Monitor for Integrity of Seams in a Shield Enclosure

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Abstract—This paper considers a new concept for measuring the performance of shield enclosures. It allows for the presence of aperture-like faults at unknown positions over the shield enclosure. The aperture penetration is measured in a manner appropriate to bounding the response of internal equipment. The technique leads to the location of the faults as well.

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I. INTRODUCTION

ONE DESIGNS practical shield enclosures for the protection of electronic equipment from various electromagnetic environments. One would also like to be able to measure the shield performance in such a way that the measured parameters were directly relatable to the electromagnetic coupling to the electronic equipment, and could be used to assure the satisfactory operation of that equipment. Furthermore, it would be helpful if this way to monitor the shield performance could be used regularly during the lifetime of the shield to assure its continued adequate performance.

For most purposes a sheet of metal is sufficiently conducting that diffusion of electromagnetic fields through it is not a significant problem [1]. Apertures (such as faulty seams) are in general more significant and will be considered here. In many situations conductor penetrations are the most important; other techniques apply to this case [2], [3].

We begin with a review of the relevant features of aperture penetration, especially for long thin apertures (slots), then the technique in [2], [3] for bounding aperture penetration is generalized to allow for multiple apertures at unknown locations over a shield enclosure. Both the existence and location of the apertures can be determined.

II. APERTURES IN SHIELDS

Generally for practical shield enclosures aperture penetration is dominant over diffusion penetration for coupling to interior equipment, especially when the equipment is located near such an aperture. One can model the penetrant fields by equivalent electric and magnetic dipoles. In this case the near fields (but not too near the aperture itself in units of aperture dimensions) are dominated by $r^{-3}$ terms [1]. The dipole moments are related to the appropriate fields on the shorted (closed) aperture by a polarizability proportional to the cube of the linear dimensions, so small apertures usually do very little while large apertures can be significant.

Better than considering the electromagnetic fields per se penetrating apertures, it is more directly related to the interference problem to consider the resulting source terms for the internal equipment. As discussed in [2], [3] an appropriate canonical problem to consider is that of a wire passing by an aperture. As in Fig. 1 there are two equivalent sources, a series (longitudinal) voltage source $V_s$ related to the short-circuit magnetic field $H_{sc}$, and a parallel (transverse) current source $I_s$ related to the short-circuit displacement $D_{sc}$ (or electric field). In the near field of the aperture (but not too near the aperture) these terms are proportional to the time derivative of the associated short-circuit fields. In spatial terms the dependence is $r^{-2}$ where $r$ is the distance from the aperture to the wire. This can be seen by integrating an $r^{-2}$ field over a distance comparable to $r$, or by direct appeal to canonical problems [1].

These sources are important in that they are part of the equations used to bound the propagation of signals through subshields as part of the general topological decomposition of complex systems as a product (or convolution in time domain) of matrices in the good-shielding approximation [2]-[5]. In the context of apertures, since the source terms depend on the distance of a conductor from the aperture, it is useful to define an exclusion volume as indicated in Fig. 1 [2], [3]. One can define this exclusion volume somewhat arbitrarily (including zero if conductors are allowed to come right up to the aperture plane), but its purpose is to provide a basis for defining the maximum values that the source terms may achieve for any allowed position of conductors near the aperture (on the transmission side). In terms of normalized source coefficients these are [1]

$$\tilde{V}_s = \frac{V_s}{aZ_0} = \text{voltage source coefficient}$$

$$\tilde{I}_s = \frac{I_s}{aE_{sc}} = \text{current source coefficient}$$

$$Z_0 = \sqrt{\frac{\mu_0}{\varepsilon_0}} = \text{characteristic impedance of free space}$$

$$a = \text{Laplace transform (time} \rightarrow \text{complex frequency)}$$

$$\varepsilon_0 = \text{some characteristic dimension of the system of concern.}$$

(1)

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Here $\mathbf{H}_{sc}$ is taken in a direction to maximize $|\mathbf{n}|$ and $\mathbf{E}_{sc}$ is, of course, the component normal to the surface. For wire-to-wire coupling through the aperture the coefficients are similarly defined [2], [3].

For present purposes let us consider the aperture of concern to be a thin slot, such as is appropriate for a fault in a seam of a shield enclosure. Such a slot might be modeled as an ellipse, for which the case of the short-circuit magnetic field parallel to the long dimension of the slot is most important. In this case, let the wire be transverse to the slot for maximum coupling. This is a key point for diagnosing the presence of such a fault in a shield; the sensing wire should run across the slot, not be parallel to it. If the slot is not resistively loaded, then for frequencies below the slot half-wave resonance the dominant coupling to the wire is proportional to the time derivative of the short-circuit magnetic field parallel to the slot (or the short-circuit surface current density transverse to the slot).

An important aspect of the development in [2], [3] is the cross-association of electric and magnetic parameters. This is reflected in (1) or in terms of wire-to-wire coupling through apertures in the form [2]

$$V_{\text{outside}} \rightarrow I_{s}\vert_{\text{inside}}$$

$$I_{\text{outside}} \rightarrow V_{s}\vert_{\text{inside}}.$$  

This cross-association of electric and magnetic parameters seems fundamental to understanding shielding. One way to look at this is that the BLT equation [6] uses wave variables that are linear combinations of voltage and current. This allows the separation of waves on one side of topological boundaries from those on the other side via a scattering matrix. Inherent in this is the possibility of electric parameters on one side producing magnetic parameters on the other side, and conversely. Another way to look at this is through the concepts of transfer impedance and transfer admittance [1], although the latter parameter may sometimes be better cast in terms of charge than voltage as the exciting term on the first side of the boundary.

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As illustrated in Fig. 2 let us start the external transmission line at \( A \) and wrap it around the shield at a uniform spacing \( a \) from the shield, returning to a position \( A' \) near \( A \). At \( A \), drive the wire with respect to the shield with a voltage \( V_{\text{in}} \), which might in general be CW or a pulse of some specified waveform. Let the other end of the wire \( A' \) be terminated to the shield in a constant resistance \( R_{\text{in}} \) equal to the characteristic impedance of this transmission line. In this case we have established a wave propagating in one direction (from \( A \) to \( A' \)) around the shield exterior. The velocity of the wave is either the speed of light \( c \) if all the dielectrics are air, or somewhat less if other insulating dielectrics are used.

As the wave propagates around the shield there is a surface magnetic field, or equivalently a surface current density that extends to both sides of the wire a distance on the order of \( a \). The actual distribution of this surface magnetic field (with apertures closed (shorted)) is readily calculated as in [10]. So a strip of width say \( 2a \) or somewhat more on the shield is effectively illuminated.

Now turn our attention to the inside of the shield. Let us begin a similar wire at \( B \) (near \( A \), except on the inside) and continue around the inside paralleling the external wire and ending at \( B' \) (near \( A' \), except on the inside). This interior wire is spaced a distance \( b \) from the shield walls and has its own characteristic impedance \( R_{\text{in}} \) with respect to the shield walls as a transmission line. The propagation velocity is also \( c \) or somewhat less depending on the use of insulating dielectrics.

Now perform a gedankenexperiment. Let a fast-rising pulse, say approximately a step function, be introduced on the external transmission line at \( a \). It propagates along the external transmission line until it reaches some fault in the shield at some position, say \( X \). At the corresponding position \( X' \) on the internal transmission line a voltage source \( V_s \) is induced sending a signal of voltage \( \pm V_s/2 \) propagating in both directions away from the fault with opposite polarities for the two waves. Assuming that the fault is a thin slot perpendicular to the two wires the coupling is maximized, and it is dominantly the external current (magnetic field) producing the internal voltage (electric field).

The two internal waves propagate away from \( X' \) until they...
reach B and B'. Let us monitor (measure) these internal waves at these positions, say by a transient recorder with two inputs, both terminating in the characteristic impedance of the internal transmission line. We now have available to us an important piece of information. Knowing the relative times of arrival of the signals at B and B' and the propagation speed of the internal transmission line we can readily locate X' (and X), the position of the fault. If the fault is an unloaded slot there is negligible resistance across it. The internal signal is basically the time derivative of the external signal (an external step function becoming ideally an internal delta function), making time-of-arrival measurements somewhat precise.

Going a step further, note that so far the technique covers four faces of a rectangular parallelepiped, but with current in only one of two possible directions on each face. So imagine three such pairs of transmission lines, each pair wrapping around in a manner orthogonal to the other two pairs. Then each face is tested in two orthogonal directions. Each pair tests four faces in one direction; three pairs cover six faces in two directions (a kind of octahedral symmetry).

But why stop here? The external transmission line from A to A' paired with the internal transmission line from B to B' covers a strip or order 2a or 2b around the shield. Suppose now that the external wire is extended from A' to A'' as a sort of helix around the shield with spacing from the wire to the previous (and subsequent) "windings" of order 2a, or perhaps more, so that coupling from one "turn" to the next is acceptably small. Continue this procedure until the appropriate shield walls (4 for a rectangular parallelepiped as in Fig. 2) are covered by this transmission line. Then terminate this transmission line at A'' in its characteristic impedance.

Considering now the internal transmission line let it be similarly extended as a quasi helix, "paralleling" the external transmission line, to B'' where it is now terminated. Comparing the signals at B and B'' the previous concept of locating the fault is still applicable. It remains to consider details of the spacing of the "turns" from each other and from the shield to optimize the technique. For example, it may be desirable to have the internal and external wires not directly opposite to each other across the shield boundary. To optimize the uniformity of coupling from external to internal transmission lines their positions might be interspersed; the position of the internal transmission line might be spaced halfway between the positions of two adjacent turns of the external transmission line.

As the title of this subsection indicates, perhaps this concept needs a name. In this case we have shield closure using transitometer ultra monitor or SCUTUM (the Latin word for shield) for short. Note the concept of transitometer (related to the transmission of a signal through a fault), similar to the concept of a reflectometer.

C. Reduction of Mutual Coupling at Crossing of Transmission Line

When one of the transmission-line wires crosses another on the same side of a face of the shield enclosure there can be significant mutual coupling. Since the wires cross at right angles no magnetic flux from one wire links the other wire leading to no induced longitudinal voltage source. However, electric field from the charge on one wire does induce charge on the other wire, leading to an induced transverse current source.

In order to reduce this electric coupling at crossings the technique in Fig. 3 can be used. Here one wire is at least partly shielded from the other at the crossing. The wire closer to the shield at the crossing goes through some sort of tunnel (or even coaxial structure if desired) in which the wire dimensions and/or spacings are adjusted to preserve the character-
istic impedance of the transmission line as it passes through the crossing. The other wire passes over this tunnel with perhaps additional conductors present, and the transmission-line impedance is also preserved through the junction in this case.

Note that this isolation structure is made of good conductors that are well bonded to the shield. Such a crossing should be made to occur at some position on a shield panel that is not too close to the seams where the shield panels are joined to each other.

IV. POSITIONING OF TRANSMISSION LINES

Consider now a pair of exterior and interior transmission lines (windings) as illustrated in Fig. 4. As previously discussed, the wires are spaced distances \( a \) and \( b \) from the shield wall. For simplicity now let

\[
a = b \quad \Rightarrow \quad \text{shield wall thickness},
\]

(4)

Note the coordinate system with \( y = 0 \) as the shield wall and the wires (transmission lines) parallel to the \( z \)-axis. Now let

\[
x = n\Delta, \quad n = 0, \pm 1, \pm 2, \ldots
\]

(5)

be the \( x \)-coordinates of the wires with

\[
y = \pm a \quad (+ \text{for } n \text{ even}, - \text{for } n \text{ odd})
\]

(6)

as the \( y \)-coordinates.

Now imagine that there is a current \( I \) in one of the wires \((n = 0)\) and consider the fields and potentials. This has been worked out in [10]. The normalized field on the shield wall (electric or magnetic) is (with value 1 at \( x = 0 \))

\[
f(x) = \left[ 1 + \left( \frac{x}{a} \right)^2 \right]^{-1}
\]

(7)

Since we wish to minimize coupling to an adjacent wire on the same side, then we need

\[
f_0 = f\left( \frac{2\Delta}{a} \right) = \left[ 1 + \left( \frac{2\Delta}{a} \right)^2 \right]^{-1} \ll 1.
\]

(8)

In terms of potential one can obtain a more accurate estimate of the potential on one wire due to a unit potential on the first [10], but the simple estimate in (8) gives the basic idea.

Now consider that there is some localized fault in the shield wall at \( x = x_1 \) and let

\[
\nu = \frac{x_1}{\Delta}.
\]

(9)

The normalized field exciting this fault from the \( n = 0 \) wire is

\[
f_1 = \left[ 1 + \left( \frac{\Delta\nu}{a} \right)^2 \right]^{-1}.
\]

(10)

The fields penetrating the fault in turn interact with the wire at \( x = \Delta \) on the other side of the shield wall. Using reciprocity considerations one can compute the relative coupling to this wire by computing the fields at \( x = x_1 \) from a current in the wire at \( x = \Delta \) giving a factor

\[
f_2 = \left[ 1 + \left( \frac{\Delta - x_1}{a} \right)^2 \right]^{-1} = \left[ 1 + \left( \frac{\Delta(1 - \nu)}{a} \right)^2 \right]^{-1}.
\]

(11)

The coupling between the \( n = 0 \) wire and the \( n = 1 \) wire is then proportional to

\[
g = f_1f_2 = \left[ 1 + \left( \frac{\Delta\nu}{a} \right)^2 \right]^{-1} \left[ 1 + \left( \frac{\Delta(1 - \nu)}{a} \right)^2 \right]^{-1}.
\]

(12)

Defining

\[
\delta = \frac{\Delta}{a}, \quad \eta = \nu - \frac{1}{2}
\]

then (12) assumes the symmetrical form

\[
g = \left[ 1 + \delta^2\left( \eta + \frac{1}{2} \right)^2 \right]^{-1} \left[ 1 + \delta^2\left( \eta - \frac{1}{2} \right)^2 \right]^{-1}
\]

\[
= \left[ 1 + 2\delta^2\left( \eta^2 + \frac{1}{4} \right) + \delta^4\left( \eta^2 - \frac{1}{4} \right)^2 \right]^{-1}.
\]

(14)

Considering that the fault may lie somewhere between one of the outside wires and one of the inside wires we take our range of interest as

\[
0 \leq x_1 \leq \Delta
\]

\[
0 \leq \nu \leq 1
\]

\[
-\frac{1}{2} \leq \eta \leq \frac{1}{2}.
\]

(15)

Differentiating \( g \) with respect to \( \eta \) we find extrema of \( g \) at

\[
\eta_1 = 0
\]

\[
\eta_2 = \pm \left[ \frac{1}{4} - \delta^2 \right]^{1/2}, \quad \text{if } \delta > 2 \text{ for } \eta \text{ real}
\]

(16)
at which
\begin{align*}
g_1 &= \left[ 1 + \left( \frac{\sigma}{2} \right)^2 \right]^{\frac{1}{2}} \\
g_2 &= \frac{\sigma}{2}, \quad \text{if } \sigma > 2.
\end{align*}
(17)

Now restricting \( \eta \) as in (15) we see that the \( \eta_2 \) solution applies in our range of interest provided \( \sigma > 2 \). Next, also take the extrema of our range of interest as
\begin{equation}
\eta_3 = \pm \frac{1}{2},
\end{equation}
(18)
for which
\begin{equation}
g_3 = \left[ 1 + \frac{\sigma^2}{2} \right]^{-1}.
\end{equation}
(19)

Then \( g_1, g_2 \) (for \( \sigma > 2 \), and \( g_3 \) represent the range of variation of \( g \) over \( \eta \) in the range of interest. The closeness of these numbers represents the uniformity of being able to detect a fault with respect to position on the shield wall.

As an example suppose we wished the coupling between wires on one side of the shield wall to be about 1 percent, giving
\begin{align*}
f_0 &= 0.01 \\
\delta &= \frac{\Delta}{a} \approx 5.
\end{align*}
(20)

Our corresponding uniformity numbers are
\begin{align*}
g_1 &= 0.019 \\
g_2 &= 0.040 \\
g_3 &= 0.0385.
\end{align*}
(21)

This represents a factor of 2 variation in the detection of the signal through the fault with respect to the location of the fault on the shield wall. This illustrates the trade-off to be made in the design.

V. SUMMARY

This gives the general outline of the SCUTUM technique for characterizing and monitoring shield enclosures. It can be operated in CW or pulse mode, but pulse mode may be more convenient for fault location.

Our discussion has centered on aperture-like faults because of their practical importance. However, diffusion through metal walls is also picked up by this technique, which can be considered a generalization of the technique in [8].

The present discussion has considered locating a single aperture (say at \( X, X' \)). The technique, however, is more general and can locate multiple apertures by considering successive pulses that arrive at \( B \), the first pulse that arrives being from the first aperture, etc. If the propagation speeds on the outer and inner transmission lines are (nearly) the same, the pulses from all the apertures arrive at \( B' \) at (nearly) the same time. The time difference of pulse arrival between \( B \) and \( B' \) can then be used to locate all apertures (if the number is not too large) excited by the transmission-line pair.

There are still various detailed problems to be considered. For example, seams at the edges of the faces of the shield enclosure have different geometries for electromagnetic coupling from the case considered here. There are also the details of the construction of the transmission lines and spacing them from other external and internal structures.

REFERENCES

[1] K. S. H. Lee, Ed., EMP Interaction: Principles, Techniques, and Reference Data. New York: Hemisphere, 1986.
[2] F. C. Yang and C. E. Baum, “Use of matrix norms of interaction supermatrix block for specifying electromagnetic performance of subshields,” Interaction Note 427, Apr. 1983.
[3] ——, “Electromagnetic topology: Measurement and norms of scattering parameters of subshields, Electromagnetics, vol. 6, no. 1, pp. 47-59, 1986.
[4] C. E. Baum, “Electromagnetic topology: A formal approach to the analysis and design of complex electronic systems,” Interaction Note 400, Sept. 1980; and also in Proc. 4th Symp. Tech. Exhibition, Electromagn. Compatib. (Zurich), pp. 209-214, Mar. 1981.
[5] ——, “On the use of electromagnetic topology for the decomposition of scattering matrices for complex physical structures,” Interaction Note 454, July 1985.
[6] C. E. Baum, T. K. Liu, and F. M. Tesche, “On the analysis of general multiconductor transmission-line networks,” Interaction Note 350, Nov. 1978, and completely contained in: C. E. Baum, “Electromagnetic topology for the analysis and design of complex electromagnetic systems,” in J. E. Thompson and L. E. Leusen, Eds., Fast Electrical and Optical Measurements. Dordrecht: Martinus Nijhoff, 1986, pp. 467-547.
[7] J. E. Bridges, “Proposed recommended practices for the measurement of shielding effectiveness of high-performance shielding enclosures,” IEEE Trans. Electromagn. Comput., pp. 82-94, Mar. 1968.
[8] P. A. A. Sevat, “Methods for calculating the shielding effect of solid-shell enclosures against EMP,” Interaction Note 457, May 1981.
[9] H. Price, G. D. Sower, and E. F. Vance, “Standard EMP test techniques for cables, connectors, and containers,” NEM Rec., p. 16, May 1982.
[10] C. E. Baum, “Impedances and field distributions for symmetrical two-wire and four-wire transmission-line simulators,” Sensor and Simulation Note 27, Oct. 1966.