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STUDY AND INVESTIGATION OF A UHF-VHF ANTENNA

QUARTERLY PROGRESS REPORT NO. 4

Period Covering January 1, 1961 to April 1, 1961

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SCOPE OF CONTRACT

For the period 1 April 1960 through 1 April 1961, the Contractor shall make available and employ its research and development facilities and personnel and provide necessary materials and services to perform research and development, study, and investigation directed to the design of an experimental UHF-VHF antenna for defensive electronic countermeasures used on advanced air vehicles, both manned and unmanned. The objective of this procurement is to advance the state-of-the-art of airborne UHF-VHF antennas wherein use is made of nonconventional radiating devices or techniques. In the past UHF-VHF designs have been largely limited to whips (monopole), isolated sections of the aircraft structure, and dielectric and/or resistive loaded slots. It is the intent of this procurement to exploit new techniques, such as, but not limited to, application of new solid-state advances to antenna design and the incorporation of the radiating element as a part of the power tube itself. This work shall include development and construction of specialized instrumentation required for any tests, together with working drawings, specifications, and operating instructions (which the Contractor prepares or has prepared).

ABSTRACT

The effort during this period was devoted to the theoretical study of plane waves incident on a spherical and cylindrical material body, formulation of a basis for comparison of antennas, tests on a ferrite loop antenna, and construction of a biconical antenna.

STUDY AND INVESTIGATION OF A UHF-VHF ANTENNA

QUARTERLY PROGRESS REPORT NO. 4

Period Covering January 1, 1961 to April 1, 1961

1. PURPOSE

This report summarizes the work done on Contract No. AF 33(616)-7180 during the period from January 1, 1961 to April 1, 1961.

The purpose of this task is to investigate the use of solid-state devices such as ferrites and dielectrics in their application to UHF-VHF antennas. More specifically, these materials are to be considered as loading devices or actual elements in the search for improvement of the following properties: (1) radiation resistance, (2) power gain and directivity, (3) broadbanding, (4) physical size, and (5) efficiency. Geometries now under consideration include dipoles, rods, slots, biconical dipoles, spirals, and yagis.

2. REPORTS, TRAVEL, AND VISITORS

No reports were issued during this period.

No travel was made during this period.

No persons visited the project during this period.

3. FACTUAL DATA

3.1 The Problem of a Plane Wave Incident on a Material Sphere

In determining the effect of materials on antenna characteristics, one significant indication would be the effect of the material above on the radio waves. If the permeability and permittivity are at all different from that of the medium in which the material is immersed, the field distribution will be altered; the geometry of the material also affects the field distribution. Diffraction theory deals with the fields external to the material, and a great deal of work has been done in the evaluation of these fields (Refs. 1, 2, 3). Thus it would be of interest to evaluate the interior fields. Using orthogonal vector functions, the case of a spherical medium has been rigorously solved in diffraction theory. An extension is made of this to treat the fields inside the sphere.

Consider a plane wave propagating in the z-direction with its electric field polarized in the x-direction; then the incident fields are

$$\begin{aligned} E_i &= \bar{a}_x E_o \, e^{jk_o z - j\omega t} \\ H_i &= \bar{a}_y E_o \sqrt{\frac{\epsilon_o}{\mu_o}} \, e^{jk_o z - j\omega t} \end{aligned} \qquad (1)$$

Following Stratton (Ref. 1) the orthogonal spherical vector functions are

$$\begin{aligned} \bar{M}^{(1)}_{\substack{o\\e}1n} = &\pm \frac{1}{\sin\theta} j_n(k_o r) \, P_n^1 (\cos\theta) \, {\cos \atop \sin} \varphi \, \bar{a}_r \\ &- j_n(k_o r) \frac{\partial P_n^1}{\partial\theta} (\cos\theta) \, {\sin \atop \cos} \varphi \, \bar{a}_\theta \end{aligned} \qquad (2)$$

$$\begin{aligned} \bar{N}^{(1)}_{\substack{o\\e}1n} = &\frac{n(n+1)}{k_o r} j_n(k_o r) \, P_n^1 (\cos\theta) \, {\sin \atop \cos} \varphi \, \bar{a}_r \\ &+ \frac{1}{k_o r} [k_o r \, j_n(k_o r)]' \frac{\partial P_n^1}{\partial\theta} (\cos\theta) \, {\sin \atop \cos} \varphi \, \bar{a}_\theta \\ &\pm \frac{1}{k_o r \sin\theta} [k_o r \, j_n(k_o r)]' P_n^1 (\cos\theta) \, {\cos \atop \sin} \varphi \, \bar{a}_\varphi \end{aligned} \qquad (3)$$

where:

$\bar{a}_i$ = unit vector in "i" direction

E_o = rms electric field magnitude

E_i = incident electric field

H_i = incident magnetic field (4)

ϵ_o = permittivity of free space

μ_o = permeability of free space

k_o = propagation constant in free space

ω = radian frequency of fields

$j_n(k_o r)$ = spherical Bessel function, defined by

$$j_n(x) = \sqrt{\frac{\pi}{2x}}\; J_{n+\frac{1}{2}}(x)$$

$P_n^1(\cos\theta)$ = associated legendre function of the first kind

The incident plane wave can be represented by:

$$\bar{E}_i = E_o\, e^{-j\omega t} \sum_{n=1}^{\infty} j^n \frac{2n+1}{n(n+1)} [\bar{M}^{(1)}_{oln} - j\, \bar{N}^{(1)}_{eln}]$$

$$\bar{H}_i = -\sqrt{\frac{\epsilon_o}{\mu_o}}\, E_o\, e^{-j\omega t} \sum_{n=1}^{\infty} j^n \frac{2n+1}{n(n+1)} [\bar{M}^{(1)}_{eln} + j\, \bar{N}^{(1)}_{oln}] \tag{5}$$

Maxwell's equations are satisfied by linear combinations of the orthogonal spherical functions. It remains now to find proper coefficients matching the boundary conditions. In order to satisfy the condition at infinity, the scattered wave must be in terms of Hankel functions. The external fields may be written:

$$\bar{E}_s = E_o\, e^{-j\omega t} \sum_{n=1}^{\infty} j^n \frac{2n+1}{n(n+1)} [a_n^s\, \bar{M}^{(3)}_{oln} - jb_n^s\, \bar{N}^{(3)}_{eln}]$$

$$\bar{H}_s = -\frac{\epsilon_o}{\mu_o} E_o\, e^{-j\omega t} \sum_{n=1}^{\infty} j^n \frac{2n+1}{n(n+1)} [b_n^s\, \bar{M}^{(3)}_{eln} + ja_n^s\, \bar{N}^{(3)}_{oln} \tag{6}$$

where the superscript (3) indicates Hankel functions of the first kind rather than Bessel functions.

The interior waves may be written

$$\bar{E}_t = E_o e^{-j\omega t} \sum_{n=1}^{\infty} j^n \frac{2n+1}{n(n+1)} [a_n^t \bar{M}_{oln}^{(1)} - jb_n^t \bar{N}_{eln}^{(1)}]$$

$$\bar{H}_t = -\sqrt{\frac{\epsilon_o}{\mu_o}} E_o e^{-j\omega t} \sum_{n=1}^{\infty} j^n \frac{2n+1}{n(n+1)} [b_n^t \bar{M}_{eln}^{(1)} + ja_n^t \bar{N}_{oln}^{(1)}] \qquad (7)$$

where the Bessel function argument is $k_1 r$ rather than $k_o r$; $[k_1 = k_o \sqrt{\mu_1 \epsilon_1} = Kk_o]$.

Let the radius of the sphere be a, and subject Eqs. 5, 6, and 7 to the following boundary conditions at r = a:

$$\bar{a}_r \times (\bar{E}_i + \bar{E}_s) = \bar{a}_r \times \bar{E}_t$$

$$\bar{a}_r \times (\bar{H}_i + \bar{H}_s) = \bar{a}_r \times \bar{H}_t \qquad (8)$$

Two pairs of inhomogeneous equations result:

$$a_n^t j_n(K\rho) - a_n^s h_n^{(1)}(\rho) = j_n(\rho)$$

$$a_n^t [K\rho\, j_n(K\rho)]' - \frac{\mu_1}{\mu_o} a_n^s [\rho h_n^{(1)}(\rho) = \frac{\mu_1}{\mu_o} [\rho j_n(\rho)]' \qquad (9)$$

$$Kb_n^t j_n(K\rho) - \frac{\mu_1}{\mu_o} b_n^s h_n^{(1)}(\rho) = \frac{\mu_1}{\mu_o} j_n(\rho)$$

$$b_n^t [K\rho j_n(K\rho)]' - Kb_n^s [\rho h_n^{(1)}(\rho)]' = K[\rho j_n(\rho)]' \qquad (10)$$

where:

$$\rho = k_o a, \; K\rho = k_1 a .$$

At this point the Ricatti Bessel functions will be introduced for convenience:

$$S_n(z) = zj_n(z)$$

$$C_n(z) = zn_n(z) \tag{11}$$

$$R_n^{(1)}(z) = zh_n^{(1)}(z) = S_n(z) + jC_n(z)$$

Solving (10) for the coefficients a_n^t and b_n^t, and defining $\mu = \mu_1/\mu_o$ and $\epsilon = \epsilon_1/\epsilon_o$,

$$a_n^t = \mu K \frac{S_n'(\rho)\, R_n^{(1)}(\rho) - S_n(\rho)\, R_n'^{(1)}(\rho)}{KS_n'(K\rho)\, R_n^{(1)}(\rho) - \mu S_n(K\rho)\, R_n'^{(1)}(\rho)}$$

$$b_n^t = \mu K \frac{S_n'(\rho)\, R_n^{(1)}(\rho) - S_n(\rho)\, R_n'^{(1)}(\rho)}{\mu S_n'(K\rho)\, R_n^{(1)}(\rho) - KS_n(K\rho)\, R_n'^{(1)}(\rho)} \tag{12}$$

The numerator reduces:

$$S_n'(\rho)\, R_n^{(1)}(\rho) - S_n(\rho)\, R_n'^{(1)}(\rho) = j$$

Thus the problem is formally solved; it remains now to evaluate the fields numerically. A simple case is for the fields at the center of the sphere. As the argument approaches zero, $z \ll 1$.

$$j_n(z) \rightarrow (2z)^n \cdot \frac{n!}{(2n+1)!}$$

$$[\rho j_n(z)]' \rightarrow (2z)^n \frac{n!}{(2n+1)!} (1+n) \tag{13}$$

From (2), (3), and (13) it can be seen that at the center of the sphere the higher modes are not sustained, and that only the $N^{(1)}_{\substack{o\\e}11}$ terms contribute to the fields; the orthogonal functions are:

$$\bar{M}^{(1)}_{\substack{o\\e}11} = \pm \frac{k_1 r}{3} \begin{matrix}\cos\\ \sin\end{matrix}\, \varphi\, \bar{a}_\theta - \frac{k_1 r}{3} \cos\theta \begin{matrix}\sin\\ \cos\end{matrix}\, \varphi\, \bar{a}_\varphi \tag{14}$$

$$\begin{aligned}\bar{N}^{(1)}_{\substack{o\\e}11} &= \frac{2}{3} \sin\theta \begin{matrix}\sin\\ \cos\end{matrix}\, \varphi\, \bar{a}_r + \frac{2}{3} \cos\theta \begin{matrix}\sin\\ \cos\end{matrix}\, \varphi\, \bar{a}_\theta \\ &\quad \pm \frac{2}{3} \begin{matrix}\cos\\ \sin\end{matrix}\, \varphi\, \bar{a}_\varphi\end{aligned} \tag{15}$$

since

$$P_1^{\prime}(\cos\theta) = \sin\theta \text{ and } \frac{\partial P_1^{\prime}(\cos\theta)}{\partial\theta} = \cos\theta \tag{16}$$

Thus at the center the fields are:

$$\begin{aligned}\bar{E}_t &= E_o\, e^{-j\omega t}\, b_1^t [\sin\theta \cos\varphi\, \bar{a}_r + \cos\theta \cos\varphi\, \bar{a}_\theta - \sin\varphi\, \bar{a}_\theta] \\ \bar{H}_t &= \sqrt{\frac{\epsilon_1}{\mu_1}}\, E_o\, e^{-j\omega t}\, a_1^t [\sin\theta \sin\varphi\, \bar{a}_r + \cos\theta \sin\varphi\, \bar{a}_\theta + \cos\varphi\, \bar{a}_\theta]\end{aligned} \tag{17}$$

which, when put in terms of x, y, z coordinates, become:

$$\begin{aligned}\bar{E}_t &= E_o\, e^{-j\omega t}\, b_1^t\, \bar{a}_x \\ \bar{H}_t &= \sqrt{\frac{\epsilon_1}{\mu_1}}\, E_o\, e^{-j\omega t}\, a_1^t\, \bar{a}_y\end{aligned} \tag{18}$$

The incident wave has a Poynting vector of

$$\frac{1}{2} \operatorname{Re} \left\{ \bar{E}_i \times \bar{H}_i^* \right\} = \sqrt{\frac{\epsilon_o}{\mu_o}}\, E_o^2\, \bar{a}_z \; . \tag{19}$$

The interior wave has the Poynting vector at the center of

$$\frac{1}{2} \operatorname{Re} \left\{ \bar{E}_t \times \bar{H}_t^* \right\} = \frac{K}{\mu} \sqrt{\frac{\epsilon_o}{\mu_o}}\, E_o^2\, a_1^{t*}\, b_1^t\, \bar{a}_z \tag{20}$$

The Poynting vector at the center differs from the original wave by the factor

$$P = \frac{K}{\mu} a_1^{t*} b_1^t \qquad (21)$$

Note that in the case of $\mu_1 = \mu_o$, $\epsilon_1 = \epsilon_o$, the expression reduces to unity, as would be expected. Thus the power density factor P is a function of the material parameters and the argument ka, which is 2π times the radius a in material wavelengths. For $\mu = \epsilon = K = 10$, the curve of P versus a/λ_m is shown in Fig. 1. The resonant peaks occur at $a/\lambda_m = .64, 1.10, 1.56$ and approach the half-wavelength points 2.0, 2.5, 3.0, and so on. These resonant points coincide with the natural oscillations of a sphere (Ref. 1, p. 556-7). Thus it appears that the plane wave is feeding energy to a spherical cavity such that the power flow near the center is a plane wave of increased power density. As a/λ_m is increased, the peaks and dips smooth out, and the curve approaches the straight line

$$\begin{aligned} P &= \mu\epsilon = K^2; \text{ i.e., as } K_\rho \gg 1 \\ a_1^t &= b_1^t \rightarrow Ke^{i\rho(K-1)} \end{aligned} \qquad (22)$$

In the limiting case $K_\rho \ll 1$,

$$a_1^t = b_1^t \rightarrow \frac{3}{(K+2)} \qquad (23)$$

and

$$P \rightarrow \frac{9}{(K+2)^2}$$

The restriction involved in the problem in using the center of the sphere as the point of investigation is fairly good as an approximation over a circular area around the center up to about $a/\lambda_m = 0.2$; that is, an area integral will yield approximately (power density) x (area) as the power flow across the area.

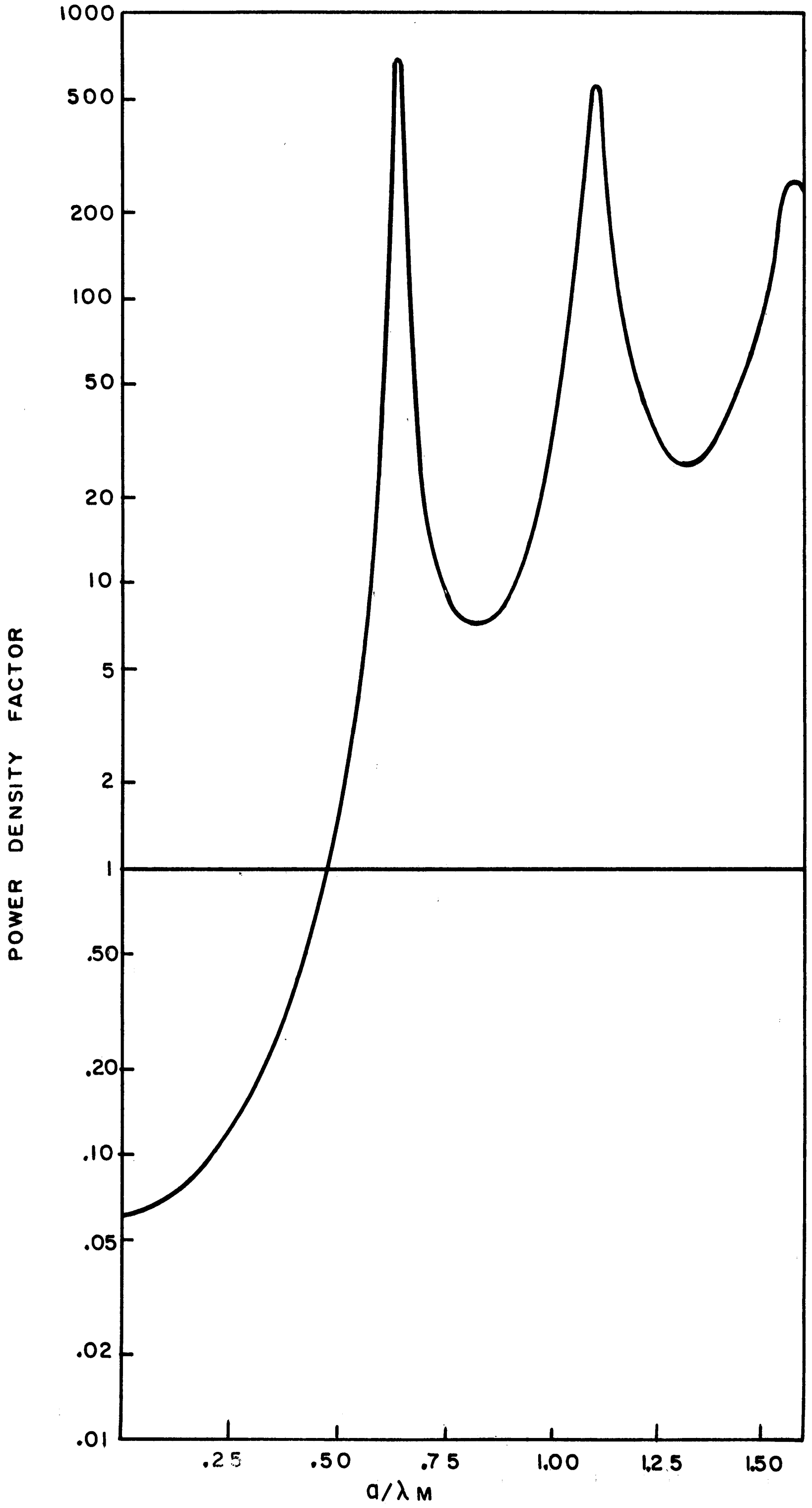


Fig. 1. Power density factor as a function of radius in material wavelengths.

The problem can be extended to deal with lossy materials by allowing μ and ϵ to be complex. Then (21) becomes:

$$P = \left(\frac{Ka_1^t}{\mu}\right)^* b_1^t \tag{24}$$

Subsequently, the cases of arbitrary μ and ϵ and lossy materials will be treated, and a more complete discussion and interpretation of the results will be given.

3.2 Scattering of a Plane Wave by an Infinitely Long Dielectric or Ferrite Cylinder

3.2.1 Basic Problem. In order to further the understanding of the various phenomena which solid-state materials can give rise to in improving the performance of antennas, a study is now under way to evaluate the fields obtainable both inside and outside a right circular cylinder with an electric and magnetic susceptibility different from that of free space.

The choice of a two-dimensional problem recommended itself for several reasons. One of the main advantages is that the amount of computation is not extremely laborious. Another is that the numerical evaluation on the computer study will not exceed the capacity of the IBM 704 computer.

This study considers the following two cases: There is an infinitely long cylinder (with both μ and ϵ possibly different from that of free space) upon which a plane wave is incident. Since an arbitrarily polarized wave can always be made up of two waves which are polarized in perpendicular planes, for the purposes of this study there is considered the case of the E field parallel to the axis of the cylinder and that of the E field perpendicular to the axis of the cylinder (see Figs. 2 and 3). The first case gives rise to TM modes and the second to TE modes.

The chief interest in this problem centers on the "collecting" power of the cylinder. If this configuration is to be made part of a useful antenna system there must exist at some points in space, either inside the cylinder or close to it on the outside, concentrations of power which exceed the power level before the scatterer was immersed in the field.

At the present time a computer program is progressing for evaluating the real and imaginary power flow inside and outside the cylinder. This program will explore a range of values of the material parameters μ and ϵ. It will also include a study of the power flow dependence on frequency and cylinder dimensions.

3.2.2 Theoretical Analysis.

(a) Incident Electric Field Parallel to the Cylinder Axis.

For this case the incident field for a plane wave is given by

$$E_z = A_o e^{i(ky-\omega_0 t)}$$

in rectangular coordinates (Fig. 1).

Expanding in cylindrical coordinates, an infinite set of TM modes is obtained for the fields inside and outside the cylinder (Ref. 1) as follows: (E parallel to the cylinder axis)

inside the cylinder ($r < a$):

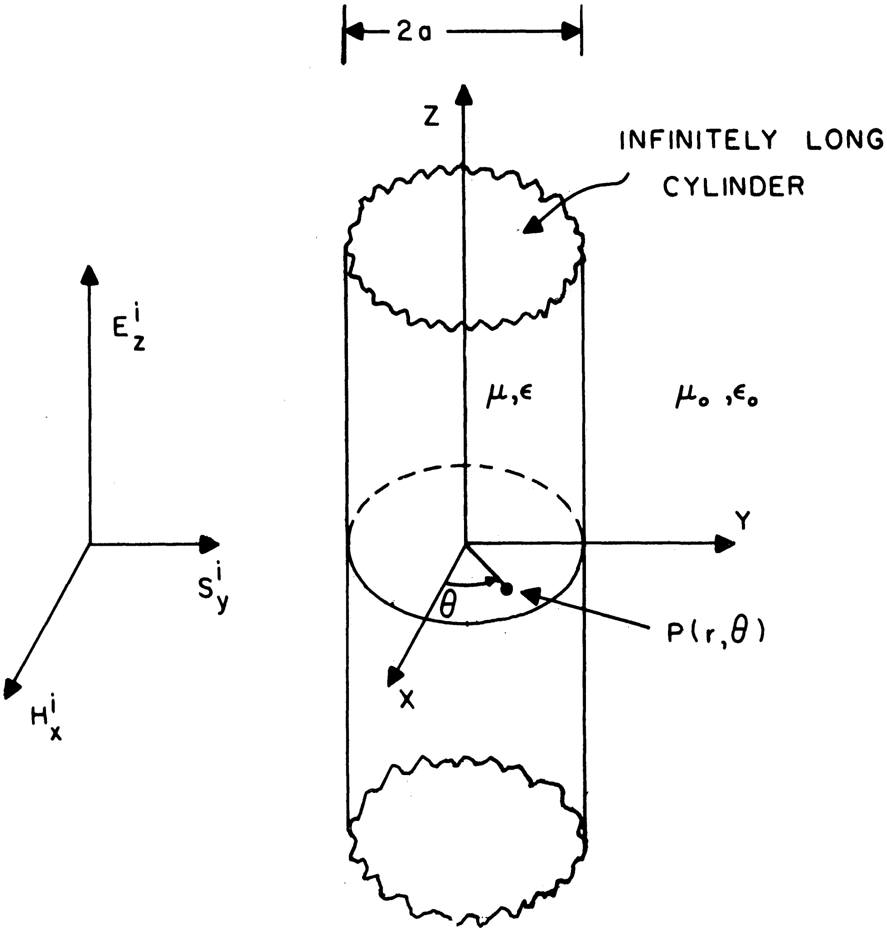


Fig. 2. Plane wave incident on cylinder, electric field parallel to cylinder axis.

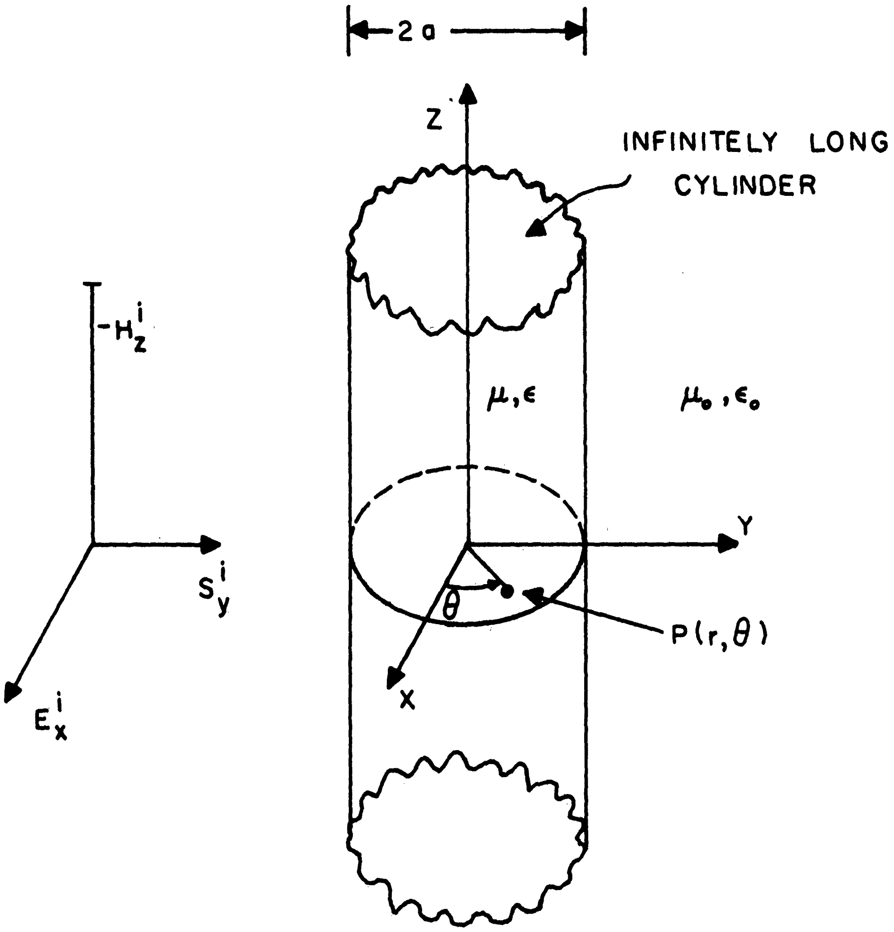


Fig. 3. Plane wave incident on cylinder, electric field perpendicular to cylinder axis.

$$E_z = k^2 \sum_{-\infty}^{\infty} b_n e^{in\theta} J_n(kr)$$

$$H_r = \frac{k^2}{\mu\omega r} \sum_{-\infty}^{\infty} b_n n e^{in\theta} J_n(kr)$$

$$H_\theta = \frac{ik^3}{\mu\omega} \sum_{-\infty}^{\infty} b_n e^{in\theta} J_n'(kr)$$

outside the cylinder $(r > a)$:

$$E_z = k_o^2 \sum_{-\infty}^{\infty} a_n e^{in\theta} H_n(k_o r) + A_o \sum_{-\infty}^{\infty} J_n(k_o r) e^{in\theta}$$

$$H_r o = \frac{k_o{}^2}{\mu_o \omega r} \sum_{-\infty}^{\infty} n a_n e^{in\theta} H_n(k_o r) + A_o \sqrt{\frac{\epsilon_o}{\mu_o}} \cos\theta \sum_{-\infty}^{\infty} J_n(k_o r) e^{in\theta}$$

$$H_\theta = \frac{ik_o^3}{\mu_o \omega} \sum_{-\infty}^{\infty} a_n e^{in\theta} H_n'(k_o r) - A_o \sqrt{\frac{\epsilon_o}{\mu_o}} \sin\theta \sum_{-\infty}^{\infty} J_n(k_o r) e^{in\theta}$$

The continuity of the tangential E and H fields (E_z and H_θ) at the cylinder-free-space boundary $(r = a)$ yields explicit expressions for the coefficients a_n and b_n:

$$b_n = \frac{\alpha A_o}{k^2} \frac{J_n(k_o a) H_n'(k_o a) - J_n'(k_o a) H_n(k_o a)}{\alpha J_n(ka) H_n'(k_o a) - J_n'(ka) H_n(k_o a)}$$

$$a_n = \frac{A_o}{k_o^2} \frac{J_n'(ka) J_n(k_o a) - \alpha J_n(ka) J_n'(k_o a)}{\alpha J_n(ka) H_n'(k_o a) - J_n'(ka) H_n(k_o a)}$$

where:

r, θ are cylindrical coordinates.

r = radial distance from axis of cylinder.

θ = angular distance, with incident wave normal in direction $\theta = \pi/2$.

A_o = amplitude of incident electric field.

μ = permeability of cylinder.

ϵ = dielectric constant of cylinder.

μ_o = permeability of surrounding medium.

ϵ_o = dielectric constant of surrounding medium.

$k = \omega\sqrt{\mu\epsilon}$ = wave number inside cylinder.

$k_o = \omega\sqrt{\mu_o\epsilon_o}$ = wave number outside cylinder.

$H_n(x) \equiv H_n^{(1)}(x)$ = Hankel function of the first kind.

$J_n'(x)$, $H_n'(x)$ = derivatives of Bessel and Hankel functions with respect to the argument.

$$\alpha = \sqrt{\frac{\mu\epsilon_o}{\mu_o\epsilon}}$$

In the evaluation of the convergence properties of the infinite series for the field components, the behavior of the coefficients a_n and b_n has to be determined.

For the TM case, for $n \gg \frac{1}{4}(ka)^2$, the formulas obtained are:

Coefficients inside cylinder:

$$b_n = \frac{2A_o}{k^2}\frac{\mu}{\mu + \mu_o}\left(\frac{k_o}{k}\right)^n$$

Coefficients outside cylinder:

$$a_n = \frac{iA_o\pi}{k_o^2 n!\,(n-1)!}\,\frac{\mu - \mu_o}{\mu + \mu_o}\left(\frac{k_o a}{2}\right)^{2n}$$

Several interesting observations can now be made regarding the coefficients of the higher-order modes: For terms outside the cylinder there is a definite difference in the frequency dependence of different modes. The expansion coefficients will ultimately converge due to the factor $(n!)^2$ in the denominator. As a matter of fact, since $n \gg \frac{1}{4}(ka)^2$, the a_n's will have to be small before the above formula will hold. In the limit $a_n = (\omega)^{2n-2}/(n!)^2 = (\omega^{n-1}/n!)^2$; so no matter how high a frequency of operation we choose the series of coefficients will ultimately converge.

In the final analysis we are interested in the value of the infinite series of modes making up the field components. Then, at any point of observation we encounter terms of the following nature, expressing the scattered fields as

$$E_z^s = \sum_n E_{zn} e^{in\theta}, \quad H_\theta^s = \sum_n H_{n} e^{in\theta}$$

For $n \gg \frac{1}{4}(ka)^2$:

$r > a$

$$E_{zn} = \frac{A_o}{n!} \frac{\mu - \mu_o}{\mu + \mu_o} \frac{1}{2^n} \frac{(k_o a)^{2n}}{(k_o r)^n}$$

$$H_{\theta n} = - A_o i \sqrt{\frac{\epsilon_o}{\mu_o}} \frac{1}{(n-1)!} \frac{\mu - \mu_o}{\mu + \mu_o} \frac{1}{2^n} \frac{(k_o a)^{2n}}{(k_o r)^{n+1}}$$

$$H_{rn} = \frac{A_o}{2k_o r} \sqrt{\frac{\epsilon_o}{\mu_o}} \frac{\mu - \mu_o}{\mu + \mu_o} \frac{1}{(n-1)!} \frac{(k_o a)^{2n}}{(2k_o r)^n}$$

$r < a$

$$E_{zn} = 2A_o \frac{\mu}{\mu + \mu_o} \frac{1}{n!} \left(\frac{kr}{2}\right)^n \left(\frac{k_o}{k}\right)^n$$

$$H_{\theta n} = iA_o \sqrt{\frac{\epsilon}{\mu}} \frac{\mu}{\mu + \mu_o} \left(\frac{k_o}{k}\right)^n \frac{1}{(n-1)!} \left(\frac{kr}{2}\right)^{n-1}$$

$$H_{rn} = \frac{2A_o}{kr} \sqrt{\frac{\epsilon}{\mu}} \frac{\mu}{\mu + \mu_o} \left(\frac{k_o}{k}\right)^n \frac{1}{(n-1)!} \left(\frac{kr}{2}\right)^n$$

Because of the n! in the denominator of all these terms, all the field components are obtained as bounded sums of the modes.

(b) Incident Electric Field Perpendicular to the Cylinder Axis.

For this case the incident field for a plane wave is given by

$$E_x = A_o e^{i(ky-\omega t)}$$

in rectangular coordinates (Fig. 2).

Expanding as before, an infinite set of TE modes is obtained as follows:

$\underline{r < a}$

$$E_r = \frac{\mu\omega}{r} \sum_{-\infty}^{\infty} n b_n e^{in\theta} J_n(kr)$$

$$E_\theta = - i\mu\omega \sum_{-\infty}^{\infty} b_n e^{in\theta} kJ_n'(kr)$$

$$H_z = k^2 \sum_{-\infty}^{\infty} b_n e^{in\theta} J_n(kr)$$

$\underline{r > a}$

$$E_r = \frac{\mu_o\omega}{r} \sum_{-\infty}^{\infty} n a_n e^{in\theta} H_n(k_o r) + A_o \cos\theta \sum_{-\infty}^{\infty} J_n(k_o r) e^{in\theta}$$

$$E_\theta = - i\mu_o\omega \sum_{-\infty}^{\infty} a_n e^{in\theta} k_o H_n'(k_o r) - A_o \sin\theta \sum_{-\infty}^{\infty} J_n(k_o r) e^{in\theta}$$

$$H_z = k_o \sum_{-\infty}^{\infty} a_n e^{in\theta} H_n(k_o r) + A_o\sqrt{\frac{\epsilon_o}{\mu_o}} \sum_{-\infty}^{\infty} J_n(k_o r) e^{in\theta}$$

The expansion coefficients are given as:

$$b_n = \frac{-A_o}{k^2}\sqrt{\frac{\epsilon_o}{\mu_o}} \; \frac{H_n(k_o a) J_n'(k_o a) + H_n'(k_o a) J_n(k_o a)}{\alpha H_n(k_o a) J_n'(ka) - H_n'(k_o a) J_n(ka)}$$

$$a_n = \frac{-A_o}{k_o^2}\sqrt{\frac{\epsilon_o}{\mu_o}} \; \frac{H_n(k_o a) J_n'(k_o a) + H_n'(k_o a) J_n(k_o a)}{\alpha H_n(k_o a) J_n'(ka) - H_n'(k_o a) J_n(ka)}$$

Again expressing the field components as:

$$E_\theta^s = \sum_n e^{in\theta} E_{\theta n}$$

Then we have:

r > a

$$E_{\theta n} = -\frac{2A_o}{(n-1)!} \frac{(k_o a)^{2n}}{(2k_o r)^{n+1}}$$

$$H_{zn} = -\frac{A_o}{n!}\sqrt{\frac{\epsilon_o}{\mu_o}} \frac{(k_o a)^{2n}}{(2k_o r)^{n}}$$

$$E_{rn} = -\frac{A_o}{k_o r}\frac{1}{(n-1)!} \frac{(k_o a)^{2n}}{(2k_o r)^{n}}$$

r < a

$$E_{\theta_n} = \frac{iA_o}{(n-1)!}\sqrt{\frac{\epsilon_o \mu}{\mu_o \epsilon}} \frac{\epsilon}{\epsilon_o + \epsilon} \left(\frac{k_o}{k}\right)^n \left(\frac{kr}{2}\right)^{n-1}$$

$$H_{z_n} = -2A_o\sqrt{\frac{\epsilon_o}{\mu_o}} \frac{\epsilon}{\epsilon + \epsilon_o} \left(\frac{k_o}{k}\right)^n \frac{1}{n!} \left(\frac{kr}{2}\right)^{n}$$

$$E_{r_n} = -\frac{2A_o}{kr}\sqrt{\frac{\epsilon_o \mu}{\mu_o \epsilon}} \frac{\epsilon}{\epsilon + \epsilon_o} \left(\frac{k_o}{k}\right)^n \frac{1}{(n-1)!} \left(\frac{kr}{2}\right)^{n}$$

Here, again, as in the TM case, because of the factoral in the denominator, the series is convergent.

3.3 Comparison of Solid-State and Conventional Antennas

In attempting to evaluate solid-state antennas, it is important to determine whether the solid state antenna represents an improvement over antennas of conventional design. Comparison can be made for a number of characteristics such as:

(1) Size

(2) Directivity

(3) Efficiency

(4) Bandwidth

(5) Impedance

(6) Power Capacity

(7) Weight

(8) Cost

In the lower portion of the frequency range of interest for the contract, antenna sizes become large and size reduction becomes of primary importance, so that a reasonable method of comparison in this range would be to compare (2) through (8) for antennas of equal size.

The natural inclination is to compare antennas of similar design, dipoles with loaded dipoles, etc. However, there are a number of circumstances under which a comparison of this sort would be invalid. First, if the antenna with solid-state material were considerably larger than the conventional antenna, the comparison would be invalid. Second, if the conventional antenna and the solid-state antenna did not represent optimum designs, or if the characteristics which had been optimized in the designs of the two antennas were not the same characteristic, then a good comparison would be difficult. In general, the characteristics of a given solid state antenna must be compared with <u>all</u> conventional antennas. The question then becomes, "Can the characteristics of this solid-state antenna be duplicated or bettered by a conventional design of equal size?"

Some of the special methods required for a good comparison of characteristics (1) through (8) are discussed below.

(1) <u>Size</u>

A strict comparison of antennas would be limited to those of approximately equal dimensions. The special space requirements of a particular application may allow comparison on a basis of equal volume, equal maximum cross-sectional area, or some such other geometric factor.

(2) Directivity and Gain

G_o = gain over isotropic (Ref. 4)

D = directivity

k = efficiency factor

$G_o = kD$

In a solid-state antenna with appreciable losses, gain will differ from directivity.

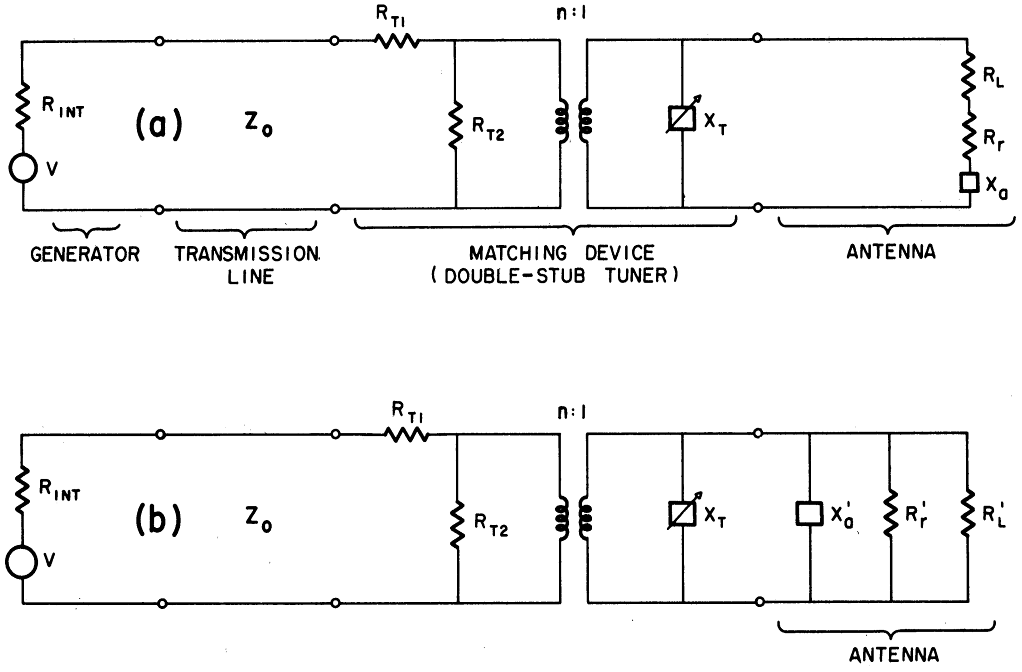
(3) Efficiency

In general the equivalent circuit of a transmitter and antenna or of a receiver and antenna can be represented as in Figs. 4 and 5. The efficiency of the antenna in all cases may be represented by $R_r/(R_r + R_L)$ or by

$$\frac{\text{total radiated power}}{\text{net input power to antenna}}$$

Total Radiated Power

The measurement of total radiated power may be made in several ways, one of which would be the use of the WADD beam pattern equipment, which integrates the radiated power over a sphere and integrates over the polarization pattern at each point. A method applicable to the equipment installed at CEL would be to take two complete measurements of the beam pattern (using the test antenna as the receiving antenna), one with vertical and one with horizontal polarization from the transmitting antenna. It can be shown (Fig. 6) that at any point the sum of the two measurements of radiated power per unit area is equal to the total radiated power per unit area at that point. Then the test antenna (in a specified orientation) would be used as a transmitter, and a field



R_{int} = internal resistance of generator

Z_o = characteristic impedance of transmission line

R_{T1}, R_{T2} = transformer losses

X_T = variable reactance

R_L = antenna losses

R_r = radiation resistance

Z_a = antenna input impedance = $(R_r + R_L) + jX_a$

$$R'_L = \frac{(R_L + R_r)^2 + (X_a)^2}{R_r}$$

$$R'_r = \frac{(R_L + R_r)^2 + (X_a)^2}{R_L}$$

$$X'_a = \frac{(R_L + R_r)^2 + (X_a)^2}{X_a}$$

$$\text{Efficiency} = \frac{R_r}{R_r + R_L} = \frac{R'_L}{R'_L + R'_r}$$

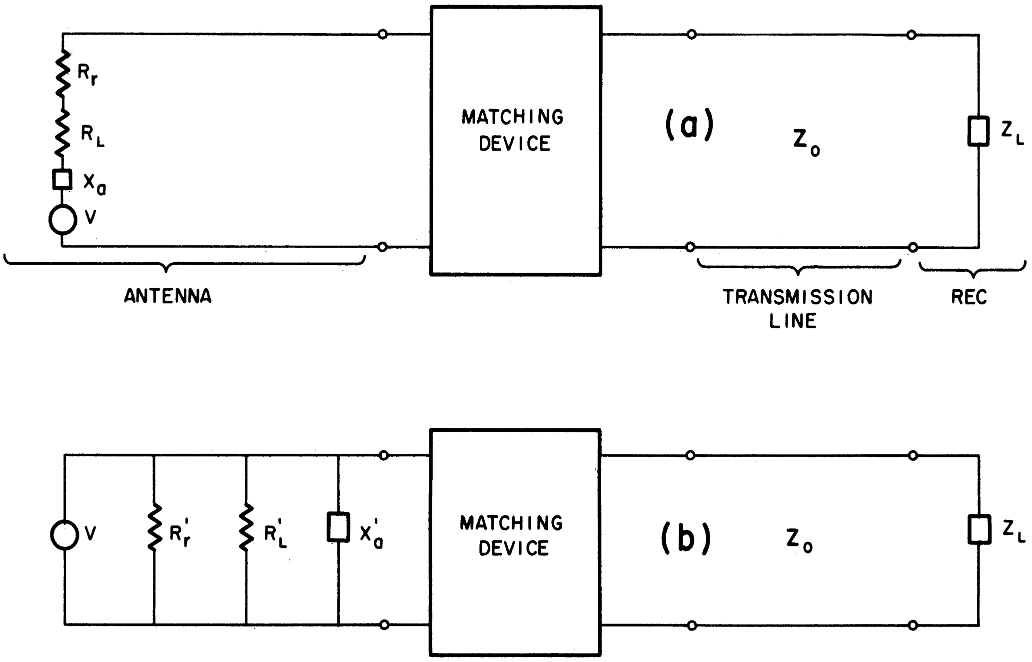
For conjugate match:

If $Z_o = R_{int}$

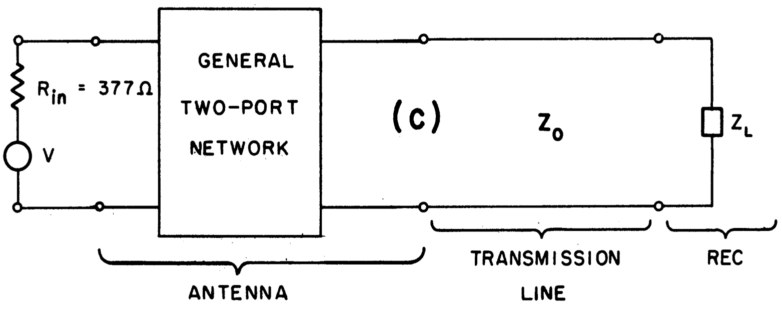
$$n = \sqrt{Z_o \cdot \frac{(R'_L + R'_r)}{R'_L R'_r}}$$

$$X_T = - X'_a$$

Fig. 4. Equivalent circuits of transmitting antenna.

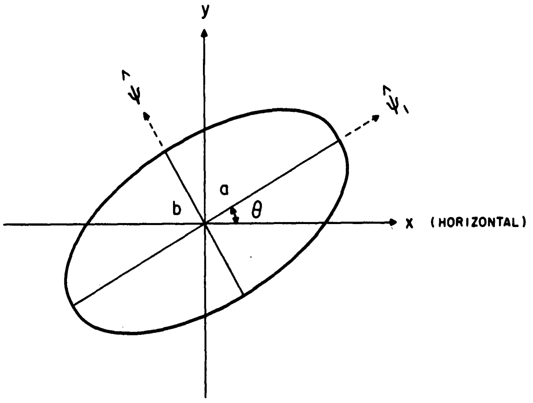


In figures 5(a) and 5(b) V is the voltage induced in the antenna by a passing wave. V is therefore dependent on the antenna parameters as well as the incident wave intensity. A receiving antenna could also be represented by the circuit in Fig. 5(c), where V is dependent only on the incident wave intensity.



$[Z_{a1}]$ = impedance matrix of antenna $\neq Z_a$ (in general). $[Z_{a1}]$ is difficult to measure.

Fig. 5. Equivalent circuits of receiving antenna.



fields may be represented as:

$$E_{\psi 1} = a \sin \omega\tau \qquad H_{\psi 2} = \frac{a}{\sqrt{\frac{u_o}{\epsilon_o}}} \sin \omega\tau$$

$$E_{\psi 2} = b \cos \omega\tau$$

$$H_{\psi 1} = - \frac{b}{\sqrt{\frac{u_o}{\epsilon_o}}} \sin \omega\tau$$

average power/area over one cycle

$$= \frac{1}{2\pi} \int_o^{2\pi} \left[E_{\psi 1} + E_{\psi 2}\right] \times \left[H_{\psi 1} + H_{\psi 2}\right] d(\omega\tau) = \boxed{\frac{a^2 + b^2}{2\sqrt{\frac{u_o}{\epsilon_o}}}} \hat z$$

where:

$$\hat\psi_1 \times \hat\psi_2 = \hat z$$

Consider measurements taken in vertical (y) + horizontal (x) planes.

$$E_x = a \cos\theta \sin\omega\tau - b \sin\theta \cos\omega\tau$$

$$E_y = a \sin\theta \sin\omega\tau + b \cos\theta \cos\omega\tau$$

$$H_x = \frac{1}{\sqrt{\frac{u_o}{\epsilon_o}}}\left[- a \sin\theta \sin\omega\tau - b \cos\theta \cos\omega\tau\right]$$

$$H_y = \frac{1}{\sqrt{\frac{u_o}{\epsilon_o}}}\left[a \cos\theta \sin\omega\tau - b \sin\theta \cos\omega\tau\right]$$

average power/area over one cycle

$$= \frac{1}{2\pi} \int_o^{2\pi} \left[E_x \hat x + E_y \hat y\right] \times \left[H_x \hat x + H_y \hat y\right] d(\omega\tau) =$$

$$= \frac{1}{2\pi} \int_o^{2\pi} \left\{ \left[(a \cos\theta \sin\omega\tau - b \sin\theta \cos\omega\tau)\, \hat x + \right.\right.$$

$$\left.(a \sin\theta \sin\omega\tau + b \cos\theta \cos\omega\tau)\, \hat y\right] \times$$

$$\sqrt{\frac{u_o}{\epsilon_o}} \left[(-a \sin\theta \sin\omega T - b \cos\theta \cos\omega\tau)\, \hat x + \right.$$

$$\left.\left.(a \cos\theta \sin\omega\tau - b \sin\theta \cos\omega\tau)\, \hat y\right]\right\} (\omega\tau) = \boxed{\frac{a^2 + b^2}{2\sqrt{\frac{u_o}{\epsilon_o}}} \hat z}$$

Fig. 6. Measurement of total radiated power with elliptical polarization.

intensity measurement would be taken at a given distance R (in the far field) from the test antenna. The field intensity measurement would be converted to watts/meter^2 and would provide a reference level on the beam patterns for the total radiated power density at the orientation (φ_o, θ_o). The total radiated power is then determined by integrating the two beam patterns over a sphere of radius R. This procedure is outlined in Fig. 7.

Net Input Power to Antenna: (See Fig. 4)

Method 1

If we assume:

1. $R_{INT} = Z_o$ (generator matched to transmission line)
2. $R_{T1} = R_{T2} = 0$ (lossless transformer)

then, if the double-stub tuner is adjusted for conjugate match to the antenna, the net input power to antenna = available power of generator $= V^2/4R_{INT}$, which can be measured.

Method 2

If we eliminate the double-stub tuner and assume:

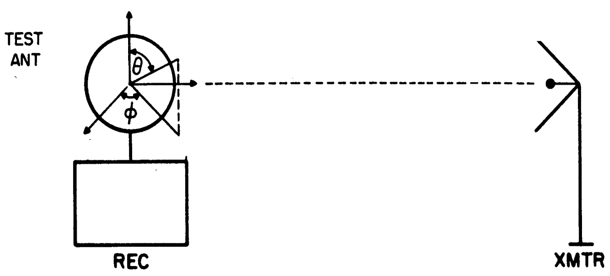
1. $R_{INT} = Z_o$

then, Γ = reflection coefficient $= (Z_a - Z_o)/(Z_a + Z_o)$

Z_a = input impedance of antenna

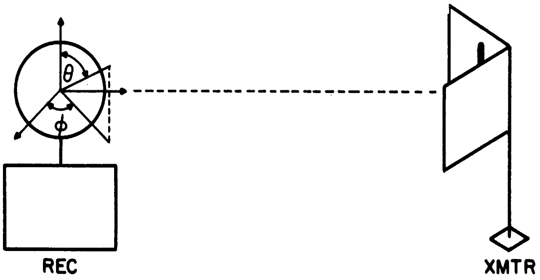
Net input power to antenna $= [1 - |\Gamma|^2] \times$ Available Power.

If $R_{INT} \neq Z_o$, then the formulas still hold if sufficient padding is added to the transmission line and available power measured after the padding.

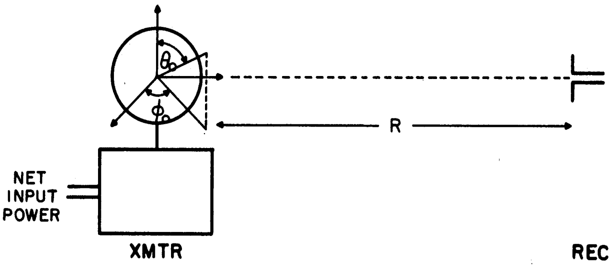


(a) Beam pattern No. 1 (horizontal)

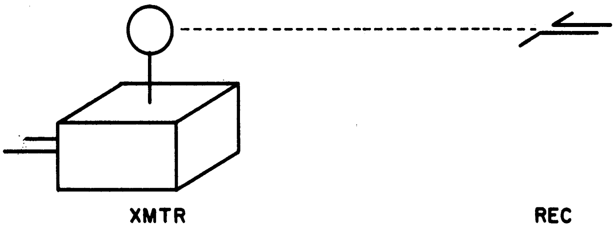
θ, φ specify orientation of test antenna



(b) Beam pattern No. 2 (vertical)



(c) Field intensity measurement No. 1 (vertical)



(d) Field intensity measurement No. 2 (horizontal)

It is necessary to have one comparison of XMTR power for horizontal + vertical measurements, either in the beam pattern or in the field intensity measurements. At φ_o, θ_o the ratio of vertical field intensity to horizontal field intensity should equal the ratio of vertical beam pattern level to horizontal beam pattern level.

Fig. 7. Measurement of total radiated power.

(4) Impedance

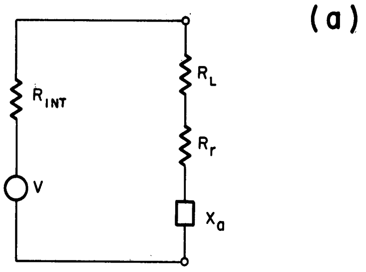
The impedance criterion depends very much on the particular application. Some of the considerations in the comparison of impedances are:

(a) If $R_{INT} = Z_o$, which impedance is more closely matched to Z_o?

(b) If a tuning device is available, which impedance maximizes $R_r/(R_r + R_L)$ (max efficiency)?

In special applications, it may be desirable to combine the impedance and efficiency characteristics and to compare radiated power when each of the two antennas is connected to a fixed transmitter system or to compare the receiver indication when each antenna is connected to a fixed receiver system. Such a comparison would make sense only if the usefulness of the antennas could be considered to be restricted to a limited class of systems (or if the interest of the investigator were limited to a particular system).

Occasionally, in the literature, antennas have been compared on the basis of equal input current or equal input voltage. This type of comparison results from a combination of the efficiency and impedance characteristics and applies only to a particular class of systems. Figur 8 shows several examples of this type. A general characteristic of these systems is that there is no possibility of matching generator to antenna or antenna to receiver. If it were possible to provide a conjugate match in the system, then a comparison would be valid only on a basis of equal input power. Moreover, if it were possible to reduce the mismatch in the system so that the inequality conditions of Fig. 8 no longer held, then a comparison on the basis of equal input current or voltage would



Case No. 1

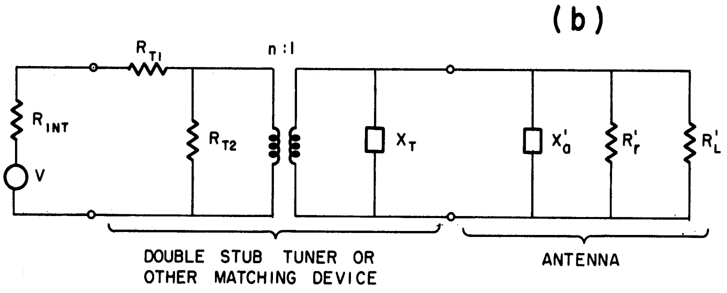
$$|R_{int}| \gg |R_L + R_r + jX_a|$$

current $\approx \frac{V}{R_{int}}$ regardless of antenna impedance. This leads to equal input current method of comparison.

Case No. 2

$$|R_{int}| \ll |R_L + R_r + jX_a|$$

Voltage across antenna $\approx V$ regardless of antenna impedance. This leads to constant voltage method of comparison.



If (1) $X_T = -X'_a$

and (2) $n^2\left(\frac{R'_r R'_L}{R'_r + R'_L}\right) \ll R_{T2}$

and (3) $n^2\left(\frac{R'_r R'_L}{R'_r + R'_L}\right) \gg R_{T1}$

Then Case No. 1:

$$\left.n^2\left(\frac{R'_r R'_L}{R'_r + R_L}\right) \ll R_{INT}\right\} \begin{array}{l}\text{current to} \\ \text{antenna} \\ = \frac{V}{R_{INT}} \times n\end{array}$$

Case No. 2:

$$\left.n^2\left(\frac{R'_r R'_L}{R'_r + R'_L}\right) \gg R_{INT}\right\} \begin{array}{l}\text{voltage at an-} \\ \text{tenna terminals} \\ = \frac{V}{n}\end{array}$$

Fig. 8. Constant current and constant voltage measurement techniques.

be invalid.

Figure 8(a) represents a simple antenna system in which a high degree of mismatch occurs between the internal impedance and antenna input impedance.

Figure 8(b) is a somewhat more complicated system, in which transformer losses (sum of lead resistance losses, core losses, and radiation losses) are important. The antenna is represented as in Fig. 4(b). In this case, a comparison may be made on the basis of current or voltage if the transformation ratio (n:1) is taken into account, provided that the transformer losses vary with n in such a way that the stated inequalities of Fig. 8(b) hold.

There are, of course, many other complex circuits which might provide, under certain conditions, other bases of comparison; however, in each case, the equal input power basis would still be valid and, in general, it is simpler to compare antennas on this basis.

The other points of comparison can be made directly and need no special interpretation.

3.4 Measurements on Loop Antenna

A loop antenna was designed and constructed (Fig. 9). The impedance of the loop was measured using the circuit shown in Fig. 10.

The GR 874-UB balun used is a balanced-to-unbalanced coaxial transformer. It is tunable to any one frequency from 54 to 1000 Mc. The impedance transformation is 4 to 1, providing an output impedance of 200 ohms. The balanced two-wire air line was designed and built in the laboratory and consists of two 1/8" diameter brass rods spaced 0.334" apart, center to center. This line is about 88 cm long and has a characteristic impedance of 200 ohms.

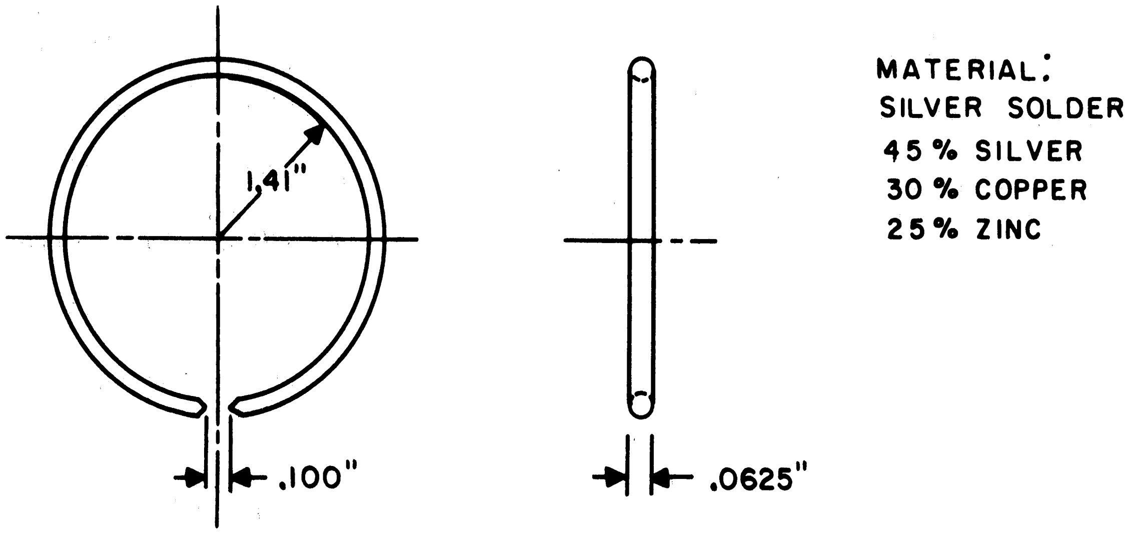


Fig. 9. Loop antenna.

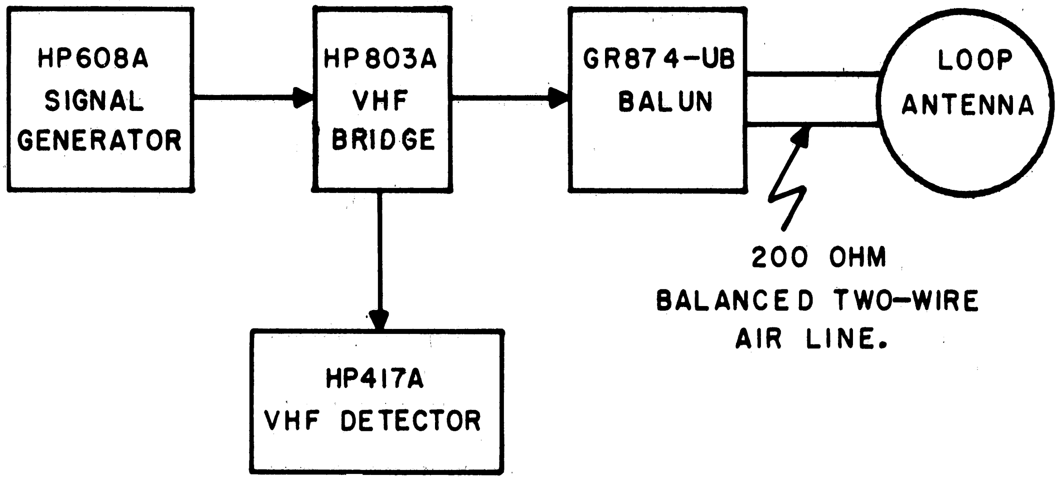


Fig. 10. Impedance measurement circuit.

Using the standard method of placing a short at the antenna terminals, measuring the impedance, and then removing the short and measuring the impedance, the actual impedance of the loop by itself was determined. Using this measurement technique and the Smith Chart, errors due to losses in the balanced two-wire line were eliminated. The loss of the two-wire line was found to be approximately 0.1 db in the range 175 Mc to 275 Mc.

The impedance and voltage reflection coefficient of the loop is shown in Table I.

The impedance and reflection of a ferrite-core loop antenna are also shown in Table I. The metal antenna is exactly the same as shown in Fig. 9. A spherical core of ferrite powder is placed within the loop as shown in Fig. 11. The ferrite powder is enclosed by a thin balsa-wood shell, as also shown in the figure.

	Air-Core Loop		Ferrite-Core Loop	
Frequency (Mc)	Impedance (Ohms)	Reflection Coefficient	Impedance (Ohms)	Reflection Coefficient
175	j188	1.0	21.2 + j347.5	0.955
200	11.3 + j260	0.96	46.0 + j477	0.925
225	17.8 + j339	0.96	12.2 + j700	0.99
250	14.8 + j424	0.98	j680	1.0
275	j472	1.0	j608	1.0

Table I. Impedance and reflection coefficient of air-core and ferrite-core loop antenna.

The ferrite used in the test is discussed in the section on materials.

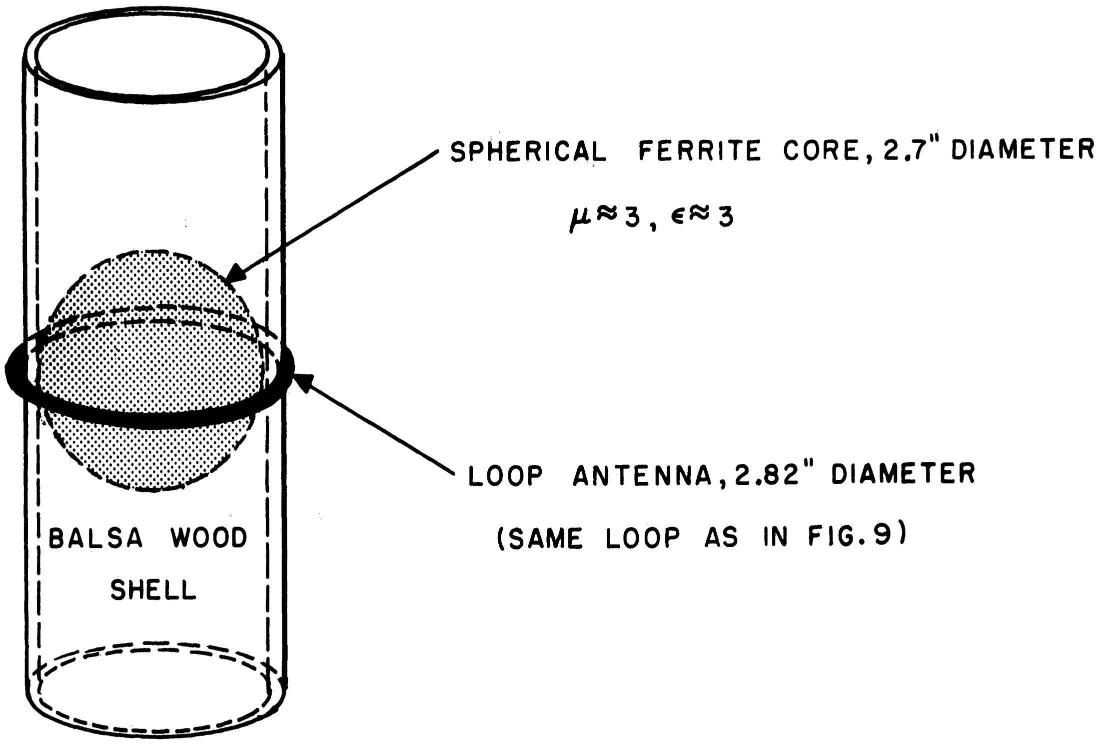


Fig. 11. Ferrite core loop antenna.

A graph of the data from Table I is shown in Fig. 12. There is an appreciable increase in the resistance of the ferrite antenna close to 200 Mc, indicating an improvement in the radiation properties of the antenna. The only way to determine whether this increase in resistance is caused by increased radiation or by increased losses in the antenna is to measure the radiated power from the two antennas. There are two reasons that make it highly probable that the peak in resistance represents an increase in radiated power. The first reason is that the loss of the ferrite in the solid form does not increase until about 250 Mc. The second reason is that a peak in radiation resistance is expected due to a resonance in the core. This resonance is probably similar to that found in the calculations of the radiation resistance for the infinitesimal

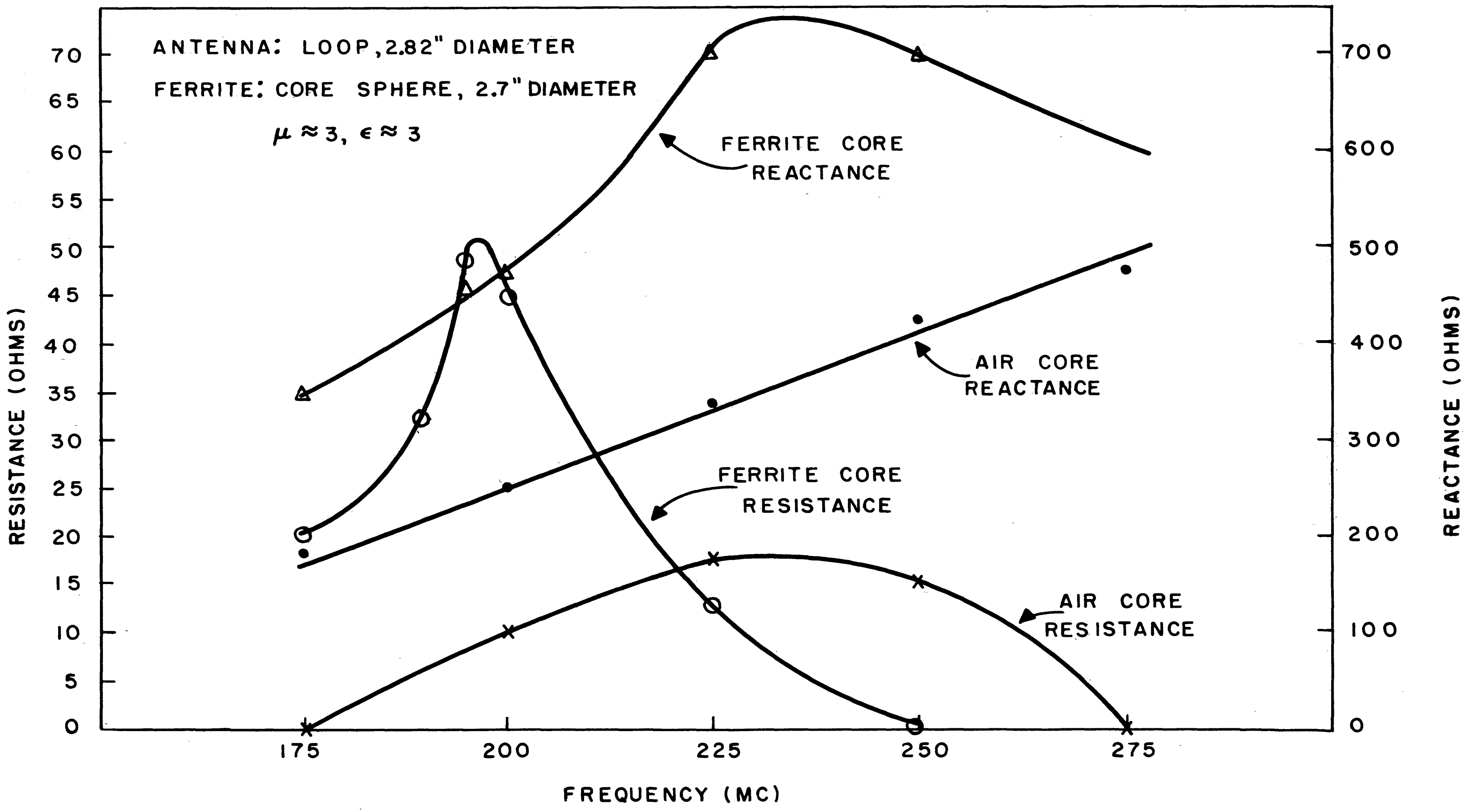


Fig. 12. Impedance of loop antenna.

dipole which were presented in an earlier report.

According to Foster's Reactance Theorem,

$$\frac{dx}{d\omega} \geq 0$$

i.e., for a lossless netowrk, the slope of the reactance vs. frequency plot is always positive. Therefore, the negative slope of the reactance indicates the onset of losses due to the ferrite material.

Several tests were made to measure the radiated power. Because of the generally poor radiation efficiency of the small loop, the power radiated from the balanced two-wire line was of the same order as that radiated by the loop, so the latter power could not be determined. A design is presently being worked on which makes use of a shielded two-wire line between the loop and the balun.

3.5 Materials

The ferrite powder used for the ferrite-core loop antenna has a permeability at 200 Mc of 6.45 in the solid form. In the powdered form, the permeability is reduced to about 3. Similarly, the dielectric constant is reduced to about 3. In the solid form, the magnetic Q of the material is greater than 600. Methods are being investigated to measure the properties of the ferrite directly in the powdered form.

3.6 Biconical Antenna Fabrication

The equipment necessary for the construction of some biconical antennas is now being fabricated. The form will be a monopole over a conducting sheet of aluminum. This sheet, with the appropriate connecting devices, will be readily adaptable to biconical monopoles of arbitrary length and flare angle. A styrofoam shell will be hollowed out to allow

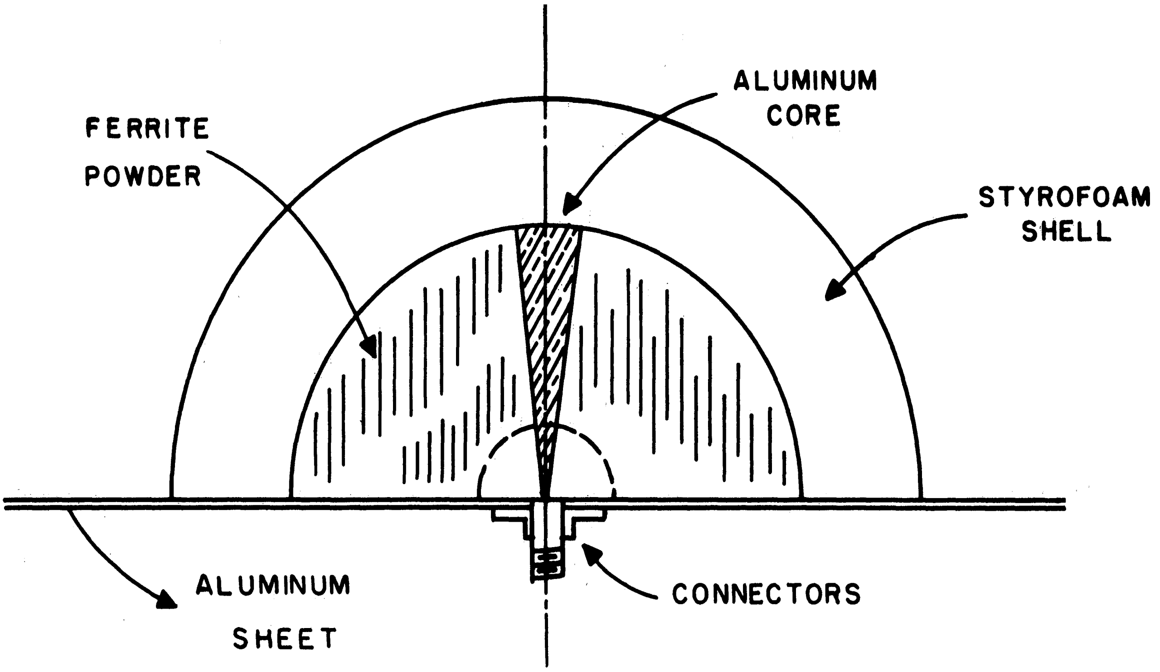


Fig. 13. Ferrite enclosed monopole.

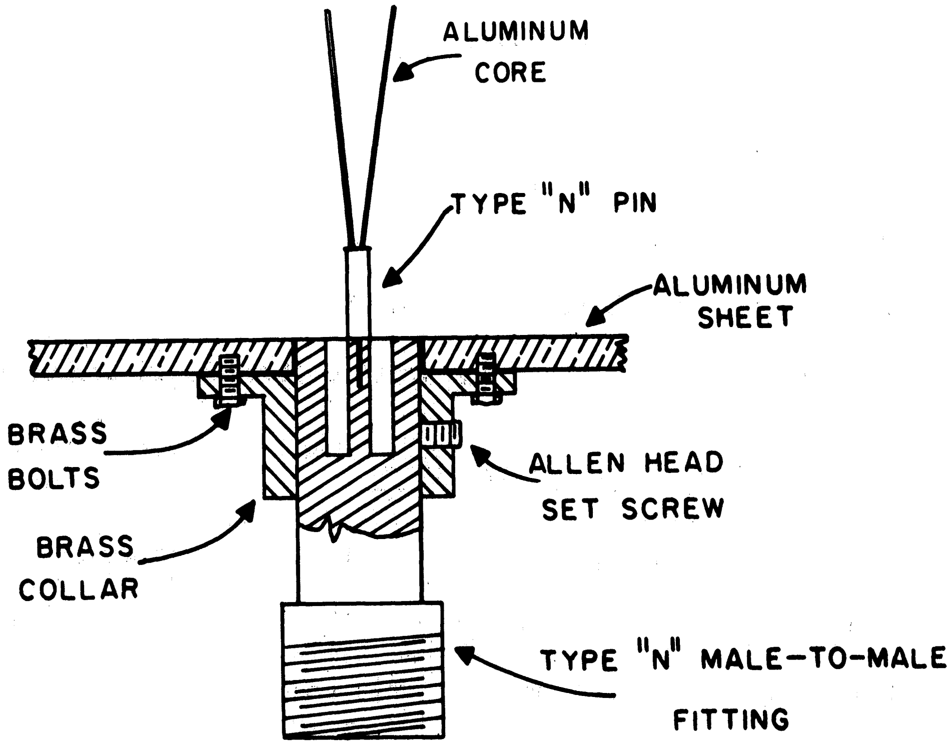


Fig. 14. Monopole terminal connections.

the powdered ferrite to take the form of a hemisphere surrounding the pole (see Fig. 13). The styrofoam has the advantages of strength, machinability, lightness of weight, and extreme transparency to radio waves.

The terminal connections, which are highly important to significant measurements, are shown in Fig. 14. The plane is a 1/8" thick aluminum sheet; the collar is brass. A male-to-male type "N" fitting is machined on one end to allow for the collar. A wooden hemisphere will be provided for strength at the base if necessary. A type "N" pin is pressed on the aluminum core to provide a continuous geometric transition from the connector to the core. Thus a coaxial cable will feed the antenna, eliminating the need for a balun.

4. ACTIVITIES FOR THE NEXT PERIOD

(1) For the spring and summer the project has been realigned in order to have additional help in the fields of RF measurement as well as analytical computations.

(2) With the coming of good weather, it is planned to fully utilize the experimental installation to obtain satisfactory data on elementary antennas as well as more sophisticated antennas utilizing ferrite materials. Antennas selected will include those which have practical radiation properties, including a high directivity, as well as a relatively high efficiency even without the use of ferrite materials. The ferrite materials will then be applied in order to determine what enhancement of characteristics, if any, is to be achieved.

(3) The computational effort, including the use of computers, has appeared to be a very fertile aspect of the over-all research program.

It has been found necessary to utilize more mathematical analysis in the preparation of the problems for the computer. For this reason additional assistance in this specialty has been acquired in terms of one added person for the project.

(4) It is contemplated that the "scimitar" antenna will be analyzed, with the possibility of adapting such to the use of ferrites.

(5) Multifeeding with appropriate phasing to simple antenna elements is one of the projected studies for this research effort. It is quite possible that slow-wave devices will be utilized in this part of the work.

5. SUMMARY

A solution to the problem of scattering of plane waves from a material sphere has been derived. Results to date indicate regions of high energy density within the sphere that may make possible the design of small antennas with large effective apertures. The results of this study will be more completely analyzed and presented in later reports.

Substantial progress has been made in the solution of scattering of plane waves from a material cylinder. Computer solutions are now in progress.

Measurements on ferrite antennas have begun and problems of RF technique have been encountered and partially solved. It is apparent that all test antennas should be as efficient as possible and all equipment be adequately shielded.

The study on a comparison of solid-state and conventional antennas indicates several techniques for determining the usefulness of any given antenna. The method used will depend on the type of antenna being tested, degree of precision needed, and available equipment.

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