\alpha(\alpha/4 - \sigma_a)} \quad (19)$$

or

$$\sigma_a + \sigma_s \leq \frac{D_s}{2\pi} \left( 1 + \sqrt{1 - \frac{D_a}{D_s}} \right) \quad (20)$$

and hence

$$\begin{aligned} \eta &= \frac{\sigma_a}{\sigma_a + \sigma_s} \geq \frac{D_a}{2D_s} \frac{1}{\left(1 + \sqrt{1 - \frac{D_a}{D_s}}\right)} \\ &= \frac{D_a}{2D_s} \frac{\left(1 - \sqrt{1 - \frac{D_a}{D_s}}\right)}{\left(1 + \sqrt{1 - \frac{D_a}{D_s}}\right)\left(1 - \sqrt{1 - \frac{D_a}{D_s}}\right)} \\ &= \frac{1}{2} \left(1 - \sqrt{1 - \frac{D_a}{D_s}}\right) \end{aligned} \quad (21)$$

which gives a lower bound on  $\eta$ . A similar derivation based on the left-hand side of (15) gives an upper bound, so finally

$$\frac{1}{2} \left(1 - \sqrt{1 - \frac{D_a}{D_s}}\right) \leq \eta \leq \frac{1}{2} \left(1 + \sqrt{1 - \frac{D_a}{D_s}}\right). \quad (22)$$

For the traditional minimum scattering antenna (MSA),  $\eta = 0.5$  and  $(D_a(-\hat{\mathbf{k}}_i)/D_s(\hat{\mathbf{k}}_i)) = 1$ , by virtue of the restrictions the MSA property imposes on the radiation and scattering patterns. For other situations, the case is illustrated in Fig. 2, which shows the upper and lower bounds on the absorption efficiency (22) as a function of the ratio between the antenna directivity and the scatter directivity ( $D_a(-\hat{\mathbf{k}}_i)/D_s(\hat{\mathbf{k}}_i)$ ). It is clear that the absorption efficiency may theoretically be as close to one as wanted for a sufficiently large scatter directivity. These are theoretical bounds valid for all antennas, which must lie inside the parabola. It is also an interesting corollary that, for any antenna, the directivity of the scattered field is higher than or equal to the antenna directivity. If this were not the case, the optical theorem would

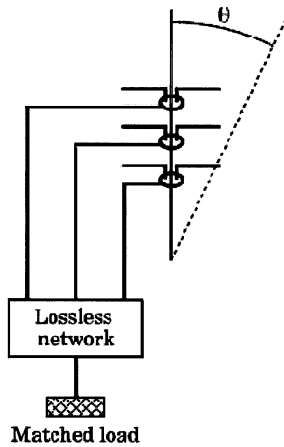


Fig. 3. Three parallel half-wave dipoles connected through a lossless network to a matched load form the antenna.

not be satisfied since the left-hand side of (14) would never be negative.

### III. EXAMPLES

#### A. An Endfire Array of Dipoles

The example is an array of parallel half-wave dipoles spaced 0.2 wavelengths and connected through a lossless network to a matched load (Fig. 3). An imaginary impedance matrix with arbitrary elements describes the network, and a search algorithm varies the elements until a solution is found which maximizes the absorption efficiency. The algorithm finds the maximum efficiency case among 10 000 random cases, then narrows the search in the next phase and so on until convergence. The incident field is incident along the array axis and copolarized with the dipoles. For each realization of the network, the scattering and absorption cross-sections are found using standard antenna theory and mutual impedances for thin dipoles, like in [1], and for the final solution the radiation and scattering patterns are determined. One optimized solution (not necessarily the global optimum) is described by the following parameters with the directivity patterns shown in Fig. 4:

$$\begin{aligned} \sigma_a &= 0.151 \lambda^2 & \sigma_s &= 0.0111 \lambda^2 \\ D_a &= 2.78 \text{ dB} & D_s &= 8.75 \text{ dB} & D_a/D_s &= 0.252 \\ \eta &= 0.93. \end{aligned}$$

The point (0.252, 0.93) is plotted in Fig. 2 and is seen to be very close to the upper bound. The antenna pattern [Fig. 4(a)] is quite wide, corresponding to the dipole-like directivity. Of course the three dipoles and the network could have been used to design a higher antenna directivity, but not without sacrificing the absorption efficiency. How to realize the network in practice has not been considered.

It is also illuminating to plot the radar cross-section (6) instead of the directivity. This is done in Fig. 5, where a single half-wave dipole is shown for comparison. The dipole has a directivity of 1.64 (2.15 dB) and a  $\sigma_s$  of  $0.13\lambda^2$ , which gives a  $\sigma_b$  of  $-6.7$  dB. The directivities of the two antennas are comparable, but the array scattering is below the dipole scattering for all angles and highest in the forward direction. In a sense this is a true “minimum scattering antenna.”

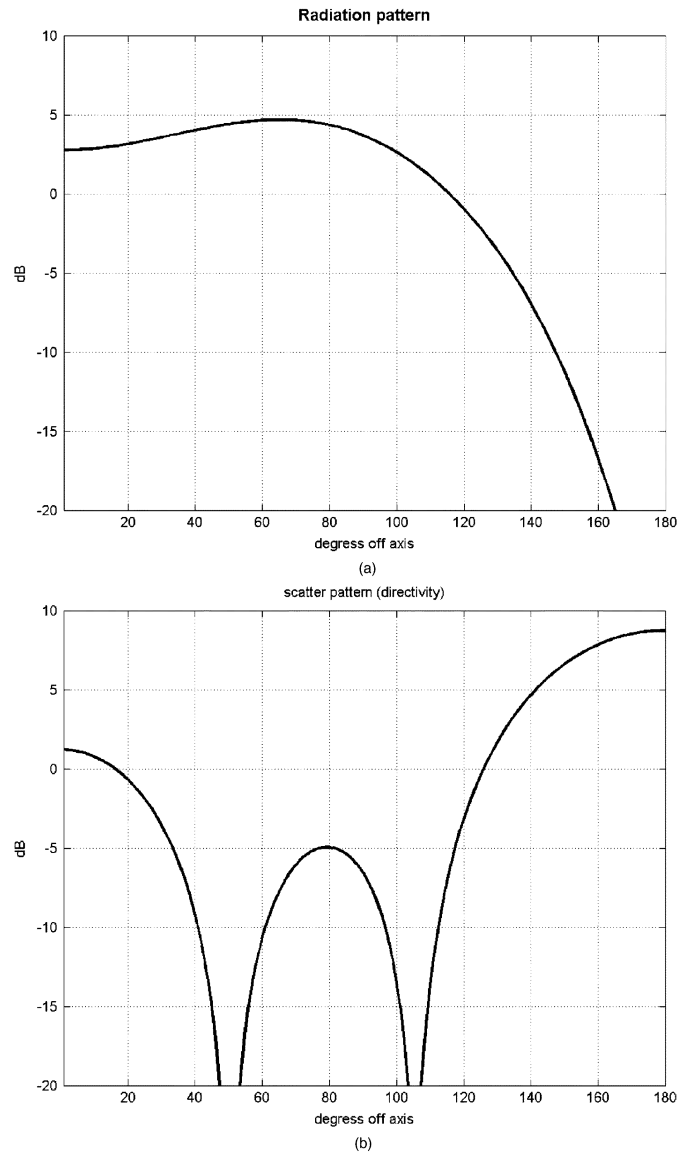


Fig. 4. (a) Radiation pattern (directivity) for an  $N = 3$  half-wave dipole array optimized for maximum absorption efficiency. (b) The scattering pattern (directivity) for the same antenna.

#### B. A Five-Element Yagi

The NEC<sup>1</sup> program has been used to calculate the parameters for a standard Yagi antenna. First the structures are excited by a source as a transmit antenna resulting in maximum directivity (and thus absorption area) and impedance. The conjugate impedance is then used as a load for an incident plane wave, resulting in scatter directivity and scattered power.

A Yagi antenna with one reflector and four directors is shown as point “Yagi” in Fig. 2. It has an absorption efficiency of 60%.

#### C. An Elementary Antenna

Green [3] has devised a matched antenna consisting of an elementary Hertzian dipole and a small loop. Properly combined, this antenna absorbs more power than it scatters, at the expense of decreasing the absorbed power, i.e., its receiving properties are not optimized. When the antenna is phased for

<sup>1</sup>SuperNEC (www.poynting.za).

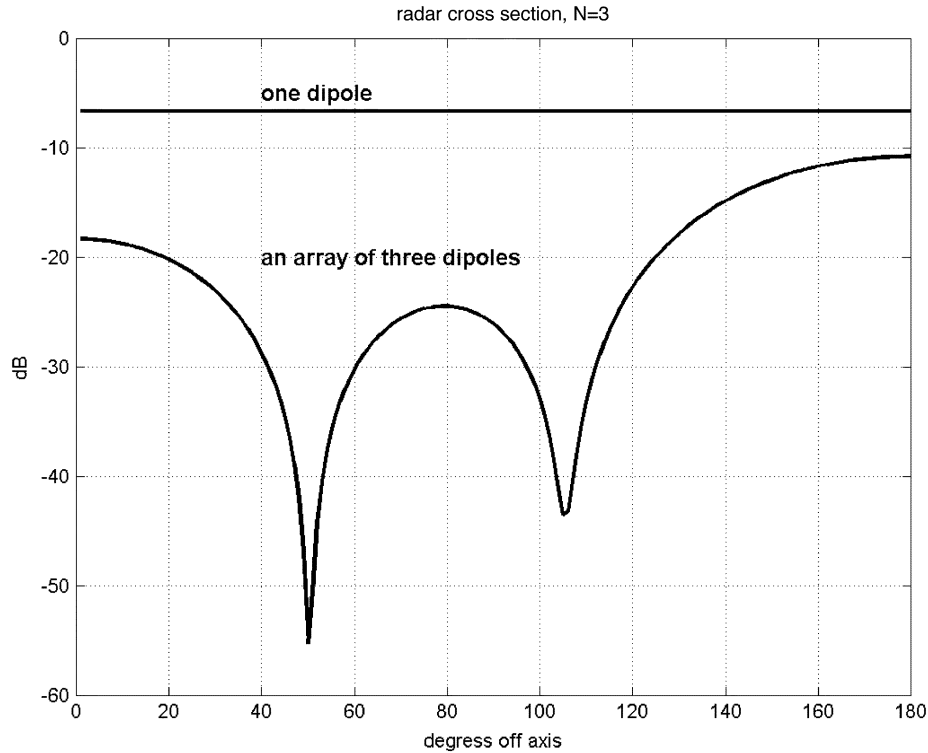


Fig. 5. Radar cross-section (6) of the three-element array compared with that of a single half-wave dipole.

maximum absorption, the scattered power equals the absorbed power. Green conveniently describes the antenna combination using scattering matrices. In doing so, two parameters  $K$  and  $e^{j\psi}$  suffice to characterize the antenna,  $0 \leq K \leq 1$  and  $-1 \leq \cos(\psi) \leq 1$ . Carrying through the analysis, one finds maximum and minimum absorption efficiencies of  $2/3$  and  $1/6$ , respectively. Specifically

$$\begin{aligned} \text{For } K = \frac{1}{3} \text{ and } \cos(\psi) = -1 : \\ \eta = \frac{2}{3} \quad D_a = \frac{24}{10} \quad D_s = \frac{27}{10} \quad D_a/D_s = \frac{8}{9} \\ \text{For } K = 0 \text{ and } \cos(\psi) = 1 : \\ \eta = \frac{1}{6} \quad D_a = \frac{3}{2} \quad D_s = \frac{27}{10} \quad D_a/D_s = \frac{5}{9} \\ \text{For } K = 1 \text{ and } \cos(\psi) = -1 : \\ \eta = \frac{1}{2} \quad D_a = 3 \quad D_s = 3 \quad D_a/D_s = 1. \end{aligned}$$

These data points are plotted in Fig. 2. It should be noted that the points lie exactly on the boundary curve for the absorption efficiency (22).

#### D. Small Conical Horn Antennas

The scattering and radiation properties of small conical horn antennas have been studied in [7] using moment methods for perfectly conducting bodies of revolution. In order to achieve a matched load condition, a “sliding short” technique was used, in which a short terminates the circular waveguide part of the horn. The horn is illuminated by a plane wave axial at incidence on the horn aperture from the direction  $\theta = 0$ . The forward direction is then at  $\theta = \pi$ . By placing the short in at least three different positions in the waveguide, and for each position calculating the

scattered field from the short-circuited horn, it is possible to extract information about the horn’s voltage standing-wave ratio, radiation pattern, and scattering pattern for arbitrary load impedances, the matched load being a special case. A convenient spacing between the short positions is  $\lambda_g/8$ , where  $\lambda_g$  is the waveguide wavelength for the fundamental  $TE_{11}$  mode. Fig. 6 shows the general geometry of a conical horn with the pertinent parameters.

Table I lists the parameters for a few conical horns, including an open ended circular waveguide. The so-called “optimum horns” are optimum in the sense that they provide the largest directivity  $D_a$  for a fixed apex length. For small apertures, this results in fairly large horn flare angles. The calculated absorption efficiencies and directivity ratios are shown to the right in Table I and also plotted in Fig. 2, with the identification tag taken from the leftmost column in Table I. It is noteworthy that, as opposed to the dipole cases in Section III-A and -B, the absorption efficiencies for these conical horns tend to be very close to the lower bound for  $\eta$ .

Fig. 7 shows the geometry of a feed horn intended for illumination of deep center-fed paraboloids. The horn is basically a choked waveguide, also sometimes referred to as a coaxial feed. The geometry is given in Table II, while the calculated results for this horn at a frequency of 8.5 GHz are

$$\begin{aligned} \eta &= 0.0824 \\ D_a(0)/D_s(\pi) &= 0.292. \end{aligned}$$

#### IV. DISCUSSION

As discussed in [1], there are several misunderstandings concerning the amount of scattered power from matched receiving

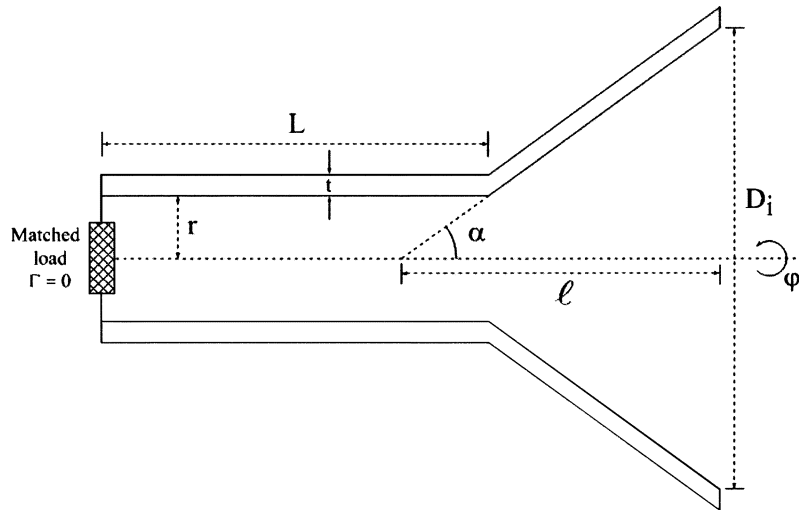


Fig. 6. Geometry of conical horn.

TABLE I  
HORN GEOMETRY PARAMETERS (Fig. 6) AND CALCULATED ABSORPTION EFFICIENCIES

Horn Identification in Figure 2	Waveguide radius $r$	Aperture diameter $D_i$	Apex length $l$	Half flare angle $\alpha$	Waveguide length $L$	Wall thickness $t$	Absorption Efficiency $\eta$	Directivity Ratio $D_a(0)/D_s(\pi)$
Open waveguide	$0.35\lambda$	$0.70\lambda$	0	0	$2.5\lambda$	$0.1\lambda$	0.245	0.729
Optimum horn # 1	$0.35\lambda$	$1.5\lambda$	$0.5\lambda$	$56.31^\circ$	$1.8\lambda$	$0.1\lambda$	0.201	0.644
Optimum horn # 2	$0.35\lambda$	$1.8\lambda$	$0.8\lambda$	$48.37^\circ$	$1.8\lambda$	$0.1\lambda$	0.199	0.615
Medium flare angle horn @ 11.816 GHz	9.0 mm	66.0 mm	45.42 mm	$36^\circ$	55.0 mm	6.0 mm	0.172	0.560
Small flare angle horn	$0.375\lambda$	$1.434\lambda$	$1.97\lambda$	$20^\circ$	$2.0\lambda$	$0.1\lambda$	0.335	0.891

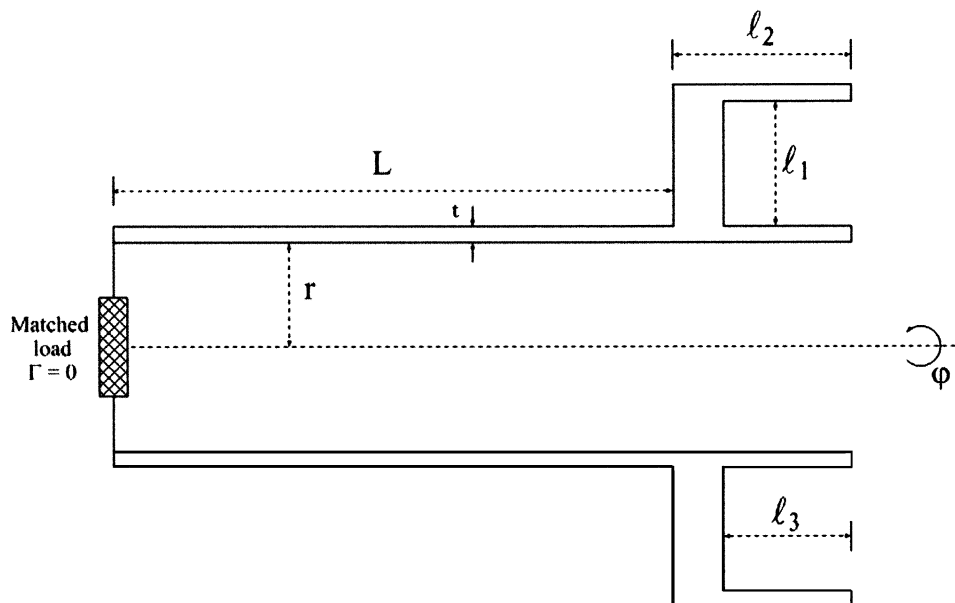


Fig. 7. Geometry of coaxial feed horn.

Table II  
GEOMETRY PARAMETERS FOR COAXIAL FEED (Fig. 7)

r	13.0 mm
$\ell_1$	15.65 mm
$\ell_2$	22.3 mm
$\ell_3$	16.0 mm
L	70.0 mm
t	2.1 mm

antennas. It is often stated that scattered power is equal to or larger than the absorbed power, which is probably due to a misconception of the notion of the so-called minimum scattering antennas, in which case it is true that the two powers are equal, leading to an absorption efficiency of 50%. It is one of the results of this paper that, literally speaking, minimum scattering antennas do not exist.

Based on fundamental concepts for scattering from objects with losses, bounds have been derived for the absorption efficiency of any antenna. The bounds depend only on the ratio between two directivities, the antenna directivity in the direction toward the source, and the directivity of the scattered field in the forward scattering direction. When the two directivities are equal, the efficiency is 50%, but may approach 100% or 0% when the scatter directivity is much larger than the antenna directivity. It should be noted that it is possible to design antennas with low backscattering with 50% efficiency [4], but here we are only concerned with the total scattered power. It is important to note that the absorption efficiency is not related to the well-known aperture efficiency of antenna theory.

It is not obvious from the theory how to design antennas with low scattered power, except that an antenna with a delta function scattering pattern in the forward direction may have zero scatter in all other directions. In this paper an example is given for an array of three half-wave dipoles coupled to a matched load via an optimized lossless network. The efficiency is 93%, and the scattered power is below that of a dipole in all directions. The directivities of the two antennas are similar. Most antennas will have absorption efficiencies below 50%, as evidenced from the many examples of conical structures, but calculations have shown that exceptions do occur. Apart from the array mentioned above, a Yagi antenna has an efficiency larger than 50%.

It is interesting to observe that, in all cases, the points lie very near or on the bounds, indicating that the imaginary part of the forward scattered field is close to being equal to the absolute value, or, in other words, the phase difference between the incident and scattered fields is close to  $90^\circ$ . For Green's antenna, this phase difference is exactly  $90^\circ$ .

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