A Comparison Between Near-Field Shielding-Effectiveness Measurements Based on Coaxial Dipoles and on Electrically Small Apertures

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Abstract—The near-field shielding effectiveness (SE) of a material may be measured by placing it between two closely spaced dipoles (electric or magnetic) and noting the resulting insertion loss. An alternative approach is to cover an electrically small aperture with the test material and to measure the resulting loaded aperture polarizability (electric or magnetic), as is done in a dual TEM cell. Expressions are developed herein which relate these two configurations.

Key Words—Small aperture, coaxial dipoles, dual TEM cell, near-field, shielding effectiveness.

Index Code—A12d, F16, F12d.

I. INTRODUCTION

Electromagnetic coupling between a pair of elementary dipoles (electric or magnetic) separated by a planar sheet is a well-studied shielding problem because it features both a simple shield geometry and fundamental antennas. In the near-field region, an electric dipole produces a high-impedance field, while a magnetic dipole generates a low-impedance field; thus, shielding can be quite different for the two cases. Shielding-effectiveness (SE) measurement routines typically involve coupling fields between a source and receiver first with no material present (the reference level), and then with the test material inserted (loaded level). The ratio of the received powers (reference/loaded), or insertion loss, gives a quantitative SE measure, usually expressed in decibels. However, developing a near-field SE measurement method based on dipoles and planar sheets is difficult. Infinite planar sheets are analytically convenient, but actual measurements necessarily involve finite samples. Unfortunately, finite sheets allow for coupling paths which do not go through the shield. These indirect paths may be removed by placing the transmitting antenna inside a metal box. Unfortunately, the field structure in a metal box is significantly different from the desired free-space distribution. Anechoic material may be used to reduce resonances; however, this approach can be expensive, it suffers at lower frequencies where absorbing material performs poorly, and it may not completely eliminate resonances. A final difficulty with placing test materials between antennas is that only a fraction of the power coupled through the shield is received. Should the shield focus energy on the receiving point, the result could be a measured gain in spite of a reduction in the total energy transmitted.

Considerations such as these are the motivation for considering SE measurements using waveguides (rather than radiating antennas) as the transducers. Coaxial transmission-line fixtures reproduce far-field conditions by propagating the TEM mode through the material under test (MUT) [1]. The dual TEM cell (DTC) simulates near-field SE by coupling energy between two TEM cells through a MUT-covered aperture [2]. The difference between the TEM modes excited in the forward and backward directions in the receiving cell of a DTC allows the electric and magnetic field coupling through the MUT to be separately determined. This yields information on both high-impedance and low-impedance (near-field) shielding capability. Clearly, the DTC method is quite different from the more direct approach of placing the test material close to an antenna. Thus, the two concepts of simulating near-field conditions need to be tied together.

One difficulty is that solutions to both problems (aperture coupling and dipoles about a planar shield) involve cumbersome integral expressions. Computer-generated parametric studies are useful but give little indication as to the relative role of important variables (such as the dipole separation or the aperture size). The purpose of this paper is to develop approximations leading to simple closed-form SE expressions. Low-frequency restrictions are natural since, for both apertures in waveguides and the near fields of antennas, physical dimensions tend to be small compared to a free-space wavelength. High conductivity will be assumed since it is typical of most shielding materials. Small-aperture theory will be discussed first, followed by a development of approximate dipole coupling expressions. The results will then be compared along with an experimental example.

II. SMALL-APERTURE THEORY

When excited by an incident field, apertures which are small compared to a wavelength radiate as if the source is a set of electric and magnetic dipoles. The strengths of these dipoles are proportional to the wavelength radiate as if the source is a set of electric and magnetic dipoles. The strengths of these dipoles are proportional to a wavelength. When excited by an incident field, apertures which are small compared to a wavelength radiate as if the source is a set of electric and magnetic dipoles. The strengths of these dipoles are proportional to a wavelength.
abilities have also been developed for thin sheets [3], [4] and adapted to the problem here [1], [5]. The respective cases will be denoted with the subscripts $E$ (electric) and $H$ (magnetic). Small-aperture theory predicts that, for a circular aperture of radius $r$, SE measurements based on the ratio of unloaded-aperture to loaded-aperture polarizabilities will give [5]

$$SE_E = 20 \log \left| \frac{8}{3\pi \eta_0 k_0 r Z_s} \right|$$

and

$$SE_H = 20 \log \left| 1 + j \frac{4}{3\pi} \frac{k_0 r \eta_0}{Z_s} \right|. \quad (1)$$

In (1) $k_0$ and $\eta_0$ are the free-space wavenumber and intrinsic impedance, and $Z_s$ is the equivalent sheet impedance (transfer impedance) of the material covering the aperture. Both expressions ignore the possible effects of contact resistance, which can be significant. Experimental data have confirmed their usefulness, particularly the magnetic shielding expression $SE_H$ where contact resistance is less of a problem. The above results may be applied to noncircular apertures by choosing an equivalent radius which gives the same polarizability value in the empty-aperture case. For example, for a square aperture of side length $s$, the conversion is $r = 0.58s$. For a planar sheet of good conductor, $Z_s$ is given approximately by

$$Z_s = \eta/j \sin kd \quad (2)$$

where $\eta = (\mu/\varepsilon)^{1/2}$ and $k = \omega(\mu\varepsilon)^{1/2}$ are the intrinsic impedance and the propagation constant of the MUT, $\mu$ and $\varepsilon = \varepsilon_0\varepsilon_r$ are the permeability and the complex permittivity for the MUT, and $d$ is its thickness. Thus, for thin shields

$$SE_E = 20 \log \left| \frac{8}{3\pi \eta_0 k_0 r \eta} \sin kd \right|$$

and

$$SE_H = 20 \log \left| 1 - j \frac{4}{3\pi} \frac{k_0 r \eta_0}{\eta} \sin kd \right|. \quad (3)$$

These expressions will be compared to the planar-shield case.

III. DIPOLE COUPLING

The dipole configuration to be analyzed is shown in Fig. 1. The shield is symmetrically located between the two dipoles, either electric or magnetic. The dipoles are coaxial along the $z$ axis of a cylindrical coordinate system ($\rho$, $\phi$, $z$), and separated by a distance $z_0$ with the transmit dipole located at $z = 0$. The coaxial configuration is chosen here because symmetry makes the analysis simple, and, consequently, it has the most relevance to previous work. The shield parameters ($k$, $\eta$, $d$) are defined as in the small-aperture case. This configuration has been analyzed recently by Yang and Mittra [6] based on plane-wave spectral representations for arbitrary dipole orientations. A number of errors appear in the transmission matrix exponentials in [6]. For the correct expressions, consult an earlier version of the same material [7]. The integral expressions from their general analysis will be approximated along the lines developed by Bannister [8] and Franceschetti and Papas [9].

The coupling equations for both the electric and magnetic dipole cases take a similar form because of duality. Let $F_\rho$ represent the $\rho$ component of the electric or magnetic field (to which the receiving dipole at $z = z_0$ will respond) due to a $z$-oriented transmitting dipole (see Fig. 1). Yang and Mittra's analysis shows that $F_\rho$ may be written

$$F_\rho(\rho, z_0) = K_{\chi_{\rho}} \int_0^\infty \frac{\tau \lambda^3}{C(\lambda)} e^{i(\rho z_0 - d)} J_0(\lambda \rho) d\lambda \quad (4)$$

where

$$C(\lambda) = (\tau^2 + \chi_{\rho}^2 \tau_0^2) \sin \tau d + 2j\chi_{\rho} \tau \tau_0 \cos \tau d$$

$$\tau_0 = (k_0^2 - \lambda^2)^{1/2}$$

$$\tau = (k^2 - \lambda^2)^{1/2} \quad (5)$$

$K$ depends on the dipole strength, $J_0$ is a zeroth-order Bessel function, $\chi_{\rho} = \varepsilon/\varepsilon_0 = \varepsilon$, for an electric dipole, and $\chi_{\rho} = \mu/\mu_0 = \mu$, for a magnetic dipole. For coaxial dipoles, only the axial component ($\rho = 0$) needs to be considered, which reduces $F_\rho$ to a function of $z_0$ alone:

$$F_\rho(z_0) = K_{\chi_{\rho}} \int_0^\infty \frac{\tau \lambda^3}{C(\lambda)} e^{i\tau(z_0 - d)} d\lambda. \quad (6)$$

It is convenient to break this integral into two parts. For $\lambda \leq k_0$, let $u = (k_0^2 - \lambda^2)^{1/2}$; and for $\lambda \geq k_0$, let $u = (\lambda^2 - k_0^2)^{1/2}$. These substitutions yield two integrals, $F_\rho = F_1 + F_2$, given by

$$F_1(z_0) = K_{\chi_{\rho}} \int_0^{k_0} \frac{\tau u(k_0^2 - u^2)e^{i\tau(z_0 - d)}}{(\tau^2 + \chi_{\rho}^2 u^2) \sin \tau d} \tau d + 2j\chi_{\rho} \tau \tau_0 \cos \tau d \quad \tau = (k^2 - k_0^2 + u^2)^{1/2}$$

for $\lambda < k_0$; and

$$F_2(z_0) = K_{\chi_{\rho}} \int_{k_0}^\infty \frac{\tau u(k_0^2 - u^2)e^{i\tau(z_0 - d)}}{(\tau^2 + \chi_{\rho}^2 u^2) \sin \tau d} \tau^2 d$$

for $\lambda > k_0$. The expressions for the magnetic field are obtained by replacing $\chi_{\rho}$ with $\chi_{\rho}$.
and
\[ F'_z(z_0) = K_{Xr} \int_0^\infty \frac{\tau u(k_0^2 + u^2) e^{-u(z_0 - d)}}{(r^2 - \chi_r^2 u^2) \sin \tau - 2 \chi_r \tau u \cos \tau \tau} \, \tau = (k_r^2 - k_0^2 - u^2)^{1/2}. \] (7)

These integrals may be evaluated directly for the case of no shield \((d = 0)\). Designating the result \(F^0_\text{t} = F'_z(0) + F^2_\text{t}\), we find that
\[ F^1_\text{t}(z_0) = -\frac{K}{2} \int_0^\infty \frac{u(k_0^2 - u^2) e^{iu(z_0 - d)}}{(k_0^2 + u^2) e^{-u(z_0 - d)}} \, du \] and
\[ F^2_\text{t}(z_0) = -\frac{K}{2} \int_0^\infty \frac{u(k_0^2 + u^2) e^{-u(z_0 - d)}}{(k_0^2 - u^2) e^{u(z_0 - d)}} \, du. \] (8)

Both integrals may be evaluated with the result that
\[ F^1_\text{t}(z_0) = K \frac{1}{2} \frac{1}{z_0} (1 - jk_0 z_0) e^{jko z_0}. \] (9)

This establishes the reference-level dipole strength (no shield present).

Most shielding materials will be good conductors. Thus, we will assume that \(|k/k_0| \gg 1\) and approximate the integrals in (7) accordingly. Note that \(\tau = k(1 - k_0^2/k_2^2) \pm u^2/k_2^2)^{1/2}\) depending upon whether \(F^1_\text{t}(+)\) or \(F^2_\text{t}(-)\) is being considered. Good conductivity implies that \(\tau = k(1 \pm u^2/k_2^2)^{1/2}\). Since \(u\) in the interval \((0, k_0)\) in \(F^2_\text{t}\), we further have \(\tau = k\). In \(F^2_\text{t}\), if the exponential is sufficient to ensure that the integrand is essentially zero before the condition \(|u/k| \ll 1\) is no longer reasonable, then \(\tau = k\) is again a valid approximation. Even small values of \((z_0 - d)\) should suffice. Applying \(\tau = k\) to (7) yields
\[ F^1_\text{t}(z_0) = K_{Xr} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 - u^2) e^{iu(z_0 - d)}}{(k_0^2 + u^2) e^{u(z_0 - d)}} \, du \] and
\[ F^2_\text{t}(z_0) = K_{Xr} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 + u^2) e^{-u(z_0 - d)}}{(k_0^2 - u^2) e^{-u(z_0 - d)}} \, du. \] (10)

The electric and magnetic dipole cases will be considered next, beginning with the latter.

A. Magnetic Dipoles

Let \(X_r = \mu_r\). If the material is not highly magnetic, the condition \(|k| \gg |\mu_r u|\) will be valid over the significant integration ranges in (10). This allows the \(\mu^2 u^2\) terms to be neglected in (10) with the result that
\[ F^1_\text{t}(z_0) = K_{\mu_r} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 - u^2) e^{iu(z_0 - d)}}{1 + 2j\mu_r u/k} \, cot kd \] and
\[ F^2_\text{t}(z_0) = K_{\mu_r} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 + u^2) e^{-iu(z_0 - d)}}{1 - 2\mu_r u/k} \, cot kd. \] (11)

The denominators may be further simplified if \(|kd|\) is either very large or small. First if \(|kd| \gg 1\), then \(\cot kd \rightarrow 1\). Thus, because we have already assumed that \(|u/k|\) is small, we have
\[ F^1_\text{t}(z_0) = K_{\mu_r} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 - u^2) e^{iu(z_0 - d)}}{k \sin kd} \, du \] and
\[ F^2_\text{t}(z_0) = K_{\mu_r} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 + u^2) e^{-iu(z_0 - d)}}{k \sin kd} \, du. \] (12)

Recalling (8) and (12), we note that
\[ F^1_\text{t}(z_0) + F^2_\text{t}(z_0) = -\frac{2\mu_r}{k \sin kd} \frac{\partial F^0_\text{t}(z_0 - d)}{\partial (z_0 - d)}. \] (13)

Then, taking the appropriate derivatives of (9), we obtain
\[ F^1_\text{t}(z_0) = 2K_{\mu_r} \frac{e^{kd(z_0 - d)}}{k \sin kd (z_0 - d)} F_H \] where
\[ F_H = 3 - 3jk_0(z_0 - d) - k_0^2(z_0 - d)^2. \] (14)

Forming the ratio of the unloaded \(F^0_\text{t}\) to loaded \(F_\text{t}\) magnetic dipole coupling, we find that \(SE_H\) is given by
\[ SE_H = 20 \log \left| \frac{-\sin kd}{2\mu_r} (z_0 - d) - \frac{1 - jk_0 z_0 (z_0 - d)^3}{F_H z_0^3} \right| \frac{\eta_0}{2} \frac{\sin kd}{\eta} \frac{\eta_{\text{J}}}{\eta_{\text{J}}} \frac{(z_0 - d)^3}{z_0^3}. \] (15)

Alternatively, if \(|kd| \ll 1\) then \(\cot kd \rightarrow 1/\langle kd\rangle\). In the limit as \(kd \rightarrow 0\), we find that \(F_\text{t} \rightarrow F^0_\text{t}\). Thus,
\[ SE_H = 0, \quad \text{for } |kd| \ll 1. \] (16)

Bannister [8] suggests combining limiting values such as these to form a hybrid expression good for a wide range of \(kd\). Because \(\mu/\mu_r = \eta/\eta_{\text{J}}\), the result is
\[ SE_H = 20 \log \left| 1 - k_0 (z_0 - d) \right| \frac{\eta_0}{2} \frac{\sin kd}{\eta} \frac{\eta_{\text{J}}}{\eta_{\text{J}}} \frac{(z_0 - d)^3}{z_0^3}. \] (17)

In the very near field, such that \(k_0 z_0 \ll 1\), \(SE_H\) reduces to
\[ SE_H = 20 \log \left| 1 - \frac{1}{6} k_0 (z_0 - d) \right| \frac{\eta_0}{2} \frac{\sin kd}{\eta} \frac{(z_0 - d)^3}{z_0^3}. \] (18)

B. Electric Dipoles

Let \(X_r = \epsilon_r\). Factoring out \(k^2\) from the denominators in (10) we have
\[ F^1_\text{t}(z_0) = K_{\epsilon_r} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 - u^2) e^{iu(z_0 - d)}}{(1 + \epsilon_r u/k^2) + 2j\epsilon_r u/k} \, cot kd \] and
\[ F^2_\text{t}(z_0) = K_{\epsilon_r} \frac{k}{2} \int_0^\infty \frac{u(k_0^2 + u^2) e^{-iu(z_0 - d)}}{(1 - \epsilon_r^2 u/k^2) + 2j\epsilon_r u/k} \, cot kd. \] (11)
and
\[ F_{\varepsilon}(z_0) = \frac{K_\varepsilon}{k \sin kd} \int_0^\infty \frac{u(k_0^2 + u^2)e^{-u(z_0-d)}}{1 - \varepsilon^2(u/k)^2 - 2\varepsilon_i(u/k) \cot kd} du. \]

(19)

Our basic assumption of high conductivity implies that \(|u/k| \ll 1\) over the significant integration ranges. However, \(\varepsilon_i/u/k\) will not be small since \(\varepsilon_i/u/k = (u/k_0)(\varepsilon_i/\mu_0)^{1/2}\). Thus, \(\varepsilon_i/u/k\) will be large except near \(u = 0\) (where the numerators in (19) are small), implying that
\[ F_{\varepsilon}(z_0) = \frac{-Kk}{\varepsilon_i \sin kd} \int_0^{z_0} \frac{(u^2 + k_0^2)e^{i(z_0-d)}}{u + ja} du \]

and
\[ F_{\varepsilon}(z_0) = \frac{-Kk}{\varepsilon_i \sin kd} \int_0^{z_0} \frac{(u^2 + k_0^2)e^{-i(z_0-d)}}{u + a} du. \]

(20)

where
\[ a = 2(k/e_i) \cot kd. \]

Rewriting the rational functions of \(u\) in the integrands yields
\[ F_{\varepsilon}(z_0) = \frac{-Kk}{\varepsilon_i \sin kd} \int_0^{z_0} \frac{(u - ja) - (k_0^2 + a^2)}{u + ja} e^{i(z_0-d)} du \]

and
\[ F_{\varepsilon}(z_0) = \frac{-Kk}{\varepsilon_i \sin kd} \int_0^{z_0} \frac{(u - a) + (k_0^2 + a^2)}{u + a} e^{-i(z_0-d)} du. \]

(21)

These may be evaluated in terms of the exponential integral \(E_1\) [10, eqs. 5.141 and 5.142] with the result that
\[ F_{\varepsilon}(z_0) = \frac{-Kk}{\varepsilon_i \sin kd} \frac{e^{i\theta_0(z_0-d)}}{(z_0-d)^2} F_E \]

where
\[ F_E = 1 - (jk_0 + a)(z_0 - d) + (k_0^2 + a^2)(z_0 - d)^2 e^{-i(\theta_0 + \theta_0_0(z_0-d))} \]

\( \cdot E_i(-j(\theta_0 - \theta_0_0(z_0-d))). \)

(22)

For \(|kd| \gg 1, |\cot kd| = 1\). Therefore, \(|a/k_0| = 2u^{1/2}/(\varepsilon_i^{1/2})|k|\) \(\ll 1\). If \(|kd| \ll 1\), then \(\cot kd = 1/kd\) and \(|a/k_0| = 2/akd|n_0|\). If the shield is sufficiently thick to ensure that \(a/k_0 \gg 1\), then the condition \(|a/k_0| \ll 1\) will again hold. Only in the limit as \(d \rightarrow 0\) will \(|a/k_0|\) not be small; however, this is the trivial no-shield case. Thus, we may expand \(E_1\) [10, eq. 5.111] and discard terms of the order \(|a/k_0|^2\) and show that
\[ E_i(-j(k_0z_0 - d)) \left( 1 + j\frac{a}{k_0} \right) = E_i(-j(k_0z_0 - d)) \]

\[ -j\frac{a}{k_0} e^{i\theta_0(z_0-d)}. \]

(23)

Inserting (23) into (22) gives
\[ F_E = 1 + (jk_0 + a)(z_0 - d) - k_0^2(z_0 - d)^2 e^{-i(\theta_0 - \theta_0_0(z_0-d))} \]

\( \cdot E_i(-j(k_0z_0 - d)) - jk_0a(z_0 - d)^2 e^{i\theta_0(z_0-d)}. \)

(24)

If \(|a|\) is neglected altogether, \(F_E\) reduces to
\[ F_E = 1 - jk_0(z_0 - d) + k_0^2(z_0 - d)^2 E_i(-j(k_0z_0 - d)). \]

(25)

Combining these with the empty-case coupling result (9), and noting that \(\varepsilon_i/k = \eta_0/\mu_0\), gives
\[ \text{SEE} = 20 \log \left| \frac{1}{k_0} \frac{\eta_0}{\eta} \frac{\sin kd \sin k_0 z_0}{z_0^2} \right|. \]

(26)

Further, if \(k_0z_0 \ll 1\), then \(F_E = 1\) and \(\text{SEE}\) reduces to
\[ \text{SEE} = 20 \log \left| \frac{1}{k_0} \frac{\eta_0}{\eta} \frac{\sin kd \sin k_0 z_0}{z_0^2} \right|. \]

(27)

IV. COMPARISON BETWEEN APERTURE AND DIPOLE SE EXPRESSIONS

Collecting results, we find that for the electric shielding case
\[ \text{SEE} = 20 \log \left| \frac{8}{3\pi} \frac{1}{\eta_0} \frac{\sin kd}{k_0} \right| \]

(aperture)

and
\[ \text{SEE} = 20 \log \left| \frac{1}{k_0} \frac{\eta_0}{\eta} \frac{\sin kd \sin k_0 z_0}{z_0^2} \right|. \]

(dipoles)

(28)

and for the magnetic shielding case
\[ \text{SEH} = 20 \log \left| \frac{1 - jk_0(z_0 - d)}{1 - jk_0(z_0 - d)} \right| \]

\( \frac{\eta_0}{\eta} \frac{\sin kd}{2} \)

\( \frac{z_0^2}{F_H} \)

\( \text{(dipoles)} \)

(29)

As the frequency goes to zero \((k_0 \rightarrow 0), \text{SEE} \rightarrow \infty\) and \(\text{SEH} \rightarrow 0\). These are the static limits which affirm that at very low frequencies a good conductor is an excellent electric field shield but a poor magnetic field shield. Also, the dipole expressions agree with similar results developed by France- schetti and Papas [9, eqs. 4.4 and 4.8] if \(z_0 - d = z_0, |a|\) is neglected, and the proper notation differences are accounted for.

These results give the desired equivalence between aperture SE measurements (as in a DTC) and near-field coaxial dipoles. Forming the difference \(\Delta\) between the two (aperture minus...
Figure 2. Measured DTC data and calculated dipole SE curves for an aluminum-on-Mylar sample.

\[ \Delta_E = 20 \log \left( \frac{8 \, z_0}{3\pi \, r} \frac{F_E}{(1 - jk_0 z_0)(z_0 - d)^2} \right) \]

and

\[ \Delta_H = 20 \log \left( \frac{8 \, r}{3\pi \, z_0} \frac{F_H}{(1 - jk_0 z_0)(z_0 - d)^3} \right) \]

where in \( \Delta_H \) the zero-frequency limit (1) has been ignored. For thin materials, \( z_0 \approx z_0 - d \), at frequencies such that \( k_z z_0 \ll 1(1 \rightarrow 1, 1 \rightarrow 3) \), these reduce to

\[ \Delta_E = 20 \log \left( \frac{8 \, z_0}{3\pi \, r} \right) \]

and

\[ \Delta_H = 20 \log \left( \frac{8 \, 3r}{3\pi \, z_0} \right) \]

This result suggests that the aperture radius and the dipole separation \( z_0 \) play similar roles in determining the electric field SE, while in the magnetic field case, the equivalent dipole separation is three times the aperture radius.

As an example of the application of the above equations, we will consider a thin layer of aluminum (on Mylar) with a measured (in a coaxial transmission-line holder) \( \alpha d \) value of 0.08 at frequencies below 1 GHz [1]. If \( \sigma = 3.72 \times 10^7 \text{ S/m} \) is assumed for aluminum, then a thickness \( d = 2.15 \times 10^{-9} \text{ m} \) follows. The aperture radius for the particular DTC used is \( r = 3.0 \text{ cm} \). Setting \( \Delta_E \) and \( \Delta_H \) equal to 0 in (31) requires that the coaxial electric dipoles be separated by 3.5 cm and that the coaxial magnetic dipoles be separated by 7.6 cm. These parameters may be used to compute \( \text{SE}_E \) and \( \text{SE}_H \) for the equivalent dipole configurations. No actual dipole measurements were attempted for the reasons given in the introduction. The results are shown in Fig. 2 along with the measured DTC data and the far-field (normally incident plane wave) level based on plane-wave transmission [9]:

\[ \text{SE}_{\text{far field}} = 20 \log \left( \frac{\eta_0 \sin kd}{\eta \, 2j} \right) \]

The calculated curves are a numerical integration of (7) for each case and the approximations given by (18) and (27) based on \( k_0 z_0 \ll 1 \). These match well with the DTC data for \( \text{SE}_H \) except around 760 MHz, where the DTC has a resonance associated with the appearance of the first higher order mode. The basic assumption in the magnetic dipole case that \( k_2 \ll \mu_0 \) is well satisfied; thus, good agreement is expected. Agreement is weaker for \( \text{SE}_E \). Above 300 MHz, curves (I) and (2) begin to separate and neither is in good agreement with the DTC data. The basic approximation in the electric dipole case that \( k_2 \ll \mu_0 \) is not well satisfied as frequency increases. The DTC data suffer from contact resistance between the MUT and the aperture plate. Contact resistance tends to degrade the measured \( \text{SE} \) data to below the predicted levels. Also, the \( \text{SE}_H \) curves tend toward the far-field SE level as measured in a coaxial transmission line [11]. Returning to (17), we see that for \( k_0 z_0 \gg 1 \)

\[ \text{SE}_H = 20 \log \left( \frac{\eta_0 \sin kd}{\eta \, 2j} \right) \]

which is just the expected result for a plane wave normally incident on an infinite sheet, as given by (32). In the electric dipole case (see (26)), we find that \( F_E \rightarrow 1 \) and

\[ \text{SE}_E = 20 \log \left( \frac{\eta_0 \sin kd}{\eta 

\[ \rightarrow \]
which is slightly above (6 dB) the predicted far-field level. The approximate expression (curve 2) tends toward this result; however, the full integral expression converges slowly.

V. CONCLUSION

Expressions have been developed which relate near-field simulation SE measurements based on small-aperture coupling, such as made in the dual TEM cell, and near-field coaxial dipole coupling SE results. Aperture coupling techniques may not intuitively appear to be near-field-type SE measurements. Thus, the expressions developed here help to relate the two approaches.

ACKNOWLEDGMENT

The author would like to thank Dr. M. T. Ma for his support of this work, and J. W. Adams and E. J. Vanzura for their valuable measurements and comments.

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