Wave Spreading Evaluation of Interconnect Systems

Haifang Liao and Wayne Wei-Ming Dai

***Abstract*—This paper derives general multiport interconnect constraints and presents a new approach, wave spreading evaluation (WSE), which uses S-parameter based network techniques to analyze coupled, multiconductor interconnect systems for high speed analog and digital integrated circuits. WSE is based on the spreading process of voltage waves with initial spreading waves created by the sources. The spreading process is independent of input sources, and every step of wave spreading meets the constrains of KCL and KVL. The continual spreading of voltage waves will create accurate results. Since the spreading voltage waves is the process of energy attenuation, the WSE method is always convergent for passive networks.**

I. Introduction

THE DESIGN of high performance VLSI and MCM systems requires accurate high frequency analysis of interconnects. Actually, the interconnect behavior has become the main factor to impose limitations on high performance systems. Because of increased signal speeds and interconnect densities, interconnect structures behave as a coupled transmission line system. Signal deterioration caused by skin effect, dispersion and discontinuities is no longer negligible. As a result, simple lumped equivalent models are inadequate for accurate analysis of complex interconnect structures [7].

The S-parameter based methods provide efficient techniques to analyze practical analog and digital integrated circuit interconnect systems that contain large number of coupled conductors and discontinuities [6], [7], [14], and [17]. Partially because it is much easier to determine the scattering parameters of any components on broad frequency bands by direct measurement [8]. Alternatively, the port parameters of many passive components (such as transmission lines) can be determine in terms of their geometric dimensions and the electrical characteristics of the materials [11]. The multiconductor transmission line segments and various discontinuities can be considered as separate components and their S-parameters can be obtained by measurement or software and stored in advance.

The earliest use of scattering parameters was the article by Campbell and Foster [4] in 1920, which dealt with properties of ideal transformers simultaneously matched at all ports. Many specific relations between the scattering coefficients were considered in that article, though the general properties of the scattering matrix were not mentioned. In the late 1930s and early 1940s, scattering concepts were mainly used to study waveguide junctions. In 1950s, the properties of scattering matrix were further studied and applied to network problems [3], [9], and [18]. Belevitch [1], [2] applied scattering concepts to lumped networks by relating the linear transformations of network voltage and current variables to the scattering matrix. Oono [15] showed that the scattering matrix technique might be applied to the frequency domain synthesis of general linear passive n-ports.

Scattering parameters can be used to analyze a single set of transmission lines [19]. However, with the advances of computer technology, scattering parameters are used to analyze large microwave networks. Monaco and Paolo [14] presented a general method to analyze biconjugate networks where the connections between component ports are restricted to the connections of pairs of ports. The interconnection relations for all the circuit component ports are described by the connection matrix whose elements are all null except the entries corresponding to pairs of adjacent ports. The problem is formulated as a linear system with sparse coefficient matrices. For a circuit consisting of many components, the computing time is too long primarily due to matrix inversion. Considerable reduction of computing time can be obtained by introducing hierarchical methods. Multiport connection method [14] divides the circuit into subcircuits, calculate the port matrix for each of the subcircuit, and, finally, the matrix relative to the ports of the complete circuit is determined. Subnetwork growth method [7], [13] combines two component and determine the S matrix of the resulting subcircuit hierarchically. Recently, Cooke *et al.* [6] applied S-parameter analysis technique to multiconjugate networks where three or more ports of components may be connected together in one interconnect junction. The general multiport interconnect constraints are based on ideal interconnect assumptions. However, some of the assumptions will result in violation of Kirchhoff's voltage law. In all of these methods, the frequency domain analysis proceeds as a series of matrix manipulations.

This paper presents a new approach—wave spreading evaluation (WSE) which uses S-parameters based network techniques to analyze coupled, high speed analog and digital integrated circuit interconnect systems. The network to be analyzed is partitioned into two parts: one consists of components which are described by S-parameters; and the other consists of the ideal interconnect topology. The general multiport interconnect constraints are derived, which show that an interconnect junction contains no preferred wave spreading paths and it can not be perfectly matched when the degree of the interconnect junction is more than two. No general sparse matrix operations are needed. The recursive computation of spreading incoming and outgoing voltage waves yields accurate results. The spreading process of voltage waves is the process of energy attenuation, thus the WSE method is always

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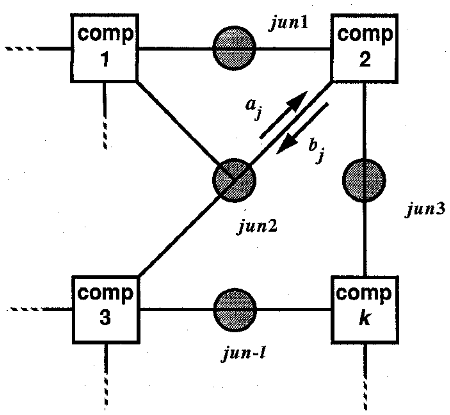


Fig. 1. Microwave circuits consist of components and ideal interconnect junctions.

convergent for passive networks.

II. Description of Circuit Networks

We model a circuit as a network which consists of two kinds of vertices: component vertices corresponding to the circuit components described by scattering (S) parameters and the junction vertices corresponding to the ideal interconnect junctions (see Fig. 1). Following a brief S-parameter description of the circuit components, we will derive scattering matrices of the ideal interconnect junctions, together with some general properties, which form the foundation for the WSE algorithm.

A. Circuit Components

It is convenient to explain WSE method when networks are described in terms of S-parameters. For the component i with p ports, a system of p equations can be written as [14]

$$\boldsymbol{B_i} = \boldsymbol{S_i A_i} \tag{1}$$

where $\boldsymbol{A_i} = [a_{i1}, a_{i2}, \cdots, a_{ip}]^T$, $\boldsymbol{B_i} = [b_{i1}, b_{i2}, \cdots, b_{ip}]^T$ represent the vectors of incoming and outgoing voltage waves at its p ports, respectively, and $\boldsymbol{S_i}$ is its $p \times p$ scattering matrix.

The voltage vector $\boldsymbol{V_i}$ and the current vector $\boldsymbol{I_i}$ are

$$\boldsymbol{V_i} = \boldsymbol{A_i} + \boldsymbol{B_i} \tag{2}$$
$$\boldsymbol{I_i} = \boldsymbol{Z_{0i}^{-1}}(\boldsymbol{A_i} - \boldsymbol{B_i}) \tag{3}$$

where $\boldsymbol{Z_{0i}} = \text{diag}\,[z_{01}, z_{02}, \cdots, z_{0p}]$ is a diagonal matrix, and z_{0j} is the reference impedance at port j ($j = 1, \cdots, p$).

B. Interconnect Junctions

For each vertex corresponding to an interconnect junction (Fig. 2), Kirchhoff's voltage law and current law hold. Assume that there are q edges connected at vertex i, we have

$$V_{i1} = V_{i2} = \cdots = V_{iq} \tag{4}$$

$$I_{i1} + I_{i2} + \cdots + I_{iq} = 0. \tag{5}$$

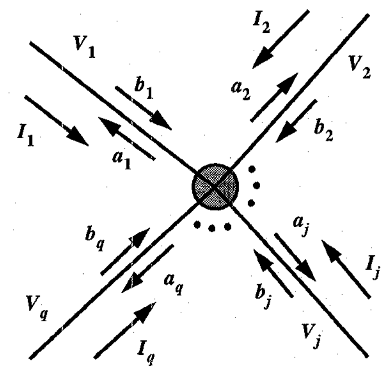


Fig. 2. An ideal interconnect junction.

Combining (4) and (5) with (2) and (3), we have

$$a_{i1} + b_{i1} = a_{i2} + b_{i2} = \cdots = a_{iq} + b_{iq} \tag{6}$$

$$\sum_{j=1}^{q} \frac{a_{ij}}{z_{0j}} = \sum_{j=1}^{q} \frac{b_{ij}}{z_{0j}} \tag{7}$$

where a_{ij} and b_{ij} are the jth element of $\boldsymbol{A}_i$ and $\boldsymbol{B}_i$, respectively. Write (6) and (7) in a matrix form

$$\begin{bmatrix} 1 & -1 & 0 & \cdots & 0 \\ 0 & 1 & -1 & \cdots & 0 \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ 0 & \cdots & 0 & 1 & -1 \\ z_{01}^{-1} & z_{02}^{-1} & \cdots & \cdots & z_{0q}^{-1} \end{bmatrix} \begin{bmatrix} a_{i1} \\ a_{i2} \\ \cdots \\ a_{iq} \end{bmatrix} = \begin{bmatrix} -1 & 1 & 0 & \cdots & 0 \\ 0 & -1 & 1 & \cdots & 0 \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ 0 & \cdots & 0 & -1 & 1 \\ z_{01}^{-1} & z_{02}^{-1} & \cdots & \cdots & z_{0q}^{-1} \end{bmatrix} \begin{bmatrix} b_{i1} \\ b_{i2} \\ \cdots \\ b_{iq} \end{bmatrix}. \tag{8}$$

To solve the system of equations, we have

$$\boldsymbol{A_i} = \boldsymbol{\Gamma_i B_i} \tag{9}$$

where $\boldsymbol{A_i} = [a_{i1}, a_{i2}, \cdots, a_{iq}]^T$, $\boldsymbol{B_i} = [b_{i1}, b_{i2}, \cdots, b_{iq}]^T$ and

$$\boldsymbol{\Gamma_i} = [\gamma_{ij}] = \frac{2}{G_i} \begin{bmatrix} \frac{1}{z_{01}} - \frac{1}{2}G_i & \frac{1}{z_{02}} & \cdots & \frac{1}{z_{0q}} \\ \frac{1}{z_{01}} & \frac{1}{z_{02}} - \frac{1}{2}G_i & \cdots & \frac{1}{z_{0q}} \\ \cdots & \cdots & \cdots & \cdots \\ \frac{1}{z_{01}} & \frac{1}{z_{02}} & \cdots & \frac{1}{z_{0q}} - \frac{1}{2}G_i \end{bmatrix} \tag{10}$$

where

$$G_i = \sum_{j=1}^{q} \frac{1}{z_{0j}}. \tag{11}$$

Notice that we use $\boldsymbol{A_i}$ as outgoing waves and $\boldsymbol{B_i}$ as incoming waves for junction vertex i, which opposite the convention for the components. Γ_i is the scattering matrix of the vertex i. In particular, when $z_{01} = z_{02} = \cdots = z_{0q}$

$$\Gamma_i = \frac{1}{q}\begin{bmatrix} 2-q & 2 & \cdots & 2 \\ 2 & 2-q & \cdots & 2 \\ \cdots & \cdots & \cdots & \cdots \\ 2 & 2 & \cdots & 2-q \end{bmatrix}. \quad (12)$$

Since $\gamma_{jk} = \gamma_{lk}$ for $j \neq k$ and $l \neq k$, we have

Theorem 1: The junction vertices contain no preferred voltage wave ports; that is, outgoing waves at all other ports due to the incoming wave at one port are equal.

For example, if port 1 has an incoming wave a_1 and other ports are perfectly matched (no incoming wave), outgoing waves on all ports other than port 1 are equal to $2G_i^{-1}z_{01}^{-1}a_1$.

The γ_{jj} is called the reflection factor at port j. We say the port j is perfectly matched (no reflection) if $\gamma_{jj} = 0$.

Theorem 2: If a junction vertex has degree greater than or equal to three, it is impossible for all its ports to be perfectly matched. That is, it is impossible that $\gamma_{jj} = 0$ for $j = 1, 2, \cdots, q$ if $q \geq 3$.

Proof: Assume that all $\gamma_{jj} = 0$ $(j = 1, 2, \cdots, q)$ for a junction vertex i, that is

$$\frac{2}{G_i z_{0j}} - 1 = 0 \quad j = 1, 2, \cdots, q. \quad (13)$$

Since $G_i \neq 0$, we can rewrite (13) as

$$\frac{2}{z_{0j}} - G_i = 0 \quad j = 1, 2, \cdots, q \quad (14)$$

the matrix form of these equations is

$$\begin{bmatrix} 1 & -1 & -1 & \cdots & -1 \\ -1 & 1 & -1 & \cdots & -1 \\ -1 & -1 & 1 & \cdots & -1 \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ -1 & -1 & -1 & \cdots & 1 \end{bmatrix}\begin{bmatrix} \frac{1}{z_{01}} \\ \frac{1}{z_{02}} \\ \cdots \\ \frac{1}{z_{0q}} \end{bmatrix} = \mathbf{0}. \quad (15)$$

Since the determinant of the coefficient matrix is $-(q-2)2^{q-1} \neq 0$ when $q \geq 3$, there exists no nontrivial solution to these equations. Therefore, the assumption that all $\gamma_{jj} = 0$ $(j = 1, 2, \cdots, q)$ can not be true in general. □

In [6], S-parameter analysis of multiconjugate networks is based on the assumption, among others, that all interconnect elements must be perfectly matched. Clearly, it contradicts to the above theorem.

Obviously, the ideal junction vertices neither store nor consume energy. We can prove this with the Γ matrix described in (10).

Theorem 3: The junction vertices neither store nor consume energy.

Proof: For junction vertex i, the input power denoted by P_{in} and output power denoted by P_{out} are

$$P_{in} = (\boldsymbol{B}_i^*)^T \boldsymbol{Z}_{0i}\boldsymbol{B}_i \quad (16)$$

$$P_{out} = (\boldsymbol{A}_i^*)^T \boldsymbol{Z}_{0i}\boldsymbol{A}_i \quad (17)$$

where $\boldsymbol{B}_i$ and $\boldsymbol{A}_i$ are input voltage waves and output voltage waves, respectively, $\boldsymbol{B}_i^*$ and $\boldsymbol{A}_i^*$ are their corresponding conjugate vectors, and $\boldsymbol{Z}_{0i} = \mathrm{diag}[z_{01}, z_{02}, \cdots, z_{0q}]$. From (9)

$$\begin{aligned} P_{out} &= [(\Gamma_i\boldsymbol{B}_i)^*]^T\boldsymbol{Z}_{0i}(\Gamma_i\boldsymbol{B}_i) \\ &= (\boldsymbol{B}_i^*)^T(\Gamma_i^*)^T\boldsymbol{Z}_{0i}\Gamma_i\boldsymbol{B}_i \\ &= (\boldsymbol{B}_i^*)^T\boldsymbol{Z}_{0i}\boldsymbol{B}_i \\ &= P_{in}. \end{aligned} \quad (18)$$

Therefore, the junction vertices neither store nor consume energy. □

III. Computation of Spreading Wave

To compute incoming and outgoing voltage waves at each port, we combine (9) and (1) with the internal source vector $\boldsymbol{C}$ added [6]

$$\boldsymbol{B} = \boldsymbol{SA} + \boldsymbol{C} \quad (19)$$

$$\boldsymbol{A} = \Gamma\boldsymbol{B}. \quad (20)$$

Solve these equations, we get

$$\boldsymbol{A} = \frac{\Gamma\boldsymbol{C}}{\boldsymbol{E} - \Gamma\boldsymbol{S}} \quad (21)$$

$$\boldsymbol{B} = \frac{\boldsymbol{C}}{\boldsymbol{E} - \boldsymbol{S}\Gamma} \quad (22)$$

where $\boldsymbol{E}$ is the identity matrix. Generally, Γ and $\boldsymbol{S}$ are sparse matrix and $\boldsymbol{S}$ is frequency dependent. First, let us expand $\boldsymbol{A}$ and $\boldsymbol{B}$ into series form

$$\boldsymbol{A} = \sum_{i=0}^{\infty}\Gamma(\boldsymbol{S}\Gamma)^i\boldsymbol{C} = \sum_{i=1}^{\infty}\boldsymbol{A}^{(i)} \quad (23)$$

$$\boldsymbol{B} = \sum_{i=0}^{\infty}(\boldsymbol{S}\Gamma)^i\boldsymbol{C} = \sum_{i=0}^{\infty}\boldsymbol{B}^{(i)} \quad (24)$$

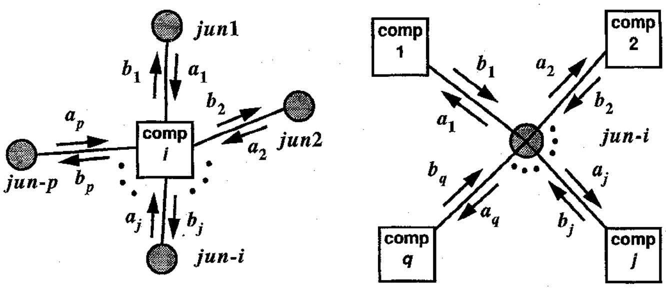
where $\boldsymbol{A}^{(i)}$ is the spreading incoming waves which start from the source and arrive at all ports after passing through i components and $\boldsymbol{B}^{(i)}$ the spreading outgoing waves which start from the source and leave from all ports after passing through i components. From (23) and (24), we have the following recursive formulas

$$\boldsymbol{B}^{(0)} = \boldsymbol{C} \quad (25)$$

$$\boldsymbol{A}^{(i)} = \Gamma\boldsymbol{B}^{(i-1)} \quad (26)$$

$$\boldsymbol{B}^{(i)} = \boldsymbol{S}\boldsymbol{A}^{(i)}. \quad (27)$$

From the above formulas, we find that ith incoming spreading waves are only dependent on the $(i-1)$th outgoing spreading waves (26); the ith outgoing spreading wave is only dependent on the ith incoming spreading wave (27); the sources only contribute to the initial conditions (25). By repeatedly using (26) and (27), we can get all $\boldsymbol{A}^{(i)}$ and $\boldsymbol{B}^{(i)}$. According to (23) and (24), the sum of them amounts to the incoming and outgoing waves at all component ports.



(a) (b)

Fig. 3. Spreading wave computation.

In order to avoid unnecessary sparse matrix operations, let us take a close look at (26) and (27). Since all components are only connected to interconnect junctions, (1) can be used for every component vertex to compute $\boldsymbol{B}_{(i)}$ instead of using (27). Similarly, (9) can be used for every junction vertex to compute $\boldsymbol{A}_{(i)}$ instead of using (26) (see Fig. 3).

Since the source only contributes to the initial condition, we replace it by a reference resistor with impedance z_0 when computing the spreading waves. Simply by setting the initial outgoing wave at the source port $b_s = 1$, we can compute the voltage V_i' at junction vertex i and incoming wave at the source port a_s. The voltage at the source port is $E' = a_s + b_s = a_s + 1$. Obviously, the actual voltage at junction i, due to the contribution of source E, is

$$V_i = \frac{V_i' E}{a_s + 1}. \tag{28}$$

IV. Convergence

Generally, reference impedances of all ports can be normalized as real numbers. For lossy and passive interconnect systems, based on the principle of energy conservation [16], we have

$$\sum_{j=1}^{n} z_{0j}^{-1} |b_j^{(i)}|^2 < \sum_{j=1}^{n} z_{0j}^{-1} |a_j^{(i)}|^2 \tag{29}$$

that is

$$[\boldsymbol{B}^{(i)*}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{B}^{(i)} < [\boldsymbol{A}^{(i)*}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{A}^{(i)} \tag{30}$$

where n is the total number of ports, $a_j^{(i)}$ and $b_j^{(i)}$ are the elements of spreading voltage waves $\boldsymbol{A}^{(i)}$ and $\boldsymbol{B}^{(i)}$, respectively. On the other hand, according to Theorem 3

$$[\boldsymbol{A}^{(i)}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{A}^{(i)} = [\boldsymbol{B}^{(i-1)*}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{B}^{(i-1)}. \tag{31}$$

So, we have

$$[\boldsymbol{A}^{(i)*}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{A}^{(i)} < [\boldsymbol{A}^{(i-1)*}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{A}^{(i-1)} \tag{32}$$

$$[\boldsymbol{B}^{(i)*}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{B}^{(i)} < [\boldsymbol{B}^{(i-1)*}]^T \boldsymbol{Z}_0^{-1} \boldsymbol{B}^{(i-1)}. \tag{33}$$

Thus, for the passive network, the process of voltage spreading is the process of energy attenuation. In other words, the spreading energy will die down to zero. That is

$$\lim_{i\to\infty} \boldsymbol{A}^{(i)} = \lim_{i\to\infty} \Gamma(\boldsymbol{S}\Gamma)^{i-1}\boldsymbol{C} = \boldsymbol{0} \tag{34}$$

$$\lim_{i\to\infty} \boldsymbol{B}^{(i)} = \lim_{i\to\infty} (\boldsymbol{S}\Gamma)^{i}\boldsymbol{C} = \boldsymbol{0}. \tag{35}$$

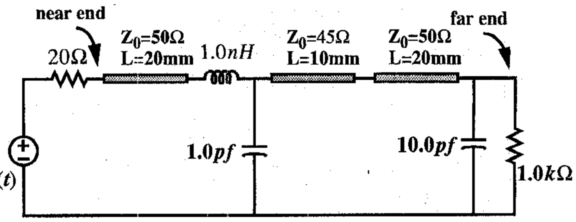


Fig. 4. Lossless transmission line network.

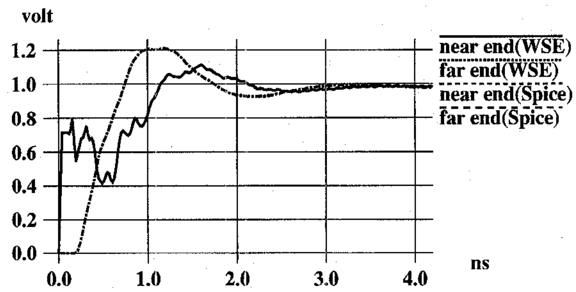


Fig. 5. Voltage waveform of lossless transmission line network.

So, (23) and (24) are always converge. And the solutions are

$$\begin{aligned} \boldsymbol{A} &= \sum_{i=0}^{\infty} \Gamma(\boldsymbol{S}\Gamma)^i \boldsymbol{C} \\ &= (\boldsymbol{E} - \Gamma\boldsymbol{S})^{-1}\boldsymbol{C} \end{aligned} \tag{36}$$

$$\begin{aligned} \boldsymbol{B} &= \sum_{i=0}^{\infty} (\boldsymbol{S}\Gamma)^i \boldsymbol{C} \\ &= (\boldsymbol{E} - \boldsymbol{S}\Gamma)^{-1}\boldsymbol{C}. \end{aligned} \tag{37}$$

Theorem 4: The computation of spreading wave based on the recursive formula (25)–(27) always converges for passive circuit networks.

V. Experimental Results

The first test circuit, shown in Fig. 4 is the model of the interconnect between MCM and PCB via a pad. It consists of three sections of lossless transmission lines together with several lumped components. The circuit and the S-parameters of the transmission lines were provided by Hewlett-Packard. The responses of the step signal with 0.04 ns rise time together with the output of SPICE3e2 are plotted in Fig. 5. The results from SPICE3e2 and WSE are identical. It took 5.6 seconds by WSE, on the SUN SPARC 1+ station, compared to 54.1 seconds by SPICE3e2.

The WSE can also be applied to transmission line structures described by measured scattering parameters. As an example, we consider a tri-conductor system given in [5] (see Fig. 6) with a piecewise linear function input, the response curves versus the result of the measurement [5] are shown in Fig. 7.

The last example is intended to demonstrate the efficiency as well as the generality of the WSE method. The test circuit contains two sets of coupled lossy transmission lines, together with some lumped elements (see Fig. 8). The input signal is the step function with 0.2 ns rise time. The resulting waveforms at the source and load points are shown in Fig. 9. It took 10.4 seconds to analyze this circuit.

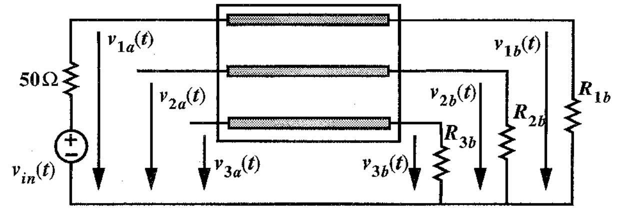


Fig. 6. A tri-conductor system.

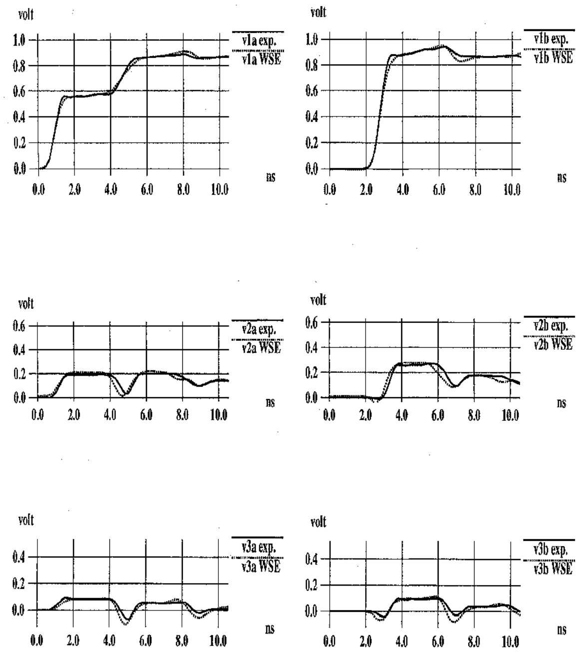


Fig. 7. Response waveforms for the tri-conductor system.

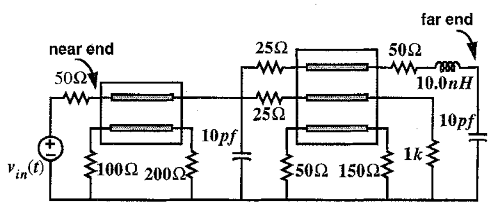


Fig. 8. A circuit containing two sets of coupled lossy transmission lines.

VI. Conclusion

We have proposed a new simulation method to analyze arbitrary interconnect structures characterized by scattering parameters. A general S-parameter description of a multiple port interconnect junction is derived. This description has been used in S-parameter macromodel based transient analysis of lumped and distributed interconnect systems [10]. The recursive computation of incoming and outgoing spreading voltage waves yields accurate results, without matrix inversion. This method is always convergent for passive networks.

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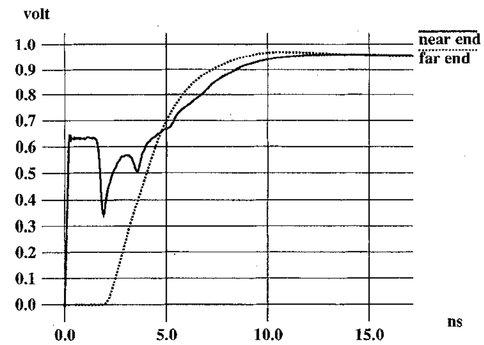

Fig. 9. Source and load waveforms for circuit containing two sets of coupled lossy transmission lines.

of Hewlett-Packard for data of the lossless transmission line network.

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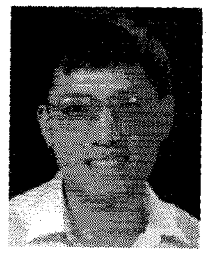

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