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Nuclear Electromagnetic Pulse Simulation by Point Source Injection Techniques for Shielded and Unshielded Penetrations

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ar Electromagnetic Pulse Simulation by Point Source Injection Techniques for Shielded and Unshielded Penetrations—by Robert F. Gray

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Nuclear electromagnetic pulse simulation by point source injection techniques were investigated for shielded and unshielded system penetrations. Circuit theory is applied to the unshielded case and a generalized injection system is suggested. To determine the relationship between point source current injection on the external shield of a cable and its response to free-field EMP excitation, general transmission line solutions were obtained in the frequency domain for each case. A numerical transform was used to obtain time domain waveforms of the external and internal currents. Both solutions were verified with experimental data collected on a braided shield coaxial cable that exhibited both types of reactive coupling (inductive and capacitive) through its shield. Shielding effectiveness transfer functions were experimentally obtained with a coaxial cable driver to determine the shield parameters. The two types of excitation were compared as a function of length and shield type indicating when single point source excitation became inadequate. Experimental results show the potential of a multiple source injection system for free-field simulation.

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1. INTRODUCTION

The overall objective of the program reported on here is to develop current injection technology for threat and lower-level system electromagnetic pulse (EMP) assessment. In general, a current injection system creates a transient current on a system penetration by means of a point source or sources coupled to the penetration either directly (resistive) or reactively (capacitance or inductive). This type of simulation is useful whenever other simulation techniques are inadequate or unrealizable from the standpoint of either peak amplitude or area of illumination.

One condition that must exist for this type of simulation to be valid is that in the event of an actual EMP, distinct points of entry excite sensitive circuits with induced signals that are due predominantly to electromagnetic responses of external conductors; consequently, it is assumed that interior electromagnetic fields within a system/subsystem enclosure do not contribute significantly to circuit upset or damage. If this assumption is not valid, it is still possible to obtain useful information from the direct-drive technique, but the data reduction becomes far more complex. Therefore, the discussion here is limited to when exterior coupling phenomena dominate the system response. Also, to accurately or adequately drive a system penetration, a greater understanding of the free-field coupling mechanisms is required than is necessary for free-field simulation. That is, before a system penetration can be driven, a description of the EMP coupled waveform and its distribution along the penetration must be obtained from either low-level free-field testing or analytical predictions.

In addition to these caveats regarding proper use of current injection, it is necessary also to demonstrate a relationship between the threat response of a system and that due to a current injection technique. To establish such a relationship, the system penetrations are divided into two cases, shielded and unshielded. A general approach for each of these cases is outlined, and then, because of the complexity of the problem, the shielded case is analyzed in more detail.

2. APPROACH

2.1 Unshielded Case

Typical examples of unshielded penetrations are incoming or outgoing ac power lines, deliberate antennas, and unshielded control or communication cables that are exterior to the system structure. In figure 1, the unshielded system penetration has an impedance Z_{in} looking into the system from points A and B and an impedance Z_{out} looking away from the system. If the impedances Z_{in} and Z_{out} are reasonably well behaved with respect to some reference conductor, G, and there is no significant coupling to the penetration further into the system, then the transient current, i(t), may be directly injected on the penetration using either a Thévenin's or Norton's equivalent source as shown in figure 2. For most unshielded cases, the Norton's equivalent should be

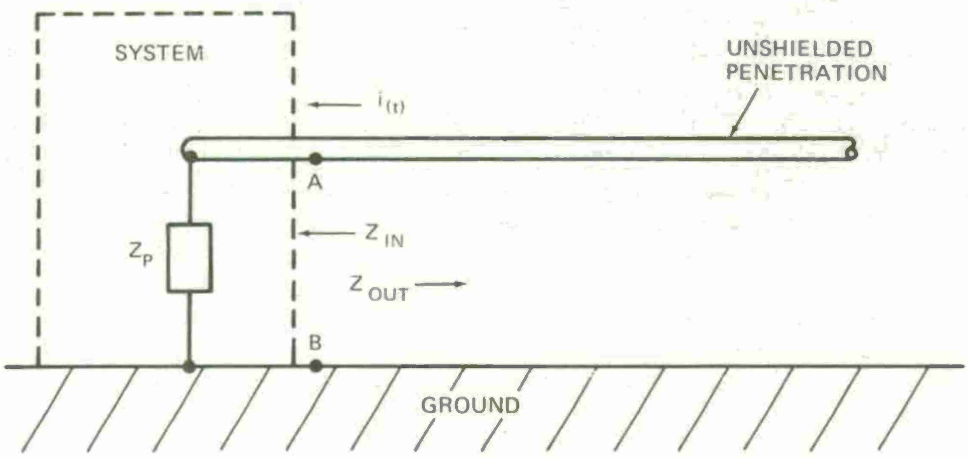


Figure 1. A system with an unshielded penetration.

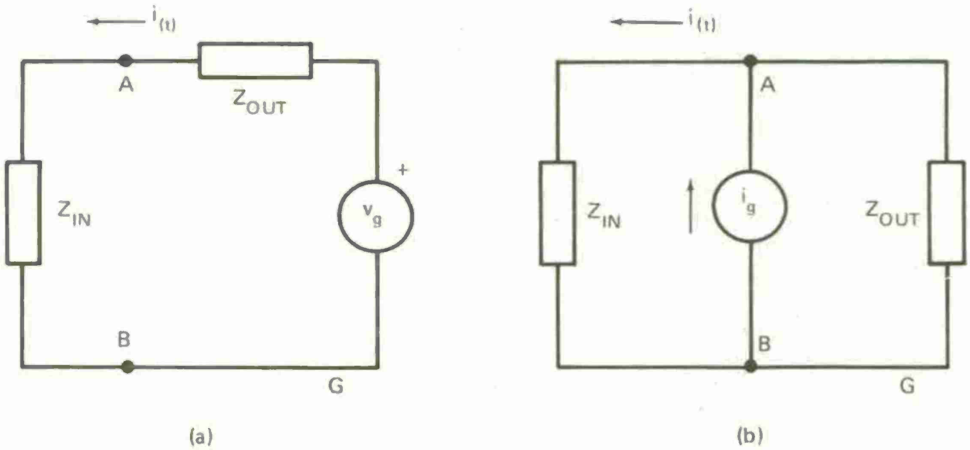


Figure 2. A Thévenin's (a) and Norton's (b) equivalent injection system.

used, because most transient high-voltage sources are an open circuit, except when discharging, and by definition, the Thévenin's voltage source, V_g, must be a short circuit when shut down. This open-circuit condition before and after discharge would unload the system at the penetration and possibly damage it. If, however, the load on the system is not critical, then a Thévenin's equivalent may be useful, because it offers a higher efficiency over a Norton equivalent source. In order not to load the system improperly, the current source, i_g, must have an internal impedance much greater than the parallel impedance of Z_{in} and Z_{out} for all frequencies of interest, so a larger source is required than for the Thévenin system. However, this inefficiency allows greater freedom in adjusting the source waveform, since a shaping impedance, Z_s, may be placed across the source without affecting the system because of the isolation impedance, Z_I, as shown in figure 3. The current source, i_g, of figure 2 is simulated in figure 3 by the voltage source, v_g, and the coupling impedances, Z_I and Z_s. Also, the use of a Norton source eliminates the need for physically altering the system, since it is necessary only to connect the source between points A and B.

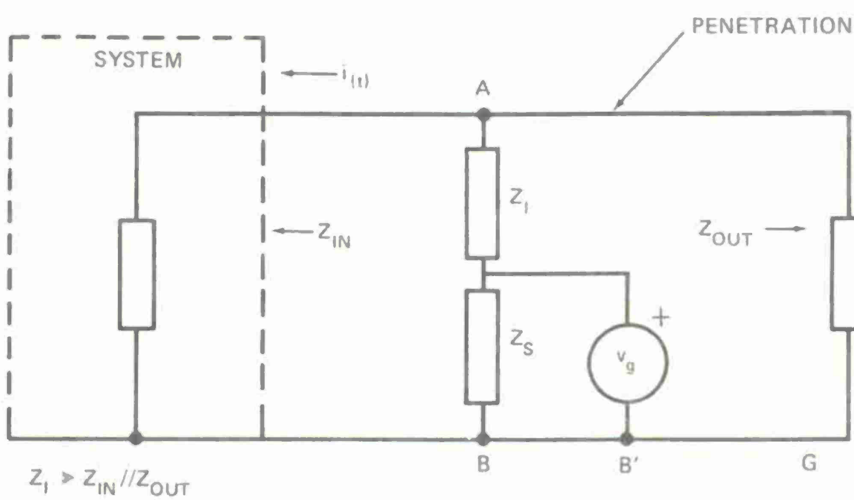


Figure 3. A current injection system for unshielded penetrations.

If no reference conductor is readily associated with the penetration, then it may still be possible to inject current on the penetration in one of two ways. As one way, a reference conductor could be added from the desired drive point to some ground point further into the system. However, care must be exercised in doing so, because it is possible to alter the system characteristics drastically (at least for the penetration of interest). Another, more preferable, technique involves inductively coupling the current with a ferrite core or similar transformer device, as shown in figure 4. The major advantage of this type of drive is that it is not physically connected to the system. This loose coupling, however, also makes the driver inefficient, and it is difficult to regulate the injected current waveshape.

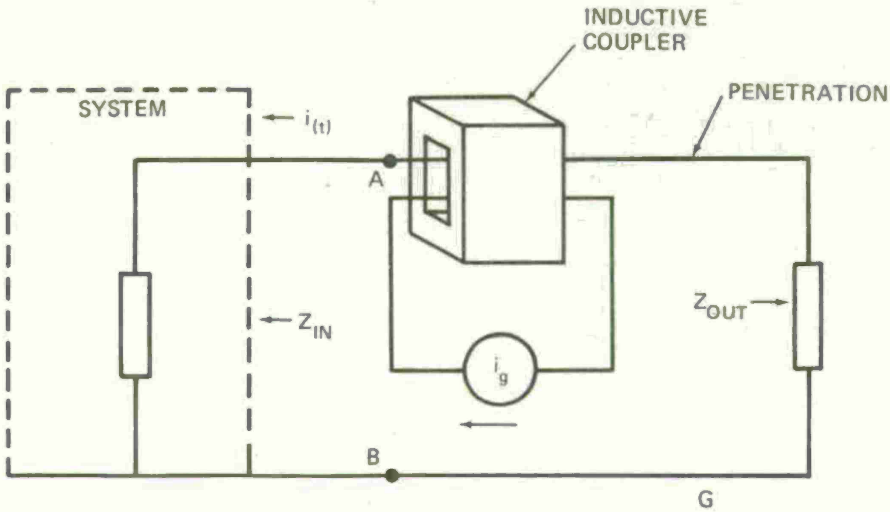


Figure 4. Inductively coupled injection system.

2.2 Shielded Case

The most common type of shielded penetration is a multi-conductor cable with an external metallic shield or a standard coaxial cable. This type of penetration is analyzed here in the absence of terminal system/subsystem enclosures. Because of the physical size of most terminal enclosures, electromagnetic scattering by these obstacles and the resulting contributions to responses of interconnecting cables cannot be included in the transmission line approach used here. Furthermore, when the penetration to be driven is a shielded cable, it is usually necessary to inject current on the external shield, rather than the internal conductors, because either the internal response is not known or the number of internal conductors is too large to consider driving them. Threat-induced signals in terminations of internal conductors can be adequately simulated via point-source excitation of the external shield, but doing so is complex, because coupled internal signals depend on the distribution and propagation of the external driving current. Therefore, the objective of this study is to describe induced internal signals due to a distributed source and the corresponding signals due to a point source. Correlation of such signals shows the adequacy of a direct-drive technique in simulating free-field induced responses for at least this basic case and gives insight to its applicability to more complex systems.

Transmission line theory is used to define the currents and voltages of interest in both cases. This approach has some limitations in the free-field case, but an exact solution for the external current and voltage would probably not alter the results significantly.

Transmission line theory is briefly outlined below. Then a circuit parameter model of the external cable shield is presented, along with a technique for measuring the parameters. The shielded model and transmission-line theory is then used to obtain general solutions for the two cases of interest.

A transmission line has length ℓ with a series impedance $Z = R + j\omega L$ and shunt admittance $Y = G + j\omega C$ per unit length such that its characteristic impedance $K = (Z/Y)^{1/2} = [(R + j\omega L)/(G + j\omega C)]^{1/2}$ and its propagation constant $\Gamma = (ZY)^{1/2} = [(R + j\omega L)(G + j\omega C)]^{1/2} = \alpha + j\beta$. If this line is terminated in an impedance Z_1 at $X = 0$ and an impedance Z_2 at $X = \ell$ and the line is excited at a point $X = \xi$ by a series generator of zero impedance (fig. 5), then from Schelkunoff,[1] the current, I_1, and the voltage, V_1, at any point along the line is

$$
\begin{aligned}
I_1(x,\xi) &= [K \cosh \Gamma x + Z_1 \sinh \Gamma x] \times \\
&\quad [K \cosh \Gamma(\ell - \xi) + Z_2 \sinh \Gamma(\ell - \xi)]/D, \quad x < \xi \\
&= [K \cosh \Gamma\xi + Z_1 \sinh \Gamma\xi] \times \\
&\quad [K \cosh \Gamma(\ell - x) + Z_2 \sinh \Gamma(\ell - x)]/D, \quad x > \xi \\
V_1(x,\xi) &= -K[K \sinh \Gamma x + Z_1 \cosh \Gamma x] \times \\
&\quad [K \cosh \Gamma(\ell - \xi) + Z_2 \sinh \Gamma(\ell - \xi)]/D, \quad x < \xi \\
&= K[K \cosh \Gamma\xi + Z_1 \sinh \Gamma\xi] \times \\
&\quad [K \sinh \Gamma(\ell - x) + Z_2 \cosh \Gamma(\ell - x)]/D, \quad x > \xi
\end{aligned}
\tag{1}
$$

[1] *S. A. Schelkunoff, Electromagnetic Waves, D. Van Nostrand Co., New York (1943).*

where

$$D = K\left[(K^2 + Z_1 Z_2)\sinh \Gamma\ell + K(Z_2 + Z_1)\cosh \Gamma\ell\right].$$

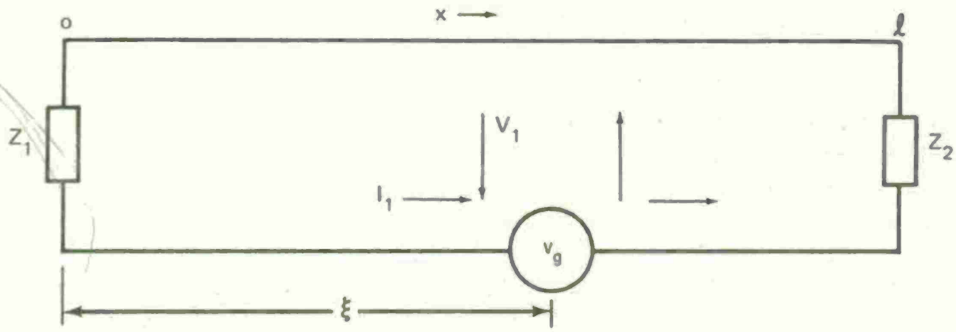


Figure 5. A transmission line excited by a series generator of zero impedance.

Similarly, if the line is excited by a shunt generator of infinite impedance, as in figure 6, the current, I_2, and the voltage, V_2, at any point along the line is given by

$$\begin{aligned} I_2(x,\xi) &= K[K\cosh \Gamma x + Z_1 \sinh \Gamma x] \times \\ &\quad [K \sinh \Gamma(\ell - \xi) + Z_2 \cosh \Gamma(\ell - \xi)]/D, \; x < \xi \\ &= -K[K \sinh \Gamma\xi + Z_1 \cosh \Gamma\xi] \times \\ &\quad [K \cosh \Gamma(\ell - x) + Z_2 \sinh \Gamma(\ell - x)]/D, \; x > \xi \\ V_2(x,\xi) &= -K^2[K \sinh \Gamma x + Z_1 \cosh \Gamma x] \times \\ &\quad [K \sinh \Gamma(\ell - \xi) + Z_2 \cosh \Gamma(\ell - \xi)]/D, \; x < \xi \\ &= -K^2[K \sinh \Gamma\xi + Z_1 \cosh \Gamma\xi] \times \\ &\quad [K \sinh \Gamma(\ell - x) + Z_2 \cosh \Gamma(\ell - x)]/D, \; x > \xi. \end{aligned} \tag{2}$$

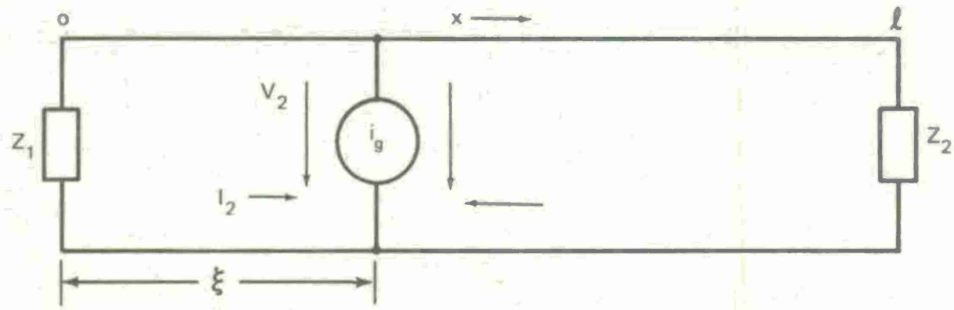


Figure 6. A transmission line excited by a shunt generator of infinite impedance.

These solutions for the discrete case can be used to determine the response due to a distributed excitation by integrating over the source region, as shown in equation (3).

$$I(x) = \int_{x_2}^{x_1} E(\xi) I_1(x,\xi)\, d\xi + \int_{x_2}^{x_1} J(\xi) I_2(x,\xi)\, d\xi$$

$$V(x) = \int_{x_2}^{x_1} E(\xi) V_1(x,\xi)\, d\xi + \int_{x_2}^{x_1} J(\xi) V_2(x,\xi)\, d\xi \quad (3)$$

where $E(\xi)$ is a series electromotive force per unit length and $J(\xi)$ is a shunt current per unit length distributed over the interval x_1, x_2. These integrals are used with the following models to define the responses for each case.

2.3 Cable Shield Model

Consider a conductor of radius r_a with a single coaxial shield of thickness t and inside radius r_b, as shown in figure 7.

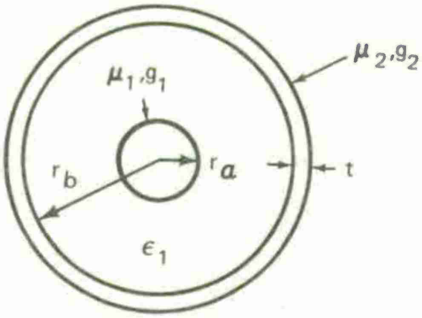


Figure 7. A conductor with a coaxial shield of thickness t.

The transfer impedance of a solid shield is given by Shelkunoff[2] to be $R_s = \eta/2\pi\left[r_b\left(r_b + t\right)\right]^{\frac{1}{2}} \sinh \sigma t$, where $\eta = j\omega\mu_2/\sigma$, $\sigma = (j\omega\mu_2 g_2)^{\frac{1}{2}}$, $\omega = 2\pi f$, f is the frequency, and g_2 and μ_2 are the conductivity and permeability of the shield, respectively.

A braided shield has two additional transfer parameters that have to be combined with the transfer impedance, R_s, given above. The longitudinal transfer impedance, Z_T, for a general cable is

$$Z_T = R_s + j\omega L_s$$

where L_s is the inductance/meter length of the shield. The lateral transfer admittance, Y_L, is

$$Y_L = j\omega C_H$$

where C_H is the capacitance/meter length between the inner conductor and an external reference conductor other than the shield. The assumed lossless dielectrics involved (inside and outside the shield) account for the purely reactive lateral transfer admittance. The parameters needed to calculate the solid shield transfer impedance, R_s, are usually

[2] *S. A. Schelkunoff, The Electromagnetic Theory of Coaxial Transmission Lines and Cylindrical Shields, Bell System Technical J. 13 (October 1934).*

well defined, but it is not possible to accurately calculate the inductive, L_s, and capacitive, C_H, terms needed for braided shields. Therefore, it is necessary to determine these two parameters experimentally.

A current, i_s, on the outside of the shield creates a distributed series electromotive force,

$$E(\xi) = Z_T i_s(\xi) = \left(R_s + j\omega L_s\right) i_s(\xi),$$

which excites the inner conductor according to equations (1) to (3). Similarly, any external voltage, $V_s(\xi)$, appearing between the cable shield and some reference conductor behaves as a shunt current source, $J(\xi)$, with the relationship

$$J(\xi) = Y_L V_s(\xi) = j\omega C_H V_s(\xi) \; .$$

The external voltages and currents need to be defined. Since the cable driver technique uses a point source at some point along the cable, equations (1) and (2) give the external voltage and current directly for any type of source. The coaxial cable driver to be considered has a point voltage source, V_s, at one end of the line $(x = 0)$. Therefore, the current, $i_s(x)$, and voltage, $v_s(x)$, along the line are

$$i_s(x) = V_s K_1 [K_1 \cosh \Gamma_1(\ell - x) + Z_2 \sinh \Gamma_1(\ell - x)] / D_1$$

$$v_s(x) = V_s K_1^2 [K_1 \sinh \Gamma_1(\ell - x) + Z_2 \cosh \Gamma_1(\ell - x)] / D_1 \qquad (4)$$

where the subscript is added to K, Γ, and D to differentiate between the external (1) and internal (2) lines.

For a distributed excitation, the integrals in equation (3) must be used to define the external voltage and current along the line. Equivalent series voltage and shunt current sources are defined by use of the incident electric and magnetic fields. Solutions for the internal current in both types of excitation may then be obtained by use of the external voltages and currents and the source models defined above.

3. COAXIAL CABLE DRIVER

3.1 Transmission Line Solution for Internal Current

The coaxial cable driver may be modeled with two separate transmission lines (fig. 8). Line ABDC is made up of the outside shell of the driver and the external cable shield. Line ABDC has a characteristic impedance of K_1 and a propagation constant of

$$\Gamma_1 = \alpha_1 + j\beta_1.$$

Line BDFE is made up of the external shield and the inner conductors and has a characteristic impedance of K_2 and a propagation constant of

$$\Gamma_2 = \alpha_2 + j\beta_2.$$

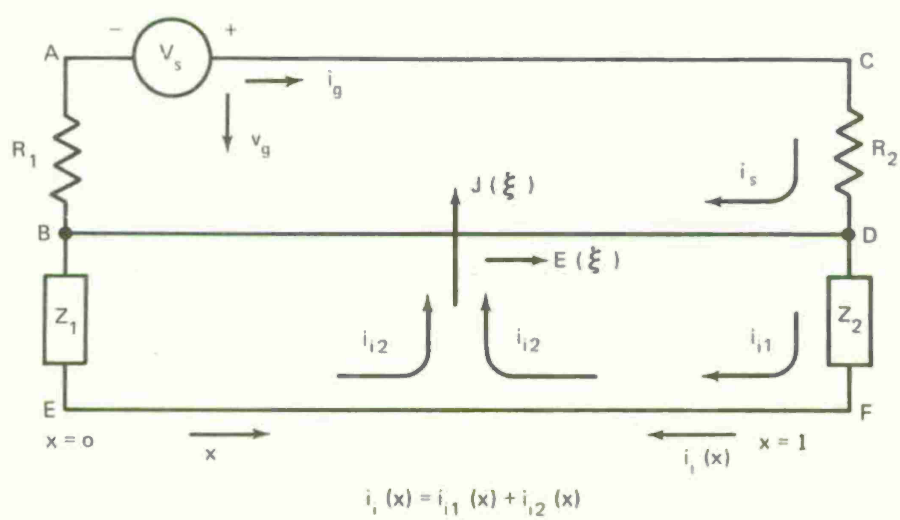


Figure 8. Circuit model of a coaxial cable driver.

V_s is the impressed source voltage, and i_g and v_g are the external transmission line current and voltage caused by V_s. $E(\xi)$ and $J(\xi)$ are distributed series electromagnetic field and shunt current sources causing internal currents i_{i1} and i_{i2}, respectively. The distributed electromagnetic field, $E(\xi)$, is produced by the sheath current, i_s, flowing on the shield, which has a transfer resistance, R_s (Ω/M), and braid inductance, L_s (h/M), as defined earlier. Therefore,

$$E(\xi) = -i_s(\xi)\left(R_s + j\omega L_s\right) = -Z_T i_s(\xi) = -Z_T\left(-i_g + i_i\right). \qquad (5)$$

But since the internal coupling is small,

$$i_s(\xi) \approx -i_g(\xi),$$

$$\therefore E(\xi) \approx Z_T i_g(\xi) .$$

The distributed shunt current source, $J(\xi)$, represents the coupling of the propagated driver voltage, v_g, through the holes in the braided shield as suggested by Frankel,[3] Vance,[4] and others. The coupling through the holes in the braid has the characteristics of capacitance C_H (per unit length) between the inner conductors and the external driver shell:

$$\therefore J(\xi) = -j\omega C_H v_g(\xi). \qquad (6)$$

---

[3] *S. Frankel, Terminal Response of Braided-Shield Cables to External Monochromatic Electromagnetic Fields, Harry Diamond Laboratories TR-1602 (August 1972).*

[4] *E. F. Vance, Comparison of Electric and Magnetic Coupling through Braided-Wire Shields, TM 18, Air Force Weapons Laboratory, Kirtland Air Force Base, Albuquerque, NM (February 1972).*

This is an approximation, because the shunt generator is really a function of the difference between the propagated driver voltage, v_g, and the internal cable voltage, v_i, but the coupling is small, so that

$$v_g - v_i \approx v_g .$$

$$i_s(\xi) \approx -i_g(\xi) = -\frac{V_s\left[K_1 \cosh \Gamma_1(\ell - \xi) + R_2 \sinh \Gamma_1(\ell - \xi)\right]}{\left(K_1^2 + R_1R_2\right) \sinh \Gamma_1\ell + K_1\left(R_1 + R_2\right) \cosh \Gamma_1\ell}, \quad (7)$$

$$v_g(\xi) = \frac{K_1V_s\left[K_1 \sinh \Gamma_1(\ell - \xi) + R_2 \cosh \Gamma_1(\ell - \xi)\right]}{\left(K_1^2 + R_1R_2\right) \sinh \Gamma_1\ell + K_1\left(R_1 + R_2\right) \cosh \Gamma_1\ell}, \quad (8)$$

$$i_{i1}(x) = -\int_0^\ell E(\xi)I_1(x,\xi)\, d\xi$$

$$= -\frac{V_sZ_T}{K_2D_1D_2}\Big\{\left[K_2 \cosh \Gamma_2(\ell - x) + Z_2 \sinh \Gamma_2(\ell - x)\right]$$

$$\cdot\int_0^x A_1(\xi)B_1(\xi)\, d\xi \quad (9)$$

$$+\left[K_2 \cosh \Gamma_2 x + Z_1 \sinh \Gamma_2 x\right]\int_x^\ell A_1(\xi)B_2(\xi)\, d\xi\Big\},$$

where

$$A_1(\xi) = K_1 \cosh \Gamma_1(\ell - \xi) + R_2 \sinh \Gamma_1(\ell - \xi)$$

$$B_1(\xi) = K_2 \cosh \Gamma_2\xi + Z_1 \sinh \Gamma_2\xi$$

$$B_2(\xi) = K_2 \cosh \Gamma_2(\ell - \xi) + Z_2 \sinh \Gamma_2(\ell - \xi)$$

$$D_1 = \left(K_1^2 + R_1R_2\right) \sinh \Gamma_1\ell + K_1\left(R_1 + R_2\right) \cosh \Gamma_1\ell$$

$$D_2 = \left(K_2^2 + Z_1Z_2\right) \sinh \Gamma_2\ell + K_2\left(Z_1 + Z_2\right) \cosh \Gamma_2\ell \, .$$

$$i_{i_2}(x) = \int_o^\ell J(\xi)I_2(x,\xi)\,d\xi = -j\omega C_H \int_o^\ell v_g(\xi)I_2(x,\xi)\,d\xi \tag{10}$$

$$= \frac{j\omega C_H K_1 V_s}{D_1D_2}\left[\left\{K_2 \cosh \Gamma_2(\ell - x) + Z_2 \sinh \Gamma_2(\ell - x)\right\}\right.$$

$$\cdot\int_o^x A_2(\xi)C_1(\xi)\,d\xi - \left\{K_2 \cosh \Gamma_2 x + Z_1 \sinh \Gamma_2 x\right\}$$

$$\left.\cdot\int_x^\ell A_2(\xi)C_2(\xi)\,d\xi\right],$$

where

$$A_2(\xi) = K_1 \sinh \Gamma_1(\ell - \xi) + R_2 \cosh \Gamma_1(\ell - \xi)$$

$$C_1(\xi) = K_2 \sinh \Gamma_2\xi + Z_1 \cosh \Gamma_2\xi$$

$$C_2(\xi) = K_2 \sinh \Gamma_2(\ell - \xi) + Z_2 \cosh \Gamma_2(\ell - \xi) \, .$$

After the integration and collection of terms, the general solution for the internal current at $x = \ell$ is found to be

$$i_i(\ell) = i_{i_1}(\ell) + i_{i_2}(\ell)$$

$$= \frac{V_s Z_T}{2D_1D_2}\left[K_2 P \left\{\frac{\sinh \Gamma_s\ell}{\Gamma_s} + \frac{\sinh \Gamma_D\ell}{\Gamma_D}\right\}\right.$$

$$+ K_2 Q \left\{ \frac{1 - \cosh \Gamma_s \ell}{\Gamma_s} + \frac{1 - \cosh \Gamma_D \ell}{\Gamma_D} \right\}$$

$$- Z_1 P \left\{ \frac{1 - \cosh \Gamma_s \ell}{\Gamma_s} - \frac{1 - \cosh \Gamma_D \ell}{\Gamma_D} \right\}$$

$$- Z_1 Q \left\{ \frac{\sinh \Gamma_s \ell}{\Gamma_s} - \frac{\sinh \Gamma_D \ell}{\Gamma_D} \right\} \Bigg] \tag{11}$$

$$+ \frac{j\omega C_H K_1 V_s K_2}{2 D_1 D_2} \Bigg[Z_1 Q \left\{ \frac{\sinh \Gamma_s \ell}{\Gamma_s} + \frac{\sinh \Gamma_D \ell}{\Gamma_D} \right\}$$

$$+ Z_1 P \left\{ \frac{1 - \cosh \Gamma_s \ell}{\Gamma_s} + \frac{1 - \cosh \Gamma_D \ell}{\Gamma_D} \right\}$$

$$- K_2 Q \left\{ \frac{1 - \cosh \Gamma_s \ell}{\Gamma_s} - \frac{1 - \cosh \Gamma_D \ell}{\Gamma_D} \right\}$$

$$- K_2 P \left\{ \frac{\sinh \Gamma_s \ell}{\Gamma_s} - \frac{\sinh \Gamma_D \ell}{\Gamma_D} \right\} \Bigg]$$

where

$$P = K_1 \cosh \Gamma_1 \ell + R_2 \sinh \Gamma_1 \ell$$

$$Q = K_1 \sinh \Gamma_1 \ell + R_2 \cosh \Gamma_1 \ell$$

$$\Gamma_s = \Gamma_1 + \Gamma_2$$

$$\Gamma_D = \Gamma_1 - \Gamma_2 \, .$$

The shielding effectiveness transfer function of a cable is defined as the ratio of the internal current at $x = \ell$ to the external current at $x = \ell$ with the external line terminated in its characteristic impedance at $x = \ell$ $(R_2 = K_1)$. This definition is useful in determining the relative shielding quality of different cable shields by use of experimental techniques to obtain the transfer functions.[5] Also, some insight into a testing technique to determine the reactive shield parameters may be gained by examining the general solution for the transfer function given in equation (12).

General solution at $x = \ell$ and $K_1 = R_2$:

$$\frac{i_i(\ell)}{i_s(\ell)} =$$

$$\frac{\left[Z_T + j\omega C_H K_2 K_1\right]\left[\frac{(K_2 + Z_1)\left(e^{\Gamma_1 \ell} - e^{\Gamma_2 \ell}\right)}{(\Gamma_1 - \Gamma_2)\ell}\right]\ell + \left[Z_T - j\omega C_H K_2 K_1\right]\left[\frac{(K_2 - Z_1)\left(e^{\Gamma_1 \ell} - e^{-\Gamma_2 \ell}\right)}{(\Gamma_1 + \Gamma_2)\ell}\right]\ell}{-2\left[\left(K_2^2 + Z_1 Z_2\right)\sinh \Gamma_2 \ell + K_2 (Z_1 + Z_2)\cosh \Gamma_2 \ell\right]} . \qquad (12)$$

If the line length, ℓ, is very small compared to the wavelength, $\lambda_i = 2\pi/\beta_i$, so that $\Gamma_i \ell \doteq 0$, then

$$\sinh \Gamma_2 \ell \rightarrow 0$$

$$\cosh \Gamma_2 \ell \rightarrow 1$$

$$\frac{e^{\Gamma_1 \ell} - e^{\Gamma_2 \ell}}{\left(\Gamma_1 - \Gamma_2\right)\ell} \rightarrow 1$$

$$\frac{e^{\Gamma_1 \ell} - e^{-\Gamma_2 \ell}}{\left(\Gamma_1 + \Gamma_2\right)\ell} \rightarrow 1$$

---

[5] *R. Gray and R. McCue, Shielding Effectiveness Tests on Typical Access Facility Telephone Cables, Harry Diamond Laboratories TM-73-3 (July 1973).*

and equation (12) simplifies to

$$\frac{i_i(\ell)}{i_s(\ell)} = -\frac{\left(Z_T + j\omega C_H K_1 Z_1\right)\ell}{Z_1 + Z_2} \quad . \tag{13}$$

The two coupling parameters may be separated by proper choice of the internal terminations Z_1 and Z_2. If Z_1 is set equal to zero, then the transfer function becomes simply

$$\frac{i_i(\ell)}{i_s(\ell)} = -\frac{Z_T \ell}{Z_2} = -\frac{\left(R_s + j\omega L_s\right)\ell}{Z_2} \tag{14}$$

which allows L_s to be determined, since R_s is well defined, as stated earlier. If the line is match terminated at both ends so that

$$Z_1 = Z_2 = K_2 \ ,$$

then the capacitive parameter, C_H, also may be determined experimentally:

$$\frac{i_i(\ell)}{i_s(\ell)} = -\frac{\left(Z_T + j\omega C_H K_1 K_2\right)}{2K_2} \ . \tag{15}$$

3.2 Experimental Verification of General Solution

To verify the general solution, shielding effectiveness tests were conducted on a cable with a braided shield that exhibited both types of reactive coupling. A 1-m length of the cable was used to determine the coupling parameters, and a 30-m length was driven for comparison with a calculated response. This cable had 25 internal conductors, some with separate inner shields, and was designed for the Pershing Missile System. The internal conductors and their shields were soldered together at each end of the test cables, to simulate a coaxial cable as required by equation (12). The external braid of this cable was woven at an angle of 30 deg from the axis and had an optical coverage of 70 percent. The dc resistance of the inner bundle and the external shield was measured to determine the actual conductivity of these nonsolid conductors. The characteristic impedances K_1 and K_2 and the velocities of propagation of each line were determined with time-domain reflectometry. Time-domain data were collected with a wide bandwidth oscilloscope, and frequency spectra of the two currents were obtained by use of a spectrum analyzer and a repetitively pulsing, capacitive discharge source. The frequency spectra were reduced to the desired transfer functions by a Hewlett-Packard 9830 calculator with a digitizer and plotter. The 1-m cable transfer functions are given in figure 9 (p 22). By these curves, the external braid inductance, L_S, was found to be 8.15×10^{-10} H/m, and the capacitive coupling parameter, C_H, was calculated to be -4.5×10^{-13} F/m. The negative capacitance indicates an error in the original choice of current directions.

The experimental and calculated cable-driver responses of a 100-ft length of the same cable are given in figures 10 and 11 (p 23,24). The minimum occurring around 0.5 MHz in the 100-ft transfer function is due to an interaction between the inductive and capacitive coupling terms and does not occur in the calculated result if only an inductive shield representation is used. This effect indicates the importance of determining both types of reactive coupling. The nearly rectangular pulse at the beginning of the time history is due to the difference in propagation velocities on the inside and outside of the cable. The correlation between the experimental and calculated response is believed to be sufficient to allow the use of calculated responses in comparing point source excitation with free-field responses.

4. FREE-FIELD SOLUTION

4.1 Transmission Line Solution

A conductor of length ℓ and radius a is over a real earth at some height, h, which is illuminated by a horizontally polarized plane wave, as shown in figure 12.

The resultant fields (incident plus reflected) in the xy plane are found by Klebers[6] to be

$$E_{xy(\omega)} = E_{o(\omega)}\left[1 + R_H e^{-jk\Delta r}\right]$$

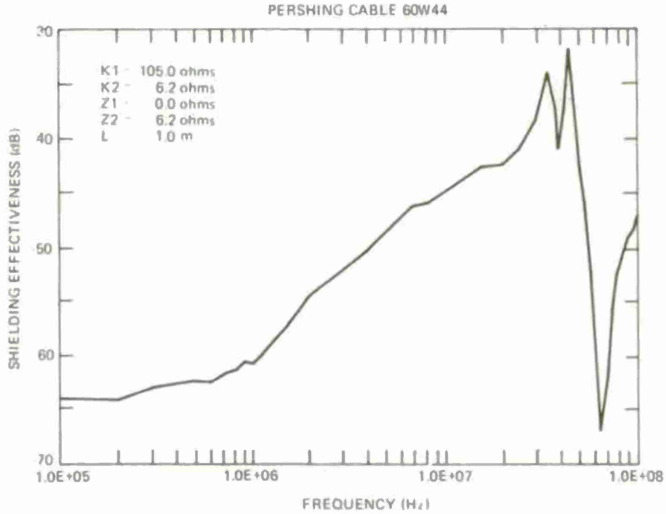
$$H_{xy(\omega)} = E_{o(\omega)}\left[1 - R_H e^{-jk\Delta r}\right] \sin(\psi)/Z_o \tag{16}$$

[6] *J. Klebers, Time Domain Analysis of the Electromagnetic Field in the Presence of a Finitely Conducting Surface, MERDC (29 January 1969).*

where

$$R_H \approx -1 + \frac{2 \sin(\psi)}{\left(\varepsilon_r - j\sigma_g/\omega\varepsilon_o\right)^{\frac{1}{2}}} \quad = \text{approximate reflection coefficient}$$

$k = \omega/c$, $\Delta r = 2h \sin(\psi)$, ψ is the angle of incidence, Z_o is the free space impedance (377 ohms), and $\varepsilon_r = \varepsilon_g/\varepsilon_o$.



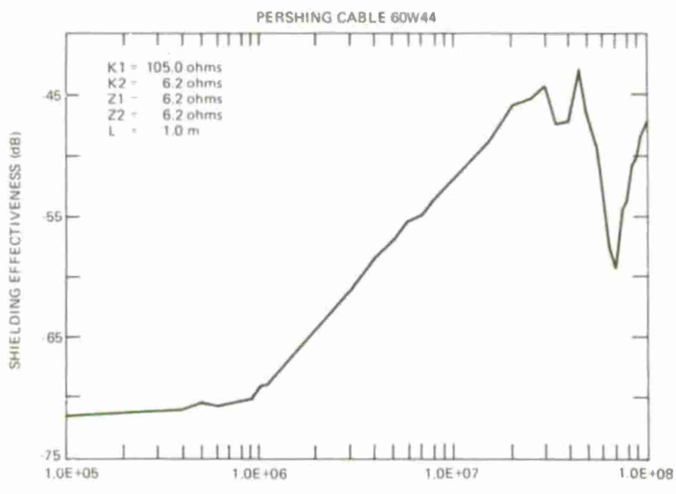
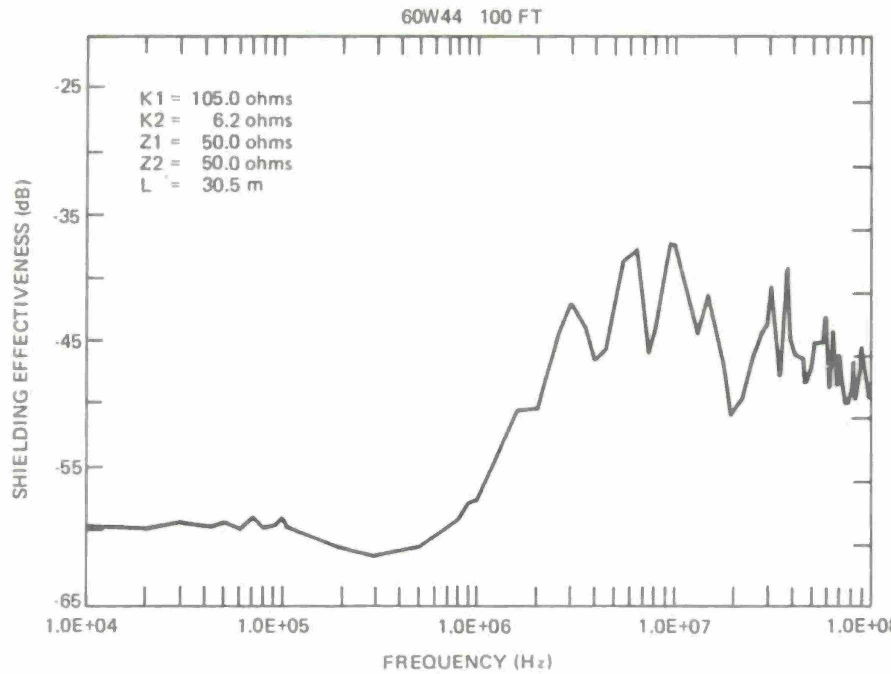


Figure 9. Shielding effectiveness curves for shield parameter definition.



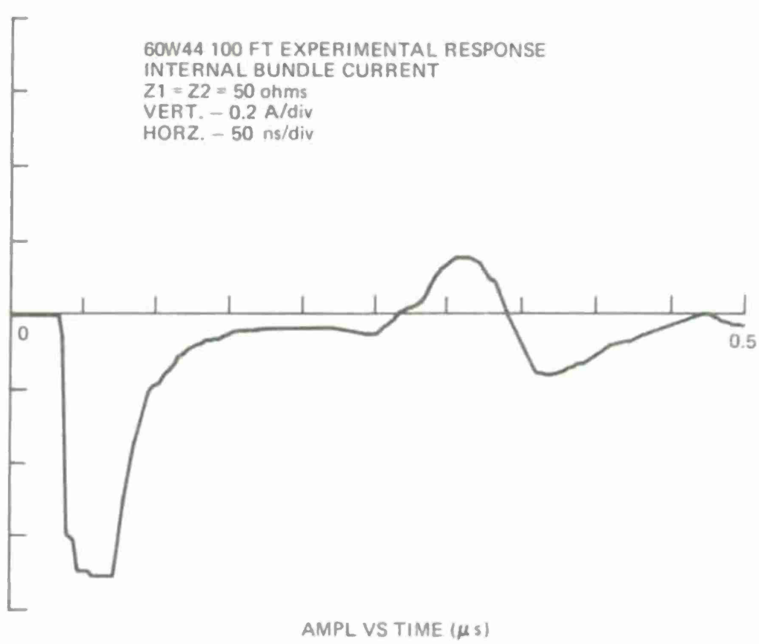


Figure 10. Experimental response of a 100-ft cable due to coaxial driver excitation.

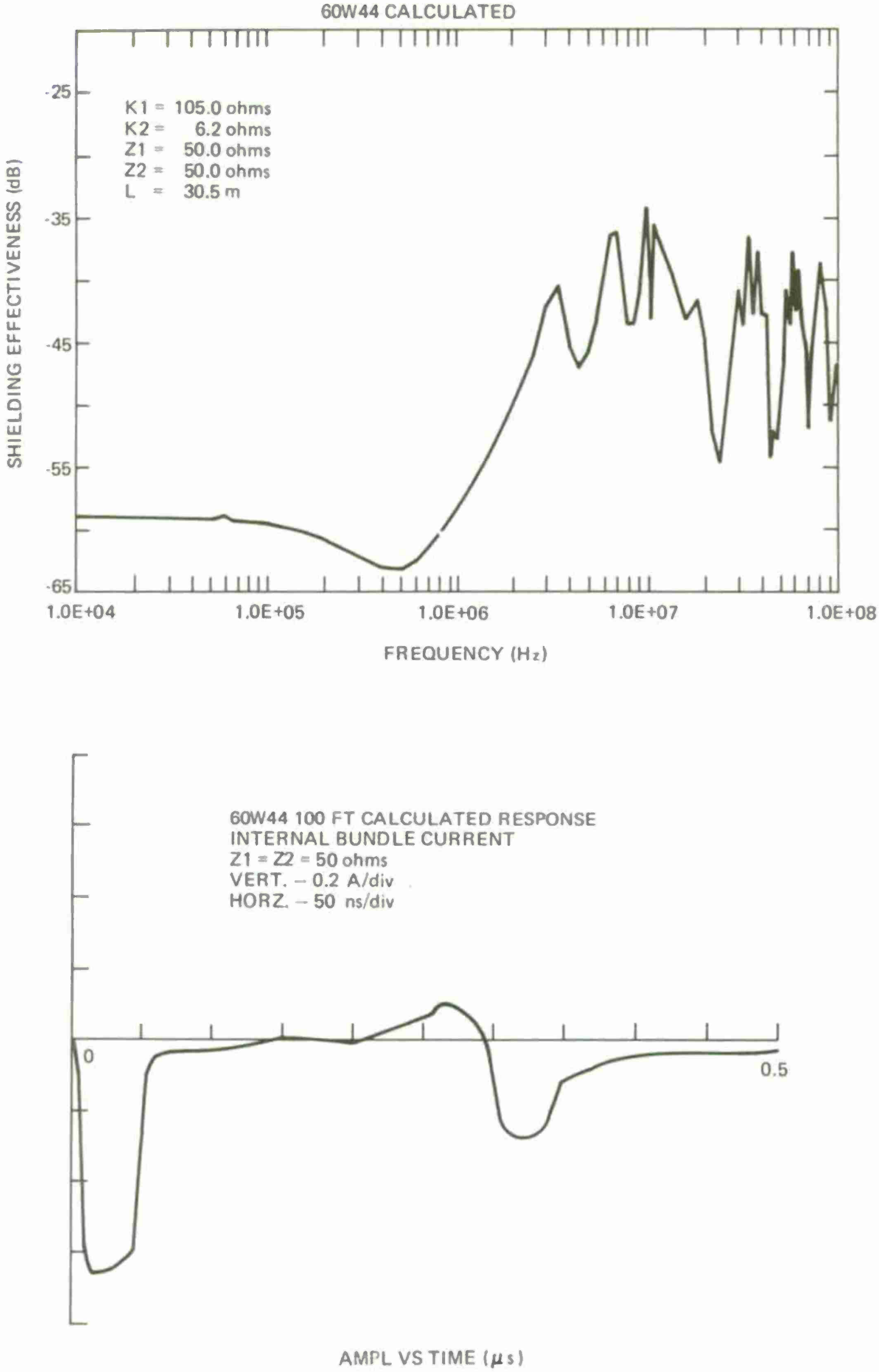


Figure 11. Calculated response of a 100-ft cable due to coaxial driver excitation.

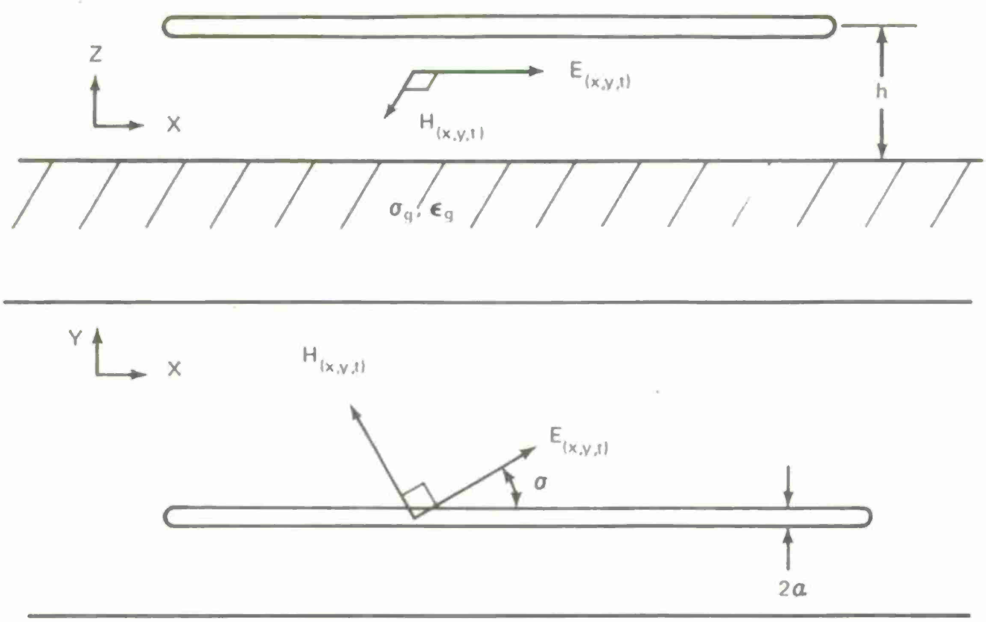


Figure 12. A conductor illuminated by a horizontally polarized wave.

The effective driving fields are the electric field, E_p, parallel to the conductor and the magnetic field, H_n, perpendicular to the conductor. In the time domain, these fields are

$$E_p(x,t) = E(t-t') \cos (\theta) \, U(t-t')$$

$$H_n(x,t) = H(t-t') \cos (\theta) \, U(t-t') \tag{17}$$

where $t' = (x/c) \sin(\theta) \cos(\psi)$ is the time delay along the line due to the wave propagating with the speed of light, c. Transformed into the frequency domain,

$$E_p(x,\omega) = E_{xy}(\omega) \cos(\theta)\, e^{-j\left[(x/c) \sin(\theta) \cos(\psi)\right]\omega}$$

$$H_n(x,\omega) = H_{xy}(\omega) \cos(\theta)\, e^{-j\left[(x/c) \sin(\theta) \cos(\psi)\right]\omega} . \tag{18}$$

The perpendicular magnetic field, H_n, may be converted to an equivalent parallel electric field by means of a coupling impedance, $Z_c = j\omega L_c$, where L_c is the inductive field coupling parameter defined by Frankel[3] to be

$$L_c = \mu_o a \left((h/a)^2 - 1\right)^{\frac{1}{2}} ,$$

for a single wire over ground. Therefore, the total equivalent electric field, E_T, driving the conductor is

$$E_T(x,\omega) = E_p(x,\omega) + j\omega L_c H_n(x,\omega)$$

$$= \left[\left(E_p(\omega) + j\omega L_c H_n(\omega)\right)\right] \cos\theta\, e^{-j\frac{\omega}{c}x \sin(\theta) \cos(\psi)}$$

$$= E_T(\omega) e^{-j\frac{\omega}{c}x \sin(\theta) \cos(\psi)} \tag{19}$$

and this field may be used to determine the current and voltage along the line by use of equation (3).

---

[3] S. Frankel, *Terminal Response of Braided-Shield Cables to External Monochromatic Electromagnetic Fields*, Harry Diamond Laboratories TR-1602 (August 1972).

The earth can be assumed to be a reasonable approximation of an infinitely conducting surface such that the conductor and its image in the earth may be treated as a two-wire transmission line. This line has an impedance

$$K_1 = 0.5(Z/Y)^{1/2} = 0.5\left((R + j\omega L)/j\omega C\right)^{1/2},$$

terminations Z_1 and Z_2 at $x = 0$ and ℓ, respectively, and a velocity of propagation, v_1. The variation in propagation velocity of the cable with height aboveground was empirically determined by measuring the current at the center of a 40-ft unterminated cable for various heights above- and belowground due to illumination by the Biconic Simulator at the Woodbridge Research Facility of the Harry Diamond Laboratories. This simulator consists of a dipole antenna that is driven in the time domain by a Marx generator at a biconical center feed section. A radiating dipole antenna is formed by extending cylindrical arms from the bicones. The experimentally determined propagation velocities are shown in figure 13. This curve was represented by the function

$$v_1 = 0.5C\left(1 + 1/\left(\varepsilon_r\right)^{1/2} + \left(1 - 1/\left(\varepsilon_r\right)^{1/2}\right) \tanh\ (4.4h - 0.449)\right)$$

and is also presented in figure 13 (p 29). This representation makes the propagation velocity frequency independent. However, this is not an unreasonable approximation when the cable is aboveground, since losses have only a second-order effect on the velocity

$$\left(v = v_o \left[1 - \frac{1}{8\omega^2}\left(\frac{R}{L} - \frac{G}{C}\right)^2\right]\right).^1$$

The external current along the conductor is found to be

$$i_s(x,\omega) = \frac{E(\omega)M(x)}{D_1}$$

$$\cdot \int_o^x \left[K_1 \cosh \Gamma_1 \xi + Z_2 \sinh \Gamma_1 \xi\right] e^{-j\left(\frac{\omega \sin (\theta) \cos (\psi)}{c}\right)\xi}\, d\xi$$

$$+\frac{E(\omega)N(x)}{D_1}$$

$$\cdot\int_x^{\ell}\Big[K_1 \cosh \Gamma_1(\ell - \xi) + Z_1 \sinh \Gamma_1(\ell - \xi)\Big]\, e^{-j\left(\frac{\omega \sin (\theta) \cos (\psi)}{c}\right)\xi}\, d\xi$$

$$= \frac{E(\omega)}{2D_1}\Bigg[\Big\{K_1 \cosh \Gamma_1(\ell - x) + Z_2 \sinh \Gamma_1(\ell - x)\Big\}$$

$$\left\{\frac{(K_1 - Z_1)\left(1 - e^{-(\Gamma_1 + j\beta')x}\right)}{\Gamma_1 + j\beta'} - \frac{(K_1 + Z_1)\left(1 - e^{(\Gamma_1 - j\beta')x}\right)}{\Gamma_1 - j\beta'}\right\} \qquad (20)$$

$$+ \varepsilon^{-j\beta'\ell}\Big\{K_1 \cosh \Gamma_1 x + Z_1 \sinh \Gamma_1 x\Big\}$$

$$\cdot\left\{\frac{(K_1 - Z_2)\left(1 - e^{(\Gamma_1 - j\beta')(x - \ell)}\right)}{\Gamma_1 - j\beta'} - \frac{(K_1 + Z_2)\left(1 - e^{-(\Gamma_1 + j\beta')(x - \ell)}\right)}{\Gamma_1 + j\beta'}\right\}\Bigg]$$

where

$$M(x) = K_1 \cosh \Gamma_1(\ell - x) + Z_2 \sinh \Gamma_1(\ell - x)$$

$$N(x) = K_1 \cosh \Gamma_1 x + Z_1 \sinh \Gamma_1 x$$

$$D_1 = K_1\Big[\left(K_1^2 + Z_1Z_2\right)\sinh \Gamma_1\ell + K_1(Z_1 + Z_2) \cosh \Gamma_1\ell\Big]$$

$$\beta' = \omega \sin (\theta) \cos (\psi)/c.$$

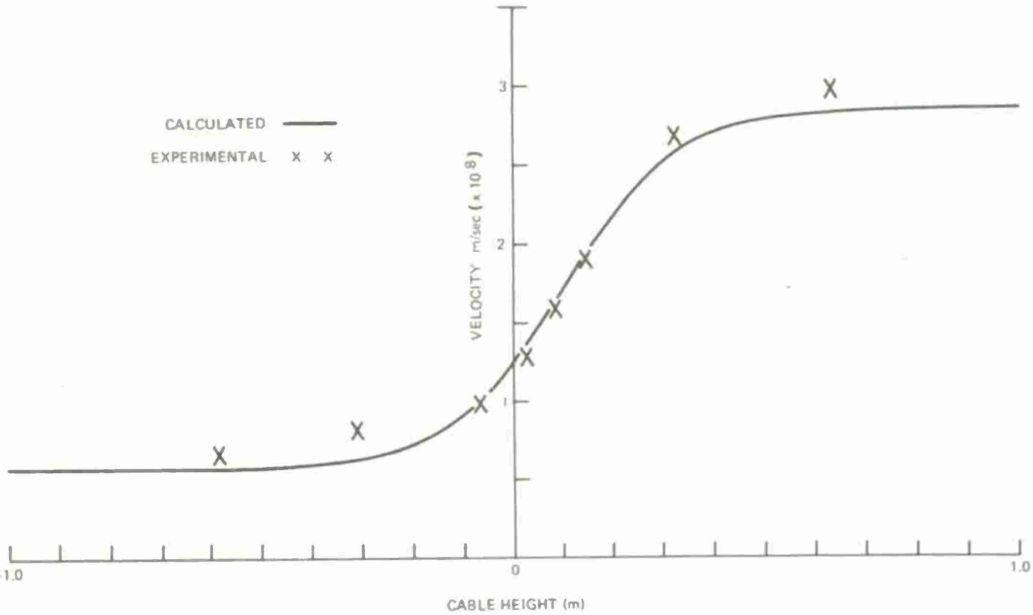


Figure 13. Propagation velocity as a function of cable height.

A similar solution is obtained for the external voltage, $v_s(x,\omega)$. The internal current is then found by use of the external voltage and current solutions and the following integration:

$$i_i(x,\omega) = \int_0^{\ell} \left[Z_T i_s(\xi,\omega) I_1(x,\xi) - j\omega C_H v_s(\xi,\omega) I_2(x,\xi)\right] d\xi \qquad (21)$$

where $I_1(x,\xi)$ and $I_2(x,\xi)$ have been defined earlier. The solutions for the internal currents are not presented here because of their bulk. However, they have been programmed (FREFLD) along with the external current solution and field definition for a threatlike incident field (double exponential representation) to run on a PDP 11/40 computer. Time-domain solutions are obtained with the same numerical inverse transform as used for the coaxial driver solution.

4.2 Experimental Verification

To validate the results of the program FREFLD, calculations were made for several cable conditions for which experimental results existed. Most of the experimental data were collected by use of the Biconic antenna. Many of the differences between theory and experiment seem to be attributable to the fact that the calculations were made with an idealized double exponential incident field, which is, at best, a rough approximation of the field radiated by the Biconic Simulator.

The external current at the center of a 40-ft cable was calculated for two different heights aboveground. The results along with the experimental data are given in figures 14 and 15. Although the calculated waveforms do not exactly duplicate the measured responses, they do exhibit the same general trends, such as the increase in the effective length of the cable and attenuation as the cable is closer to the earth.

The next set of data (fig. 16) demonstrates the effect of the angle of rotation, θ, between the incident electric field and the axis of the cable. These calculations were made for a 1500-ft cable that was terminated in a slightly inductive short to earth at each end. Internal current was calculated by use of a solid shield found to be representative of that used in telephone systems.[5] The internal termination at $x = 0$ was a matched impedance ($Z_3 = K_2$), and at $x = \ell$, the calculation point, the inner bundle was shorted to the shield. The external current peak amplitude increased considerably, whereas its duration decreased (due to a shortening of its electrical length) as the angle of rotation varied from broadside to near end-fire illumination. This effect was found in scale-model measurements for similar conditions. The solid-line graph in figure 17 plots the measured initial current peaks relative to the peak at broadside incidence. Also shown (with x's) are the relative calculated peak amplitudes, including some additional angles not shown earlier. The internal current peak increases only slightly and then decreases with increasing angle of rotation.

A set of field measurements was made on the same 100-ft cable that was used to validate the coaxial driver work. The Biconic antenna was used to illuminate the cable with a 0-deg angle of rotation (θ = 0 deg) and a 7-deg incidence with the earth (ψ = 7 deg). The cable was supported at 1 m aboveground, and the measurements were made with both open- and short-circuit terminations at each end:

$$Z_1 = Z_2 = 0 \text{ and } Z_1 = Z_2 = \infty .$$

The external current was measured at the center of the cable, and the internal current was measured with both ends matched terminated:

$$Z_3 = Z_4 = K_2 .$$

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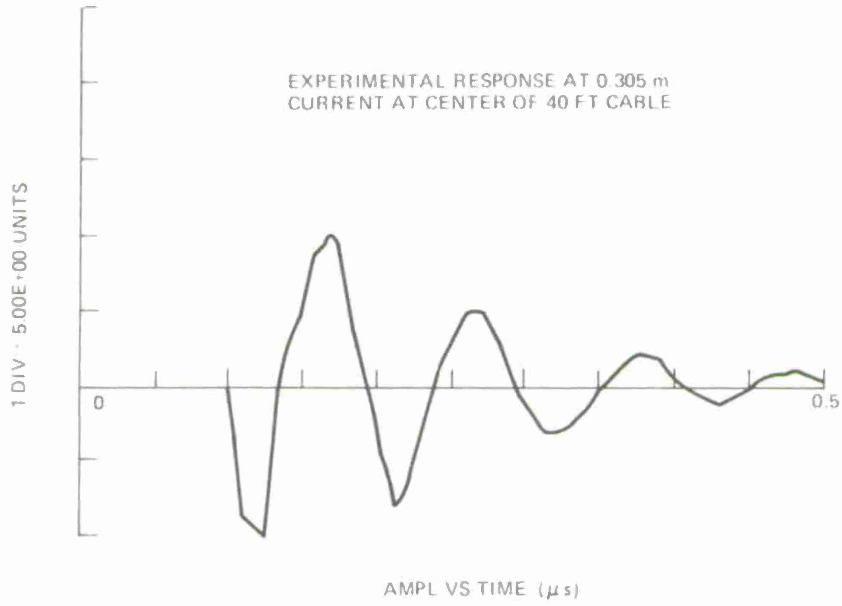
[5] *R. Gray and R. McCue, Shielding Effectiveness Tests on Typical Access Facility Telephone Cables, Harry Diamond Laboratories TM-73-3 (July 1973).*



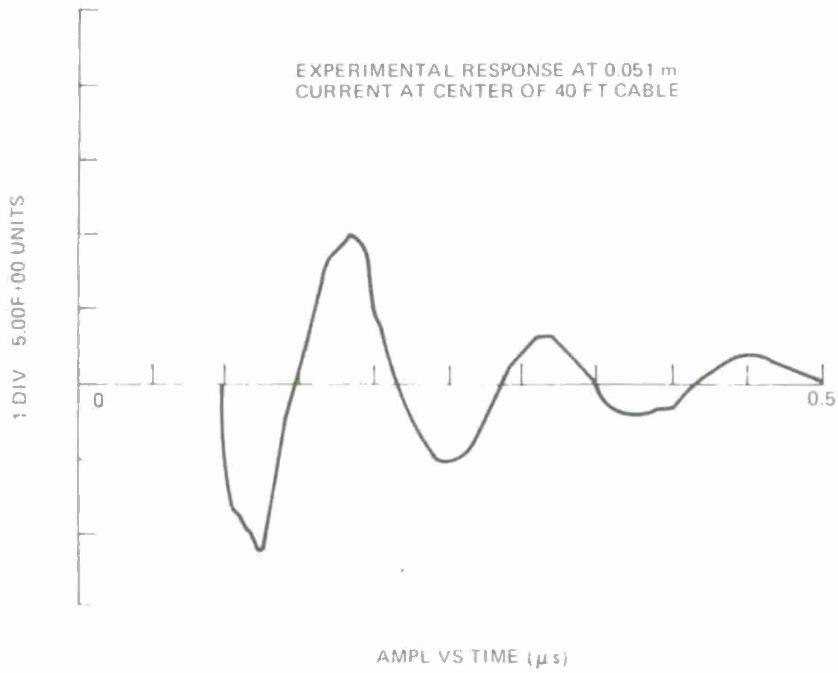


Figure 14. Experimental responses of cable at different heights aboveground.

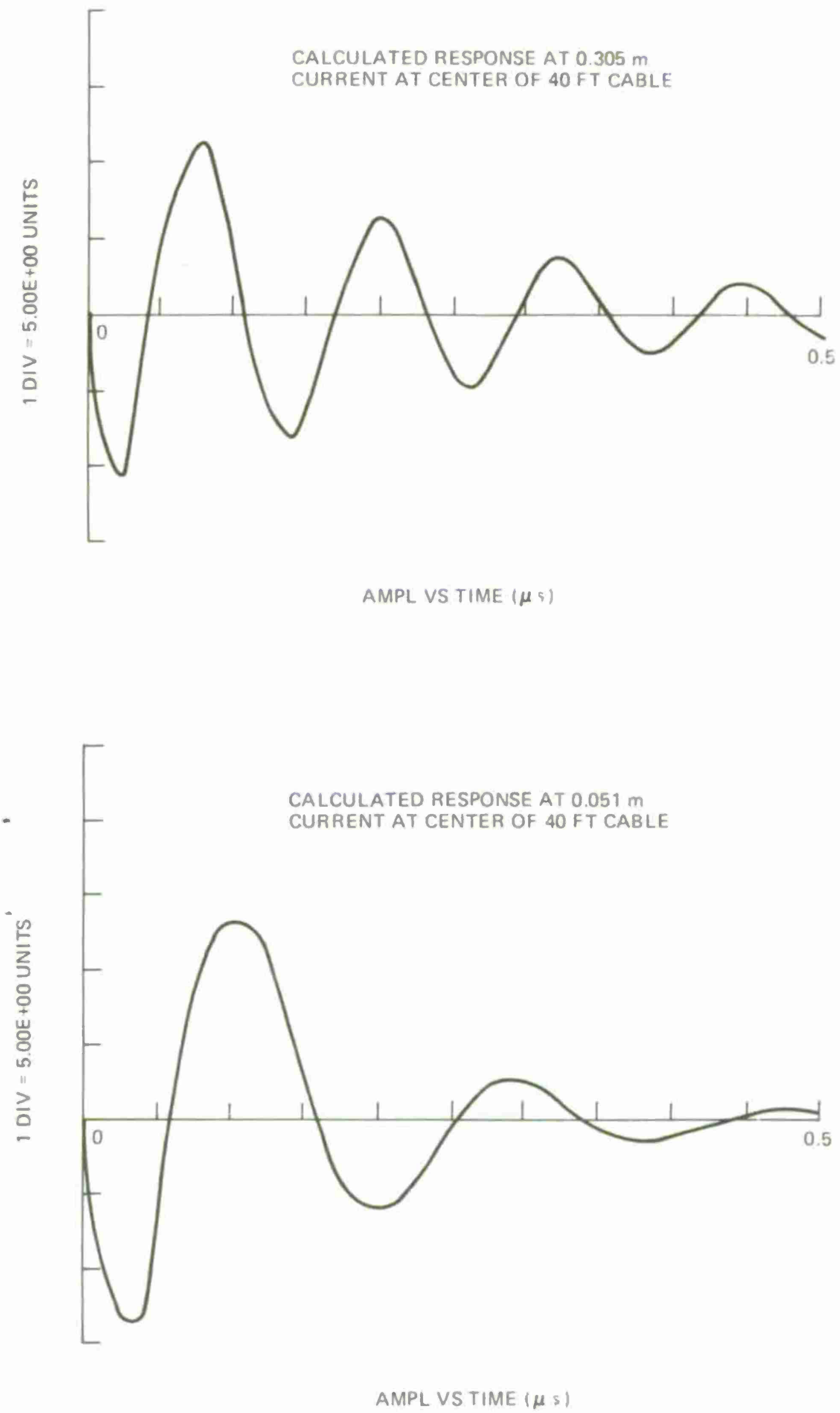


Figure 15. Calculated responses of cable at different heights aboveground.

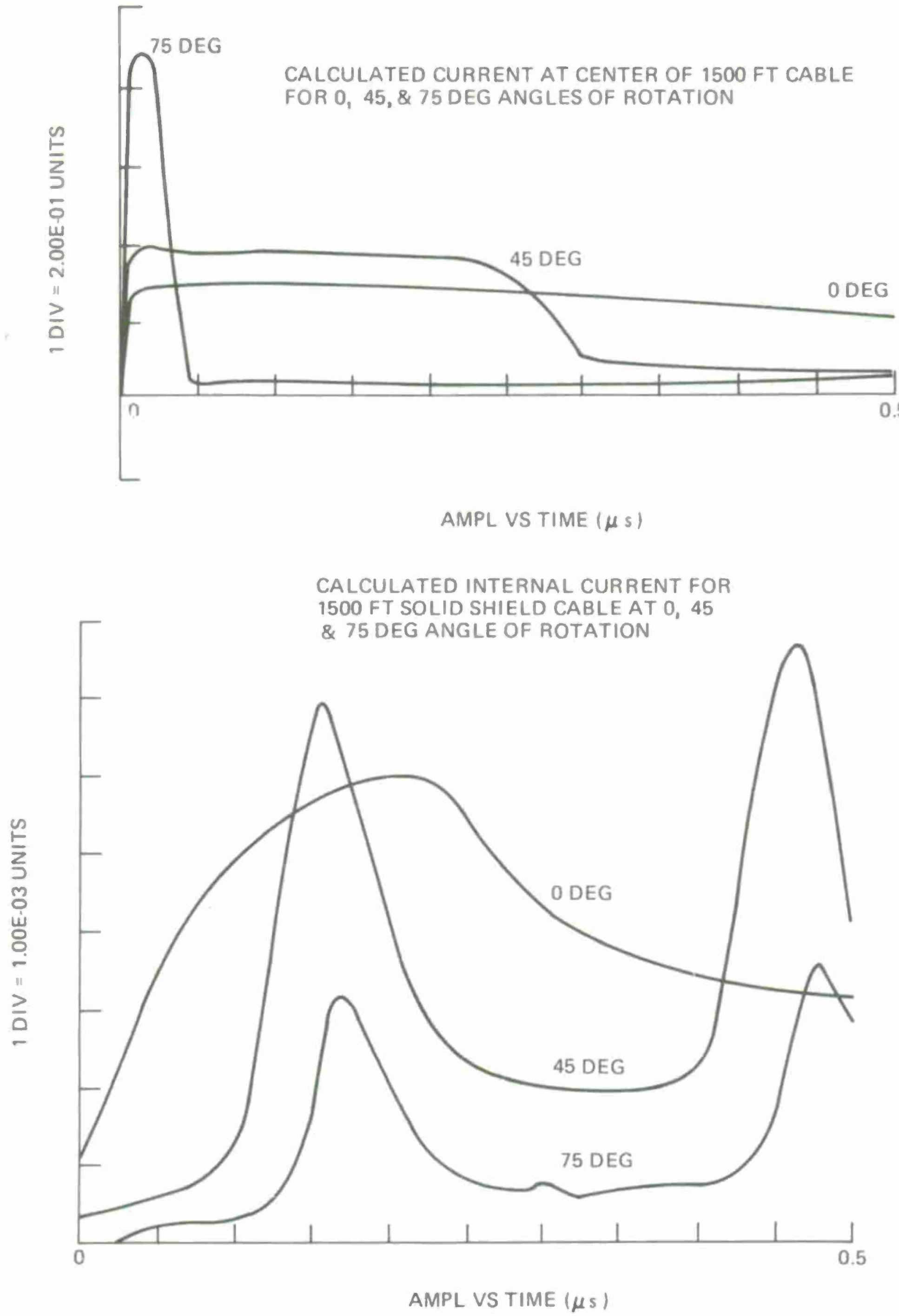


Figure 16. Calculated currents for different angles of rotation.

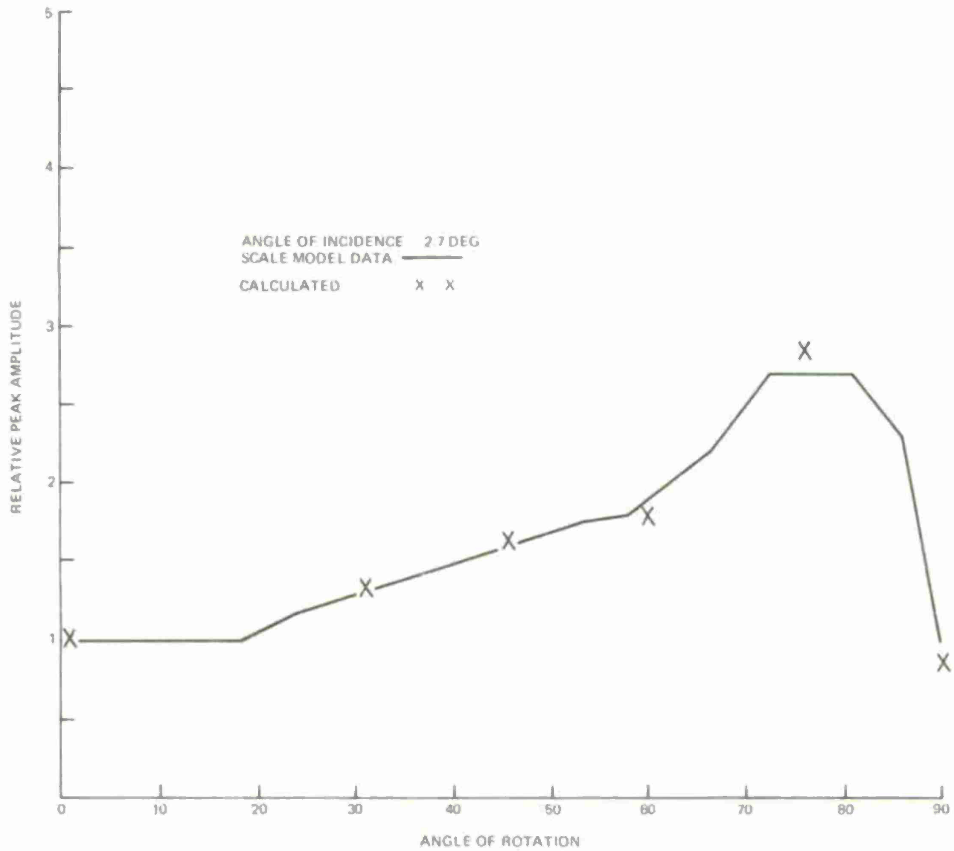
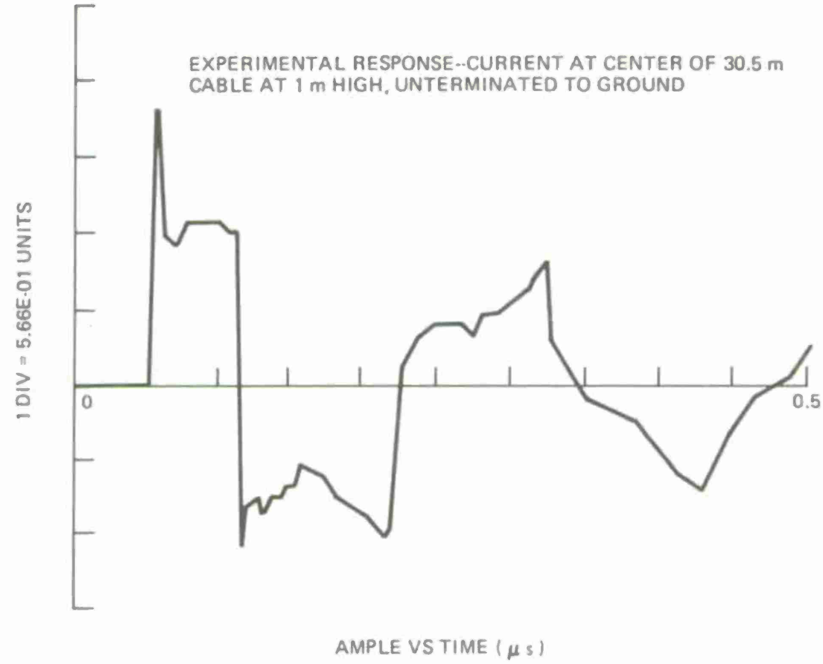


Figure 17. Relative peak amplitude as a function of angle of rotation.

These measured responses and the corresponding calculated results are presented in figures 18 to 21. The two results do not correlate exactly, but the calculations represent the experimental responses enough to allow use of the code to determine where cable-driver and free-field responses diverge. They do so especially since most of the differences are probably due to error in representing the external shield terminating impedances Z_1 and Z_2. The internal responses erred the most at or after a reflection was seen in the shield current (time shift about 50 ns between the two response points, i_s at $\ell/2$ and i_i at ℓ). With the short-circuit termination, the problem lies in defining the inductance of the ground straps used and the impedance of the grounding stakes. For the open-circuit termination, there is a capacitance to ground from the 1-m^3 aluminum box used on the measurement end to house the instrumentation. This box was removed when the external current was measured.



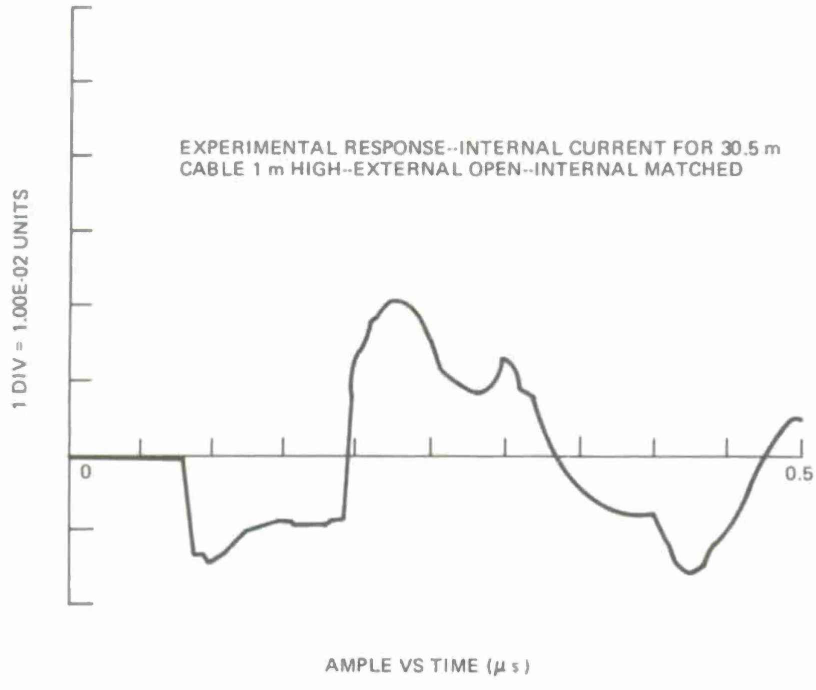
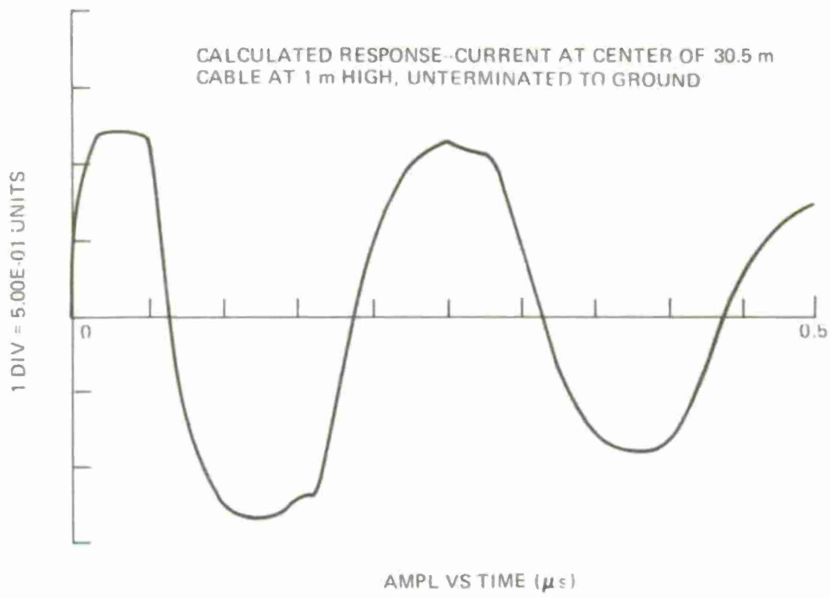


Figure 18. Experimental responses with cable shield unterminated.



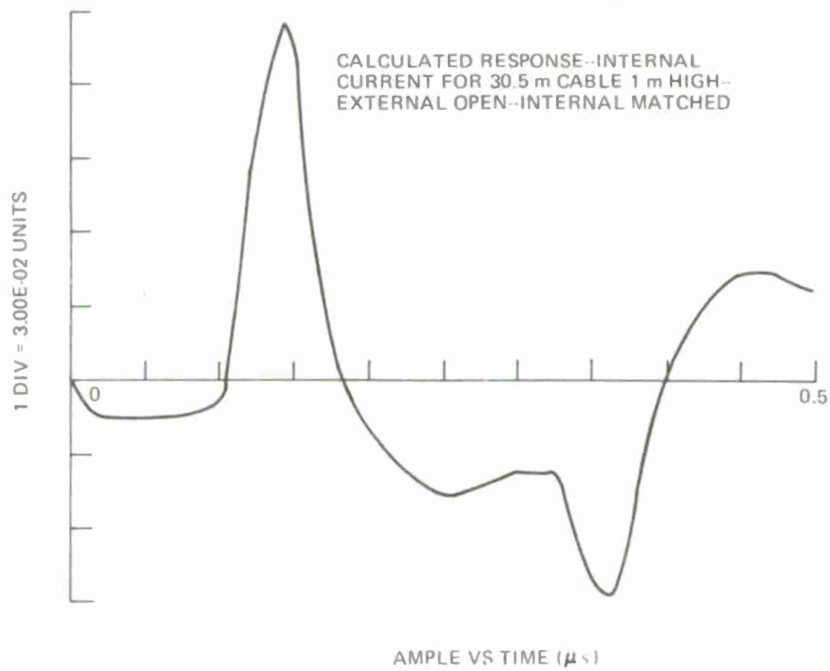
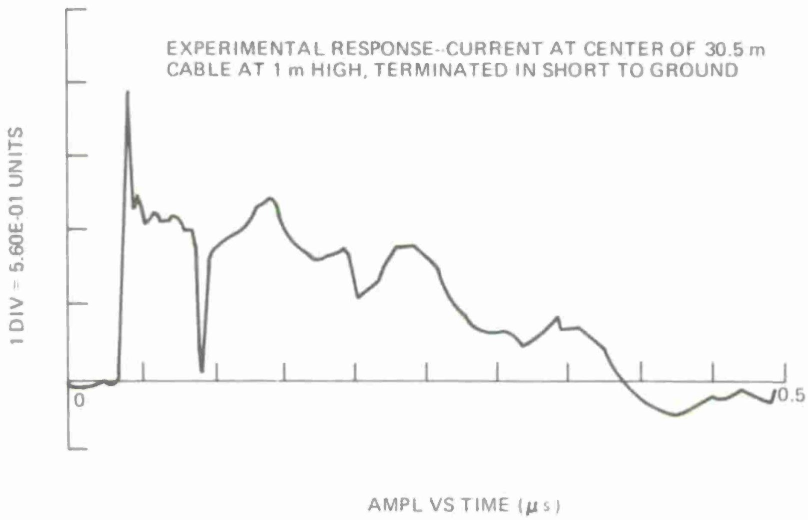


Figure 19. Calculated responses with external shield unterminated.



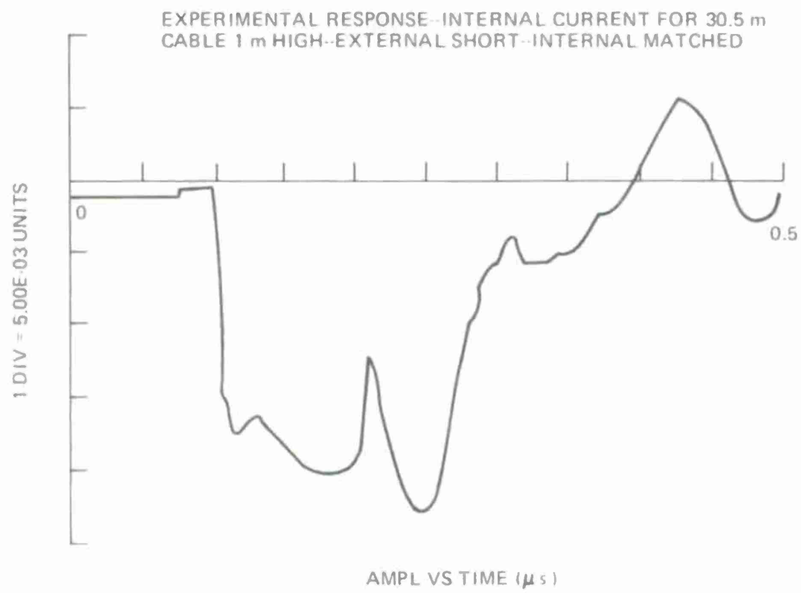
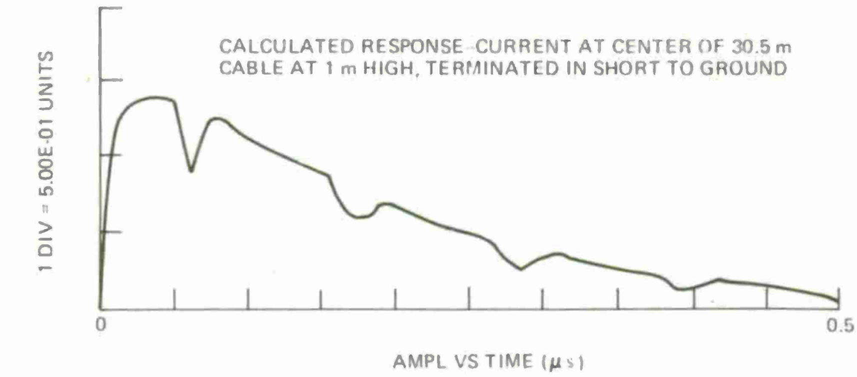


Figure 20. Experimental responses with cable shield terminated to the earth.



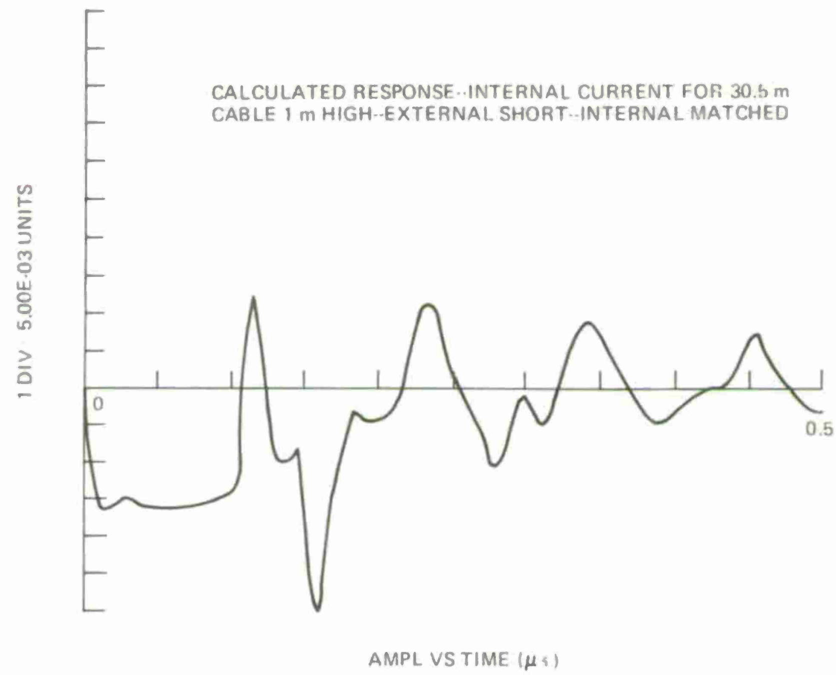


Figure 21. Calculated responses with cable shield terminated to the earth.

Therefore, considering these uncertainties in external terminating impedances and the error due to the field representation used, the correlation between the experimental and calculated responses is sufficient to validate the code for its use here.

5. CABLE DRIVER VERSUS FREE FIELD

5.1 Calculated Comparison

In the previous sections, a general cable shield model was presented, and transmission line solutions for both types of excitation were developed and validated. These solutions are now used to obtain at least an initial idea of how well, if at all, a point source injection system simulates the response of a shielded cable illuminated by an EMP.

Internal current must be calculated for both excitations, with the same internal terminations and the external current waveforms as similar as possible. Since the coaxial-driver solution treats only an overdamped, capacitive-discharge driving source, the free-field calculations have to be limited to conditions that produce a double exponential, external-current waveform. The condition necessary to produce this response is broadside incidence with both ends of the cable terminated in idealized short circuits to the earth.

Calculations were made for three different lengths, 2, 5, and 8 m, of the braided-shield cable. The external current for the coaxial driver was adjusted to match the external current of the free-field calculation. The internal current responses with both ends match-terminated are given in figures 22 and 23. The peak amplitudes for each excitation are about the same, but the durations are consistently longer for the cable-driver calculation. This difference is caused by the fact that the current due to the driver is propagated from one end of the cable to the other, and the current due to the free field occurs simultaneously along the line. The pulse duration, T_f, for the free-field calculation varies proportionately to the one-way propagation time of the internal line, $T_f \sim \ell/v_2$. The driver pulse duration, T_c, is proportional to the sum of the internal and external transit times, T_c $(1/v_1 + 1/v_2)\ell$. The internal current response at each end of the cable is identical for the free-field calculation, but quite different for the coaxial driver, as shown in figure 24. This difference is caused by the difference in phasing for the two excitations. Similar calculations were made for an 8-m solid-shield

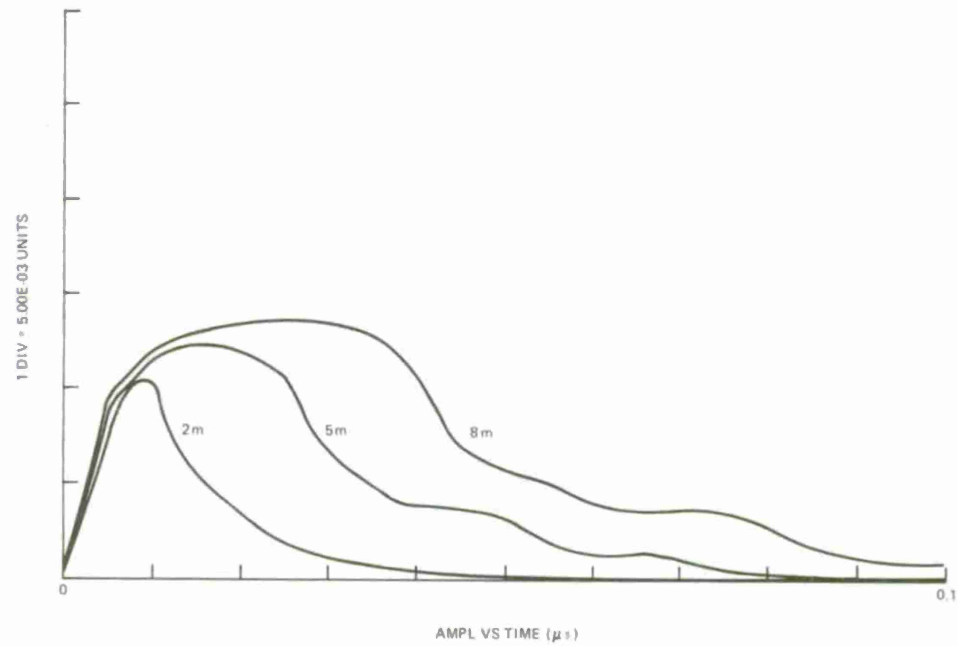


Figure 22. Calculated internal current responses of braided-shield cable for free-field excitation.

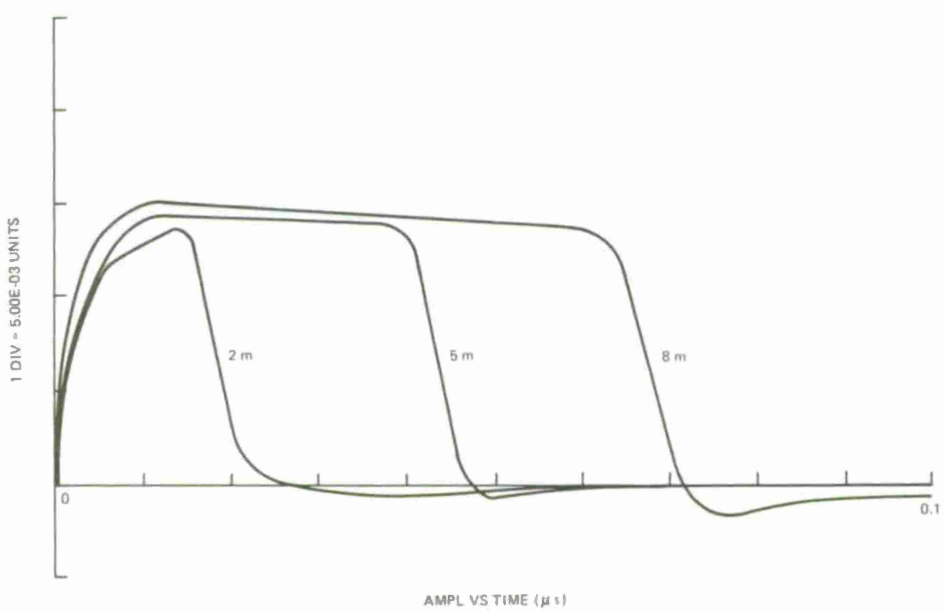


Figure 23. Calculated internal current responses of braided-shield cable for coaxial driver excitation.

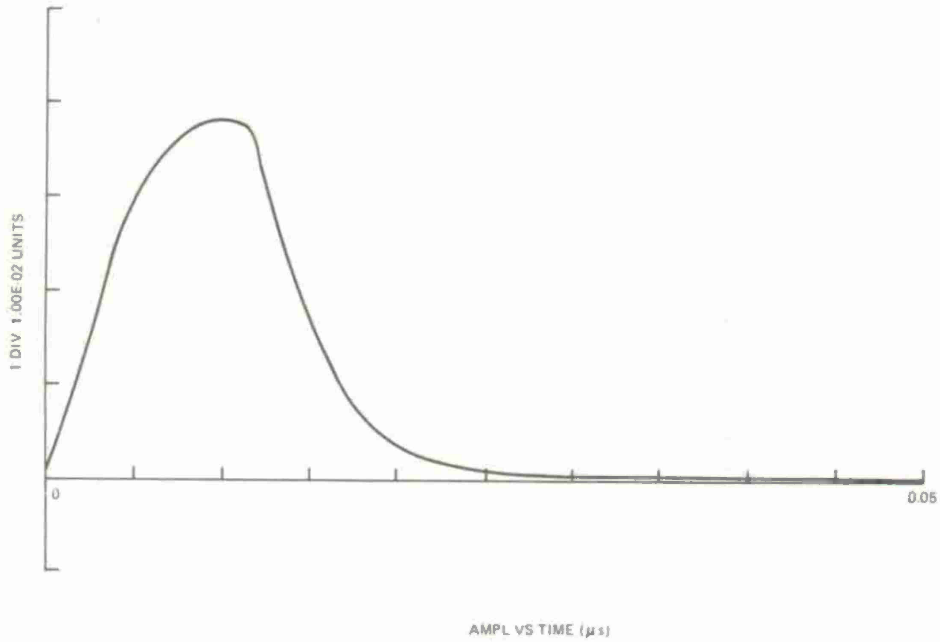


Figure 24. Calculated internal current response at other end of 8-m braided-shield cable for coaxial driver excitation.

cable, and the responses are given in figures 25 and 26. Here, the much greater high-frequency attenuation of the solid shield caused the internal responses to be far less sensitive to the phase differences of the external current.

The differences observed in the calculations indicate that the adequacy of any point source injection system depends on the type of cable shield being driven. It also suggests that the performance of an injection system could be improved by making the phasing of the external current closer to that caused by free-field illumination. It may be made so by use of several distributed point sources along the line, rather than just one, as considered earlier.

5.2 Multiple Point Source Injection System

The same 100-ft braided-shield cable that was used in the free-field experiments was also driven by an injection system inductively coupled to the shield via multiple outputs. The cable was tested in the field with the same terminations as those in the test with the Biconic Simulator. The inductive coupler is a transformer made up

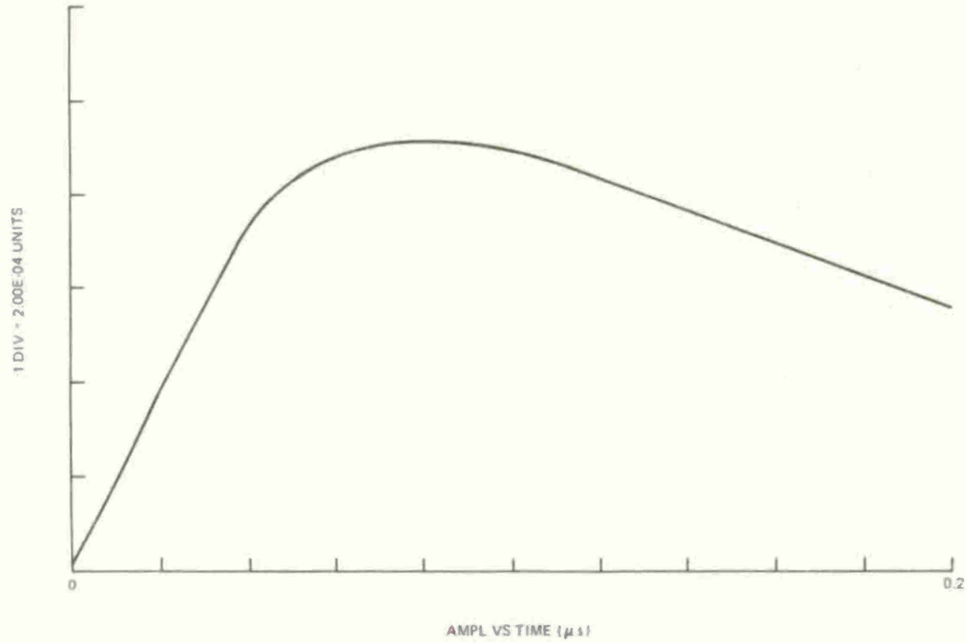


Figure 25. Calculated internal current response of 8-m solid shield cable for free-field excitation.

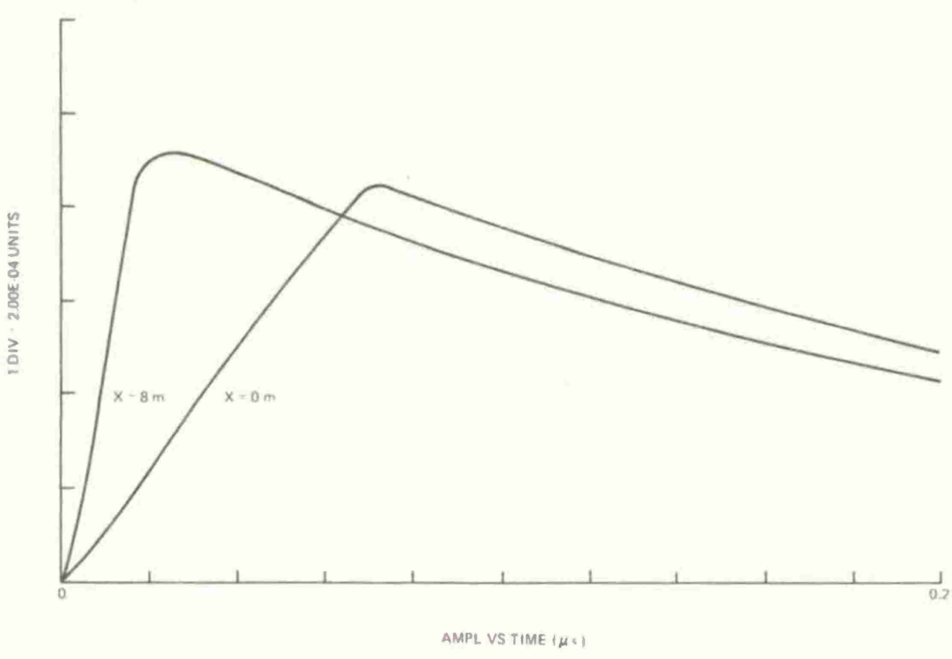
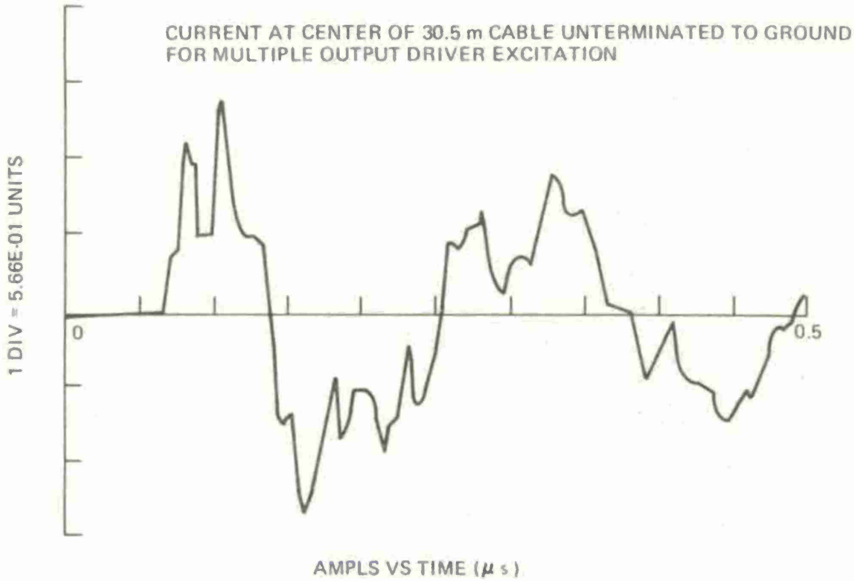


Figure 26. Calculated internal current responses at each end of 8-m solid shield cable for coaxial driver excitation.

of a square ferrite core that is split in the middle, a one-turn primary winding that is driven by a capacitive discharge source, and a one-turn secondary winding that is the cable to be driven. This type of system was used because of the ease in connecting the couplers to the cable. Also, this type of coupler behaves as a series voltage source along the cable, whereas most other types of physically obtainable coupling methods would act like shunt current sources. The major disadvantage of this type of coupling is that both efficiency and secondary waveforms depend very much on the secondary impedance. It is possible to control the secondary response somewhat by use of different types of core material. The type of core material available for this test would only couple a narrow (~30 ns) pulse onto the cable shield due to its high impedance (>300 ohms). This narrow pulse width affected the test results, since it was not possible to couple the desired waveform, but with this problem uncovered, it is possible to factor it out of the results.

Five separate coupling units were used with a spacing of 25 ft along the cable. These units were driven by one source through 50 ft of 50-ohm coaxial line each. The external and internal responses for both external termination conditions are given in figures 27 and 28. Unlike the single source driver, the internal current response was the same at each end of the cable. The narrow current pulse induced by each coupler was the reason for the jagged response. The overall envelopes for the open-circuit termination, though, are very close to those seen for free-field illumination (fig. 18). However, the waveforms for the short-circuit termination do not agree as well with the free-field data (fig. 20). The short-circuit condition has a greater dependence on the late-time characteristic of the driving field. Therefore, the comparisons would probably be improved if a better core material were used. If some imagination is used to smooth out the first half of the internal waveform, some similarity with the free-field response may be seen. The free-field response is basically a rectangular pulse followed by a narrower pulse due to the sheath current reflection from the inductive ground straps. This response also occurs in the cable-driver experiment, but after the smaller pulse, the external current pulse is very different from the free-field case.

In comparing the relative amplitudes of the free-field and multiple driver responses, it is difficult to obtain an exact ratio of the internal current to the external current for the cable-driver experiment because of its jagged waveform. However, using average values for the driver-excited currents indicates that the driver responses are about 30 percent lower than the free-field currents. Better inductive couplers may improve this situation, since the calculated comparisons with near-identical external waveforms were very close.



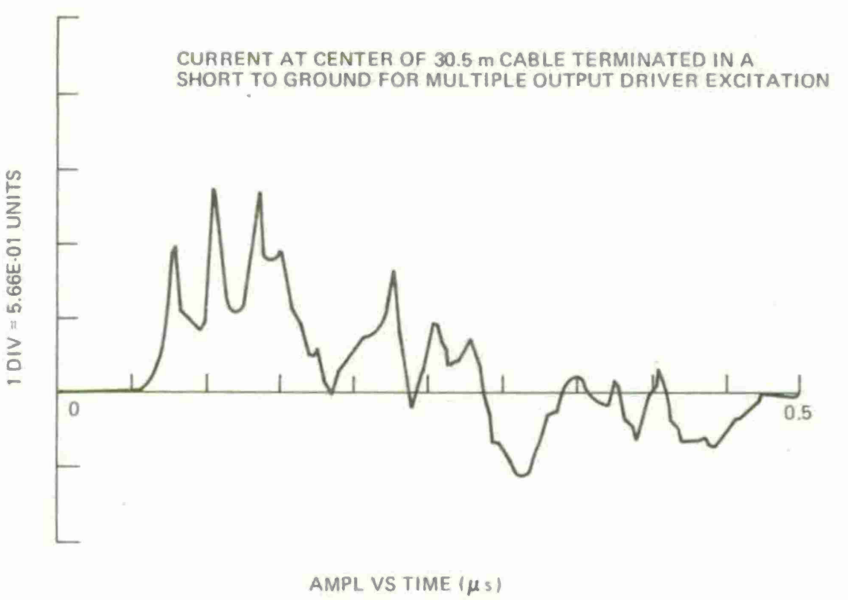
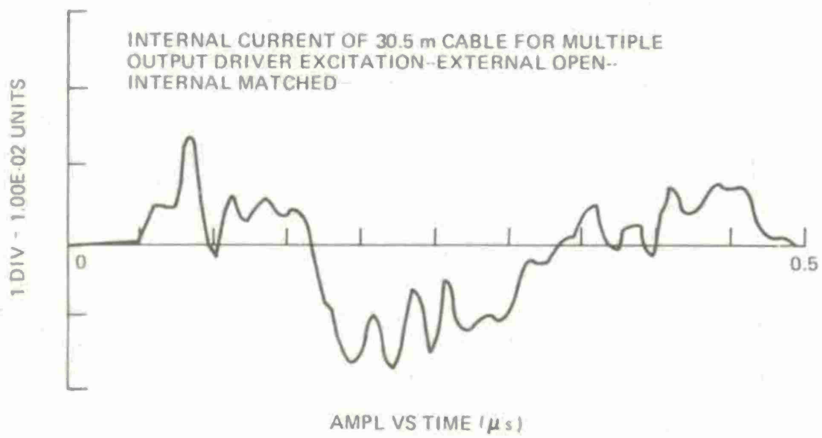


Figure 27. External currents for multiple output driver excitation.



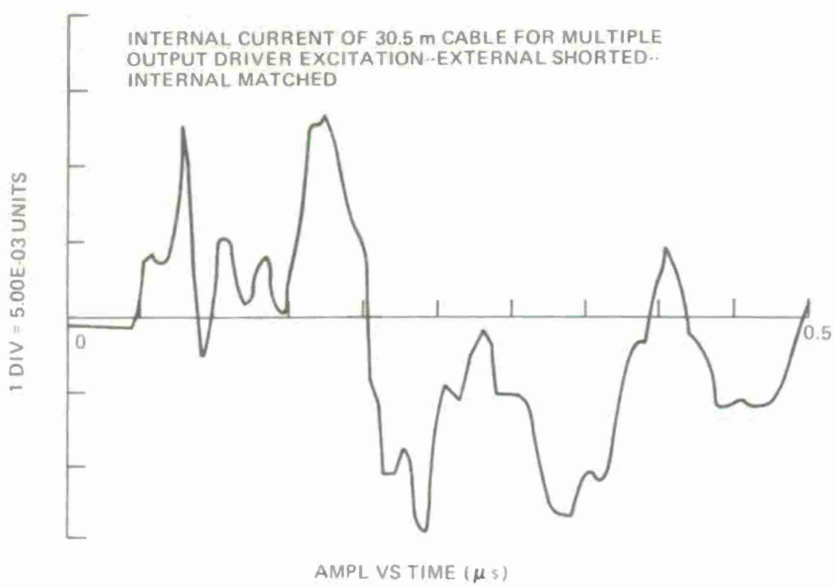


Figure 28. Internal currents for multiple output driver excitation.

6. CONCLUSIONS AND RECOMMENDATIONS

Some basic design techniques have been explored for point source simulation of EMP-produced currents on both shielded and unshielded system penetrations. A high-impedance, shunt-current generator was found to be the most advantageous method of point-source exciting simple unshielded penetrations. A more extensive study of shielded cable excitations was found to be necessary. A general shield coupling model was presented and used in obtaining transmission line solutions for cable responses due to point-source and distributed-source (free-field) excitation of the external current. These solutions were verified with various experimental data and then used to determine some basic areas of applicability of point-source injection simulation.

The validity of this type of simulation was found to be very dependent on the type of shield involved (solid or braided), but not sensitive to the length of the penetration. A multiple-output, inductively coupled injection system was postulated to increase the viability of point-source simulation for braided-shield cables. Experimental results showed that this technique improved the distribution of the external current along the cable shield, thereby better simulating an actual EMP. Some basic problems need to be examined to determine the most efficient type. Also, the transmission of energy to the couplers from the source would need to be improved, since the coaxial lines used in these experiments leaked enough energy to cause unwanted excitation of the test cable. Once these problems were resolved, then a system level validation of the technique would be required before it could be generally applied.

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