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Sensor and Simulation Notes

Note 205

September 1974

Direct Voltage and Current Injection

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Abstract

Methods of directly coupling sources to cables for EMP testing are reviewed. Effects of cable impedance, cable shielding, and structure in the vicinity of the cable are examined. Several practical coupling mechanisms are discussed.

Acknowledgement

This work sponsored by the Defense Nuclear Agency under Subtask R99QAXEC091-03.

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## 1. INTRODUCTION

Systems that use long interconnecting cables and transmission lines between parts of the system are susceptible to energy penetration through the cables because cables represent large antennas with correspondingly large intercept area. Thus, it is often found that these long conductors are the paths of strongest EMP penetration into the various system enclosures. Many major system responses to EMP therefore can be simulated by exciting cables with waveforms that simulate the transients induced in the cables by the EMP fields. This method of simulating system response to EMP has been successful on systems that have long, buried, shielded cables or aboveground power transmission lines.

The advantage of being able to drive the cable or transmission line directly to investigate system responses, rather than to illuminate the entire system with an EM wave of sufficient intensity to simulate the EMP fields, is apparent when the energy storage and coupling structures for these two approaches are compared. In EM simulation, a large volume of space around the system is illuminated with the simulated EMP, but only a small fraction of the energy in this wave is collected by the system (the remainder is scattered or absorbed by the ground). By directly injecting the transient onto the cable, however, only the small fraction of the incident energy that is collected by the cable need be stored and coupled. Moreover, the requirements of the injection system are often much less stringent than those of the EM simulator. For example, in systems using buried cables with solid shields, it may not be necessary to strive for very fast waveform rise times in the direct injection system because the soil and the cable shields filter out these rapidly varying components. Finally, the time and cost required to design, fabricate, and use a direct-injection test system are usually only a small fraction of that required for a full-scale, large-area, EM simulator.

The ability to perform full-scale, source-relatable, direct-injection experiments depends on knowing the pulse shape that is

to be injected. For simple antenna structures<sup>1</sup> (dipoles, loops) and for long cables and transmission lines<sup>2</sup>, analytical techniques are available for predicting the induced currents and voltages from the incident EM field. More complicated structures may require low-level, EM experiments to determine the pulse shape or spectrum that should be injected.

One of the limitations of the direct injection scheme for simulating the system response to the EMP is the uncertainty in reconstituting the entire system response from its responses to several injected stimuli. Suppose, for example, that the system of interest is found to have strong EMP energy penetration on commercial power lines, long communication cables, and large radio communication antennas. Direct injection tests can be performed on each of these systems at high levels but if each of these penetrations has a comparable effect on a particularly critical component of the system, the question arises as to how these individual responses combine at the critical component. If the system is nonlinear, in general, linear circuit analysis techniques may not be used to combine the responses (although in many special cases such techniques are adequate).

One approach to determining the combined effect of several penetrations is to synchronize the injectors so that all major penetration paths are driven simultaneously. This approach requires sophisticated trigger systems to sequence each direct injection at known times with low trigger jitter. It also presumes that the reconstitution, which was questioned at the critical system component level, can be performed with adequate

confidence at the injection points. (This approach can be verified experimentally by using various firing sequences.) An obvious approach to eliminating the problem of reconstituting the responses at a critical component, however, is simply to harden the system until no combination of individual responses can possibly affect the critical component. If the component is indeed critical and if its response to the injected pulses is close enough to a malfunction or damage threshold so that 6 to 10 dB increases from combining the responses to two or three penetrations are of concern, then the safety margin is probably unacceptable, and additional hardening will be required. If this approach is taken, the question of how to reconstitute the responses to individual penetrations becomes academic and the expensive and time-consuming test approach of synchronizing several excitation systems is avoided.

There are several methods of coupling a signal onto a cable or transmission line. The one that is chosen for a particular test problem will depend on such factors as the configuration of the cable circuit, its impedance, the type of shielding, the degree of simulation required, and the budget for the test. In this chapter, the injection methods are classified as current injection or voltage injection according to whether the quantity simulated is current (as in a cable shield) or voltage (as between a cable conductor and the equipment ground). The driving techniques may thus be classified according to whether the cable is shielded or unshielded. In most shielded cable systems in which the shield is to be driven, the shield current is of primary interest; for unshielded conductors or for injection of the conductors inside the shield, the voltage between the conductor and system ground may be of primary interest. Because there will be a voltage associated with the shield current and a current associated with the conductor-to-ground voltage, these classifications indicate the focus of interest rather than the characteristics of ideal generators.

## 2. CURRENT INJECTION ON CABLE SHIELDS

The object of driving cable shields is usually to produce a transient current in the shield and observe the response of the system to this current. The requirements placed on the current waveform often include specific wave shapes, rise times, and

uniformity along the length of the shield. To control these properties of the shield current, it may be necessary to design wave shaping circuits and current return paths as well as an energy storage unit. The current return path is an important element of the coupling structure because it sets the size and uniformity of the impedance that must be driven by the energy source. Several coupling structures that may be used to obtain a known, uniform characteristic impedance between the cable shield and its current return are described below. Some of these schemes take advantage of the system or cable construction to minimize the amount of additional hardware that must be installed, and some permit the pulse to be shaped by the natural impedances of the system structure. It should also be pointed out that while this section is concerned with shielded cables, the techniques discussed apply equally well to similar structures by any other name (e.g., armor, conduit, and pipeline).

## 2.1 COAXIAL TRANSMISSION LINE CONFIGURATIONS FOR DRIVING CABLE SHIELDS

One of the simplest concepts for obtaining a uniform current density and characteristic impedance in a cable shield is to make the cable shield the center conductor of a coaxial transmission line as illustrated in Figure 1. With this configuration, the characteristic impedance of the transmission line formed by the cable shield, and its concentric current return path is:

$$Z_o = \frac{60}{(\epsilon_r)^{1/2}} \log \frac{r_o}{r_s} \quad (1)$$

where  $r_o$  is the inside radius of the return path,  $r_s$  is the outside radius of the shield, and  $\epsilon_r$  is the dielectric constant of the insulation between the shield and the return path. If this structure is terminated in the impedance  $Z_o$  and driven by discharging a capacitor bank into the transmission line, then the RLC pulser circuit discussed in Chapter Two describes the shield current. When the current rise time caused by the energy source inductance is neglected the current in the shield for a step-function voltage will be:

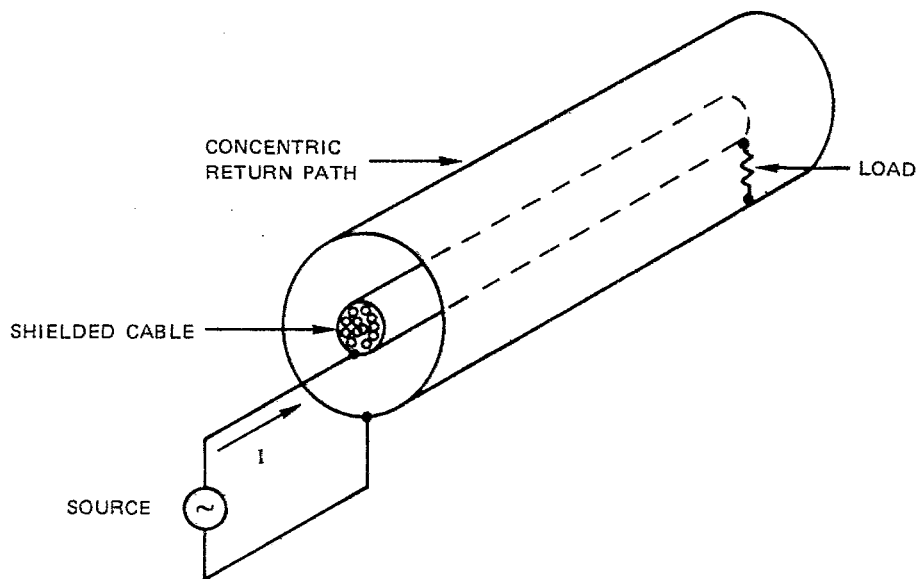


FIGURE 1 CONCENTRIC CYLINDER CURRENT INJECTION COUPLING STRUCTURE IN A COAXIAL TRANSMISSION LINE

$$I(t) = \frac{V}{Z_0} e^{-t/CZ_0} \quad (2)$$

where  $V$  is the initial voltage of the capacitor bank and  $C$  is its capacitance. The peak voltage between the shield and the cylindrical return path is the initial voltage  $V$  of the capacitor bank. The spectrum of the current pulse is:

$$I(\omega) = \frac{CV}{1 + j\omega CZ_0} \quad (3)$$

The magnitude of the spectrum is  $CV$ , and the spectrum is flat within 3 dB from dc to  $f = 1/2\pi CZ_0$ . The bandwidth, spectral magnitude, and peak value of the driving current are related to the capacitance, characteristic impedance, and operating voltage as follows:

$$\begin{aligned} \text{Bandwidth} &= 1/2\pi CZ_0, \text{ in Hz} \\ \text{Spectral magnitude} &= CV, \text{ in ampere-seconds} \\ \text{Peak current} &= V/Z_0, \text{ in amperes.} \end{aligned} \quad (4)$$

The exponential current pulse obtained from the capacitor discharge into a terminated transmission line has the desirable characteristics that—(1) it is a fair simulation of the pulse shape that is induced in buried cables by the EMP and (2) its spectrum is continuous — that is, it contains no dominant zeros or poles where the current spectrum is very small or very large. (For example, the rectangular pulse spectrum contains many zeros, and the lightly damped sinusoid spectrum is dominated by a narrow band.) Although the flat bandwidth of the exponential pulse is  $1/2\pi CZ_o$ , its usable bandwidth is much greater because the spectrum is well behaved even when its magnitude is decreasing as  $1/f$ . The usable spectrum is determined by the rise time of the pulse which was assumed to be zero in Equation (2).

The capacitance, voltage, and characteristic impedance can be manipulated to obtain the desired spectral magnitude, peak current, or bandwidth as shown in Equation (4) within the ranges permitted by available capacitors and voltage breakdown limits of the coaxial configuration. To maximize the peak current, for example,  $Z_o$  must be made small, which implies making  $r_o/r_s$  in Equation (1) approach unity. This approach requires that the gap between the inner and outer conductors become small, however, and encroaches on the ability of the gap to hold off the peak voltage  $V$ . The peak field strength  $E_s$  at the surface of the shield is:

$$E_s = \frac{V}{r_s \log \frac{r_o}{r_s}} \quad (5)$$

This field strength, which is critical to breakdown considerations, is the minimum for a given applied voltage  $V$  and outer conductor radius  $r_o$  when  $r_o/r_s = e = 2.72$ . In our case, however, we are more likely to have a fixed shield radius  $r_s$ , with freedom to choose the diameter of the return current path. With these conditions, the maximum peak current can be expressed as:

$$\begin{aligned} I_{\max} &= \frac{V}{Z_o} = \frac{r_s \log \left( \frac{r_o}{r_s} \right) E_{\text{breakdown}}}{\frac{60}{(\epsilon_r)^{1/2}} \log \left( \frac{r_o}{r_s} \right)} \\ &= \frac{(\epsilon_r)^{1/2} r_s E_{\text{breakdown}}}{60} \end{aligned} \quad (6)$$

For air at atmospheric pressure  $E_{\text{breakdown}} = 3 \times 10^6$  V/m and  $\epsilon_r = 1$ ,  $I_{\text{max}}$  is:

$$I_{\text{max}} = 5 \times 10^4 r_s \quad (7)$$

Thus, the maximum current that can be developed in a terminated coaxial structure having a 2-inch diameter cable with atmospheric air dielectric is about 1 kA. By using oil dielectric, for which  $\epsilon_r \approx 2.4$  and  $E_{\text{breakdown}} \approx 1.5 \times 10^7$  V/m, the maximum current can be increased to about 8 kA. For polyethylene ( $\epsilon_r = 2.4$ ,  $E_{\text{breakdown}} = 4 \times 10^7$  V/m), the maximum current is about 20 kA. It should be noted that these are fundamental limits on the currents that can be injected on semi-infinite cables or cables terminated in their characteristic impedance. These currents can be achieved only when the driving terminals and termination structure are carefully designed to avoid concentrations of the electric field that might reduce the breakdown voltage. It should also be pointed out that the transmission line geometry of concentric circular cylinders produces an electric field strength around the cable shield that is uniform in the azimuthal direction. Any other configuration, such as parallel cylinders, produces a nonuniform, azimuthal distribution of field strength so that a transmission line of a given characteristic impedance formed with any other geometry will break down at a lower voltage than a line of that impedance formed by concentric circular cylinders. To achieve currents larger than those given above, therefore, the terminating impedance must be made smaller than the characteristic impedance. Reducing the terminating impedance produces a mismatch in the transmission line that results in a damped oscillatory current in the shield, however.

Coaxial cylinders can be formed from the two outer shields of a doubly-shielded cable to drive a current in the inner shield. This scheme, as illustrated in Figure 2, is very efficient in terms of the pulse driver requirements because only the current in the inner shield must be simulated and this current is often much smaller than the total cable current induced by an incident EM wave. Whether or not this driving technique can be used depends on the characteristics of the shield system and the inner shield current waveform. To apply this technique or any direct injection technique, the designer must have prior knowledge of the inner shield current waveform (from either test or analysis) to determine the coupling between the incident EM wave and the inner shield.



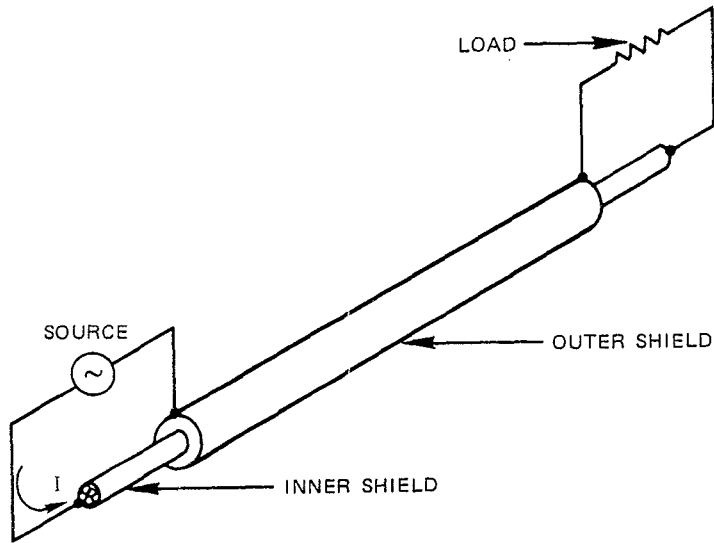


FIGURE 2 SHIELD AS A COUPLING STRUCTURE FOR CURRENT INJECTION

A second variation of the coaxial cylinder coupling geometry, as illustrated in Figure 3, also makes use of the natural environment of the cable system. In this case, the shielded cable is buried in the soil and is insulated from the soil by its plastic jacket. Since the metal shield, plastic jacket, and soil form a natural coaxial geometry, they can be used as the coupling structure for producing current in the shield. If this driving scheme is to be effective, the insulating jacket on the cable must be free of penetrating cuts or abrasions throughout the length of the cable that is to be driven. It will also be necessary to establish a low impedance connection to the soil at the driving point so that an acceptable fraction of the source voltage is applied to the transmission line. Finally, the wave propagating on the transmission line must not be severely attenuated by the soil return path so that it is dissipated before a sufficient length of cable has been excited.

The characteristic impedance of the shield/soil transmission line is:

$$Z_o \approx \left( \frac{j\omega L + Z_s}{j\omega C} \right)^{1/2} \quad (8)$$

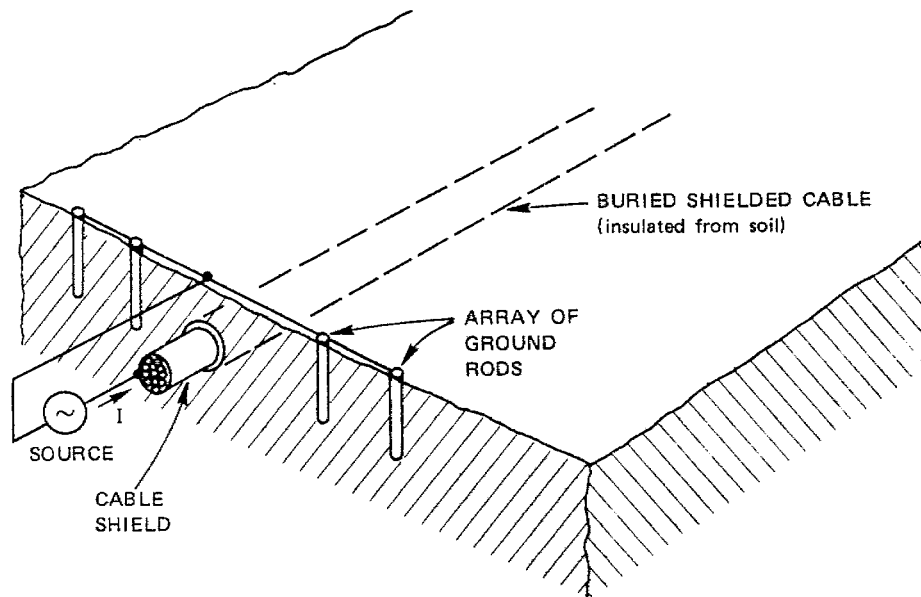


FIGURE 3 GROUND RETURN AS A COUPLING STRUCTURE FOR CURRENT INJECTION

and the propagation factor is:

$$\begin{aligned} \gamma &= \left[ j\omega C (j\omega L + Z_s) \right]^{1/2} \\ &= \alpha + j\beta \end{aligned} \quad (9)$$

where  $C$  is the capacitance per unit length and  $L$  is the inductance per unit length associated with the insulating gap, and  $Z_s$  is the impedance per unit length of the return path through the soil. (It is assumed that  $Z_s$  is much greater than the internal impedance of the cable; otherwise the internal impedance of the cable must be added to  $Z_s$ ). The inductance, capacitance, and soil impedance are given by:

$$L = \frac{\mu_0}{2\pi} \log \frac{r_i}{r_s} \quad (10)$$

$$C = \frac{2\pi\epsilon}{\log \frac{r_i}{r_s}} \quad (11)$$

$$Z_s = -\frac{j\Gamma r_i}{2\pi r_i^2 \sigma} \frac{H_0^{(1)}(j\Gamma r_i)}{H_1^{(1)}(j\Gamma r_i)} \quad (12)$$

$$\approx \frac{1}{4\sigma\delta^2} + j \frac{1}{\pi\sigma\delta^2} \log \frac{1.414\delta}{\gamma_o r_i}$$

where  $r_s$  is the radius of the shield and  $r_i$  is the outside radius of the insulating jacket (or the radius of the hole in the soil). In Equation (12)  $\sigma$  is the soil conductivity,  $\gamma_o$  is Euler's constant (1.7811), and  $\Gamma$  is the propagation factor in the soil:

$$\Gamma = \left[ j\omega\mu_o (\sigma + j\omega\epsilon_s) \right]^{1/2} \quad (13)$$

$$\approx \frac{1+j}{\delta} \quad (\sigma \gg j\omega\epsilon_s)$$

The approximation for  $Z_s$  in Equation (12) is based on the assumption that  $|\Gamma r_i| \ll 1$ , which implies that the radius of the insulation is small compared with a skin depth in the soil.

A plot of the characteristic impedance,  $Z_o$ , attenuation constant,  $\alpha$ , and propagation factor,  $\beta$ , for a 2-inch diameter cable with a 100-mil thick polyethylene jacket buried in soil with a conductivity of  $10^{-2}$  mho/m is shown in Figure 4. It can be seen from the figure that the characteristic impedance of the cable when soil is the return conductor is larger than the characteristic impedance for a perfect return conductor by a factor of almost 10. That is because of the large internal inductance of the soil return path. It is also apparent that frequencies above 1 MHz are attenuated at rates of 10 dB per 100 meters of propagation, or greater, so that these higher frequency components are rapidly lost if long cable segments are driven by this method. This limitation on high-frequency components will not be severe if the cable shields are solid-walled cylindrical shields of moderate thickness, however, since these shields do not permit significant electromagnetic penetration at frequencies above 100 kHz. For leaky shields, such as braided wire shields, this limitation might be unacceptable.

The impedance of the ground rods that connect one side of the generator to the soil (see Figure 3) must be made small compared with the characteristic impedance

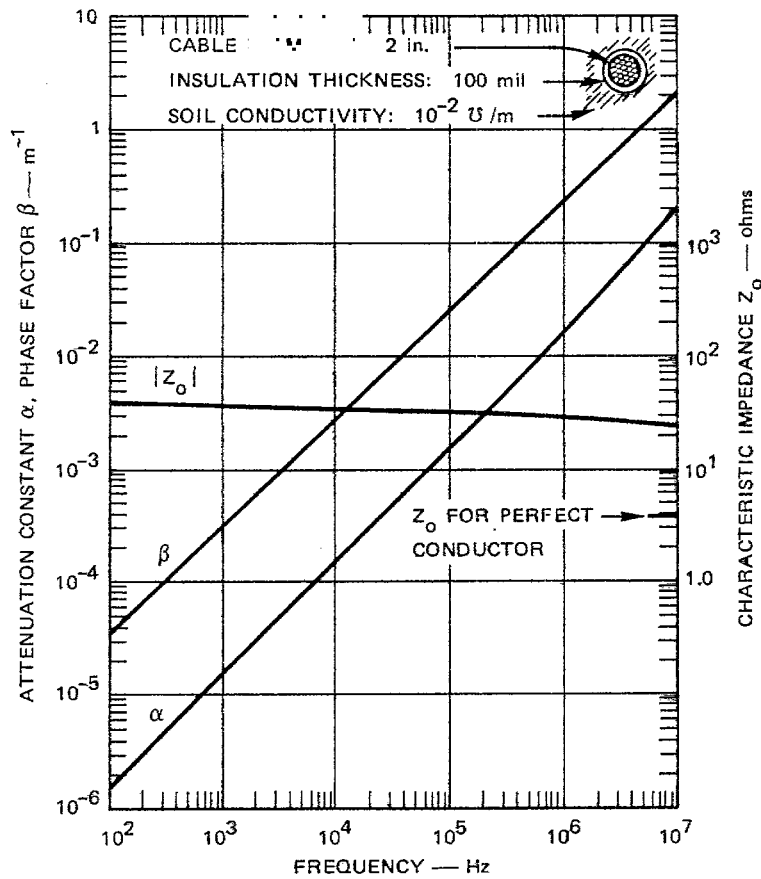


FIGURE 4 CABLE CHARACTERISTICS FOR A 2-INCH DIAMETER CABLE IN SOIL

$Z_0$ . This condition is not too difficult to achieve, inasmuch as the characteristic impedance is fairly large (several tens of ohms) and impedances of 10 ohms or less can be achieved with single ground rods that are driven 1 to 2 meters into the soil.

## 2.2 PARALLEL WIRE CONFIGURATIONS FOR DRIVING CABLE SHIELDS

In practice, the required bandwidth of the current pulse spectrum and the voltage limitations of the terminating resistor and capacitor bank require that the characteristic impedance be made as small as possible. The lowest characteristic impedances

are available in coaxial transmission lines; however, it is difficult to construct such a line (except in those cases discussed above where the natural geometry of the system can be used), if the test cable is more than about a hundred feet long, particularly, if the outer shield of the test cable is not insulated for high voltages. In spite of its very desirable electrical features (i.e., low characteristic impedance and uniform current distribution), this method of forming the transmission line has limited application because of the mechanical problems of drawing long cables through pipes and providing high-voltage insulation between the cable and the pipe.

An alternative to the coaxial line is the parallel-wire transmission line. The characteristic impedance of a parallel-wire line with unequal diameters as illustrated in Figure 5 is:<sup>3</sup>

$$Z_o = \frac{120}{(\epsilon_r)^{1/2}} \cosh^{-1} \left( \frac{4D^2 - d_1^2 - d_2^2}{2 d_1 d_2} \right) \quad (14)$$

where  $D$  is the wire spacing,  $d_1$  and  $d_2$  are the wire diameters, and  $\epsilon_r$  is the dielectric constant of the insulating medium. When allowance is made for high-voltage insulation, it is difficult with a single driving conductor to obtain characteristic impedances of less than 100 ohms. It is possible to reduce this impedance by nearly 50 percent, however, by using two conductors in parallel (as illustrated by the second conductor indicated by the dashed lines in Figure 5) to drive the test cable. This arrangement

also produces a more uniform distribution of the current in the outer shield of the test cable and reduces magnetic coupling to the core of the cable.

It is fairly easy to construct a long, low-impedance uniform test line using the parallel-wire configuration as illustrated in Figure 6.

High-voltage lines are used to drive the test cable, and the test cable is used as the low-voltage return for

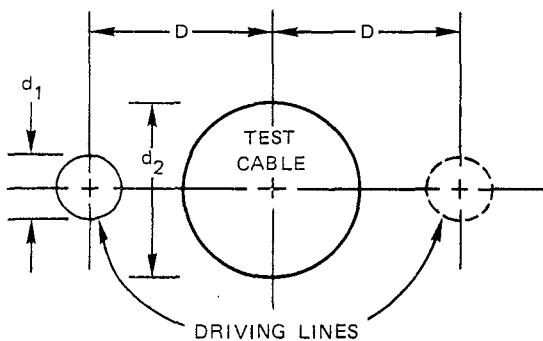


FIGURE 5 PARALLEL-WIRE TRANSMISSION LINE WITH UNEQUAL WIRE DIAMETERS

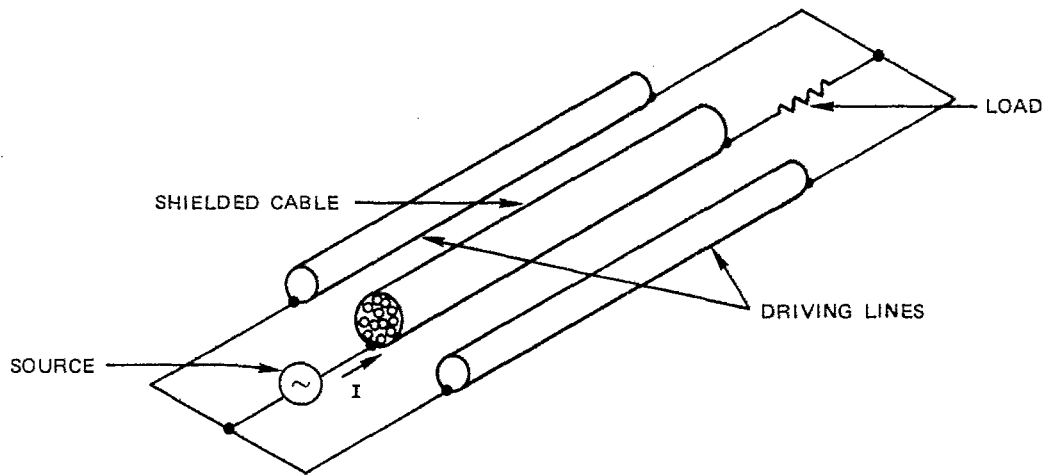


FIGURE 6 PARALLEL DRIVING LINES AS A COUPLING STRUCTURE FOR CURRENT INJECTION

the parallel-wire line. If necessary, the test cable can be insulated against the soil with wood or similar low-cost, low-voltage insulation to prevent the return current from flowing into the soil. Since the high voltage is confined to the well-insulated driving lines, this approach also has advantages in terms of the safety of the operating personnel.

The maximum current that can be produced in a semi-infinite parallel wire transmission line is even more severely limited than the current for a coaxial configuration because the electric field is concentrated between the driving lines and the cable shield. Both the characteristic impedance and the azimuthal uniformity of the field can be improved, however, by increasing the number of driving lines.

A variation of the parallel-wire driving structure is useful for driving shielded cables with insulating jackets that are routed along metal structure or laid in metal cable trays as illustrated in Figure 7. This configuration has the characteristics of a symmetrical two-wire transmission line, since the ground plane can be replaced by an image conductor to form a two-conductor line having twice the characteristic impedance of the conductor and ground plane. For the conductor of diameter  $d$ , whose center is a height  $h$  above the ground plane, the characteristic impedance is:

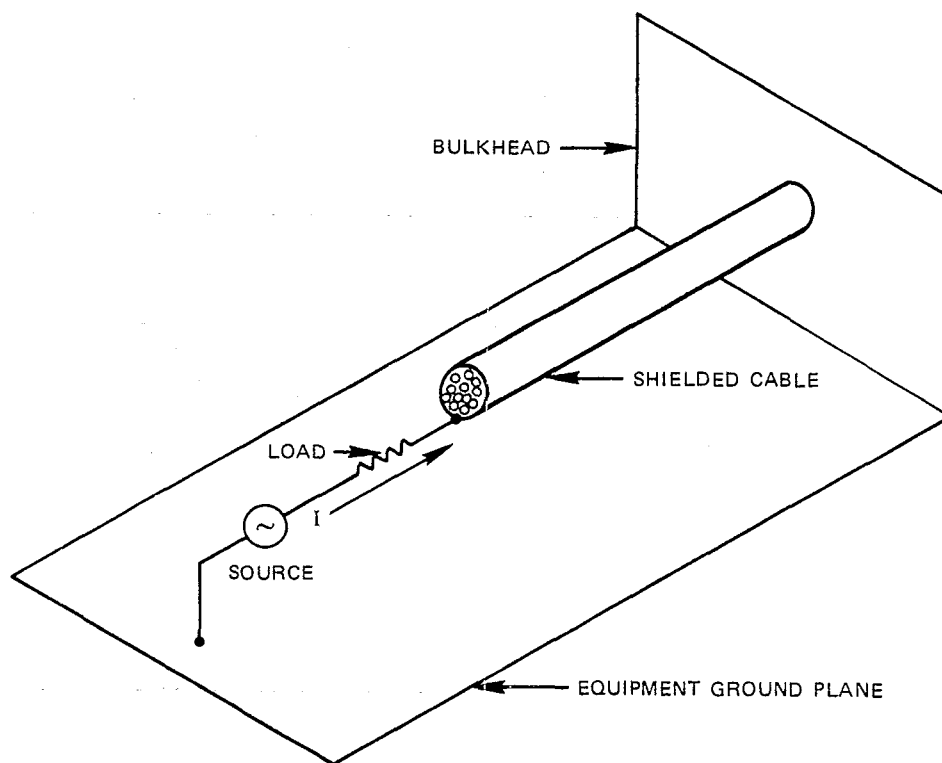


FIGURE 7 EQUIPMENT GROUND AS A COUPLING STRUCTURE FOR CURRENT INJECTION

$$Z_o = \frac{60}{(\epsilon_r)^{1/2}} \cosh^{-1} \left( \frac{2h}{d} \right) \quad (15)$$

This method of driving the shield is limited to applications where one end of the shield can be removed from the ground and connected to the energy source. In cases where simulation is required, it has the advantage that much of the pulse shaping is accomplished by the system structure, if the cable length and terminations are preserved.

If both ends of the cable shield are grounded to the structure in the system and it is desired to preserve this transmission line configuration so that the geometry of the system will shape the current waveform, then the current may be injected by means of a current transformer constructed as illustrated in Figure 8. The toroidal core can be split and clamped around the cable without disturbing the cable system. The

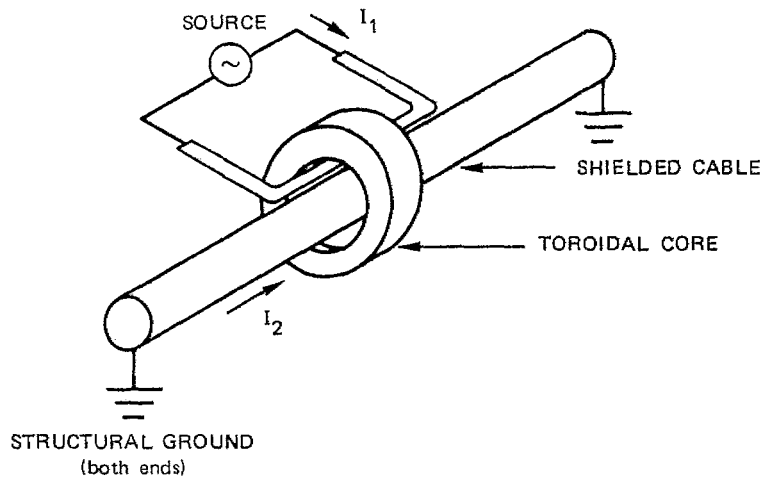


FIGURE 8 CURRENT TRANSFORMER TO INJECT CURRENT ON A GROUNDED CABLE SHIELD

equivalent circuit of an  $N$  turn primary and single-turn secondary current transformer referenced to the primary is illustrated in Figure 9 where  $L_1$  and  $C_1$  are the primary inductance and stray capacitance,  $R_c$  is the core loss resistance,  $L_c$  is the core leakage inductance, and  $L_2$  and  $C_2$  are the secondary inductance and stray capacitance. The core loss resistance is:

$$R_c = \frac{0.4\pi N^2 A}{\ell \times 10^8} \nu \quad (16)$$

where  $A$  is the cross section area in  $\text{cm}^2$  of the toroidal core material,  $\ell$  in  $\text{cm}$  is the mean magnetic path length in the toroidal core, and  $\nu$  is the core material loss resistance factor. The core loss resistance factor is a function of frequency, core saturation induction, and heat dissipation in the core. The core leakage inductance is:

$$L_c = \frac{0.4\pi N^2 A}{\ell \times 10^8} \mu_c \quad (17)$$

where  $\mu_c$  is the pulse permeability of the toroidal core material.



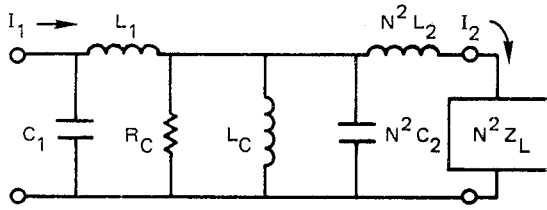


FIGURE 9 EQUIVALENT CIRCUIT OF A TOROIDAL CORE CURRENT TRANSFORMER

(with a characteristic impedance  $Z_0$ ) for at least a few ns. If the current transformer is installed in the middle of a cable 10 feet long, therefore, the initial load impedance will be approximately  $2Z_0$  for about 10 ns, as is shown in the early time equivalent circuit in Figure 10. This equivalent circuit neglects the primary and secondary inductances and is valid for pulse rise times greater than a few nanoseconds. In practice, the loss resistance  $R_c$  is frequently smaller than the load impedance, and thus the current transfer efficiency is low (5 to 50% is typical). Loss resistance can be increased by using higher  $\nu$  core material; but this approach suffers from the typically inverse relationship between  $\nu$  and frequency and may adversely affect the rise time characteristics.

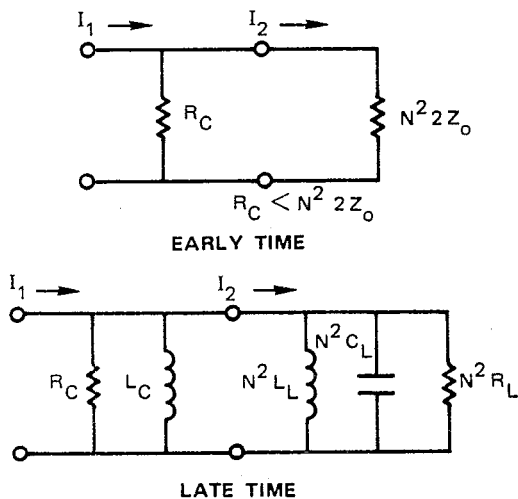


FIGURE 10 EARLY AND LATE TIME EQUIVALENT CIRCUITS

In a typical toroidal core current-injection application, the primary and secondary windings consist of one turn each so that  $L_1$ ,  $C_1$ ,  $L_2$ , and  $C_2$  are all small compared to  $R_c$  and  $L_c$  which are characteristic of the toroidal core material. The  $Z_L$  must be determined from the cable geometry and terminations. Frequently, the cable may be presented as a uniform transmission line

Loss resistance can also be increased by increasing the cross-sectional area  $A$  while maintaining (or decreasing) the mean magnetic path length  $\ell$  (see Equation 16). This is readily accomplished by using multiple (or smaller) cores, and has only a second-order effect on the rise time due to the increase in  $L_1$ .

For late times, the load impedance will consist of lumped inductance, capacitance, and resistance parameters as is shown in the late time equivalent circuit in Figure 10. The late time current in the cable will have a response

that is predominately determined by the load characteristics. The core loss,  $R_c$  and  $L_c$ , determine the current efficiency as in the early time equivalent circuit and the loss inductance will slightly modify the cable response.

### 2.3 NATURAL AND UNMATCHED CABLE TERMINATIONS

In most of the cases described above, it is implied that the transmission line used for injecting current into the cable shield should be terminated with its characteristic impedance to prevent the current from reflecting at the end of the driven structure. In the discussion of the technique illustrated in Figure 8, however, it was pointed out that the designer could take advantage of the natural structure to shape the injected pulse since this approach would greatly simplify the design of the driving pulse source required to simulate a free-field-induced current in the cable shield.

The use of unmatched terminations for the driving structure, particularly short-circuits, has other advantages that may contribute to the simplicity and economy of the direct-injection test. As was pointed out in Equation (6) the maximum current that can be injected on a transmission line of a given size that is terminated in a matched load is limited to a fairly moderate level by the dielectric strength of the insulation. However, if the transmission line is terminated in a short circuit and if the current rise time is considerably longer than the round-trip transit time between the driving source and the short-circuit, much larger currents can be produced without approaching dielectric breakdown voltages. Moreover, high-current, short-circuit terminations are very easy to produce and are very inexpensive; high-current resistive loads are often expensive and usually must be designed with considerable skill to avoid high-voltage flashover problems without compromising high-frequency impedance characteristics.

Because of these advantages, short-circuit terminations are well suited to applications where large currents must be produced but where precise control of the current pulse shape is not critical — for example, for the injection of current on a very long cable that has a solid tubular shield. Because the cable is very long, the current that must be simulated may be very large, but, because it has a solid shield, only the low frequencies penetrate the shield so that only the low-frequency spectrum of the current pulse is of interest.

As an example of how to utilize the natural characteristics of the cable, consider a case such as that illustrated in Figure 2 in which the current is injected into the inner shield of a long, doubly-shielded cable and the outer shield is used as the current return path. The load between the shields at the end opposite the driving source is a short circuit. For this example, the energy source is a charged capacitor that is connected between the shields and discharged into the shield-to-shield transmission line. No deliberate shaping of the leading edge of the current pulse is used. The spectrum of the current pulse is:

$$I(\omega) = \frac{CV}{1 + j\omega CZ} \quad (18)$$

where  $Z$  is the input impedance of the shorted cable. For the shorted transmission line, the input impedance is:

$$Z = Z_0 \tanh \gamma \ell \quad (19)$$

where  $Z_0$  is the characteristic impedance of the coaxial shield-to-shield line,  $\gamma$  is the propagation factor for the line, and  $\ell$  is the line length between the input terminals and the short circuit. This impedance is plotted in Figure 11 for a 640 m-long cable with a 20-mil-thick copper outer shield and a 10-mil-thick steel ( $\mu_r = 620$ ) inner shield.<sup>4</sup> At frequencies below 1 kHz, the input impedance is essentially the dc resistance of the shields and the first impedance maximum occurs between 40 and 50 kHz.

A plot of the current spectrum in Equation (18) is shown in Figure 12 for the discharge into the line of a 2.5  $\mu$ F capacitor charged to 10 kilovolts. The spectrum is flat out to 1 kHz with a slight peak at 4 kHz. This peak results from the resonance of the driving capacitance and inductance of the shorted line. The higher order resonances are over 30 dB below the low-frequency magnitude and the high-frequency spectrum approaches a 1/f falloff such as would be expected from a capacitor discharging into a resistance. (In this case, the resistance is the high-frequency characteristic impedance of the line.) The wave form of the current in the inner shield is shown in Figure 13. The initial current is the capacitor voltage divided by the high-frequency characteristic impedance as the line begins to become charged. The current in the shields decreases rapidly during the first 10  $\mu$ s with a time constant

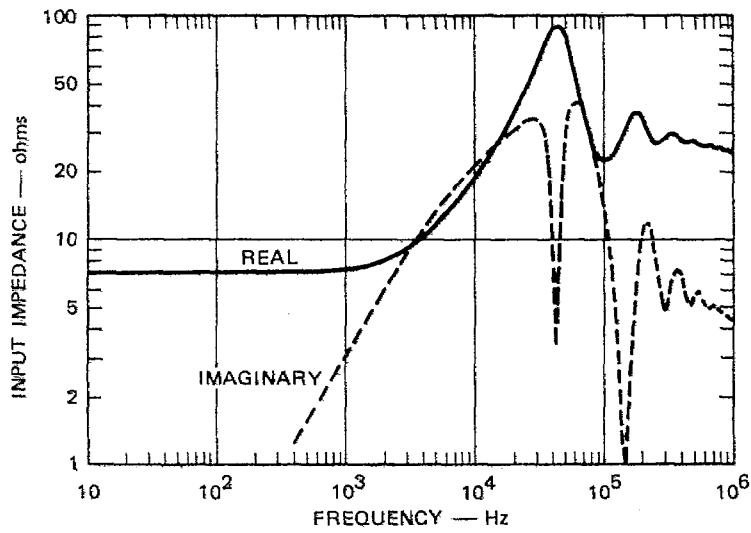


FIGURE 11 INPUT IMPEDANCE OF A SHORTED SHIELD-TO-SHIELD TRANSMISSION LINE

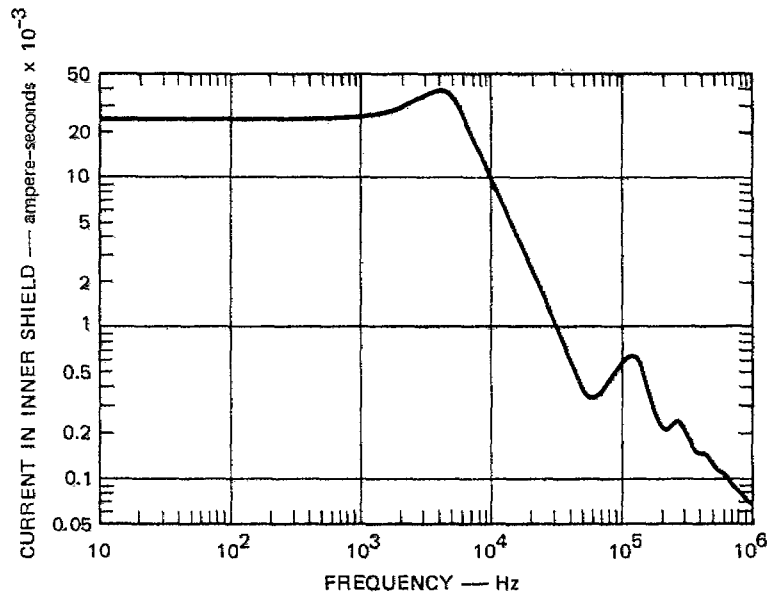


FIGURE 12 CURRENT SPECTRUM FOR A SHORTED SHIELD-TO-SHIELD TRANSMISSION LINE

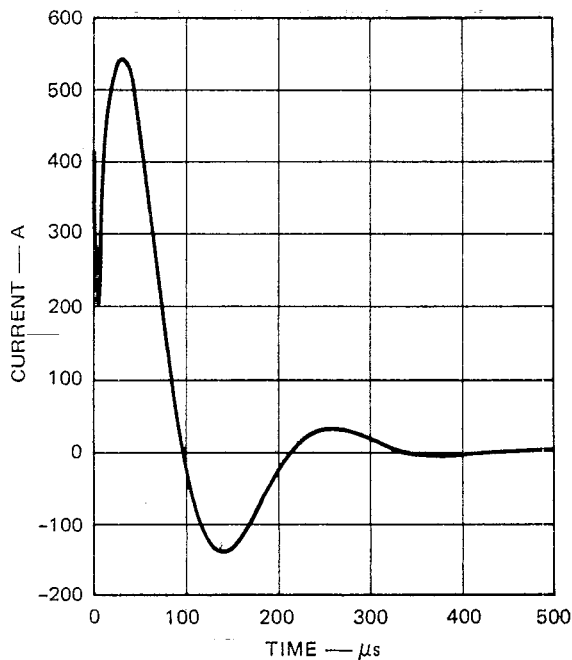


FIGURE 13 COMPUTED INNER SHIELD CURRENT FOR A SHORTED SHIELD-TO-SHIELD TRANSMISSION LINE

impedance of the reflected short circuit. Consequently, it may be desirable to increase the rise time of the current pulse so that it is greater than the round-trip transit time of the transmission line.

The use of the unshaped pulse to drive the line was postulated on the premise that the inner shield would filter out the high-frequency components of the pulse. This approach is illustrated by the spectra of the core-to-shield voltage shown in Figure 14, in which the core-to-shield voltage spectra are plotted at various distances from the driving point. Note that the first and second resonances of the driver current are suppressed and that the higher order resonances are more than 100 dB below the low-frequency magnitude of the spectra. Thus, only the low-frequency spectrum of the driving current (see Figure 12) penetrates the inner shield, and only this portion of the current spectrum is of interest for test purposes. However, in spite of the fact that the transmission line is terminated in a short-circuit, the current spectrum shown in Figure 12 does not display high-Q resonances because the line is fairly long and

$CZ_0$  as the pulse propagates down the line. After a few microseconds, however, the reflection from the shorted end reaches the driver and the current begins to increase again. Thereafter, there is a damped oscillation of the driving capacitance with the inductance of the shorted shields. Superimposed on the slow oscillation is a low-amplitude oscillation at the second resonance (see Figure 12), but this condition is barely detectable on the leading edge of the slow oscillation.

The leading edge of the current waveform may contribute to excessive voltage between the shields, since this current is flowing through the characteristic impedance of the transmission line rather than through the low

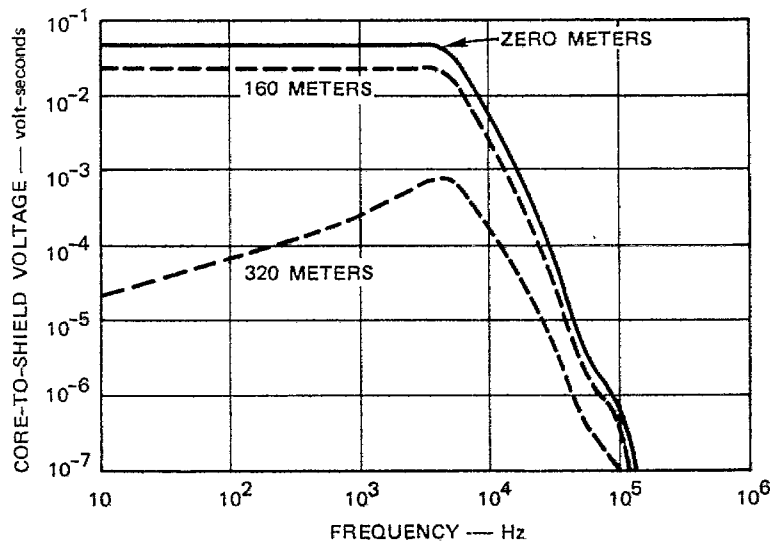


FIGURE 14 CORE-TO-SHIELD VOLTAGE SPECTRA FOR A SHIELD-TO-SHIELD DRIVE

is made of materials having finite conductivity. The propagation losses therefore are large enough (particularly at high frequencies) that the natural resonances of the line are quickly damped out as indicated by the current waveform shown in Figure 13. For longer cables, damping will be even greater.

### 3. DIRECT INJECTION ON SIGNAL-CARRYING CONDUCTORS

The injection of the test signal into signal-carrying conductors, such as the core conductors of a shielded cable or the conductors of an unshielded cable, usually requires a more carefully designed experiment than the injection of test signals into cable shields. Most tests in which the cable shield is driven utilize loose coupling between the driving source and the signal-carrying conductors because the shield usually provides more than 20 dB of isolation. Because of this loose coupling, the system impedances that affect the responses of the signal-carrying conductors are not significantly affected by the driving system and the responses essentially occur from natural excitation of the cable shield. When the test signal is injected directly into the signal-carrying conductors, however, the driving source and coupling-system impedances may alter the system

response. Thus, additional effort may be required to evaluate the effect of these differences — that is, to determine the system response had the effects of the injection system not been present or to place bounds on the possible effects of the injection system.

Direct injection of test signals on signal-carrying conductors of shielded cables also requires a more comprehensive understanding of the interaction of the system with the electromagnetic pulse, inasmuch as the designer must be able to determine (either through analysis or experiment) the EMP signal that couples to and penetrates through the cable shield. For unshielded cables, antennas, and power lines directly exposed to electromagnetic radiation, however, the problem of specifying the pulse shape may be no more difficult than that of specifying the pulse shape for shielded cables. The quality of simulation required (or the analysis needed to justify the quality of simulation used) may also be considerably greater for direct injection into signal-carrying conductors because all pulse shaping must be accomplished by the driving source and coupling network. The designer cannot take advantage of the pulse-shaping and loose coupling characteristics of the shield as is the case when testing with injection of currents on cable shields.

For direct injection into signal-carrying conductors, there are usually many more individual signal-carrying conductors than shielded cables; hence, there are more parameters to simulate. One common approach to simulating voltage injection into unshielded cable bundles is to induce (by some coupling means discussed later) currents that simulate the common-mode current that was measured in a low-level test of the complete system and hope that the currents will assume the proper distribution in the individual wires in the bundle. Although this approach is useful, particularly in the early stages of the system test, the results can be misleading. Interconnecting cables between equipment cabinets within a shielded enclosure are often excited by currents conducted along a cable or along some of the wires in the cable rather than by coupling to ambient EM fields. Consequently, the current in the individual wires is not necessarily determined by specifying the total current in the bundle and the individual wire terminations. For example, assume that a bundle with many conductors contains some conductors that are connected to circuits outside a shielded enclosure as well as to some that are interconnecting conductors for circuits totally within a shielded enclosure. Thus, the interconnecting bundle contains some conductors that are tightly coupled to the

conductors in the external bundle. In such a case, it is probable that the excitation of the interconnecting bundle will be dominated by the signal on those wires from outside the shield and that the signal on the remainder of the conductors in the interconnecting bundle will result primarily from mutual coupling among the conductors. A bulk common-mode current injected into the interconnecting wire-bundle, therefore, will not produce the same system response in the equipment as the bulk current produced by the EMP environment.

In the case described above, the bulk-current injection would produce a much more effective simulation if it were made on the wire bundle entering from the external equipment. If it were not possible to inject the test pulse into the external conductors (for example, only the interconnecting bundle was easily accessible), each conductor in the interconnecting bundle might have to be driven individually. Usually, it is possible to rank the importance of the individual wires in terms of the susceptibility of the circuits served by the wire so that precise simulation is not necessary for every wire in the bundle.

### 3.1 INJECTION OF COMMON-MODE VOLTAGES

For the case where one end of a cable is accessible for injection of test signals, a common-mode test voltage can be directly injected at the end through an impedance matrix as illustrated in Figure 15. The impedance matrix may simulate the impedances usually connected between the conductors and between the conductors and the cable shield or system ground. In the case of a very long cable, the voltage delivered to the conductors is:

$$V_{in} = \frac{Z_{in}}{Z + Z_{in}} V \quad (20)$$

where  $V$  is the source voltage,  $Z_{in}$  is the input impedance of the cable with its normal load on the right-hand side (see Figure 15), and  $Z$  is the common mode characteristic impedance of the terminating resistors between the source and the cable. It should be noted that, if the right-hand end of the cable is also terminated in its characteristic



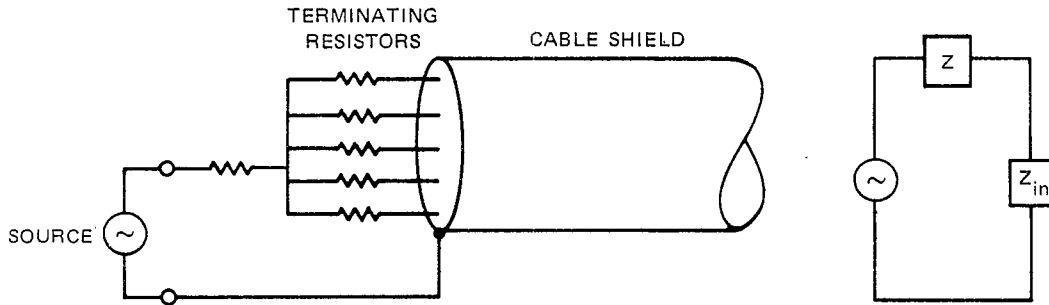


FIGURE 15 IMPEDANCE MATRIX USED TO INJECT COMMON-MODE VOLTAGES

impedance, only half of the source voltage is applied to the cable.

This method of driving cable conductors is perhaps the most straightforward and commonly used of all the direct injection methods. It can also be used with unshielded cables that are routed along a metal structure (such as an aircraft wing) or are placed in metal cable trays. With unshielded cables of this type, the current is driven against the metal structure or trays rather than against the shield. One disadvantage of this method is that the cable being driven must be disconnected at one end; hence, the system may not be operating in its normal state during the test.

When one end of the cable is not accessible, another injection method must be used. Such cases arise where disconnecting the cable precludes operating the system in its normal mode — for example, disconnecting the main power leads to inject a signal on these leads precludes operating the system from power supplied through the leads. In these cases, it may be necessary to accept some compromise in the quality of the simulation to perform tests economically. One approach that can be used under certain conditions is illustrated in Figure 16. At some suitable junction in the cable system, the energy source is capacitively coupled to the conductors and permitted to drive them with respect to the local ground or chassis. As illustrated in the figure, however, the current injected at this point is divided into two parts, one flowing in each direction from the injection point. Because this method of distributing the current differs radically from the current distribution that would have resulted from the EMP excitation of the system, some care is required in designing a valid test using this approach.

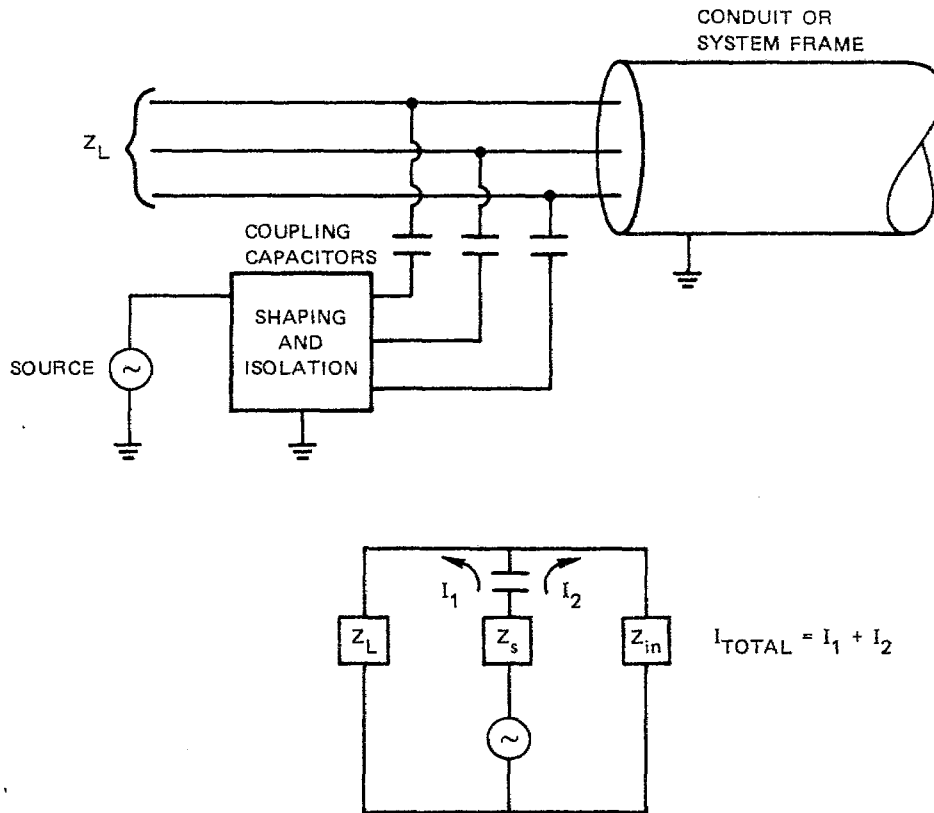


FIGURE 16 CAPACITORS USED TO INJECT COMMON-MODE VOLTAGES

The validity of a test using the injection method shown in Figure 16 depends on not significantly affecting system response by the attachment of the energy source or by injection of unwanted currents. To clarify the latter point, assume that the portion of the system to the right of the injection point in the figure is of primary interest in the test. Thus, for the test to be valid, the portion of the current that flows to the left from the injection point must not be reflected back into the right-hand circuitry during the period when the system is being observed to determine its response to the EMP. This condition implies that no significant reflections should return from the left end of the circuit, which in turn implies that this circuit be a very long (around-trip transit time that is longer than viewing time) or very short (around-trip transit time that is shorter than any response of interest) or that it be terminated in a matched load (no reflections).

The stipulation that attachment of the energy source should not significantly affect the system response usually implies that the coupling between the direct-injection source and the system conductors be loose so that the system impedances are not significantly affected. This stipulation can also be met by having a sufficient length of uniform cable between the injection point and the system test point so that reflections from the injection point do not return to the test point during the desired viewing time. It is important to observe that the loose coupling requirement applies to the differential-mode impedances as well as to the common-mode, or line-to-ground, impedances — that is, the attachment of the energy source should, in general, disturb neither the line-to-ground impedances nor the line-to-line impedances of the cable at the injection point. The loose-coupling requirement usually requires that the source voltage be much larger than the voltage that is to be injected, since most of the source voltage must be dissipated in the coupling network if loose coupling is to be achieved.

#### 4. EXAMPLES OF DIRECT-INJECTION SYSTEMS

The first direct-injection system considered here injects the current on the inner shield of a long, double shielded cable. The second system injects a pulse directly into the power conductors at the service entrance where commercial power enters the facility under test. The third system injects a common-mode voltage on the multiple conductors of an intra-system unshielded cable. In all of these systems, compromises are made to facilitate the physical requirements of the tests or to make the tests economically feasible.

##### 4.1 SHIELDED CABLE DRIVER

This shielded cable driving system is to inject current into the inner shield of a long double-shielded, buried cable. The cable consists of a core of twisted pairs covered by an insulating jacket. The inner shield (solid 10-mil-thick steel) and outer shield (solid 20-mil-thick copper) are insulated from each other by a polyethylene jacket between the shields. The outer jacket over the copper shield is a conducting material

that permits the outer shield to be in electrical contact with the soil when the cable is buried. The test requirement is to drive the cable shields in such a manner that large, naturally induced, common-mode, core-to-shield voltages will be simulated.

Because the cable shield system is fairly effective (solid shields), the desired core-to-shield voltages would require that huge shield currents be produced. Moreover, because the outer shield is in electrical contact with the soil, the injection of large currents in the outer shield requires the injection of large currents into the surrounding soil, which produce serious personnel hazards and operating problems, as well as greatly increase in the size and complexity of the simulation system. To avoid these problems, the current is injected on the inner shield of the cable and the outer shield is used as the return path for this current (in the manner illustrated in Figure 2).

The driving scheme consists of a charged capacitor, a switch, and a long coaxial transmission line formed by the inner and outer shields of the cable. Because about 95 percent of the induced EMP cable current is carried by the outer shield, the current in the inner shield that is required to produce the desired core-to-shield voltage is much smaller than the bulk cable current being simulated. Moreover, because the coaxial cable geometry is amenable to accurate analysis, the relationship between the bulk cable current and the inner shield current can be established with confidence. Furthermore, the solid inner shield permits only the low-frequency portion of the pulse spectrum to penetrate to the cable core, so that the rise time of the leading edge of the injected-current pulse is not critical. Moreover, the length of the cable between driving point and short circuit termination can be utilized to simplify the test system. Because the shield-to-shield transmission line has loss, for a long cable, the current will not oscillate significantly at its natural frequency. Thus, the problem of designing and installing high-current terminating resistors is eliminated. Finally, since most of the EM energy is confined to the region between the cable shields, personnel hazards are limited to the energy source.

The driver for this injection system consists of a 3  $\mu$ F, 50 kV capacitor bank charged from a 50 kV dc power supply and switched onto the cable with a 40 kV ignitron. The complete system includes a power supply, controls, capacitors, and a trigger circuit for the ignitron switch. The essential components of the driver and the end of the cable are illustrated in Figure 17. At a point several thousand feet along

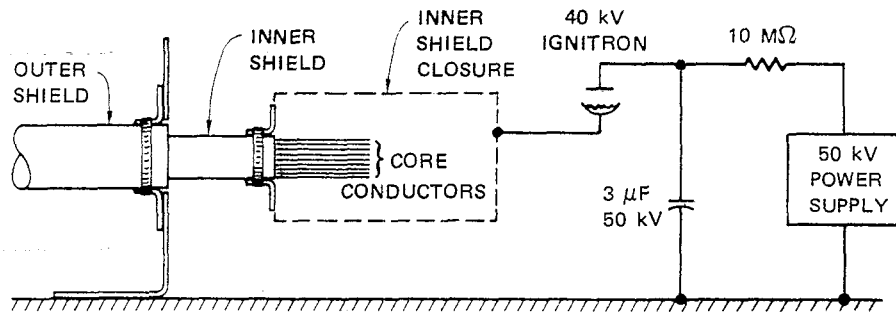


FIGURE 17 DIRECT-INJECTION PULSER FOR DRIVING INNER SHIELD OF DOUBLE-SHIELDED CABLE

the cable from the driving point, the inner and outer shields are connected by a shorting strap located in a cable splice case.

Because it is desirable to drive the core wires through the inner shield, allowing it to shape the pulse developed on the wires, the inner shield has to be well shielded at the driving end of the cable to prevent direct coupling to the core wires at this point. Thus, the inner shield is extended with a metal cylinder as illustrated in Figure 17 by the inner shield closure (indicated by the dashed line) to completely enclose the core conductors. For this length of cable, the system circuits several thousand feet away are insensitive to the impedance between the core conductors and the shield. Therefore, the core conductors are left open as undriven conductors. The core wires are driven by the inner shield of the cable as is desired to simulate the EMP excitation.

In spite of system constraints, such as the outer shield being in contact with the soil and the difficulty in obtaining any termination between the shields other than a short circuit, it is possible to excite the core conductors in a manner closely approximating the natural response excitation with relatively unsophisticated driving and coupling techniques.

#### 4.2 POWER LINE DRIVER

For this example the test pulse is to be injected directly into the signal-carrying

conductors — in this case, the power conductors supplying commercial power to a facility. The pulser injects its output on the power lines between the conductors and the conduit at the service entrance weatherhead. The characteristic impedance of the conductors entering the weatherhead is approximately 40 ohms. A 50-kV transient injected at this point is equivalent to a 500-kV pulse on the power lines driving the weatherhead, since the impedance mismatch between aboveground power lines and the weatherhead conduit permits only about 10 percent of the incident voltage to be transmitted to the conduit circuit.

The design criteria for this power line driver are as follows:

|                     |                   |
|---------------------|-------------------|
| Peak voltage        | $\geq 50$ kV      |
| Rise time           | $\leq 10$ ns      |
| Pulse shape         | exponential decay |
| Decay time constant | $\geq 500$ ns     |
| Repetition rate     | $\geq 1$ pps      |

The configuration selected is indicated schematically in Figure 18. The driver is constructed with a pulse unit that is located away from the power line and a coupler unit that is located at the commercial power weatherhead. Figure 19 shows the coupler unit attached at the weatherhead. The pulser unit output is delivered to the coupler unit through four 50-ohm coaxial cables. These cables are connected in

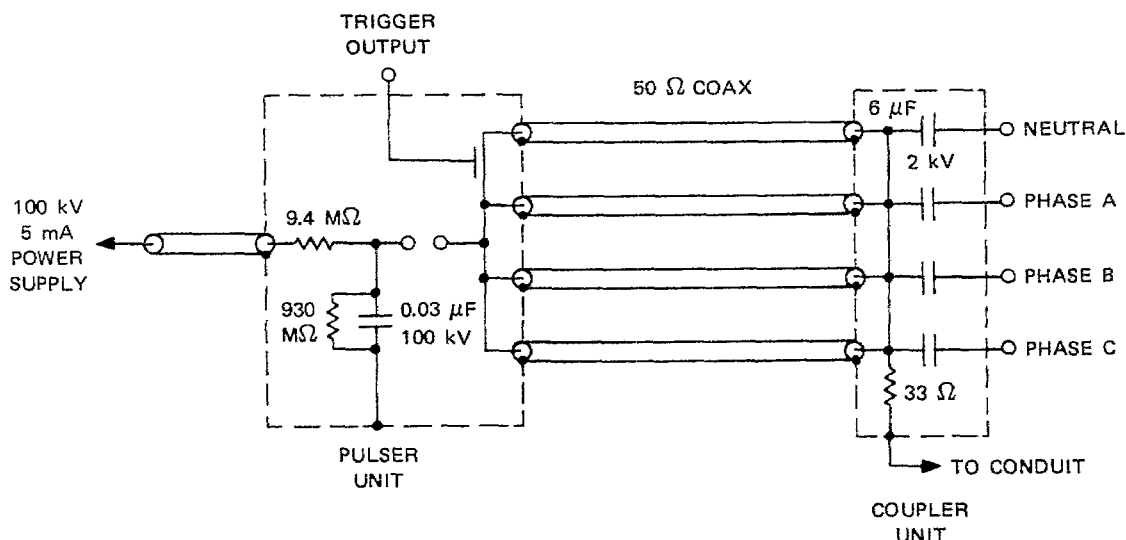


FIGURE 18 SCHEMATIC DIAGRAM OF POWER LINE PULSER AND COUPLER

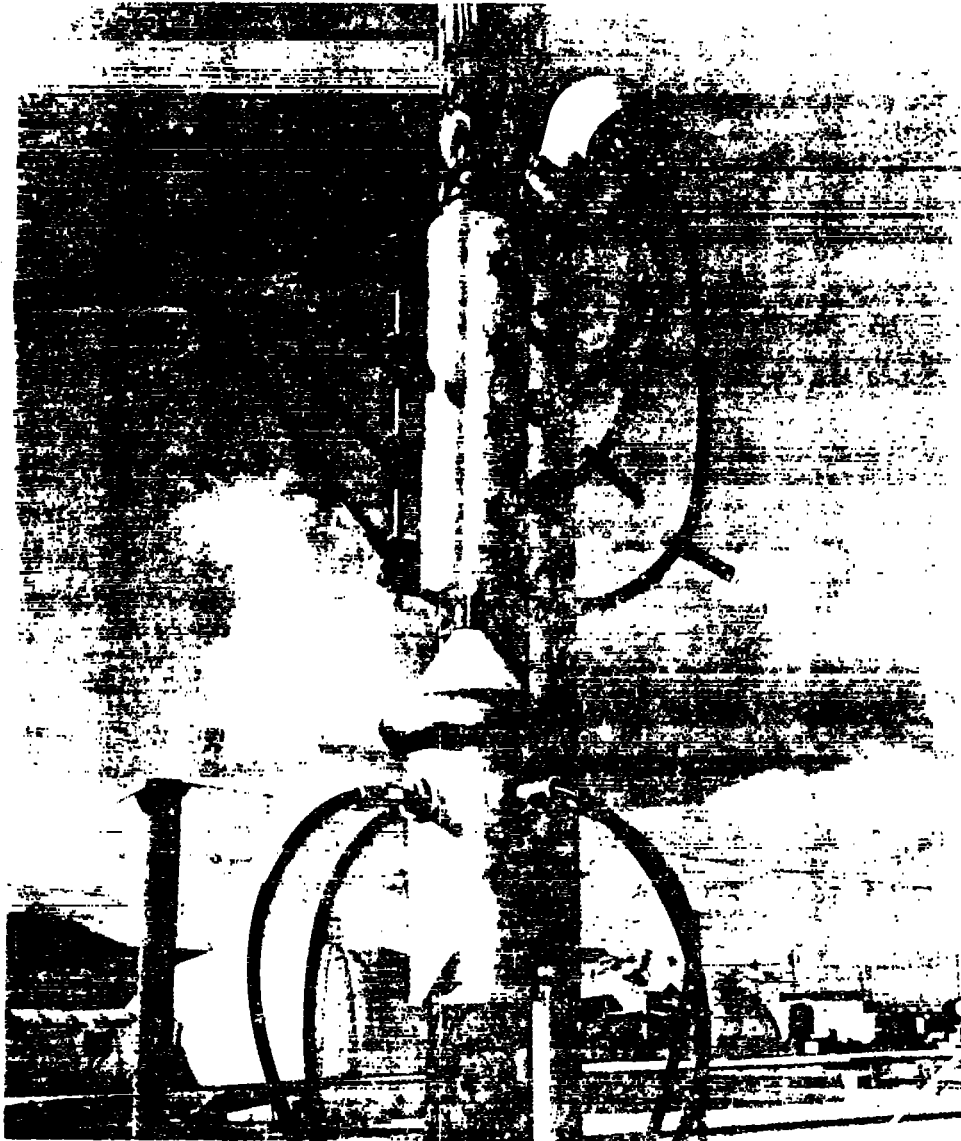


FIGURE 19 POWER-LINE COUPLER UNIT

parallel at the coupler unit so that a single cable could be used. However, for pulser unit flexibility, separate cables are provided for each power line conductor. (Variation in the number and impedance of these cables can be used to vary pulser characteristics.)

To provide the desired pulse shape at the input to the 40-ohm weatherhead conduit circuit, a 50-kV low-inductance capacitor is required whose capacitance is:

$$C = \frac{\tau}{Z_o} = \frac{500 \times 10^{-9}}{40} = 0.0125 \mu\text{F} \quad (21)$$

Because the conduit circuit is not terminated in its characteristic impedance, however, its input impedance is not 40 ohms at all times. It is therefore necessary to use a somewhat larger capacitance and to load the pulser in the coupling unit to make the pulse shape somewhat less dependent on the characteristics of the conduit circuit. Therefore, a total capacitance of 0.03  $\mu\text{F}$  was selected in the form of a series stack of low-inductance pancake capacitors mounted in a concentric cylindrical housing. This capacitor bank is charged through a series resistance as illustrated in Figure 18 and discharged into the load (through the output cables and coupler) when a spark-gap switch is overvolted.

The coupler permits the output of the pulser to be delivered to the power lines without altering the functioning of the 60-Hz power system. An additional function of the coupler is to load the pulser so that the shape of the delivered pulse is not determined solely by the commercial power system.

Coupling capacitors are used to allow the fast pulse delivered by the pulser to flow onto the power lines, yet prevent excessive 60-Hz current from flowing into the pulser. The four 6- $\mu\text{F}$  capacitors operate in parallel to deliver the pulse equally to all four conductors of the power system. The common mode impedance of the four conductors (with respect to the conduit) is approximately 40 ohms when reflections from the power system loads are neglected. Thus, the capacitors will pass that part of the pulse spectrum above 200 Hz with little distortion. At 60 Hz, however, the reactance of the capacitors is 440 ohms, so that the 60-Hz current flowing through the phase capacitors is only 0.27 ampere. Since the coupling capacitors are approximately 1,000 times larger than the storage capacitance in the pulser, only about 1/1000th of the source voltage will appear across the coupling capacitors. For this reason, relatively low-voltage capacitors can be used for this function. The capacitors used are 6- $\mu\text{F}$ , 2 kV dc pulse capacitors.

Figure 18 also shows a 33-ohm load resistor in the coupler that is in parallel with the power-line, common-mode impedance. This resistor, while not essential to



injecting the pulse on the power line, is convenient because it permits the coaxial cables to be terminated so that the pulser can be operated with no load (other than the coupler) to measure open-circuit voltage at the coupler output terminals. Also, without the dc current path provided by the resistor, the cables would charge to the bank voltage during the first pulse and the spark-gap switch would not fire on subsequent cycles. The 33-ohm resistor is in parallel with the common mode impedance in the coupler unit to provide an approximate terminating impedance for the four coaxial cables. This division of the loads provides pulser operation that is not greatly affected by reflections from the power system.

Because of the inductance of the leads between the coupling capacitors and the weatherhead, the rise time of the pulse actually delivered to the power conductors may be somewhat greater than the 10 ns achieved for the pulser and coupler. However, there is enough additional stray inductance in the normal power system wiring so that the degradation of the pulse rise time is of little consequence. Thus, in spite of some apparent shortcomings of this direct injection system, tests conducted with this system can be valid and extremely informative.

#### 4.3 COMMON-MODE VOLTAGE DRIVER

The third example of a direct injection system is one in which a common-mode voltage is injected onto the conductors of an unshielded cable that connects an operating console with remote equipment cabinets. To drive the conductors in such a manner that the system response to the conductor-coupled signals is simulated, it is necessary to produce bulk currents similar to those induced by the EMP field, whether the cable is shielded or not. When the cable is not shielded, however, the problem of coupling a driving source to the cable is more difficult, because tight coupling to the cable may alter the cable response and invalidate the test. Conversely, loose coupling may not produce a large enough cable excitation to be useful. The approach used here is to capacitively couple the signal to the unshielded cable by passing the cable conductors through a large conductive pipe that is excited by a capacitor discharge. This arrangement is illustrated in Figure 20.

a section of transmission line that will oscillate unless properly terminated, the resistor  $R_T$  at the end of the pipe opposite the driving capacitor is used to terminate the pipe.

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