Power Loss for Multimode Waveguides and Its Application to Beam–Waveguide System

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***Abstract*— The conventional way of expressing power loss in decibels/meter for a multimode waveguiding system with finite wall conductivity (such as a beam–waveguide (BWG) system with protective shroud) can be incorrect and misleading. The power loss (in decibels) for a multimode waveguiding system is, in general, not linearly proportional to the length of the waveguide. New power-loss formulas for multimode system are derived in this paper for arbitrarily shaped conducting waveguide tubes. In these formulas, there are factors such as $[\exp(jx)-1]/(jx)$, where $x=(\beta_a-\beta_b)\ell$, with β_a and β_b being the propagation constants of the different propagating modes and ℓ being the distance from the source plane to the plane of interest along the guide. For a large BWG supporting many propagating modes, β_a's are quite close to β_b's, thus the mode coupling terms remain important for a very long distance from the source plane. The multimode power-loss formula for a large circular conducting tube has been verified by experiments. This formula was also used to calculate the additional noise temperature contribution due to the presence of a protective shroud surrounding a millimeter-wave BWG system.**

***Index Terms*—Beam waveguides, cylindrical waveguide, noise measurement, waveguide theory.**

I. Introduction and the Consideration of a Fundamental Concept

In textbooks on electromagnetics and guided waves, the perturbation technique is used to calculate the attenuation factor of a given propagating mode in a slightly lossy and highly conducting hollow metallic waveguide. Based on this technique, the attenuation constant for the mth mode $\alpha^{(m)}$ due to conductor loss in a general cylindrical hollow metallic waveguide is found to be [1]–[6]

$$\alpha^{(m)} \simeq \frac{R\oint_c \left[\underline{H}^{(m)} \cdot \underline{H}^{(m)*}\right] d\ell}{2\mathrm{Re}\iint_s \left[\underline{E}^{(m)} \times \underline{H}^{(m)*}\right] \cdot \underline{e}_z \, dA} \tag{1}$$

where R denotes the surface resistance of the metal walls, $\underline{E}^{(m)}$ and $\underline{H}^{(m)}$ are the unperturbed electric and magnetic fields for the mth propagating mode in this waveguide with perfectly conducting walls, Re denotes the real part of the integral, $\underline{e}_z$ is the unit vector in the z-propagating direction, * denotes the complex conjugate of the integral, C is the contour around the cross section of the waveguide, and S is the cross-sectional area of the waveguide. Here, $\alpha^{(m)}$ expressed in nepers/meter is the attenuation constant for the mth propagating mode per unit length of the waveguide.

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A more accurate determination of the attenuation constant $\alpha^{(m)}$ can be obtained through the boundary-value-problem approach. Here, the fields in different regions [i.e., the metal region characterized by $(\in, \mu, \sigma)$ and the vacuum region characterized by $(\in_0, \mu_0)$] of the waveguide are matched at the boundary, yielding a dispersion relation from which the complex propagation constant for each mode may be determined. For this approach, in general, all field components must be assumed to be present. In other words, for a hollow circular metal pipe, the field components (E_z, E_r, E_ϕ, H_z, H_r, H_ϕ) will all be present when circular symmetry of the mode is not present. Here, the circular cylindrical coordinates (r, ϕ, z) are assumed. This was the approach (called the hybrid-mode approach) used by Chou and Lee to calculate modal attenuation in multilayered coated waveguides [7]. Other improved versions of the perturbation formula of (1) for the attenuation constant of a single mode were given by Gustincic [8] and Collin [2].

The intent of this paper is not to improve the power-loss calculation for a single mode. One notes that, for the small loss case, this improvement is negligible. The intent of this paper is to provide a correct way of finding the power loss for the multimode case.

In all of the above considerations, the power loss has always been expressed by $\alpha^{(m)}$ for each mth mode in nepers/meter.

It is the limitation of this way of expressing power loss that we wish to address in the following sections.

When a single mode, say the mth mode, is propagating in this hollow waveguide, the following expression is normally used to represent the power carried by this mode along this waveguide structure:

$$P^{(m)}(z) = P_0^{(m)} e^{-2\alpha^{(m)} z} \approx P_0^{(m)}[1 - 2\alpha^{(m)} z] \tag{2}$$

where $P_0^{(m)}$ is the initial input power of the mth mode and z is the distance along the guide. That this expression is valid if and only if a single mode is propagating alone in this waveguide is usually glossed over in the textbooks. Furthermore, (1) and (2) offer the impression that the power loss in a given waveguide may be expressed by the attenuation constant $\alpha^{(m)}$ in nepers/meter. From (2), for small attenuation, the power loss $P_L^{(m)}$ is

$$\begin{aligned} P_L^{(m)} &= P_0^{(m)} \text{ (Power Input)} - P^{(m)}(z) \text{ (Power Output)} \\ &\simeq P_0^{(m)} 2\alpha^{(m)} z. \end{aligned}$$

Consequently, one may obtain the mistaken impression that since the modes are orthogonal, the total loss is additive when more than one mode is present simultaneously in the waveguide; after all, we know that the total power is additive. For the multimode propagation case, the total power loss should not be expressed through an attenuation constant as certain nepers/meter (or decibels/meter). Indeed, due to the contributions of the cross-product terms in $\underline{J} \cdot \underline{J}^*$ where $\underline{J}$ is the total surface current, and the total power loss in the multimode case is no longer a linear function of the length of the guide, as in the single-mode case.

For example, assume that a given source in an infinitely long hollow conducting waveguide excites two equal amplitude lowest order propagating modes. Further assume that the waveguide can only support these two lowest order propagating modes. The walls of the waveguide are made with highly conducting (but not perfectly conducting) metal. Let us find the total power loss at a distance d from the source plane.

According to the classical textbook formula (1), the attenuation constant for each mode can be calculated using this formula. Say the answer for mode 1 is $\alpha_1 = 0.001$ (nepers/meter) and for mode 2 is $\alpha_2 = 0.002$. (Even if we use the more exact way of calculating the attenuation constant by the boundary-value-problem approach (or the hybrid-mode approach) described in [1], due to the highly conducting nature of the walls, the attenuation constants for these two modes would not deviate much from the given values). Let P_0 be the input power for mode 1 as well as for mode 2. Thus, the power of mode 1 after propagating for a distance z in the waveguide is $P_0 \exp[-2\alpha^{(1)}z]$ and for mode 2 is $P_0 \exp[-2\alpha^{(2)}z]$. Since the power is additive, the total power loss is

$$\begin{aligned} P_{\text{Total Loss}} &= P_{\text{Input}} - P_{\text{Output}} \\ &= 2P_0 - P_0\{\exp[-2\alpha^{(1)}z] + \exp[-2\alpha^{(2)}z]\} \\ &= 2P_0[\alpha^{(1)} + \alpha^{(2)}]z. \end{aligned} \tag{3}$$

Extending this concept to n modes would yield

$$P_{\text{Total Loss}} = 2P_0(\alpha^{(1)} + \alpha^{(2)} + \cdots + \alpha^{(n)})z.$$

Thus, according to the above formula, no matter how small z or α is, $P_{\text{Total Loss}}$ is proportional to z. To demonstrate that this concept is incorrect, consider the following: a high-gain horn radiating inside a waveguide with boresite along the axis of the waveguide. If the modes are considered to be uncoupled, then the loss for each mode can be independently computed and summed. Therefore, the power loss per unit length would be independent of the position in the waveguide. A simple thought experiment should be sufficient to conclude that if the diameter became larger and larger, one would certainly expect the loss per unit length in a region near the plane of the horn aperture (where there is virtually no radiation from the horn) to be quite different from the loss at a distance where the radiation pattern from the horn would illuminate the waveguide walls. This is very similar to the experiment described later in this paper. A correct interpretation of the perturbation theory would be to apply it to the total tangential **H** field on the waveguide wall. Since the power loss formula uses tangential **H** squared, the modal fields are thus coupled through this term resulting in a waveguide loss which varies as a function of the axial dimension. The results are precisely what one would expect for the horn example, i.e., very little loss very near the aperture plane and increasing significantly when the radiation pattern of the horn intersects the waveguide wall. Using this approach, the resultant theoretical/numerical data compares very favorably with the measured experimental data.

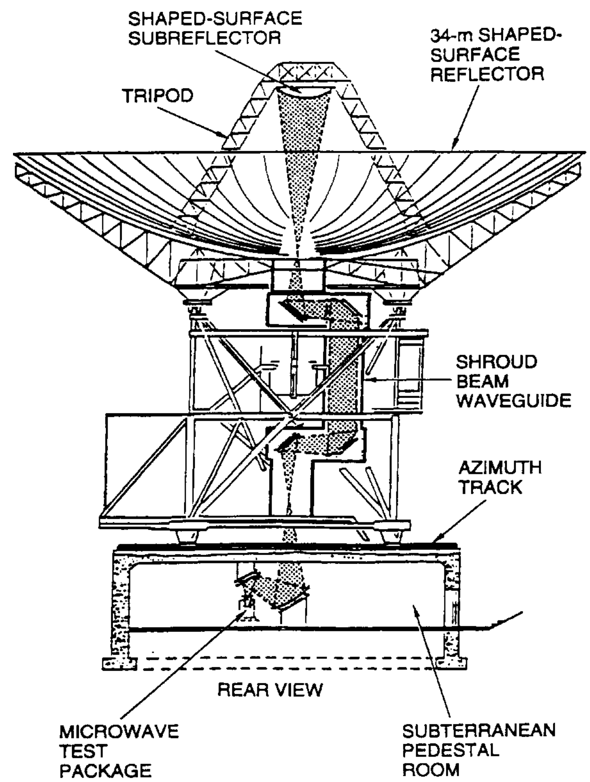


Fig. 1. View of Jet Propulsion Laboratory's (JPL's) DSS-13 BWG antenna.

Therefore, the purpose of this paper is to address the power-loss problem when more than one mode is simultaneously present in the waveguide. This effort is motivated by our desire to verify the measured data for a millimeter-wave beam–waveguide (BWG) with a protective shroud consisting of sections of a round conducting tube, as shown in Fig. 1. Solution of this problem is of great importance in optimizing the design to yield minimum noise temperature for the NASA/Deep Space Network's low-noise microwave receiving system [9].

II. Formulation of the Problem and Formal Solution

Shown in Fig. 2 is the geometry of the canonical problem. A uniform conducting waveguide of arbitrary cross section with its axis aligned in the z-direction has a length ℓ. In the $z = 0$ plane, the transverse electric field $\underline{E}_t(x, y)$ is assumed to be given. Thus, the amplitudes of all the modes (propagating and evanescent modes) can be calculated [1]–[5] and are assumed to be known. We wish to calculate the power loss of the fields due to the imperfect conductivity of the wall with intrinsic wave resistance (surface resistance) R.

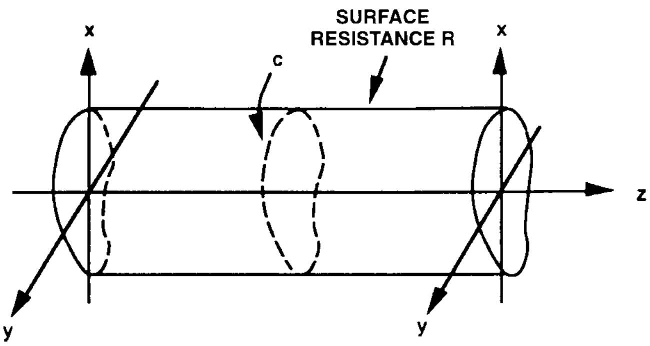


Fig. 2. Geometry of the problem.

From Ohm's Law and Poynting's vector theorem, the power loss is given by [1]–[6]

$$P_L = \tfrac{1}{2} R \iint_A (\underline{J}_s \cdot \underline{J}_s{}^*)\, dA \qquad (4)$$

where

$\underline{J}_s$ $\underline{n} \times \underline{H} =$ surface current density on the wall;
$\underline{n}$ unit vector normal to the wall surface;
$\underline{H}$ total magnetic field in the waveguide;
A surface area of the wall

or

$$P_L = \tfrac{1}{2} R \iint_A (\underline{H}_\tau \cdot \underline{H}_\tau^*)\, dA \qquad (5)$$

where $\underline{H}_t$ is the component of the total magnetic field which is tangential to the wall surface. It is known that in a hollow arbitrarily shaped uniform waveguide with a conducting wall there can exist two sets of eigenmodes: [1]–[6] transverse electric (TE) modes and transverse magnetic (TM) modes with a specific propagation constant for each mode. The total fields for TE modes are

$$\underline{E}^{(TE)}(x, y, z) = \sum_{m=1}^{\infty} A_m^{(\mathrm{TE})} \underline{E}_{mt}^{(\mathrm{TE})}(x, y) e^{j\beta_m^{(\mathrm{TE})} z} \qquad (6)$$

$$\underline{H}^{(\mathrm{TE})}(x, y, z) = \sum_{m=1}^{\infty} A_m^{(\mathrm{TE})} \Big[\underline{H}_{mt}^{(\mathrm{TE})}(x, y) + H_{mz}^{(\mathrm{TE})}(x, y) \underline{e}_z \Big] e^{j\beta_m^{(\mathrm{TE})} z} \qquad (7)$$

where $\underline{E}_{mt}^{(\mathrm{TE})}$ and $(\underline{H}_{mt}^{(\mathrm{TE})} + H_{mz}^{(\mathrm{TE})} \underline{e}_z)$ are connected through the Maxwell's equations and $\beta_m^{(\mathrm{TE})}$ is the propagation constant of the mth TE eigenmode, and the total fields for TM modes are

$$\underline{E}^{(\mathrm{TM})}(x, y, z) = \sum_{m=1}^{\infty} A_m^{(\mathrm{TM})} \Big[\underline{E}_{mt}^{(\mathrm{TM})}(x, y) + E_{mz}^{(\mathrm{TM})}(x, y) \underline{e}_z \Big] e^{j\beta_m^{(\mathrm{TM})} z} \qquad (8)$$

$$\underline{H}^{(\mathrm{TM})}(x, y, z) = \sum_{m=1}^{\infty} A_m^{(\mathrm{TM})} \underline{H}_{mt}^{(\mathrm{TM})}(x, y)\, e^{j\beta_m^{(\mathrm{TM})} z} \qquad (9)$$

where $\underline{H}_{mt}^{(\mathrm{TM})}$ and $(\underline{E}_{mt}^{(\mathrm{TM})} + E_{mz}^{(\mathrm{TM})} \underline{e}_z)$ are connected through Maxwell's equations and $\beta_m^{(\mathrm{TM})}$ is the propagation constant of the mth TM eigenmode. $A_m^{(\mathrm{TE})}$ and $A_m^{(\mathrm{TM})}$ are arbitrary amplitude coefficients for TE and TM modes. The subscript t indicates the transverse components of the field (transverse to the z-direction). The index m is used to tally the modes—it does not necessarily correspond to mode order. One notes that β_m may take on negative values, indicating modes propagating in the opposite direction.

Substituting (6)–(9) into (5) yields

$$\begin{aligned} P_L = \tfrac{1}{2} R \Bigg\{ &\sum_{m=1}^{M} \sum_{n=1}^{M} A_m^{(\mathrm{TE})} A_n^{(\mathrm{TE})*} \oint_c [H_{mc}^{(\mathrm{TE})} H_{nc}^{(\mathrm{TE})*} \\ &+ H_{mz}^{(\mathrm{TE})} H_{nz}^{(\mathrm{TE})*}]\, dc \int_0^\ell e^{j(\beta_m^{(\mathrm{TE})} - \beta_n^{(\mathrm{TE})})\, z}\, dz \\ &+ \sum_{m'=1}^{M'} \sum_{n=1}^{M} A_{m'}^{(\mathrm{TM})} A_n^{(\mathrm{TE})*} \oint_c [H_{m'c}^{(\mathrm{TM})} H_{nc}^{(\mathrm{TE})*}]\, dc \\ &\cdot \int_0^\ell e^{j(\beta_{m'}^{(\mathrm{TM})} - \beta_n^{(\mathrm{TE})})\, z}\, dz \\ &+ \sum_{m=1}^{M} \sum_{n'=1}^{M'} A_m^{(\mathrm{TE})} A_{n'}^{(\mathrm{TM})*} \oint_c [H_{mc}^{(\mathrm{TE})} H_{n'c}^{(\mathrm{TM})*}]\, dc \\ &\cdot \int_0^\ell e^{j(\beta_m^{(\mathrm{TE})} - \beta_{n'}^{(\mathrm{TM})})\, z}\, dz \\ &+ \sum_{m'=1}^{M'} \sum_{n'=1}^{M'} A_{m'}^{(\mathrm{TM})} A_{n'}^{(\mathrm{TM})*} \oint_c [H_{m'c}^{(\mathrm{TM})} H_{n'c}^{(\mathrm{TM})*}]\, dc \\ &\cdot \int_0^\ell e^{j(\beta_{m'}^{(\mathrm{TM})} - \beta_{n'}^{(\mathrm{TM})})\, z}\, dz \Bigg\}. \end{aligned} \qquad (10)$$

Here, c is the contour around the inner surface of the waveguide, which is also normal to the z-axis (see Fig. 2). The subscript c represents the component of the transverse field that is tangential to the contour c, M is the number of TE propagating modes, M' is the number of TM propagating modes and m, m', n, n' are mode indices. Simplifying (10) gives

$$P_L = [\text{Part 1}] + [\text{Part 2}] \qquad (11)$$

with (12) and (13), shown at the bottom of the following page, where

$$\begin{aligned} I_m^{(\mathrm{TE})} &= \oint_c [|H_{mc}^{(\mathrm{TE})}|^2 + |H_{mz}^{(\mathrm{TE})}|^2]\, dc \\ I_{m'}^{(\mathrm{TM})} &= \oint_c |H_{m'c}^{(\mathrm{TM})}|^2\, dc \\ I_{mn}^{(\mathrm{TE})} &= \oint_c [H_{mc}^{(\mathrm{TE})} H_{nc}^{(\mathrm{TE})*} + H_{mz}^{(\mathrm{TE})} H_{nz}^{(\mathrm{TE})*}]\, dc \\ I_{m'n'}^{(\mathrm{TM})} &= \oint_c [H_{m'c}^{(\mathrm{TM})} H_{n'c}^{(\mathrm{TM})*}]\, dc \\ I_{m'n}^{(\mathrm{TM})(\mathrm{TE})} &= \oint_c [H_{m'c}^{(\mathrm{TM})} H_{nc}^{(\mathrm{TE})*}]\, dc \\ I_{mn'}^{(\mathrm{TE})(\mathrm{TM})} &= \oint_c [H_{mc}^{(\mathrm{TE})} H_{n'c}^{(\mathrm{TM})*}]\, dc. \end{aligned} \qquad (14)$$

It should be noted that P_L is always purely real.

One should point out that all the field components used in the above expressions are assumed to be the field components for a perfectly conducting waveguide. The use of surface resistance R and Ohm's Law to calculate the total power loss is an application of the perturbation technique.

Using the orthogonality properties of these field components, one can show that the total power carried in a multimode waveguide is the sum of the power carried by each propagating mode in this multimode waveguide. On the other hand, (11) shows that power losses or attenuation of different simultaneously existing modes are not simply additive, as indicated by the first bracketed term [Part 1]. The correct expression must include the second bracketed term [Part 2], which shows the cross-product terms. Indeed, the use of an attenuation constant to describe power loss in a waveguide should be limited to the single-mode unidirectional propagation case only, because only for this case is the power loss linearly dependent on the length of the guide. For the multimode propagation case, the power loss varies with the length of the guide in a rather complicated manner, as shown in (11). Equation (11) vividly demonstrates the importance of the modal coupling term. Since the factor

$$f(x) = [\exp(jx) - 1]/(jx) \tag{15}$$

(where $x = (\beta_1 - \beta_2)\ell$, β_1, and β_2 are the propagation constants for the coupling modes, and ℓ is the distance from the entrance of the waveguide to the point of interest along the guide) determines the importance of the coupling term, let us now examine this factor closely. The function $|f(x)|$ is largest when $x \to 0$ and begins to diminish and approaches zero when x increases. This means that the cross-product terms in (11) (i.e., [Part 2]) are important when the difference between the propagation constants of the propagating modes that are excited in the waveguide is small and/or when ℓ is small, such that the product $(\beta_1 - \beta_2)\ell$ is small. This condition is particularly true when the transverse dimensions of the waveguide are very large, such as in the BWG case that we considered. It is also noted that under certain conditions, [Part 2] can be negative. This means that the total power loss can be less than that given by [Part 1], the part representing only the additive aspect of power loss by each mode in a multimode waveguide.

When $|x| \gg 1$, then $f(x) \to 0$, and the coupling terms in (11) [Part 2] approach zero. This means that when $\ell \to \infty$, [Part 2] $\to 0$, and the usual decoupled result given by [Part 1] in (11) becomes valid. Thus, as $\ell \to \infty$, the power loss for each mode in the multimode waveguide is additive.

III. Application to BWG Noise Temperature Computations

We shall now apply the above theory to calculate the conductivity loss (power loss) in a large BWG tube. The noise temperature contributed by the conductivity loss in a BWG can then be easily computed. Computed results are compared with measured data from an experiment, validating the theory.

Large BWG-type ground-station antennas are generally designed with metallic tubes enclosing the BWG mirrors. The scattered field from a BWG mirror is obtained by the use of a physical optics integration procedure with a Green's function appropriate to the circular waveguide geometry [10]. In this manner, the coefficients $A_m^{(\mathrm{TE}),\,(\mathrm{TM})}$ of the circular waveguide modes that are propagating in the oversized waveguides are determined.

Knowing the coefficients $A_m^{(\mathrm{TE}),\,(\mathrm{TM})}$, one may calculate the tangential magnetic fields for the TE and TM modes from (7) and (9). The total tangential magnetic field is the sum of these tangential magnetic fields. Substituting the total tangential magnetic field into (5) and carrying out the integral in (5) numerically, one may readily obtain the total power loss P_L. This numerical technique is quite general; it can be applied to a metal tube waveguide of arbitrary shape. Another way may also be used: knowing $A_m^{(\mathrm{TE}),\,(\mathrm{TM})}$ for the modes in a circular metal tube (sleeve) waveguide or in a rectangular

$$[\text{Part 1}] = \tfrac{1}{2}R\ell\left[\sum_{m=1}^{M}|A_m^{(\mathrm{TE})}|^2 I_m^{(\mathrm{TE})} + \sum_{m'=1}^{M'}|A_{m'}^{(\mathrm{TM})}|^2 I_{m'}^{(\mathrm{TM})}\right] \tag{12}$$

$$\begin{aligned}[\text{Part 2}] = \tfrac{1}{2}R\ell\Bigg(&\sum_{m=1}^{M}\sum_{\substack{n=1\\n\neq m}}^{M} A_m^{(\mathrm{TE})}A_n^{(\mathrm{TE})*}I_{mn}^{(\mathrm{TE})}\left\{\frac{e^{j[\beta_m^{(\mathrm{TE})}-\beta_n^{(\mathrm{TE})}]\ell}-1}{j[\beta_m^{(\mathrm{TE})}-\beta_n^{(\mathrm{TE})}]\ell}\right\}\\
&+\sum_{m'=1}^{M'}\sum_{\substack{n'=1\\n'\neq m'}}^{M'} A_{m'}^{(\mathrm{TM})}A_{n'}^{(\mathrm{TM})*}I_{m'n'}^{(\mathrm{TM})}\left\{\frac{e^{j[\beta_{m'}^{(\mathrm{TM})}-\beta_{n'}^{(\mathrm{TM})}]\ell}-1}{j[\beta_{m'}^{(\mathrm{TM})}-\beta_{n'}^{(\mathrm{TM})}]\ell}\right\}\\
&+\sum_{m'=1}^{M'}\sum_{n=1}^{M} A_{m'}^{(\mathrm{TM})}A_n^{(\mathrm{TE})*}I_{m'n}^{(\mathrm{TM})(\mathrm{TE})}\left\{\frac{e^{j[\beta_{m'}^{(\mathrm{TM})}-\beta_n^{(\mathrm{TE})}]\ell}-1}{j[\beta_{m'}^{(\mathrm{TM})}-\beta_n^{(\mathrm{TE})}]\ell}\right\}\\
&+\sum_{m=1}^{M}\sum_{n'=1}^{M'} A_m^{(\mathrm{TE})}A_{n'}^{(\mathrm{TM})*}I_{mn'}^{(\mathrm{TE})(\mathrm{TM})}\left\{\frac{e^{j[\beta_m^{(\mathrm{TE})}-\beta_{n'}^{(\mathrm{TM})}]\ell}-1}{j[\beta_m^{(\mathrm{TE})}-\beta_{n'}^{(\mathrm{TM})}]\ell}\right\}\Bigg)\end{aligned} \tag{13}$$

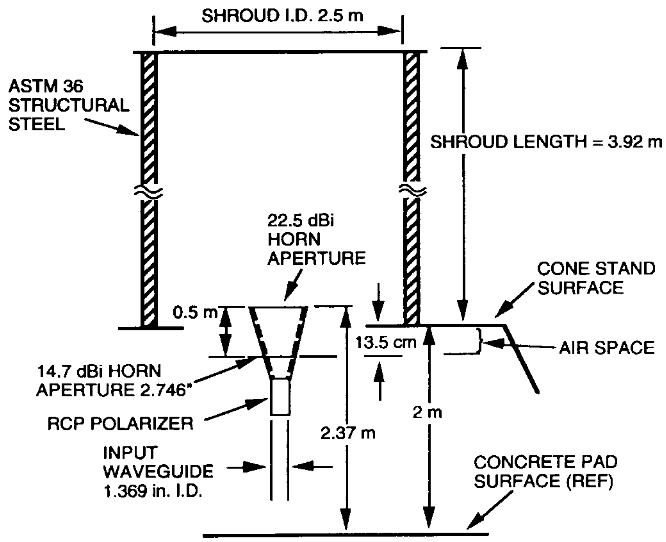


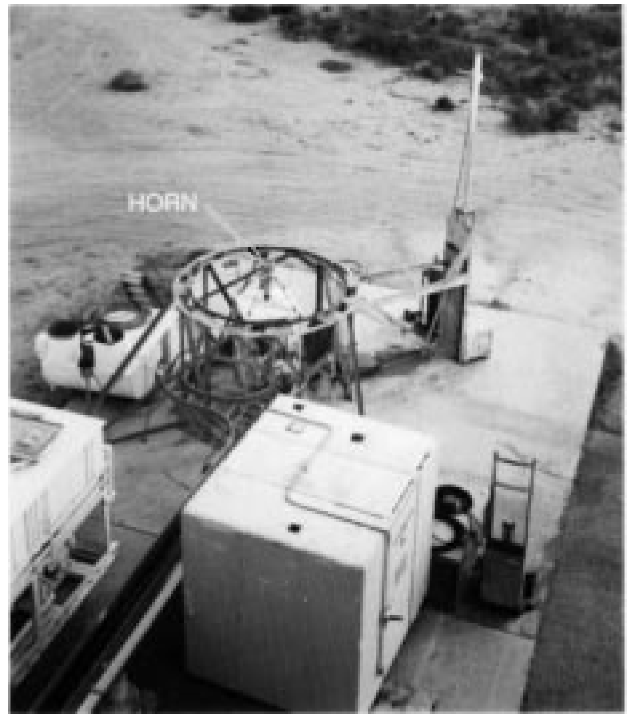
Fig. 3. Experimental setup.

metal tube waveguide, one may derive analytic expressions for the total power loss P_L.

The above numerical approach was used to calculate the conductivity loss of a short length of BWG tube [11]. The experiment utilized a 3.92-m-long 2.5-m-diameter structure steel tube and a very sensitive noise temperature measuring radiometer (see Figs. 3 and 4). Noise temperature comparisons were made between several different horns radiating in free space and radiating into the BWG tube. The experiment also included measurements with the steel tube and the tube lined with aluminum sheets. Utilizing the measured conductivity of the aluminum and steel [12] (see Table I) and the computed modes in the BWG tube, a conductivity loss was computed and converted into a noise temperature prediction. For the 14.7-dBi gain horn, the TE_{1p} and TM_{1p} modes to $p = 50$ were included, and for the 22.5 dBi gain horn, modes to $p = 22$ were included. The following formula was used for the conversion:

$$\text{Noise Temperature in K} = (P_L/P_T)T_0 \qquad (16)$$

where P_L is the total power loss, P_T is the given total input power, and T_0 is the ambient temperature in K (for room temperature, $T_0 = 293.1$ K). A comparison of the measurement with both the new theory (11) and the textbook theory (3) is shown in Table II. The most dramatic difference was with the higher gain (22.5-dBi) horn. It was this experimental result which showed that the result obtained according to (3) was incorrect. The measurement was 0.1 K $\pm$ 0.1 K and there was no question that the calculation of 2.6 K from (3) was significantly outside the range measurement uncertainty. The explanation can be seen in Fig. 5, which plots the attenuation loss as a function of tube size. Because the high-gain horn does not "illuminate" the wall until further down the tube from its aperture plane, there is only a very small loss near the aperture. This clearly demonstrates the fact that the power loss is not linearly dependent on z, and thus validates the analysis.



(a)

(b)

Fig. 4. Measurement setup. (a) Horn in free space. (b) Horn with BWG tube.

IV. Conclusions

The concept of expressing power loss along a given uniform waveguide in nepers/meter must be used with caution. This concept is only generally true for single-mode unidirectional propagation. When more than one mode exists simultaneously, the power loss is no longer linearly proportional to the length of waveguide. Depending on the differences for the

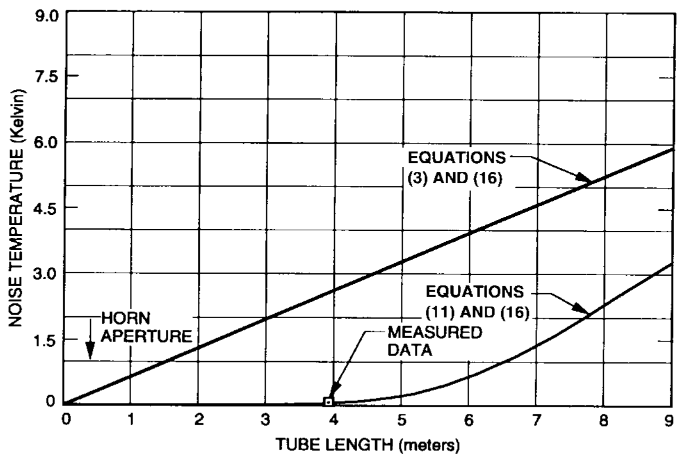


Fig. 5. Noise temperature versus tube length for 22.5-dBi gain horn.

TABLE I
ELECTRICAL CONDUCTIVITIES OF SHROUD MATERIALS [12]

Material	Effective Conductivity mhos/meter
BWG antenna shroud ASTM A36 steel	0.003×10^7
0.064 in. thick 6061 aluminum sheet	2.2×10^7
0.024 in. thick galvanized steel	1.2×10^7
High-conductivity copper	5.66×10^7

TABLE II
COMPARISON OF EXPERIMENTAL AND THEORETICAL RESULTS. THEORETICAL RESULTS ARE CALCULATED FROM (11) AND (16) AND FROM (3) AND (16)

	Measured, K	Calculated	
		New Method (11), K	Textbook Method (3), K
22.5-dBi gain horn with steel tube	0.1 ± 0.1	0.1	2.6
14.7-dBi gain horn with steel tube	2.5 ± 0.4	2.3	3.0
14.7-dBi gain horn with aluminum tube	0.2 ± 0.1	0.09	0.11

propagation constants of the coexisting propagating modes and the length of the waveguide, the total power loss may be more than, equal to, or less than the proportional sum of the power losses for each mode propagating separately, as shown in (11).

Accurate formulas for the total power loss by taking mode coupling into account can be derived for an arbitrarily shaped conducting tube, circular conducting tube, and rectangular conducting tube. The factor $f(x)$ [see (15)]—where $x = (\beta_a - \beta_b)\ell$, with $(\beta_a - \beta_b)$ being the differences between the propagation constants of various modes propagating simultaneously in the conducting tube, and ℓ being the length of the waveguide from the source plane to the plane of interest along the guide—appears to be the governing factor that controls the importance of mode coupling between mode a and mode b in affecting the total power loss calculation. Since the factor $\sin x/x$ approaches zero as x approaches infinity, the effect of the term containing this factor approaches zero, indicating the diminishing effect of mode coupling on the total power-loss calculation. Since $x = (\beta_a - \beta_b)\ell$ in order that x may approach a large value quickly, two possibilities exist.

1) If β_a is close to β_b, as in the case of a very large guide, then ℓ must be very long in order that x may be large, indicating that the mode-coupling effect can affect the total loss calculation for a very long distance from the source plane.
2) If β_a is not close to β_b, as in the case of a smaller guide, then ℓ can be relatively short for x to be large enough so that the $f(x)$ term may be negligible, indicating that the mode coupling term only affects the total loss calculation for a relatively short distance from the source plane.

When applied to the JPL millimeter-wave BWG case, one notes that β_a is very close to β_b. The newly developed loss formula for an oversized circular conducting tube was thus used to calculate the additional noise temperature contribution due to the presence of a protective shroud surrounding a millimeter-wave BWG.

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