

## **SPICE MODELS FOR RADIATED AND CONDUCTED SUSCEPTIBILITY ANALYSES OF MULTICONDUCTOR SHIELDED CABLES**

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**Abstract**—This paper presents SPICE models to analyze the radiated and conducted susceptibilities of multiconductor shielded cables in the time and frequency domains. These models, which can be used directly in the time and frequency domains, take into account the presence of both the transfer impedance and admittance, and allow the transient analysis when the termination is nonlinear or time-varying. The radiated and conducted susceptibilities are studied by using an incident plane-wave electromagnetic field and an injection current on the cable shield as the source, respectively. Results obtained by these models are in good agreement with those obtained by other methods.

### **1. INTRODUCTION**

Shielded cables are usually used in wired communication systems to protect signal transmission from external interference. Multiconductor shielded cables are more appealing compared with coaxial shielded cables due to the capability to transmit differential signals. Nevertheless, there exists coupling between the exterior and interior of the shield due to the imperfect nature of the shield. Consequently, prediction of the susceptibility of multiconductor shielded cables will be helpful for the optimization of system design.

SPICE equivalent circuit models for multiconductor transmission lines without shields have drawn much attention, and lots of researches have done on them and their applications [1–9]. Recently, some interest has been addressed to the development of SPICE models for the susceptibility analysis of shielded cables. SPICE models to analyze the

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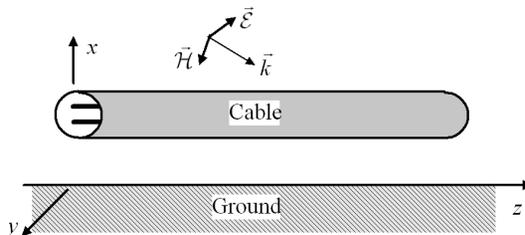
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conducted immunity of both lossless and lossy coaxial cables have been presented in [10], and the inverse Fourier transform (IFT) is needed to get the time domain results. Then, some SPICE models have been proposed for the time and frequency domain analyses of bulk current injection test on lossless shielded cables [11, 12]. However, the radiated immunity can not be directly analyzed with these models. Recently, some circuit models have been developed to analyze the conducted and radiated immunity of shielded cables [13], and the IFT is required for the models to obtain the time domain results. After then, some lossless models have been proposed to analyze the radiated and conducted susceptibilities of coaxial shielded cables [14]. These models can be employed directly in the time and frequency domains and take into account both the transfer impedance and admittance of the shield.

This paper presents compact SPICE models for the radiated and conducted susceptibility analyses of multiconductor shielded cables. These models are derived from the transmission-line equations of multiconductor shielded cables. They do not need the discretization of shielded cables and allow the transient analysis when there are nonlinear or time-varying loads at the termination. Multiconductor shielded cables over a ground excited by an incident plane-wave electromagnetic field or a lumped current source on the shield are studied with the proposed models. The differential and common mode voltages obtained by these models agree well with those from other methods.

## 2. DEVELOPMENT OF SPICE MODELS FOR MULTICONDUCTOR SHIELDED CABLES

A shielded cable over an infinite and perfectly conducting ground, as shown in Figure 1, can be considered as two transmission line systems, when its cross section and height over the ground are electrically small



**Figure 1.** A multiconductor shielded cable over an infinite and perfectly conducting ground.

compared with the wavelength [15, 16]. The exterior of the shield and the ground compose the outer system, while the interior of the shield and the inner wires form the inner system. These two systems are coupled through the transfer impedance and admittance of the shield.

### 2.1. SPICE Models for Radiated Susceptibility Analysis

A multiconductor shielded cable with  $n$  parallel wires located inside the shield above the ground, excited by an incident plane-wave electromagnetic field, can be described by [13]

$$\frac{d}{dz} \begin{bmatrix} \hat{V}_{out} \\ \hat{\mathbf{V}}_{in} \end{bmatrix} + \begin{bmatrix} \hat{Z}_{out} & -\hat{\mathbf{Z}}_t^T \\ -\hat{\mathbf{Z}}_t & \hat{\mathbf{Z}}_{in} \end{bmatrix} \begin{bmatrix} \hat{I}_{out} \\ \hat{\mathbf{I}}_{in} \end{bmatrix} = \begin{bmatrix} \hat{V}_{sout} \\ \mathbf{0} \end{bmatrix} \quad (1a)$$

$$\frac{d}{dz} \begin{bmatrix} \hat{I}_{out} \\ \hat{\mathbf{I}}_{in} \end{bmatrix} + \begin{bmatrix} \hat{Y}_{out} & \hat{\mathbf{Y}}_t^T \\ \hat{\mathbf{Y}}_t & \hat{\mathbf{Y}}_{in} \end{bmatrix} \begin{bmatrix} \hat{V}_{out} \\ \hat{\mathbf{V}}_{in} \end{bmatrix} = \begin{bmatrix} \hat{I}_{sout} \\ \mathbf{0} \end{bmatrix}, \quad (1b)$$

where  $\hat{V}_{out}$  is the shield-to-ground voltage,  $\hat{I}_{out}$  is the current flowing between the external shield and the ground,  $\hat{\mathbf{V}}_{in} = [\hat{V}_{in1}, \hat{V}_{in2}, \dots, \hat{V}_{inn}]^T$  is the inner voltage vector whose  $k$ th element  $\hat{V}_{ink}$  is the voltage of the  $k$ th wire versus the internal part of the shield, and  $\hat{\mathbf{I}}_{in} = [\hat{I}_{in1}, \hat{I}_{in2}, \dots, \hat{I}_{inn}]^T$  is the inner current vector whose  $k$ th element  $\hat{I}_{ink}$  is the current of the  $k$ th wire.  $\hat{Z}_{out}$  and  $\hat{Y}_{out}$  are the per-unit-length (p.u.l.) impedance and admittance of the outer system, respectively.  $\hat{\mathbf{Z}}_t = [\hat{Z}_{t1}, \dots, \hat{Z}_{tn}]^T$  is the transfer impedance vector whose  $k$ th element  $\hat{Z}_{tk}$  is the shield transfer impedance relative to the  $k$ th wire, while  $\hat{\mathbf{Y}}_t = [\hat{Y}_{t1}, \dots, \hat{Y}_{tn}]^T$  the transfer admittance vector of which  $k$ th element  $\hat{Y}_{tk}$  refers to the shield transfer admittance relative to the the  $k$ th wire. The submatrices  $\hat{\mathbf{Z}}_{in}$  and  $\hat{\mathbf{Y}}_{in}$  are the p.u.l. impedance and admittance matrices of the inner system, respectively.  $\hat{V}_{sout}$  and  $\hat{I}_{sout}$  are the distributed voltage and current sources of the outer system, respectively, generated by the incident electromagnetic field [17, 18].

For the sake of simplicity, some assumptions are made. These assumptions are: 1) the loss of the multiconductor shielded cable is neglected, which may be not reasonable when the cable is extremely long and the frequency is very high [19], but ensures the worst case condition; 2) the energy transferred from the inner system to the outer system through the shield is neglected, which implies the good shielding approximation; 3) the transfer impedance  $\hat{Z}_{tk}$  ( $k = 1, \dots, n$ ) is taken

as  $\hat{Z}_{tk}(\omega) = R_{dck} + j\omega L_{tk}$  where  $R_{dck}$  is the constant p.u.l. transfer resistance of the shield relative to the  $k$ th wire, which is greater than its true value but the worst case is considered [11], and  $L_{tk}$  is the p.u.l. transfer inductance of the shield relative to the  $k$ th wire. Under these assumptions, Equation (1) can be written as

$$\frac{d}{dz}\hat{V}_{out} + j\omega L_{out}\hat{I}_{out} = \hat{V}_{sout} \quad (2a)$$

$$\frac{d}{dz}\hat{I}_{out} + j\omega C_{out}\hat{V}_{out} = \hat{I}_{sout} \quad (2b)$$

$$\frac{d}{dz}\hat{\mathbf{V}}_{in} + j\omega \mathbf{L}_{in}\hat{\mathbf{I}}_{in} = \hat{\mathbf{Z}}_t\hat{I}_{out} \quad (2c)$$

$$\frac{d}{dz}\hat{\mathbf{I}}_{in} + j\omega \mathbf{C}_{in}\hat{\mathbf{V}}_{in} = -\hat{\mathbf{Y}}_t\hat{V}_{out}, \quad (2d)$$

where  $L_{out}$  and  $C_{out}$  are the p.u.l inductance and capacitance of the outer system, respectively, while  $\mathbf{L}_{in}$  and  $\mathbf{C}_{in}$  the p.u.l inductance and capacitance matrices of the inner system.

The responses of the outer loads can be obtained with some already developed SPICE models [1, 2], so the solutions to the response of the inner loads are the interest of this paper. Equations (2c) and (2d) can be decoupled with a similarity transformation. Define the transformation to mode quantities as

$$\hat{\mathbf{V}}_{in} = \mathbf{T}_V\hat{\mathbf{V}}_{inm} \quad (3a)$$

$$\hat{\mathbf{I}}_{in} = \mathbf{T}_I\hat{\mathbf{I}}_{inm}. \quad (3b)$$

Substituting (3) into (2c) and (2d) gives

$$\frac{d}{dz}\hat{\mathbf{V}}_{inm} + j\omega \mathbf{L}_{inm}\hat{\mathbf{I}}_{inm} = \hat{\mathbf{V}}_{Finm} \quad (4a)$$

$$\frac{d}{dz}\hat{\mathbf{I}}_{inm} + j\omega \mathbf{C}_{inm}\hat{\mathbf{V}}_{inm} = \hat{\mathbf{I}}_{Finm}, \quad (4b)$$

where

$$\mathbf{L}_{inm} = \mathbf{T}_V^{-1}\mathbf{L}_{in}\mathbf{T}_I \quad (5a)$$

$$\mathbf{C}_{inm} = \mathbf{T}_I^{-1}\mathbf{C}_{in}\mathbf{T}_V \quad (5b)$$

$$\hat{\mathbf{V}}_{Finm} = \mathbf{T}_V^{-1}\hat{\mathbf{Z}}_t\hat{I}_{out} \quad (5c)$$

$$\hat{\mathbf{I}}_{Finm} = -\mathbf{T}_I^{-1}\hat{\mathbf{Y}}_t\hat{V}_{out}. \quad (5d)$$

Both  $\mathbf{L}_{inm}$  and  $\mathbf{C}_{inm}$  are diagonal matrices, and the matrices  $\mathbf{T}_V$  and  $\mathbf{T}_I$  are frequency independent when the loss is neglected. From (4), the relation between the terminal mode voltages and currents can be

written as

$$\begin{aligned} & \hat{\mathbf{V}}_{inm}(0) - \mathbf{Z}_{Cinm} \hat{\mathbf{I}}_{inm}(0) \\ &= e^{-j\omega\Lambda\mathcal{L}} \left[ \hat{\mathbf{V}}_{inm}(\mathcal{L}) - \mathbf{Z}_{Cinm} \hat{\mathbf{I}}_{inm}(\mathcal{L}) \right] + \hat{\mathbf{V}}_{0inm} \end{aligned} \quad (6a)$$

$$\begin{aligned} & \hat{\mathbf{V}}_{inm}(\mathcal{L}) + \mathbf{Z}_{Cinm} \hat{\mathbf{I}}_{inm}(\mathcal{L}) \\ &= e^{-j\omega\Lambda\mathcal{L}} \left[ \hat{\mathbf{V}}_{inm}(0) + \mathbf{Z}_{Cinm} \hat{\mathbf{I}}_{inm}(0) \right] + \hat{\mathbf{V}}_{\mathcal{L}inm}, \end{aligned} \quad (6b)$$

where  $\mathcal{L}$  is the length of the cable,  $\mathbf{Z}_{Cinm}$  is the mode characteristic impedance diagonal matrix of the inner system, and  $\mathbf{\Lambda}$  is a diagonal matrix defined by  $\mathbf{\Lambda} = \sqrt{\mathbf{L}_{inm} \mathbf{C}_{inm}}$ .  $\hat{\mathbf{V}}_{0inm}$  and  $\hat{\mathbf{V}}_{\mathcal{L}inm}$  are the effect of the distributed source of the inner system, given by

$$\begin{aligned} \hat{\mathbf{V}}_{0inm} &= - \int_0^{\mathcal{L}} e^{-j\beta_{in}z} \\ & \left\{ \mathbf{R}_{dc}^{eq} I_{out}(z) + j\omega \left[ \mathbf{L}_t^{eq} I_{out}(z) + \mathbf{Z}_{Cinm} \mathbf{C}_t^{eq} V_{out}(z) \right] \right\} dz \end{aligned} \quad (7a)$$

$$\begin{aligned} \hat{\mathbf{V}}_{\mathcal{L}inm} &= \int_0^{\mathcal{L}} e^{-j\beta_{in}(\mathcal{L}-z)} \\ & \left\{ \mathbf{R}_{dc}^{eq} I_{out}(z) + j\omega \left[ \mathbf{L}_t^{eq} I_{out}(z) - \mathbf{Z}_{Cinm} \mathbf{C}_t^{eq} V_{out}(z) \right] \right\} dz, \end{aligned} \quad (7b)$$

where

$$\beta_{in} = \omega \mathbf{\Lambda} \quad (8a)$$

$$\mathbf{R}_{dc}^{eq} = \mathbf{T}_V^{-1} \mathbf{R}_{dc} \quad (8b)$$

$$\mathbf{L}_t^{eq} = \mathbf{T}_V^{-1} \mathbf{L}_t \quad (8c)$$

$$\mathbf{C}_t^{eq} = \mathbf{T}_I^{-1} \mathbf{C}_t. \quad (8d)$$

Equation (6) can be considered as  $n$  independent single-wire transmission lines, and each element can be written in the time domain as

$$\begin{aligned} & [\mathbf{V}_{inm}(0, t) - \mathbf{Z}_{Cinm} \mathbf{I}_{inm}(0, t)]_i \\ &= [\mathbf{V}_{inm}(\mathcal{L}, t - T_i) - \mathbf{Z}_{Cinm} \mathbf{I}_{inm}(\mathcal{L}, t - T_i)]_i + [\mathbf{V}_{0in}(t)]_i \end{aligned} \quad (9a)$$

$$\begin{aligned} & [\mathbf{V}_{inm}(\mathcal{L}, t) + \mathbf{Z}_{Cinm} \mathbf{I}_{inm}(\mathcal{L}, t)]_i \\ &= [\mathbf{V}_{inm}(0, t - T_i) + \mathbf{Z}_{Cinm} \mathbf{I}_{inm}(0, t - T_i)]_i + [\mathbf{V}_{\mathcal{L}in}(t)]_i, \end{aligned} \quad (9b)$$

where  $T_i$  is the one-way delay of the  $i$ th mode line. The elements of  $\hat{\mathbf{V}}_{0inm}$  and  $\hat{\mathbf{V}}_{\mathcal{L}inm}$  are similar to those of the coaxial cable, so  $[\mathbf{V}_{0in}(t)]_i$  and  $[\mathbf{V}_{\mathcal{L}in}(t)]_i$  can be written as [14]

$$\begin{aligned} [\mathbf{V}_{0in}(t)]_i &= -\frac{1}{2Z_{Cout}} \left( [\mathbf{R}_{dc}^{eq}]_i + B_i \frac{d}{dt} \right) E_{i1}(t) \\ &+ \frac{1}{2Z_{Cout}} \left( [\mathbf{R}_{dc}^{eq}]_i + C_i \frac{d}{dt} \right) E_{i2}(t) \end{aligned} \quad (10a)$$

$$\begin{aligned} [\mathbf{V}_{\mathcal{L}in}(t)]_i &= \frac{1}{2Z_{Cout}} \left( [\mathbf{R}_{dc}^{eq}]_i + C_i \frac{d}{dt} \right) E_{i3}(t) \\ &\quad - \frac{1}{2Z_{Cout}} \left( [\mathbf{R}_{dc}^{eq}]_i + B_i \frac{d}{dt} \right) E_{i4}(t), \end{aligned} \quad (10b)$$

where the coefficients  $B_i$  and  $C_i$  are given by

$$B_i = [\mathbf{L}_t^{eq}]_i + Z_{Cout} [\mathbf{Z}_{Cinm}]_i [\mathbf{C}_t^{eq}]_i \quad (11a)$$

$$C_i = [\mathbf{L}_t^{eq}]_i - Z_{Cout} [\mathbf{Z}_{Cinm}]_i [\mathbf{C}_t^{eq}]_i, \quad (11b)$$

and the voltages  $E_{ik}$  ( $k = 1, \dots, 4$ ) is composed of two parts  $E_{tik}(t)$  and  $E_{sik}(t)$ .  $E_{tik}(t)$  ( $k = 1, \dots, 4$ ) are the outer system's terminal voltage and current's contribution to  $E_{ik}$ , and are given by [14]

$$E_{tik}(t) = \int_{-\infty}^t F_{tik}(t') dt', \quad k = 1, \dots, 4 \quad (12)$$

where

$$\begin{aligned} F_{ti1}(t) &= \frac{\mathcal{L}}{T_i + T_{out}} \{ [V_{out}(0, t) + Z_{Cout} I_{out}(0, t)] \\ &\quad - [V_{out}(0, t - T_i - T_{out}) + Z_{Cout} I_{out}(0, t - T_i - T_{out})] \} \end{aligned} \quad (13a)$$

$$\begin{aligned} F_{ti2}(t) &= \frac{\mathcal{L}}{T_{out} - T_i} \{ [V_{out}(\mathcal{L}, t - T_i) - Z_{Cout} I_{out}(\mathcal{L}, t - T_i)] \\ &\quad - [V_{out}(\mathcal{L}, t - T_{out}) - Z_{Cout} I_{out}(\mathcal{L}, t - T_{out})] \} \end{aligned} \quad (13b)$$

$$\begin{aligned} F_{ti3}(t) &= \frac{\mathcal{L}}{T_{out} - T_i} \{ [V_{out}(0, t - T_i) + Z_{Cout} I_{out}(0, t - T_i)] \\ &\quad - [V_{out}(0, t - T_{out}) + Z_{Cout} I_{out}(0, t - T_{out})] \} \end{aligned} \quad (13c)$$

$$\begin{aligned} F_{ti4}(t) &= \frac{\mathcal{L}}{T_i + T_{out}} \{ [V_{out}(\mathcal{L}, t) - Z_{Cout} I_{out}(\mathcal{L}, t)] \\ &\quad - [V_{out}(\mathcal{L}, t - T_i - T_{out}) - Z_{Cout} I_{out}(\mathcal{L}, t - T_i - T_{out})] \}. \end{aligned} \quad (13d)$$

Here  $T_{out}$  is the one-way delay of the outer system. Thus the voltage  $E_{tik}$  ( $k = 1, \dots, 4$ ) can be realized through integral circuits, delay lines and controlled sources. When  $T_i = T_{out}$ , (13b) and (13c) seem to be undefined, however, they can be realized by differential circuits.  $E_{sik}(t)$  ( $k = 1, \dots, 4$ ) are the contributions related to external fields. When the incident electromagnetic field is a plane wave, and its incident direction and polarization are defined as in [1, 2],  $E_{sik}(t)$  ( $k = 1, \dots, 4$ ) can be written as [14]

$$E_{sik}(t) = \int_{-\infty}^t F_{sik}(t') dt', \quad k = 1, \dots, 4 \quad (14)$$

where

$$F_{si1}(t) = 2h\mathcal{L} \frac{e_x T_z - e_z T_x - e_x T_{out}}{T_z - T_{out}} \left\{ \frac{\mathcal{E}_0(t) - \mathcal{E}_0(t - T_i - T_{out})}{T_{out} + T_i} - \frac{\mathcal{E}_0(t) - \mathcal{E}_0(t - T_z - T_i)}{T_z + T_i} \right\} \quad (15a)$$

$$F_{si2}(t) = -2h\mathcal{L} \frac{e_x T_z - e_z T_x + e_x T_{out}}{T_z + T_{out}} \left\{ \frac{\mathcal{E}_0(t) - \mathcal{E}_0(t - T_z - T_i)}{T_z + T_i} + \frac{\mathcal{E}_0(t - T_z - T_i) - \mathcal{E}_0(t - T_z - T_{out})}{T_i - T_{out}} \right\} \quad (15b)$$

$$F_{si3}(t) = 2h\mathcal{L} \frac{e_x T_z - e_z T_x - e_x T_{out}}{T_z - T_{out}} \left\{ \frac{\mathcal{E}_0(t - T_z) - \mathcal{E}_0(t - T_i)}{T_z - T_i} - \frac{\mathcal{E}_0(t - T_i) - \mathcal{E}_0(t - T_{out})}{T_i - T_{out}} \right\} \quad (15c)$$

$$F_{si4}(t) = 2h\mathcal{L} \frac{e_x T_z - e_z T_x + e_x T_{out}}{T_z + T_{out}} \left\{ \frac{\mathcal{E}_0(t - T_z) - \mathcal{E}_0(t - T_i)}{T_z - T_i} + \frac{\mathcal{E}_0(t - T_z) - \mathcal{E}_0(t - T_z - T_i - T_{out})}{T_i + T_{out}} \right\}. \quad (15d)$$

Here  $h$  is the height of the cable over the ground,  $\mathcal{E}_0$  is the expression for the electric field,  $e_x$  and  $e_z$  are the  $x$  and  $z$  components of the incident electric field vector, and  $T_x$  and  $T_z$  are defined as  $T_x = \mathcal{L}/v_x$  and  $T_z = \mathcal{L}/v_z$ , where  $v_x$  and  $v_z$  are the velocities of the wave propagation along the  $x$  and  $z$  axes. According to (14) and (15),  $E_{sik}(t)$  ( $k = 1, \dots, 4$ ) can be realized through integral circuits, delay lines, controlled sources, and differential circuits when they are needed.

To allow loads to be connected between the inner wires and the ground, the transformation of the voltages and currents of the outer and inner systems to the voltages (referenced to ground) and currents of the shield and the wires is needed [10], which are denoted by  $V_0, V_1, \dots, V_n$ , and  $I_0, I_1, \dots, I_n$ . With (3), the transformation can be written as

$$\begin{bmatrix} \hat{V}_0 \\ \hat{\mathbf{V}}_I \end{bmatrix} = \begin{bmatrix} 1 & \mathbf{0}^T \\ \mathbf{1} & \mathbf{T}_V \end{bmatrix} \begin{bmatrix} \hat{V}_{out} \\ \hat{\mathbf{V}}_{inm} \end{bmatrix} \quad (16a)$$

$$\begin{bmatrix} \hat{I}_{out} \\ \hat{\mathbf{I}}_{inm} \end{bmatrix} = \begin{bmatrix} 1 & \mathbf{1}^T \\ \mathbf{0} & \mathbf{T}_I^{-1} \end{bmatrix} \begin{bmatrix} \hat{I}_0 \\ \hat{\mathbf{I}}_I \end{bmatrix}, \quad (16b)$$

where  $\hat{\mathbf{V}}_I = [V_1, \dots, V_n]^T$ ,  $\hat{\mathbf{I}}_I = [I_1, \dots, I_n]^T$ ,  $\mathbf{0}$  is a  $n \times 1$  vector with each element equals 0, and  $\mathbf{1}$  is a  $n \times 1$  vector whose elements are 1. Equation (16) can be realized in SPICE by using controlled sources.

It should be noted that the responses obtained by this model should be divided by 2 when the wave propagates parallelly to the ground [14].

The SPICE circuit for a multiconductor shielded cable excited by an incident plane-wave electromagnetic field is shown in Figure 2, where  $R_1$  and  $R_2$  are large resistances (like  $1\text{G}\Omega$ ) to avoid floating nodes according to the SPICE syntax. The controlled voltage sources  $E_{0out}$ ,  $E_{\mathcal{L}out}$ ,  $E_{i0}$ , and  $E_{i\Lambda}$ , are given by

$$E_{0out}(t) = -2h(e_x - e_z T_x / (T_z + T_{out})) [V(400) - V(403)]$$

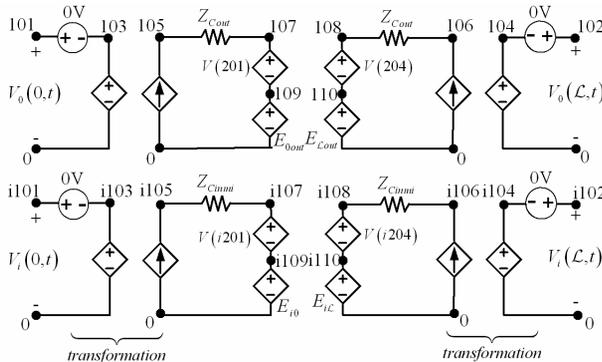
$$E_{\mathcal{L}out}(t) = \begin{cases} 2h(e_x - e_z T_x / T_z - T_{out}) [V(401) - V(402)] & (T_z \neq T_{out}) \\ 0 & (T_z = T_{out}) \end{cases}$$

$$E_{i0} = \frac{1}{2Z_{Cout}} \{ [\mathbf{R}_{dc}^{eq}]_i [V(i321, i322) - V(i311, i312)] - B_i V(i312) + C_i V(i322) \}$$

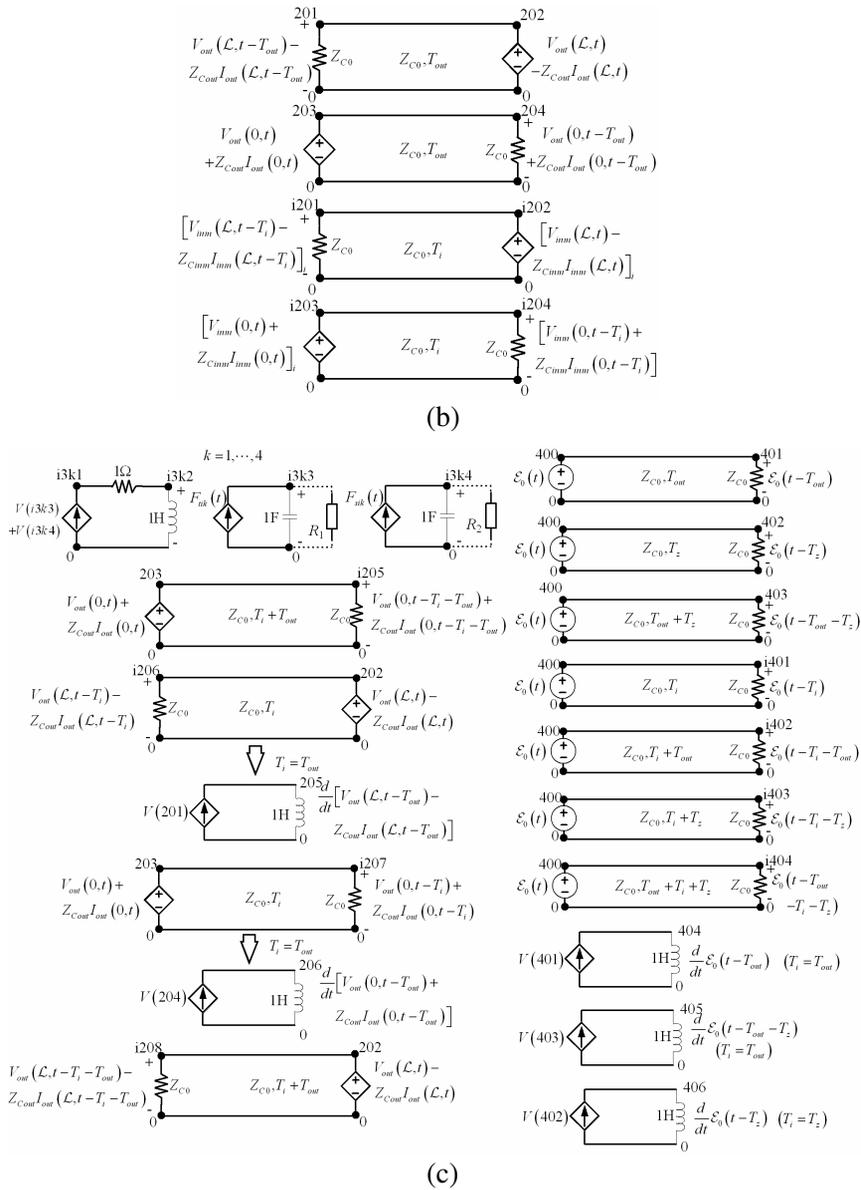
$$E_{i\mathcal{L}} = \frac{1}{2Z_{Cout}} \{ [\mathbf{R}_{dc}^{eq}]_i [V(i331, i332) - V(i341, i342)] - B_i V(i342) + C_i V(i332) \}$$

**2.2. SPICE Models for Conducted Susceptibility Analysis**

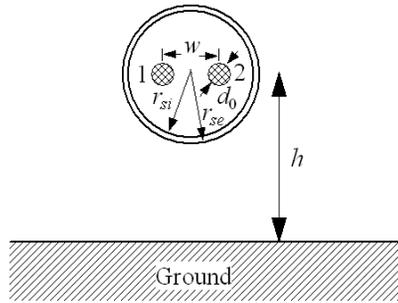
The development of the SPICE model for conducted susceptibility analysis is similar to and much simpler than that for radiated susceptibility analysis. This model can be easily obtained by setting  $\hat{V}_{sout}$  and  $\hat{I}_{sout}$  to zero, and modeling the excitation with a lumped source at the end of the cable. The SPICE circuit for conducted susceptibility analysis can be realized by removing the controlled sources  $E_{0out}$ ,  $E_{\mathcal{L}out}$ ,  $F_{sik}$ , and their realization circuits in Figure 2.



(a)



**Figure 2.** The SPICE circuit for multiconductor shielded cables excited by an incident plane-wave electromagnetic field. (a) The main circuit. (b) The realization circuits of the controlled voltage sources  $V(201)$ ,  $V(204)$ ,  $V(i201)$ , and  $V(i204)$ . (c) The realization circuits of the controlled sources  $E_{0out}$ ,  $E_{Lout}$ ,  $E_{i0}$ , and  $E_{iL}$ .



**Figure 3.** The cross section of a two-parallel-wire shielded cable.

**Table 1.** Parameters of the shield.

$C$	$P$ ( $\text{mm}^{-1}$ )	$N$	$d$ (mm)	$\alpha$ (degree)
16	0.17	12	0.18	27.6

### 3. NUMERICAL VALIDATIONS OF SPICE MODELS

The analysis is carried out on a two-parallel-wire shielded cable as shown in Figure 3. The outer radius  $r_{se}$  and the inner radius  $r_{si}$  of the braided shield are 7.49 mm and 7.13 mm, respectively. The characteristics of the braided shield can be defined in terms of the number of carriers  $C$ , the picks  $P$ , the ends  $N$ , the wire diameter  $d$ , and the weave angle  $\alpha$  [15, 16]. The values of these parameters are given in Table 1. The two inner wires with the diameter  $d_0 = 2$  mm and the distance  $w = 5.64$  mm are located symmetrically.

The differential and common mode disturbances at the termination due to a plane-wave electromagnetic field or a lumped current source are computed. The differential voltages  $V_{DMi}$  ( $i = 1, 2$ ) are the voltages across the resistances  $R_{Di}$  ( $i = 1, 2$ ), and the common mode voltages  $V_{CMi}$  ( $i = 1, 2$ ) are defined as

$$V_{CMi} = (I_{i1} + I_{i2}) \frac{R_{i1}R_{i2}}{R_{i1} + R_{i2}} \quad (i = 1, 2), \quad (17)$$

where  $I_{i1}$  and  $I_{i2}$  are the current through  $R_{i1}$  and  $R_{i2}$ , and their positive direction is defined as the direction towards the loads.

#### 3.1. Radiated Susceptibility Analysis

In the radiated susceptibility analysis, an incident electromagnetic wave with its electric field modeled by a biexponential pulse  $\mathcal{E}_0(t) = kE_0[\exp(-\beta t) - \exp(-\alpha t)]$ , where  $k = 1.3$ ,  $E_0 = 50$  kV/m,  $\alpha =$

$6.0e8s^{-1}$ , and  $\beta = 4.0e7s^{-1}$ , is used for the time domain analysis, while the electric field of 1 V/m magnitude for the frequency domain analysis.

The first configuration for the radiated susceptibility analysis is shown in Figure 4, where the shield is grounded at both ends (e.g.,  $R_1 = R_2 = 0$ ),  $R_{11}$ ,  $R_{12}$ , and  $R_{D1}$  are 200  $\Omega$ , 50  $\Omega$ , and 100  $\Omega$ , respectively, and the loads at the right termination has the same value as those at the left end. The length and height of the cable are 3 m and 2 cm, respectively. The incident wave propagates perpendicularly to the ground with the electric field parallel to the  $z$  axis. Figure 5 and Figure 6 show the differential and common mode voltages in the time and frequency domain analyses, respectively, where “multiconductor” means the multiconductor method [20] in which (1) is solved and the assumptions mentioned above have not been adopted,

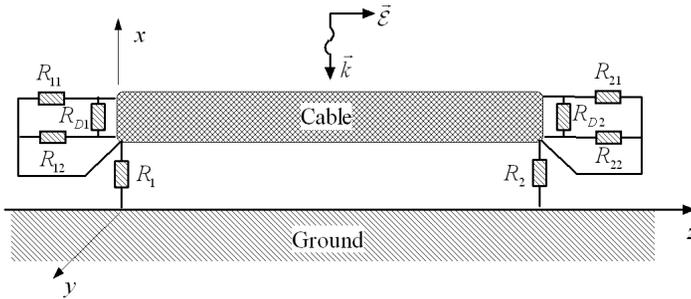


Figure 4. The configuration for the radiated analysis.

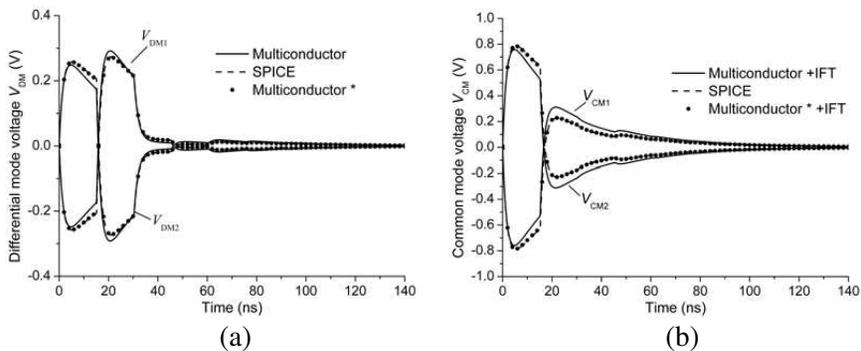
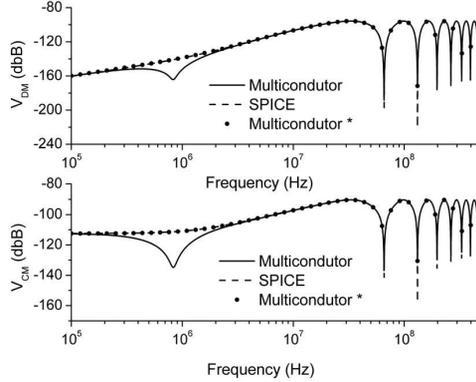


Figure 5. The disturbances at the cable ends in the time domain analysis with the incident wave propagating in the  $-x$  direction. (a) Differential mode voltages. (b) Common mode voltages.



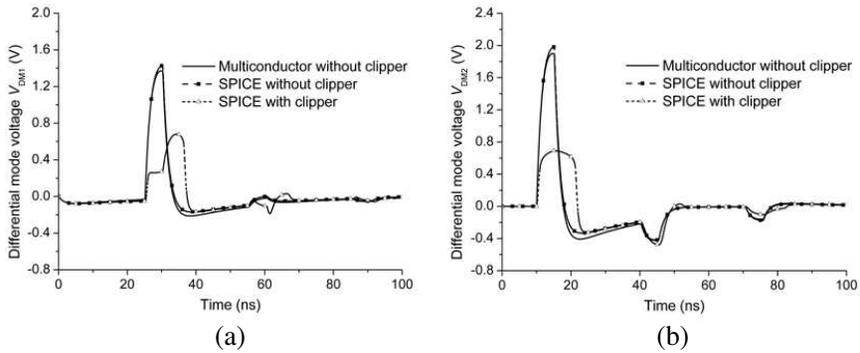
**Figure 6.** The disturbances at the cable ends in the frequency domain analysis with the incident wave propagating in the  $-x$  direction.

and “multiconductor \*” means the multiconductor method but with the assumption  $\hat{Z}_{tk}(\omega) = R_{dck} + j\omega L_{tk}$  ( $k = 1, \dots, n$ ) used. The results show that the voltages at the two ends have the same size but opposite polarity, and the differences between the results from the SPICE model and the multiconductor method are primarily due to the assumption  $\hat{Z}_{tk}(\omega) = R_{dck} + j\omega L_{tk}$  ( $k = 1, \dots, n$ ).

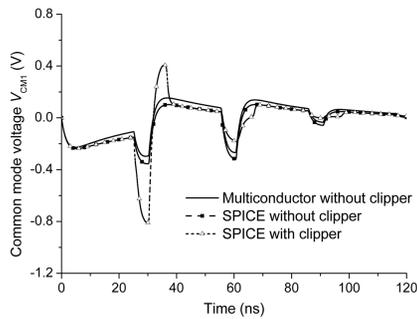
The second configuration considered is similar to the first one but with the incident wave propagating along the  $z$  axis, the electric field parallel to the  $x$  axis, and  $R_{21}$ ,  $R_{22}$ , and  $R_{D2}$  set to  $1\text{G}\Omega$ ,  $0\Omega$ , and  $1\text{G}\Omega$ , respectively. As mentioned above, the SPICE model can also be used when the loads are nonlinear. Thus, a voltage clipper, which is formed by two anti-parallel 1N4148 diodes, is added to the right termination and is in parallel with the load  $R_{D2}$  in this configuration. Figure 7 and Figure 8 show the differential and common mode voltages at the cable ends in presence and in absence of the voltage clipper, respectively. But only the common mode voltage  $V_{CM}$  at the left termination is given, because the right end is grounded, namely,  $R_{22} = 0\Omega$ . The differences of the curves in presence and in absence of the voltage clipper reveal the effects of the nonlinear device, which depresses the differential mode voltages and increases the common mode voltage.

### 3.2. Conducted Susceptibility Analysis

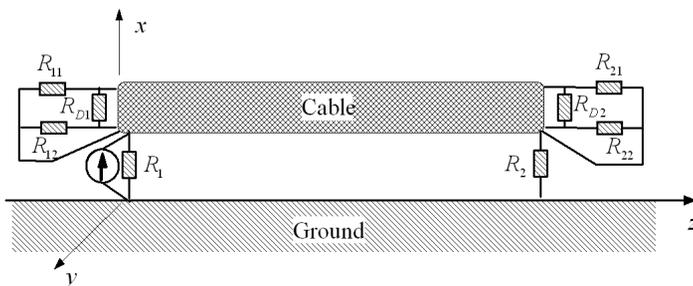
The configuration used for the conducted susceptibility analysis is shown in Figure 9, where a lumped current source is connected to



**Figure 7.** The differential mode voltage induced by the incident wave in presence and in absence of the voltage clipper. (a) The differential mode voltage  $V_{DM1}$  at the left termination. (b) The differential mode voltage  $V_{DM2}$  at the right termination.



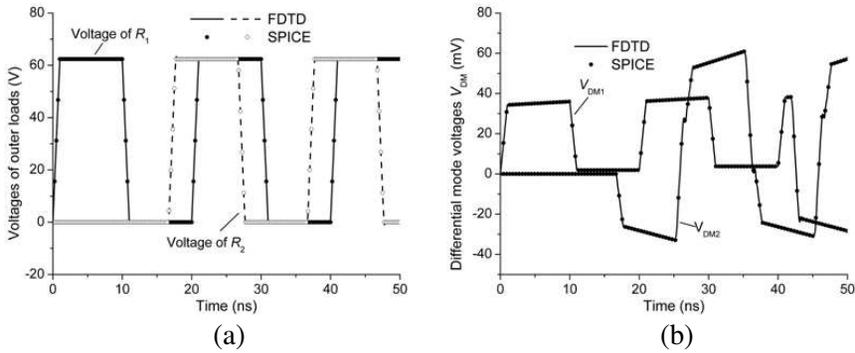
**Figure 8.** The common mode voltage  $V_{CM1}$  induced by the incident wave at the left termination.



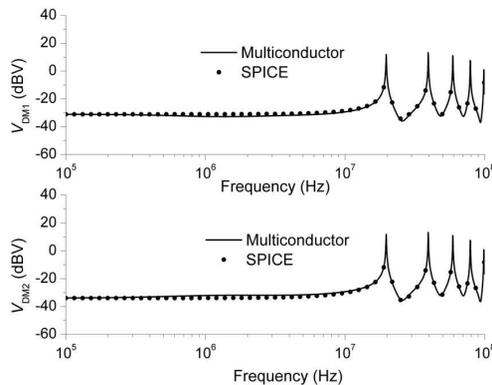
**Figure 9.** The configuration for the conducted susceptibility analysis.

the cable shield at the left end. The length and the height of the cable are set to 5 m and 3 cm, respectively. In this configuration, the inner loads  $R_{D1} = \infty$ ,  $R_{12} = \infty$ , and  $R_{11} = 0$ . The outer load  $R_1$  equals to the characteristic impedance of the outer system. The loads at the right end have same value as those at the left end. Other parameters of the cable are the same as those in the radiated susceptibility analysis.

The current source adopted for the time domain analysis is a clock wave of unit amplitude characterized by period  $T_{clk} = 20$  ns, rise and fall time  $T_r = T_f = 1$  ns, and duty cycle  $\eta = 0.5$ , while a AC current source of 1 A for the frequency domain analysis. The outer terminal voltages and the differential mode voltages obtained by the proposed



**Figure 10.** The responses obtained in the time domain analysis. (a) The voltages of the outer termination. (b) The differential mode voltages at the cable ends.



**Figure 11.** The voltage responses in the frequency domain analysis.

model in the time domain analysis are shown in Figure 10, where the “FDTD” means the finite difference time domain solution to the transmission-line equations of the cable. The differential mode voltages in the frequency domain analysis are shown in Figure 11. The results show that the solutions from different methods are in good agreement.

#### 4. SUMMARY AND CONCLUSION

Compact SPICE models for the radiated and conducted susceptibility analyses of multiconductor shielded cables in the time and frequency domains have been proposed, and their numerical validations have been given. Based on the four main assumptions: 1) multiconductor shielded cables are lossless; 2) the energy transferred from the inner to the outer system can be neglected; 3) the resistance part of the transfer impedance is frequency independent; 4) the height of cables over the ground is electrically small, the proposed models are derived from the transmission-line equations of multiconductor shielded cables. These SPICE models do not need any discretization of cables and allow the transient analysis when there are nonlinear or time-varying loads at the termination. Results obtained by these models agree well with those from other methods.

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