Techniques for Measuring the Electromagnetic Shielding Effectiveness of Materials: Part II—Near-Field Source Simulation

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Abstract—This paper continues to discuss the topic of measurements of electromagnetic shielding effectiveness of materials by simulating a near-field source. Two specific measurement approaches, the use of a dual TEM cell and the application of an apertured TEM cell in a reverberating chamber, are studied. In each case we also consider the system frequency range, test sample requirements, test field type, dynamic range, measurement time required, and analytical background, and present data taken on a common set of materials.

Key Words—dual TEM cell, high-impedance wave, low-impedance wave, near-field source simulation, reverberating chamber, shielding effectiveness, shielding materials, small aperture coupling.

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I. INTRODUCTION

THE SHIELDING effectiveness (SE) measurement techniques discussed in Part I of this two-part paper simulate incident plane waves and thus are far-field approaches. If a material is brought into the near field of a source, its shielding characteristics may be quite different. Part II of this paper addresses this near-field SE measurement problem. Near-field sources arise frequently when shielding against unwanted emissions is a concern, such as the FCC requirements for computing devices. Both high-impedance (electric field dominant) and low-impedance (magnetic field dominant) waves need to be considered. The basic requirement is again to measure the insertion loss (IL) as in Part I ([11]). Definitions of the material parameters \( \eta, \eta_0, k, k_0, d, \) etc.) are also the same.

Ideally, near-field SE could be measured by placing the test material (preferably an infinite sheet) between closely spaced antennas, such as dipoles. If a finite sample must be used, there will necessarily be signal components arriving at the receiving antenna that do not go through the material. These signals that arrive by indirect paths may be eliminated by placing the transmitting antenna in a shielded box but at the expense of perturbing the desired field distribution. Anechoic material may be used to suppress reflections from the shielded box. Unfortunately, anechoic material is expensive and cumbersome for this application, and has frequency dependence. These considerations provide the motivation for making SE measurement based on waveguides rather than radiating antennas.

The dual TEM cell (DTC), considered in Section II, allows us to simultaneously measure the electric and magnetic polarizabilities of a small aperture covered with the test material. These in turn represent a measure of the high-impedance and low-impedance shielding capabilities of the material. The results of aperture polarizability can be shown to be equivalent to those obtained by using coaxial dipoles to measure SE of an infinite sheet.

The DTC fixture essentially uses one cell to drive another through a common aperture. The cells need to be fairly large to allow the internal mounting of the sample and proper joining of the cells. As a result the upper frequency of DTC usage is limited by the appearance of resonances associated with higher-order modes. An alternative approach is to use a small, apertured TEM cell as the receiver to be driven by a high-frequency source. This is the idea behind an apertured TEM cell in a reverberating chamber, as discussed in Section III. The reverberating chamber allows us to generate a high-frequency, statistically known field while the TEM-cell receiver again allows us to consider separately the shielding of electric- and magnetic-field components. The analysis of this system is similar to that of the DTC. In fact, half of the DTC is used as the apertured TEM cell in the present study. Thus, the high-frequency limit is the same and the results here primarily establish the feasibility of such an approach.

Section IV gives some brief information on other methods that have been considered such as the transfer-impedance technique and the dual-box method.

II. DUAL TEM CELL

A. Test Configuration

A TEM cell is a section of expanded 50-\( \Omega \) rectangular coaxial transmission line (RCTL) tapered at each end to match ordinary 50-\( \Omega \) coaxial line [1]. It is very similar to the coaxial holders discussed in Part I except that the cross section is now rectangular rather than circular. Clearly, the TEM cell could be used in a similar fashion [2]; however, no real advantage results. Instead, let us consider two cells that are coupled via an aperture in a shared wall. This forms the dual TEM cell (DTC) concept [3]-[5] depicted in Fig. 1. The aperture transfers power from the driving cell (shown as the upper cell fed at Port 1) to the receiving cell (shown as the lower cell). The DTC is unique in that it has two output signals (Ports 2
Fig. 1. The dual TEM cell.

Fig. 2. The cross section of a rectangular coaxial transmission line.

B. Analytical Background

The DTC has been analyzed previously [5]. We need only quote the necessary results. Referring to the \( x - y \) axes shown in Fig. 2, we will assume that the aperture is centrally located \( (x = 0) \) and parallel to the inner conductor. In the lower cell the excitation coefficients \( a^{' \pm} \) for the TEM mode, propagating away from the aperture in the forward \((+)\) and backward \((-)\) directions, are given by [5, eq. 12]

\[
a^{' \pm} = j \frac{k_0}{2\eta_0} \left[ \alpha_{ep} \pm \alpha_{mx} \right] E_{0y}^2(0, b) \tag{1}
\]

where \((x, y) = (0, b)\) is the aperture location in the lower cell (as implied in Fig. 2), \( \alpha_{ep} \) and \( \alpha_{mx} \) are respectively the electric and magnetic polarizabilities applicable to TEM mode excitation, and \( E_{0y} \) is the \( y \)-component of the TEM-mode electric field. For a small gap TEM cell \((g/a \ll 1)\)

\[
E_{0y}(x, y) = \text{sign}(y) \frac{2}{a} Z_0^{1/2} \sum_{m=1,3,5,\ldots} \frac{\cosh M(b-y \text{sign}(y))}{\cosh Mb} \cos Mx \sin MaJ_0(Mg) \tag{2}
\]
and

\[ Z_e/\eta_0 = \frac{\pi}{8} \left\{ \ln \left( \frac{8a}{\pi g} \right) - \sum_{m=1,3,5,\ldots}^{\infty} \frac{2(1 - \coth Mb)}{m} \right\}^{-1} \]

where the coordinates are as in Fig. 2, the summations are over the odd integers, \( M = m\pi/2a \), \( J_0 \) is a zero-order Bessel function, and \( Z_0 \) is the characteristic impedance of the cell (typically 50 Ω).

As mentioned, the ± between terms in (1) implies asymmetric coupling to the two receiving ports. If we form the sum (\( \Sigma \)) and difference (\( \Delta \)) of the two outputs in (1) we find that

\[ a(\Sigma) = a^+ + a^- = j \frac{k_0}{\eta_0} \alpha_{eg} E_{eg}^2(0, b) \]

and

\[ a(\Delta) = a^+ - a^- = j \frac{k_0}{\eta_0} \alpha_{mg} E_{mg}^2(0, b) \] (3)

In practice the two output signals from a TEM cell are mixed in a hybrid junction to produce \( a(\Sigma) \) and \( a(\Delta) \). The cables connecting the TEM cell outputs to the hybrid junction should have equal phase and attenuation to insure accurate results. Equation (3) indicates that the sum signal depends on the normal electric-field coupling while the difference signal depends on the tangential magnetic-field coupling. This approach is equivalent to methods for determining the emission level from an unknown radiator placed in a TEM cell [6] or the unknown polarizabilities of gaskets and joints placed in a DTC [3]. Insertion-loss data based on the sum and difference signals will yield

\[ IL(\Sigma) = 20 \log \left| \frac{\alpha_{eg}}{\alpha_{eg}} \right| = IL_E \]

and

\[ IL(\Delta) = 20 \log \left| \frac{\alpha_{mg}}{\alpha_{mg}} \right| = IL_H \] (4)

where it is assumed that only the polarizabilities are changed by the material presence.

The unloaded (no test material) aperture polarizabilities have been studied for a variety of shapes (cf. [7], [8]). The loaded (material covered) aperture problem is more difficult and to our knowledge only circular apertures have been considered. Let \( r \) denote the radius of a circular aperture. The ratios in (4) for electrically small apertures may be approximately given by [9, 10]

\[ IL_E = 20 \log \left| \frac{8}{3\pi} \frac{\eta_0 \alpha_{eg}}{k_0 r} Z_e \left( 1 + 8\pi \frac{Z_e}{Z_i} \right)^{-1} \right| \]

and

\[ IL_H = 20 \log \left| 1 + j \frac{4}{3\pi} \frac{\eta_0}{k_0 r} \frac{Z_e}{Z_i} \left( 1 + 2\pi \frac{Z_e}{Z_i} \right)^{-1} \right| \] (5)

where \( Z_e \) is the equivalent sheet impedance of the material, and \( Z_i \) is the contact impedance between the material and the aperture conductor. Contact impedance tends to degrade the measured SE as discussed in Part I. It is also typically unknown. For a well designed system \( Z_e \) should be small and may be dropped from (5). For a conductive sheet [11]

\[ Z_e = -j\eta_0/\sin kd \] (6)

where \( k \) is the wave number in the material. Aperture polarizabilities are not overly sensitive to shape. Thus, (5) may be applied to noncircular, convex apertures (such as the square aperture of our DTC) by choosing an equivalent radius that gives the same polarizability value in the unloaded case. For example, for a square aperture of side \( s \), \( \alpha_{eg} = 0.258s^3 \) while \( \alpha_{mg} = 4r^3/3 \) for a circular aperture [12]. Thus, the conversion is \( r = 0.579s \). For \( \alpha_{eg} \), the respective polarizabilities are \(-0.114s^3 \) and \(-2r^3/3 \), and the conversion is \( r = 0.555s \).

The loaded aperture expressions (5) may be related to the dipole near-field SE reasessment depicted in Fig. 3. For the latter case, both coaxial electric and magnetic dipoles have been considered [13] based on low-frequency approximations. The result is

\[ IL_E = 20 \log \left| \frac{1}{k_0 Z_0} \frac{\eta_0}{\eta} \frac{\sin kd}{z_0^2} \frac{1 - jk_0 (z_0 - d)^2}{F_E} \frac{z_0^2}{z_0^3} \right| \]

and

\[ IL_H = 20 \log \left| 1 - k_0 (z_0 - d) \frac{\eta_0}{\eta} \frac{\sin kd}{2} \frac{1 - jk_0 (z_0 - d)^3}{F_H} \frac{z_0^3}{z_0^3} \right| \] (7)

where

\[ F_E = 1 - (jk_0 + a)(z_0 - d) + (k_0^2 + a^2)(z_0 - d)^2 \]

\[ \cdot e^{-jk_0 (z_0 - d)a}E_1[-(jk_0 - a)(z_0 - d)] \]

\[ F_H = 3 - 3jk_0 (z_0 - d) - k_0^2 (z_0 - d)^2 \] (8)

where \( a = 2(k_0/c) \cot kd \), and \( E_1 \) is the exponential integral ([14, eq. 5.141-5.142]). The relationships in (5) and (7) give the desired approximate (low-frequency zero contact impedance) equivalence between SE measurements by aperture (as in the DTC) and by near-field coaxial dipoles. Forming the difference \( \Delta \) between (5) and (7) and neglecting \( Z_e \) yield

\[ \Delta_E = 20 \log \left| \frac{8}{3\pi} \frac{\eta_0}{k_0 r} \frac{Z_e}{Z_i} \left( 1 + 8\pi \frac{Z_e}{Z_i} \right)^{-1} \right| \]

and

\[ \Delta_H = 20 \log \left| \frac{8}{3\pi} \frac{r}{z_0} \frac{1 - jk_0 (z_0 - d)^3}{F_H} \frac{z_0^3}{z_0^3} \right| \] (9)

In \( \Delta_H \) we have also neglected the zero-frequency limit of 1 which is equivalent to only considering good shields. For thin
Fig. 3. Coaxial dipoles used to measure insertion losses of a planar shield.

Fig. 4. Insertion loss of the gold-Mylar sample measured in the dual TEM cell.

Fig. 5. Insertion loss of the aluminum-Mylar sample measured in the dual TEM cell.

Fig. 6. Insertion loss of the plastic-aluminum-plastic sample measured in the dual TEM cell.

materials \((z_0 - d = z_0)\) at frequencies such that \(k_0 z_0 < 1\) \((F_E \rightarrow 1, F_H \rightarrow 3)\), (9) reduces to

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

and

\[
\Delta_H = 20 \log \left| \frac{3r}{3\pi z_0} \right|.
\]

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. Equation (7) also allows near-field shielding to be predicted if the material parameters \(k\) and \(d\) are known or measured. For electrically thin \([\sin kd \approx kd\) in (7)], highly conductive \((\sigma/\omega\varepsilon_0 \gg 1)\) materials only the product \(ad\) need be known to evaluate (7). This is an important simplification since the thickness \(d\) is often not well known for very thin depositions on an insulating substrate. The product \(ad\) can often be determined by measurements using the flanged coaxial holder discussed in Part I of this paper.

C. Data and Discussion

The characteristics of empty DTC coupling and single-port IL measurements on the plastic-aluminum-plastic sample have been reported previously [5], and data support the above equations. Figs. 4–6 show the IL data for our present samples, beginning with the gold-Mylar sample (Fig. 4). The curves plotted are the sum power \((I_{LE})\) and the difference power \((I_{LH})\), both measured and theoretical (5) with \(ad = 0.144\) determined by (3) of Part I using the flanged coaxial holder, and the far-field source \(I_L\) of approximately 29 decibels from Part I. In general \(I_{LE} > I_L\) (far field) \(> I_{LH}\) as expected for a good conductor. Also, the measured \(I_{LH}\) agrees very well with the theoretical curve. The sharp spike that appears in both of the measured curves at around 760 MHz is due to a resonance in the TEM cell. Clearly, the \(I_{LE}\) data are more variable than
the $I_{H}$ data. This is a combination of lesser dynamic range, contact impedance, and more sensitivity to discontinuities in the transmission lines. It is hoped that a better system will evolve and that the $I_{E}$ data may be improved.

Figs. 5 and 6 show the data for the aluminum-Mylar and plastic-aluminum-plastic samples. Because the materials are similar to the gold-Mylar sample the same comments apply. Attempts to measure the graphite composite sample failed because the present dynamic range of the system was exceeded.

III. Apertured TEM Cell in a Reverberating Chamber

A. Test Configuration

An alternative to using a TEM cell to drive a second TEM cell by means of aperture coupling (as is the case with the DTC) is to employ a reverberating chamber as the field source. In essence, the advantages of the DTC are retained, namely, the system is RF tight, and the electric- and magnetic-field couplings can be considered separately. The major difference is that we now have a statistically well-behaved field driving the sensing cell. A reverberating chamber begins as a simple shielded room with one significant modification. A paddle is added that may be turned continuously (mode stir) or in small discrete steps (mode tune). The paddle presence allows us to take advantage of a basic shielded room drawback, namely, multimoding. In any given paddle position we simply have a shielded room with an added boundary condition. As the paddle positioning is varied, the field generated inside the test zone of the chamber yields a wave impedance whose average approaches the free-space value $\eta_{0}$ [15]. Thus, in a statistical sense the reverberating chamber may be used to generate an equivalent plane-wave incident field. Proper average-field convergence depends on exciting a large number of modes in the chamber. Thus, this technique, at least from the source point of view, works best at higher frequencies. The present NBS chamber is 2.74 x 3.05 x 4.57 m and is used above 200 MHz into the gigahertz range. The TEM cell used as the receiver in the present study is half the DTC discussed in Section II with the same square aperture. Thus, the upper frequency limit for the present system will again be approximately 1 GHz. A smaller cell is necessary to take advantage of the higher frequency potential. The statistical field distribution is a good test since real-world shields will no doubt encounter a range of incident field types. Unwanted scattering near the sample, due to mounting hardware for example, is not overly important since the reverberating chamber strives for a complicated boundary condition environment anyway. The dynamic range, 90--100 dB, is quite good. The chamber is a high-Q cavity, thus high-field levels can be generated using only moderate input power. The primary difficulties are that the chamber will not give meaningful results if too few modes are present (in this case below 200 MHz), and that the technique can be very time consuming. At frequencies below 1 GHz, where the chamber must be operated in the mode-tune procedure, each frequency tested requires measurements at 200 paddle positions. Although the operation at NBS has been automated and computer-controlled, a typical curve presented here still takes up to 4 h to generate. At higher frequencies (above 1 GHz) where the mode-stir technique is applicable and less input-output power normalization is required, the measurement time is considerably less. Nonetheless, this method is very data intensive and is likely to remain so.

B. Analytical Background

The analysis of an apertured TEM cell in a reverberating chamber is similar to that for the DTC except that the source field is distinctly different [16]-[17]. Another consideration is the lack of symmetry about the aperture. Polarizabilities are usually derived for the case of an aperture located in an infinite ground plane subject to a static impressed field. However, as is pointed out by Collin [18], in the case of coupling between dissimilar regions the geometry can significantly affect the dipole moments. Clearly, the reverberating chamber and a coaxial line will not be symmetric about the aperture; thus Collin’s correction may be important. However, calculations show that in this case the corrections are negligible [17, Section VIII].

As with the DTC, the basic advantage to the apertured TEM cell probe in the reverberating chamber is that we can monitor power at both ends, take sum and difference from the outputs with a hybrid junction, and thereby separate the electric- and magnetic-field couplings. Any apertured waveguide would allow us to apply this basic approach; however, coaxial lines are convenient since they couple to standard connectors and 50-Ω lines, and the rectangular geometry allows us to mount flat panel samples to the exterior of an apertured TEM cell.

Typically, two types of power measurement are performed in the reverberating chamber: peak output and average output. However, regardless of whether we monitor peak or average power, the insertion loss should behave according to (4) if the peak and average driving fields are the same in each case (loaded and unloaded) [17]. Actual measurements indicate that IL is nearly independent of whether the peak or average is monitored. We will typically consider data based on averages. Peak power data may have an advantage where dynamic range is a problem.

C. Data and Discussion

The IL data obtained by the apertured TEM cell in the reverberating chamber are presented in Figs. 7-10 along with theoretical curves based on (5) and an approximate far-field source level based on flanged coaxial-holder measurements from Part I. In Fig. 7 the agreement between the measured and the theoretically predicted $I_{H}$ for the gold-Mylar sample is not good. The sample was poorly mounted; this led to a large contact impedance. Fig. 8 shows the equivalent aluminum-Mylar data. Here the results are similar to the DTC case. Again, $I_{H}$ is well behaved but $I_{E}$ is both variable and below the theoretically expected levels. The plastic-aluminum-plastic sample yields the results shown in Fig. 9. In all of the above curves the TEM cell begins to show resonances above approximately 750 MHz although the $I_{E}$ data seem to be more sensitive to their appearance than do the $I_{H}$ data, as was the case in the DTC. The multimoding in the reverberating chamber is weak below 200 MHz. Thus our present data have
Before considering other samples let us compare the apertured TEM cell data to that from the DTC. Because the gold-Mylar sample was mounted improperly, there is not much agreement (Figs. 4 and 7). The DTC is symmetric about the sample so the side orientation is not important. The IL of DTC is much higher. The aluminum-Mylar data (Figs. 5 and 8) are almost identical, particularly the $I_L$ curves. The $I_E$ data fluctuate but their basic shape is the same. This is also the case for the plastic-aluminum-plastic sample (Figs. 6 and 9).

Fig. 10 shows the graphite composite IL data. In general $I_L > I_E$, but not to the extent of the previous materials. This material shows significantly higher shielding levels especially at the lower frequencies leading one to suspect that the composite has some magnetic reflecting capability. The sample is also significantly thicker than are the others so we may be seeing an appreciable amount of absorption. The IL range found with this technique, 60-80 dB, is similar to that found in the flanged coaxial holder (Part I, Fig. 12). Thus, the graphite composite material appears to be a very effective shield against both near and far fields.

Finally, we tested a material that consists of a copper-ferrite-copper layering. We expect this material to have very good electric and magnetic shielding properties. Fig. 11 shows the IL data and we definitely see here $I_L$ exceeding $I_E$, although only slightly. This material proved to be an excellent shield and in fact tended to exceed the dynamic range of the system. Thus, instead of using averages, peak readings were used to determine the IL since these are more likely to be above the noise floor.

Separate measurements have shown that peak and average measurements tend to closely track each other at a difference of about 7–8 dB [19]. Thus, an insertion-loss measurement should be independent of whichever is used since two signals are subtracted. Fig. 12 shows the sum and difference insertion losses based on both the average- and peak-power measurements for the plastic-aluminum-plastic sample. As may be seen they track quite well.

The basic characteristics of the methods considered in
Sections II and III are summarized in Table I for easy reference and comparison purposes.

IV. OTHER METHODS

The SE test methods studied in the preceding sections by no means form an exhaustive list. Other methods often mentioned include measuring transfer impedances [20]-[22] and the dual box fixture [23], [24]. In the context of the present study these additional techniques are briefly discussed.

A. Surface Transfer Impedance Measurements

A fixture for measuring surface transfer impedance is pictured in Fig. 13. The system was originally designed to evaluate the SE of shielded cables and gaskets but may also be

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**TABLE I**

**SUMMARY OF TWO SE TEST METHODS USING A NEAR-FIELD SOURCE SIMULATION (LBE: LIMITED BY EQUIPMENT)**

<table>
<thead>
<tr>
<th>SE Test Method</th>
<th>Frequency</th>
<th>Field Type Simulated</th>
<th>Sample Material Requirement</th>
<th>Theoretical Support For Data</th>
<th>Dynamic Range [dB]</th>
<th>Repeatability</th>
<th>Time Required For Meas. (Hrs)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dual TEM Cell</td>
<td>1 MHz (LBE)</td>
<td>1 GHz Near-Field or Glazing Plane Wave</td>
<td>Cover An Aperture</td>
<td>Good</td>
<td>50-60</td>
<td>Good</td>
<td>0.5</td>
</tr>
<tr>
<td>Apertured TEM Cell In Rever. Chamber</td>
<td>200 MHz</td>
<td>1 GHz Near-Field</td>
<td>Cover An Aperture</td>
<td>Good</td>
<td>90-100</td>
<td>Good</td>
<td>3-4</td>
</tr>
</tbody>
</table>
The surface transfer impedance may be measured by comparing the voltage \( V \) (as shown in Fig. 13) across the test sample to the driving current \( I \), and indeed is equivalent to the sheet impedance given in (6). The idea is similar to that of the coaxial holders discussed in Part I. In fact, the circuit diagram for the transfer impedance fixture is the same as given in Part I, Fig. 5. A difficulty is that we again have contact impedances in series with the load impedance. The fixture does allow for adequate contact to be formed between the sample and the fixture. Air pressure is used to aid in creating contact. The reported frequency range is from dc to 700 MHz whereupon fixture resonances begin to occur. The dynamic range approaches 100 dB [20]. It should be possible to automate this system to shorten the measurement time.

Surface transfer impedance data, determined by either measurement or theory, can be used to predict SE performance. Birkin et al. [25] discuss approximations relating \( Z_s \) to SE for a set of basic geometries, both for high-\((Z_H)\) and low-impedance \((Z_L)\) fields. They find that

\[
Z_s = E_t / J.
\]

(11)

The surface transfer impedance is defined as the ratio of the tangential electric field \( E_t \) induced on the interior of the shield to the surface current density \( J \) excited on the exterior of the shield by external sources [25].

\[
Z_s = E_t / J.
\]

(11)

The transfer-impedance measurement system used to investigate the SE of a washer shaped sample. The surface transfer impedance \( Z_s \) is defined as the ratio of the tangential electric field \( E_t \) induced on the interior of the shield to the surface current density \( J \) excited on the exterior of the shield by external sources [25].

Fig. 13. The transfer-impedance measurement system.

Surface transfer impedance data, determined by either measurement or theory, can be used to predict SE performance. Birkin et al. [25] discuss approximations relating \( Z_s \) to SE for a set of basic geometries, both for high-\((Z_H)\) and low-impedance \((Z_L)\) fields. They find that

\[
Z_s = E_t / J.
\]

(11)

which are more complicated, are given as

\[
Z_H = (V/S)\omega \mu_0 \]

and

\[
Z_L = \omega \mu_0 (r_1 + r_2)/6
\]

for flat plate excited by identical coaxial loops (14) where in the latter expression \( r_1 \) and \( r_2 \) are the distances from the loops to the flat sheet, and it is assumed that \( (r_1 + r_2) \gg \) the loop radius and that \( (r_1 + r_2)d \gg 0.03 \), where \( d \) is the shield thickness.

\[
IL_H = 20 \log |Z_H/Z_s| = 20 \log k_0 z_0 \frac{\eta_0 \sin k d}{\eta}
\]

(15)

which is the same as the low-frequency limit \((k_0 z_0 \ll 1)\) of (7) for a good conductor and loops not too close \((z_0 - d \approx z_0)\). There is no ready equivalence for the electric-dipole case.

B. The Dual Box

The dual box has also been proposed to perform high-impedance and low-impedance IL measurements. The dual box consists of two small metal chambers coupled through a rectangular aperture as shown in Fig. 14. Each chamber contains an antenna, either a dipole for high-impedance data or a small loop for low-impedance data. In essence, however, the dual box is simply a scaled version of the shielded room and thus suffers some of the same problems known to the shielded room. By keeping the chambers small, the users can avoid multimoding, but specifying the field distribution exciting the sample remains difficult. No theoretical justification is available. It may be possible, through very rigid standardization of the technique, to obtain repeatable data, but interpreting results and relating them to other methods is still formidable.

IV. CONCLUSION

Theoretical and measured insertion losses for a few sample materials have been presented as a function of frequency using
two different methods of simulating the near-field source. The method of using a dual TEM cell is very good in principle. Its practical value is limited by the presence of contact resistance between the fixture and sample material, and the upper-frequency bound. The method of reverberating chamber with an apertured TEM cell yields measured results that must be interpreted in the average sense. It is most suitable for a higher frequency range. Both methods are capable to separate the shielding effectiveness of materials due to electric-field coupling from that due to magnetic-field coupling. Also, useful frequency ranges for both techniques can be adjusted by modifying the TEM cell size.

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