A New Method for the Reduction of Crosstalk and Echo in Multiconductor Interconnections

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***Abstract*—Crosstalk and echo can be reduced in multiconductor interconnections, using (truly) matched terminations and a different modal variable for each transmission channel. We first study a conventional technique for reducing reflections, using grounded linear two-terminal circuit elements. Using the concepts of modal voltages and modal currents, we define a new method for the reduction of crosstalk and echo, which involves specific terminations, specific transmitting circuits to send signals, and specific receiving circuits to receive signals. We establish the design equations and show that the new method is related to a particular choice of eigenvectors, called associated eigenvectors. Simulations of two examples of implementation of this method confirm that it provides reduced crosstalk and echo. We also discuss the implementation of the new method with an interconnection having identical propagation constants for all modes. Finally, we compare the new method with a different concept based on modal variables and unspecified terminations.**

***Index Terms*—Crosstalk, interconnection, multiconductor cable, signal integrity, transmission.**

I. Introduction

In this paper, we consider interconnections used for sending several signals, such as found in multiconductor cables, printed circuit board and integrated circuit. We have shown in Fig. 1 such an interconnection with four transmission conductors and a reference conductor, connected to line transmitters and line receivers spread over the length of the interconnection. The transmission conductors may be the traces of a printed circuit board, the reference conductor being a ground plane. This type of interconnection supports bidirectional transmission, and could be used for sending digital signals, using a data bus architecture. Such an interconnection could also be used to send analog signals.

The degradation of signals transmitted through the interconnection is usually described as the result of three frequency-dependent phenomena: echo, crosstalk, and the variation of attenuation and propagation velocity with frequency. In this paper, we will call *echo* the detrimental phenomenon by which a signal propagating in a given direction, in one of the transmission channels, produces a noise on the same transmission channel, propagating in the opposite direction. Crosstalk is the detrimental phenomenon by which a signal sent on one of the transmission channels produces noise on other transmission channels. We distinguish between *echo* and the more general concept of *reflections*, because reflections occurring for instance at an end of the interconnection may be responsible for echo and/or crosstalk. Often, when the type of interconnection is chosen (for instance when a type of multiconductor cable has been chosen), echo and crosstalk limit the maximum length of the interconnection for a given bandwidth, or the maximum bandwidth for a given length.

The main purpose of this paper is to present a new method for the reduction of crosstalk, which also provides a reduction of echo and of the distortions caused by propagation. However, it is also our intention to show how this finding is related to a particular (and unusual) implementation of modal analysis, based on a particular choice of eigenvectors, called *associated eigenvectors*. The definition of associated eigenvectors will be given in Section II below, together with other definitions and results used in the paper.

We will first look at the most widely used technique for reducing echo and crosstalk, namely terminating the interconnection with grounded linear two-terminal circuit elements, in Section III. We will then define the new method for the reduction of crosstalk in Section IV, which uses $n+1$ conductors to provide n transmission channels. This is unlike conventional techniques such as the use of differential circuits or the use of an individual screen (i.e., shield) for each transmission channel, both techniques requiring at least $2n$ conductors. The design equations for the new method will be derived in Section V. In Section VI, we will show that this method is related to modal transforms producing associated eigenvectors. In Sections VII and VIII, we will apply the new method to the cancellation of crosstalk and echo in two different interconnections. In Section IX, we will discuss the simplification of circuits which can be obtained when the MTL is completely degenerate, i.e., when the propagation constants of all modes are substantially equal. Finally, we will compare the new method with other concepts of modal transmission, in Section X.

II. Notations and Basic Results

Let us consider an interconnection with n transmission conductors placed close to a reference conductor. Let us number these conductors from 0 to n, the number 0 being attributed to the reference conductor (ground). We will locate any point along the interconnection of length L with a real curvilinear abscissa z, the interconnection extending from $z = 0$ to $z = L$. Any integer j such that $1 \leq j \leq n$ will be used as an index for defining the current i_j flowing on the transmission conductor j, and the voltage v_j between this transmission conductor and the reference conductor. These n currents and n voltages will respectively be called the *natural currents* and the *natural voltages*. The column vector $\mathbf{I}$ of the natural currents $i_1, \ldots, i_n$ and

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the column vector $\mathbf{V}$ of the natural voltages $v_1, \ldots, v_n$ both depend on z.

The interconnection may be described with the model of the multiconductor transmission line (MTL) if it is characterized with a per-unit-length (p.u.l.) impedance matrix $\mathbf{Z}$ and a p.u.l. admittance matrix $\mathbf{Y}$, and if its electrical behavior is described in the frequency domain by the telegrapher's equations

$$\begin{cases} \frac{d\mathbf{V}}{dz} = -\mathbf{ZI} \\ \frac{d\mathbf{I}}{dz} = -\mathbf{YV}. \end{cases} \tag{1}$$

For instance, the $(n+1)$ conductor MTL model can be applied to interconnections made of parallel conductors, provided the frequency is not too high [1]. The matrices $\mathbf{Z}$ and $\mathbf{Y}$ are symmetrical square matrices of order n, and they are frequency dependent. The MTL is said to be *uniform* if the matrices $\mathbf{Z}$ and $\mathbf{Y}$ are independent of z. It is well known that (1) may in this case easily be solved using a suitable diagonalization [1], [2] of the matrices $\mathbf{ZY}$ and $\mathbf{YZ}$. More precisely, $\mathbf{Z}$ and $\mathbf{Y}$ being symmetrical matrices, $\mathbf{ZY}$ and $\mathbf{YZ}$ have the same eigenvalues and the same characteristic polynomial, and we shall use $\mathbf{T}$ and $\mathbf{S}$ to denote two regular matrices such that

$$\begin{cases} \mathbf{T}^{-1}\mathbf{YZT} = \mathbf{D} \\ \mathbf{S}^{-1}\mathbf{ZYS} = \mathbf{D} \end{cases} \tag{2}$$

where

$$\mathbf{D} = \mathrm{diag}_n\left(\gamma_1^2, \ldots, \gamma_n^2\right) \tag{3}$$

is the diagonal matrix of order n of the eigenvalues of $\mathbf{ZY}$ and $\mathbf{YZ}$, each of these eigenvalues being written as the square of a propagation constant γ_i chosen with a positive imaginary part.

Assuming that matrices $\mathbf{T}$ and $\mathbf{S}$ solutions of (2) and (3) exist, they define a *modal transform* for the natural currents and for the natural voltages, and the results of this transform are called the *modal currents* and the *modal voltages*. If we use $\mathbf{I}_M$ to denote the vector of the n modal currents $i_{M1}, \ldots, i_{Mn}$ and if we use $\mathbf{V}_M$ to denote the vector of the n modal voltages $v_{M1}, \ldots, v_{Mn}$, we get (by definition)

$$\begin{cases} \mathbf{V} = \mathbf{SV}_M \\ \mathbf{I} = \mathbf{TI}_M. \end{cases} \tag{4}$$

Consequently, we shall call $\mathbf{S}$ the *transition matrix from modal voltages to natural voltages*, and we shall call $\mathbf{T}$ the *transition matrix from modal currents to natural currents*. The modal voltages have the remarkable property of being able to propagate along the transmission line without coupling to one another when they have a different index. This also applies to the modal currents. Also, for any integer α such that $1 \leq \alpha \leq n$, a modal current $i_{M\alpha}$ and a modal voltage $v_{M\alpha}$ propagate with the same propagation constant γ_α toward the far end, and with the opposite propagation constant $-\gamma_\alpha$ toward the near-end.

We note that (2) means that the column vectors of $\mathbf{S}$ (respectively, of $\mathbf{T}$) are linearly independent eigenvectors of $\mathbf{ZY}$ (respectively, of $\mathbf{YZ}$), and that consequently $\mathbf{S}$ and $\mathbf{T}$ are not defined in a unique manner by (2) and (3) alone, because: first, the order of the eigenvalues in (3) is arbitrary, and second, the choice of eigenvectors corresponding to a degenerate eigenvalue is arbitrary for $\mathbf{S}$ and for $\mathbf{T}$. It is therefore possible to introduce an additional (and well chosen!) condition to solve (2), but such a choice is not unique.

The characteristic impedance matrix $\mathbf{Z}_C$ is defined by [3]

$$\begin{aligned} \mathbf{Z}_C &= \mathbf{S\Gamma}^{-1}\mathbf{S}^{-1}\mathbf{Z} = \mathbf{S\Gamma S}^{-1}\mathbf{Y}^{-1} \\ &= \mathbf{Y}^{-1}\mathbf{T\Gamma T}^{-1} = \mathbf{ZT\Gamma}^{-1}\mathbf{T}^{-1} \end{aligned} \tag{5}$$

where

$$\mathbf{\Gamma} = \mathrm{diag}_n\left(\gamma_1, \ldots, \gamma_n\right) \tag{6}$$

is the diagonal matrix of order n of the propagation constants γ_i. $\mathbf{Z}_C$ is a symmetrical matrix independent of the choice of the matrices $\mathbf{T}$ and $\mathbf{S}$. The reflection occurring at an end terminated by a linear $(n+1)$-terminal device presenting an impedance matrix $\mathbf{Z}_L$ to the MTL is given by

$$\begin{aligned} \mathbf{V}_- &= (\mathbf{Z}_L - \mathbf{Z}_C)(\mathbf{Z}_L + \mathbf{Z}_C)^{-1}\mathbf{V}_+ \\ &= (\mathbf{Z}_L + \mathbf{Z}_C)^{-1}(\mathbf{Z}_L - \mathbf{Z}_C)\mathbf{V}_+ \\ &= \mathbf{Z}_C(\mathbf{Z}_L + \mathbf{Z}_C)^{-1}(\mathbf{Z}_L - \mathbf{Z}_C)\mathbf{Z}_C^{-1}\mathbf{V}_+. \end{aligned} \tag{7}$$

As a consequence, no reflection occurs for incident waves if and only if the impedance matrix $\mathbf{Z}_L$ of the termination is equal to the characteristic impedance matrix $\mathbf{Z}_C$. In this case, the termination is said to be *matched* to the MTL. A matched termination for a lossless line requires a network of $n(n+1)/2$ resistors or less [4]. We note that some authors improperly refer to particular terminations made of n grounded linear circuit elements (discussed in Section III) as matched terminations.

The matrices $\mathbf{Z}$ and $\mathbf{Y}$ being symmetrical, we observe that if we determine, with a diagonalization of the matrix $\mathbf{YZ}$, a matrix $\mathbf{T}$ satisfying the first line of (2), then

$$\mathbf{S} = {}^t\mathbf{T}^{-1} \tag{8}$$

is a solution of the second line of (2). Eigenvectors obtained with (8) will be referred to as *bi-orthonormal eigenvectors* [5]. Most authors use bi-orthonormal eigenvectors. If a matrix $\mathbf{T}$ satisfying the first line of (2) is known, another possible choice to obtain a solution $\mathbf{S}$ for the second line of (2) is

$$\mathbf{S} = j\omega c_K \mathbf{Y}^{-1}\mathbf{T} \tag{9}$$

where c_K is an arbitrary nonzero scalar depending eventually on the angular frequency ω and having the dimensions of p.u.l. capacitance. This choice is not as usual as (8), but it has for instance been used in [6]–[9]. In order to indicate that a matrix $\mathbf{S}$ and a matrix $\mathbf{T}$ are defined by (2), (3), and (9) we shall say that they are *associated*, and that the eigenvectors contained in $\mathbf{S}$ and $\mathbf{T}$ (i.e., their column vectors) are associated. With associated eigenvectors, for a wave propagating in a given direction and for any integer α such that $1 \leq \alpha \leq n$, we have

$$v_{M\alpha} = \frac{\varepsilon}{j\omega c_K}\gamma_\alpha i_{M\alpha} \tag{10}$$

where ε is equal to 1 if the wave propagates toward the far-end, or to -1 if the wave propagates toward the near-end. It implies that the propagation of the modal voltage $v_{M\alpha}$ and of the modal current $i_{M\alpha}$ can be viewed as the propagation on a ficticious

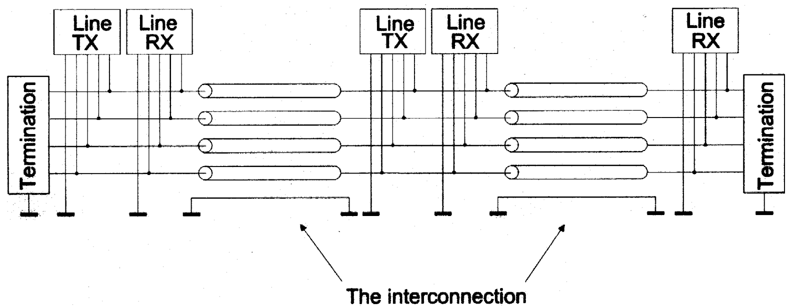

Fig. 1. Interconnection having four transmission conductors, used for bidirectional signals.

2-conductor transmission line having the propagation constant γ_α and the characteristic impedance $\gamma_\alpha / j\omega c_K$. As a result, we say that associated eigenvectors provide a *total decoupling* of the telegrapher's equation, since it allows to define an equivalent circuit for the $(n+1)$ conductor MTL, comprising n independent two-conductor transmission lines.

III. Optimal Terminations

In the following, we will only consider interconnections with n transmission conductors and a reference conductor, providing n transmission channels, as in Fig. 1. The state of the art as regards fighting against reflections in such an interconnection of given cross-section implements terminations made of n linear two-terminal circuit elements, having each a terminal connected to a different transmission conductor, the other terminal being grounded. Their impedance is chosen to obtain reduced reflected waves. This directly reduces echo, and indirectly (and marginally) reduces crosstalk, since reflected waves also give rise to some crosstalk. The impedance matrix $\mathbf{Z}_L$ of such a termination is diagonal. Since the characteristic impedance matrix $\mathbf{Z}_C$ of the MTL is not diagonal (unless the transmission conductors are uncoupled), some reflection will always occur at such a termination, according to (7). Applying the adjective "matched" to it is therefore improper. We will use the wording *pseudo-matched impedance* to designate the impedance attributed to each linear two-terminal circuit element of such a termination, with the minimization of the detrimental effects of reflections in mind.

There are obviously several ways of computing pseudo-matched impedances. However, this question does not seem to have been discussed in the literature. In most available documents, the pseudo-matched impedance for a transmission conductor seems to be determined as the characteristic impedance of an equivalent transmission line with a single transmission conductor, thereby neglecting the propagation on the other transmission conductors. This point of view is an approximation which can only be justified when the couplings between transmission conductors are very weak. When the coupling is not very weak, the pseudo-matched impedances should be computed using the properties of the MTL regarding reflections, contained in $\mathbf{Z}_C$. Let $\mathbf{P}$ be the matrix of the voltage reflection coefficients ρ_{ij} defined by

$$\mathbf{P} = (\mathbf{Z}_L - \mathbf{Z}_C)(\mathbf{Z}_L + \mathbf{Z}_C)^{-1} \tag{11}$$

for a given impedance matrix $\mathbf{Z}_L$ of the termination.

A first way of defining pseudo-matched impedances is to use the diagonal elements of $\mathbf{Z}_C$. The Authors have already used this definition [7], [8], but it does not correspond to an outstanding property of the reflected wave. This definition may therefore be regarded as arbitrary.

A second way of defining pseudo-matched impedances is to require that the diagonal elements of $\mathbf{P}$ are equal to zero. In this case, if the incident wave exists on a single transmission conductor, there is no reflected wave on this conductor: reflections are present but there is no echo!

A third way of defining pseudo-matched impedances is to require that the maximum absolute column sum norm $||\mathbf{P}||_1$ of the $\mathbf{P}$ matrix be minimized. This norm [10] is defined by

$$||\mathbf{P}||_1 = \max_j \sum_{i=1}^{n} |\rho_{ij}|. \tag{12}$$

Since it is the natural norm induced by the $\mathbf{L}_1$-norm for vectors, defined by

$$||\mathbf{V}||_1 = \sum_{i=1}^{n} |v_i| \tag{13}$$

this norm is such that

$$\frac{||\mathbf{V}_-||_1}{||\mathbf{V}_+||_1} \le \max_{\mathbf{V}_+ \neq 0} \frac{||\mathbf{V}_-||_1}{||\mathbf{V}_+||_1} = ||\mathbf{P}||_1 \tag{14}$$

$\mathbf{V}_-$ and $\mathbf{V}_+$ being the vector of the natural voltages of the reflected or of the incident wave respectively. Thus, minimizing $||\mathbf{P}||_1$ amounts to minimizing $||\mathbf{V}_-||_1 / ||\mathbf{V}_+||_1$.

A fourth way of defining pseudo-matched impedances is to require that the maximum absolute row sum norm $||\mathbf{P}||_\infty$ of the $\mathbf{P}$ matrix be minimized. This norm [10] is defined by

$$||\mathbf{P}||_\infty = \max_i \sum_{j=1}^{n} |\rho_{ij}|. \tag{15}$$

Since it is the natural norm induced by the $\mathbf{L}_\infty$-norm for vectors, defined by

$$||\mathbf{V}||_\infty = \max_i |v_i| \tag{16}$$

this norm is such that

$$\frac{||\mathbf{V}_-||_\infty}{||\mathbf{V}_+||_\infty} \le \max_{\mathbf{V}_+ \neq 0} \frac{||\mathbf{V}_-||_\infty}{||\mathbf{V}_+||_\infty} = ||\mathbf{P}||_\infty. \tag{17}$$

Thus, minimizing $||\mathbf{P}||_\infty$ amounts to minimizing $||\mathbf{V}_-||_\infty / ||\mathbf{V}_+||_\infty$.

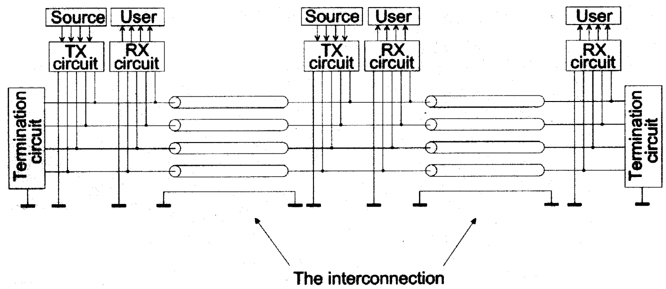

Fig. 2. New method for crosstalk reduction applied to the interconnection of Fig. 1.

All these quantities are defined as a function of frequency. However, in the special case where losses are neglected, they become frequency-independent, and the pseudo-matched impedances are resistances.

In practical electronic design, the maximum amplitude of every wanted incident signal is usually known, and it is important to keep the maximum level of the unwanted reflected wave as low as possible, because it can readily be compared with noise margins. The fourth way of computing pseudo-matched impedances, based on a minimization of $||\mathbf{P}||_\infty$, looks the most appropriate design option for this purpose. However, the elements of $\mathbf{P}$ capable of taking on the largest values being obviously the diagonal elements, minimizing $||\mathbf{P}||_\infty$ somehow puts the emphasis on the reduction of echo, and the termination so designed is likely to create crosstalk. The second way of computing pseudo-matched impedances is of course the worse in this respect. As a consequence, when a low crosstalk is important, we might prefer not to use these two definitions of pseudo-matched impedances, and rather minimize $||\mathbf{P}||_1$.

Note that, in order to limit the signal power consumed by each grounded linear two-terminal circuit element, designers often replace it with two resistors respectively connected to ground and to the power supply voltage of a digital circuit (split termination), or with a resistance in series with a capacitor [11].

IV. New Method for Crosstalk and Echo Reduction

A. *Definition of the New Method*

A new method for transmission with reduced crosstalk [12] through an interconnection with n transmission conductors and a reference conductor is shown on the example of Fig. 2. It provides, in a given frequency band, n transmission channels each corresponding to a signal to be transferred between the input of at least one transmitting circuit and the output of at least one receiving circuit. The method comprises the following steps.

- Modeling the interconnection, taking into account the lumped impedances seen by the interconnection and caused by the circuits connected to it elsewhere than at its ends, as a uniform MTL.
- Determining, in the frequency band of interest, the characteristic impedance matrix $\mathbf{Z}_C$ and a transition matrix from modal electrical variables to natural electrical variables (i. e. $\mathbf{S}$ or $\mathbf{T}$).
- Placing at at least one end of the interconnection a termination circuit having an impedance matrix approximating $\mathbf{Z}_C$ (i.e., a matched termination circuit).
- Combining the input signals in a transmitting circuit according to linear combinations defined by the transition matrices from modal electrical variables to natural electrical variables, so as to obtain at the output (connected to the interconnection) of this transmitting circuit the generation of modal electrical variables, each being proportional to one of the input signals.
- Combining in a receiving circuit, the input of which is connected to the interconnection, the signals present on the transmission conductors, according to linear combinations defined by the inverse of the transition matrices from modal electrical variables to natural electrical variables, so as to obtain at the output of this receiving circuit n output signals each corresponding to one of the transmission channels, each of these signals being proportional to one of the modal electrical variables.

The new method uses a superposition of waves being each composed of a unique modal electrical variable corresponding to a single channel, because these waves have the following properties mentioned in Section II: firstly the wave of a modal electrical variable propagates along the MTL without being coupled to other modal electrical variables of a different index, secondly at one end of the MTL connected to a matched termination circuit, the wave of a modal electrical variable is absorbed, without giving rise to any significant reflected wave.

These properties show that the propagation of waves each corresponding to a single modal variable, produced with a suitable conversion in one of the transmitting circuits and used with an inverse conversion in one of the receiving circuits, enables transmission without crosstalk between the channels. For this reason the new method will be called ZXtalk, for "zero crosstalk", though it also provides "zero echo."

In Fig. 2, any of the n natural voltages or natural currents being a linear combination of the n modal voltages or modal currents respectively, it appears that the value of a natural electrical variable *a priori* depends on the value of each of the signals present on each of the n channels. This is radically different from the behavior obtained in Fig. 1. In other words, the new method uses one modal electrical variable (therefore several transmission conductors) for each channel, whereas previous methods used one transmission conductor (therefore several modal electrical variables) for each channel.

The circuit in Fig. 2 implements a data bus architecture, though the signals and the circuits needed to control the active state of at most one transmitting circuit at a given time are not shown. This data bus architecture suggests that a digital circuit is represented. However, the new method is applicable to analog signals and to digital signals. In Fig. 2, two termination circuits are used because waves coming from the interconnection may be incident on both ends. When this is not the case, it is possible to use only one termination circuit.

The linear combinations realized in a transmitting circuit and the linear combinations realized in a receiving circuit may be implemented by an analog circuit, or by digital processing, D/A conversion and A/D conversion [13]. However, in the schematics presented below, for simplicity, we will only use analog circuit elements to perform these linear combinations.

B. Explanations on the Uniformity of the MTL

In order that the new method provides the desired characteristics, it is important that the interconnection behaves like a uniform MTL, because an inhomogeneity such as a variation of the characteristic impedance matrix with respect to z, may produce detrimental couplings between the channels, that is to say, crosstalk. Also, the circuits connected to the interconnection should be such that they do not disturb the propagation along the interconnection. We see that this result can for instance be obtained by:

- using transmitting circuits and/or receiving circuits connected in series with the conductors of the interconnection, and presenting a low series impedance to the interconnection;
- using transmitting circuits and/or receiving circuits connected in parallel with the conductors of the interconnection, and presenting a high parallel impedance to the interconnection.

In Fig. 2 for instance, the transmitting circuits and the receiving circuits, being connected in parallel with the interconnection, must present a high impedance to the interconnection.

C. On the Losses of the MTL

It is in many cases possible to consider that, when computing the matrices $\mathbf{Z}_C$, $\mathbf{S}$ and $\mathbf{T}$ of the MTL, the losses are negligible in some frequency bands, for instance for frequencies greater than 100 kHz. In this case, we consider that $\mathbf{Z}_C$ is real and frequency-independent and the matrices $\mathbf{S}$ and $\mathbf{T}$ chosen may be real and frequency-independent. Consequently, the matrix $\mathbf{Z}_C$ may be realized with a network of $n(n+1)/2$ resistors or less. At lower frequencies, losses are often not negligible and the matrix $\mathbf{Z}_C$ cannot be considered as real, which obviously leads to a more complex implementation of the new method. However, this question can often be disregarded, because the crosstalk and echo at low frequencies may in many cases be ignored.

However, the frequency above which losses may eventually be neglected for the computation of the matrices $\mathbf{Z}_C$, $\mathbf{S}$ and $\mathbf{T}$ of the MTL depends on the size of the conductors, on the distance between the conductors and, of course, on their conductivity. We note that in the case of on-chip interconnects, this frequency may be above 1 GHz [14].

V. Design Equations

A. Introduction to the Proportioning Problem

We are going to establish some equations for proportioning a circuit according to the ZXtalk method. Since many implementations are possible, we need to introduce a parameter a equal to the number (1 or 2) of termination circuits.

- If only one end of the interconnection is connected to a termination circuit with an impedance matrix approximating $\mathbf{Z}_C$, then the other end is connected to the transmitting circuit (there is a single transmitting circuit in this case), consequently the output of the transmitting circuit sees an impedance matrix near $\mathbf{Z}_C = a\mathbf{Z}_C = \mathbf{Z}_C/a$, with $a = 1$.
- If both ends of the interconnection are connected to a termination circuit with an impedance matrix approximating $\mathbf{Z}_C$, then $a = 2$ and the output of a transmitting circuit sees an impedance matrix near $\mathbf{Z}_C/a$ if the transmitting circuit is connected in parallel with the interconnection, or near $a\mathbf{Z}_C$ if the transmitting circuit is connected in series with the interconnection.

Let $\mathbf{X}_I$ be the column vector of the n input signals $x_{I1}, \ldots, x_{In}$ of a transmitting circuit, and let $\mathbf{X}_O$ be the column vector of the n output signals $x_{O1}, \ldots, x_{On}$ of a receiving circuit. These signals may for instance be voltages or currents. We are looking for relationships between $\mathbf{X}_I$ and the voltages or currents at the output of a transmitting circuit, and relationships between the voltages or currents received by a receiving circuit and $\mathbf{X}_O$. The reason for seeking these relationships is that they define the linear combination to be performed in transmitting circuits and receiving circuits.

B. If the Modal Electrical Variables are Voltages

Let us first use modal voltages as modal electrical variables. There is a proportionality between each modal voltage produced by a transmitting circuit and the input signal of the corresponding channel. Using a suitable numbering of the input signals, we may therefore write:

$$\mathbf{V}_M = \mathrm{diag}_n\left(\alpha_1, \ldots, \alpha_n\right) \mathbf{X}_I \tag{18}$$

where $\mathbf{V}_M$ is the column vector of the modal voltages produced by the transmitting circuit, and where $\mathrm{diag}_n(\alpha_1 \ldots, \alpha_n)$ is the diagonal matrix of the nonzero proportionality coefficients α_i. The dimensions of each of these coefficients depend on the dimensions of the input signals: if for instance these input signals are voltages, the coefficients α_i will be dimensionless. Therefore, using (4), we see that the transmitting circuit must produce, on each conductor, at its point of connection to the interconnection, the natural voltages of the column vector $\mathbf{V}$ given by

$$\mathbf{V} = \mathbf{S}\, \mathrm{diag}_n\left(\alpha_1, \ldots, \alpha_n\right) \mathbf{X}_I. \tag{19}$$

Equation (19) may be used to define a transmitting circuit connected in series with the conductors of the interconnection, providing a low series impedance. In this case, we note that the column vector $\mathbf{V}_T$ of the voltages across each pair of output terminals of the transmitting circuit must be

$$\mathbf{V}_T = \pm a\mathbf{S}\, \mathrm{diag}_n\left(\alpha_1, \ldots, \alpha_n\right) \mathbf{X}_I \tag{20}$$

in which the sign depends on the side of the interconnection with respect to the point of connection, where the value given by (19) is expected. This is because in the case $a = 2$, $\mathbf{V}_T$ gives rise to two voltage waves of opposite polarity propagating in opposite directions, absorbed at each end of the interconnection.

If, on the contrary, a transmitting circuit is connected in parallel with the conductors of the interconnection, showing a high parallel impedance, the designer might prefer to consider its output terminals as current sources. If $\mathbf{I}_T$ is the column vector of the current injected by the output terminals of the transmitting circuit, using (19), we get

$$\mathbf{I}_T = a\mathbf{Z}_C^{-1}\mathbf{S}\,\mathrm{diag}_n\left(\alpha_1,\ldots,\alpha_n\right)\mathbf{X}_I. \tag{21}$$

Moreover, given that, for each channel, a receiving circuit produces at its output a signal proportional to the modal voltage corresponding to this channel, we may write

$$\mathbf{X}_0 = \mathrm{diag}_n\left(\beta_1,\ldots,\beta_n\right)\mathbf{V}_M \tag{22}$$

where $\mathbf{V}_M$ is the vector of the modal voltages received by the receiving circuit, and where $\mathrm{diag}_n(\beta_1\ldots\beta_n)$ is the diagonal matrix of the nonzero proportionality coefficients β_i. The dimensions of these coefficients depend on the dimensions of the output signals: if for instance the output signals are currents, the β_i will have the dimensions of admittance. We see that the receiving circuit must read the set of conductors, to obtain the modal voltages by applying (4). Therefore, if, at the connection point of the receiving circuit to the interconnection, the vector of the natural voltages is $\mathbf{V}$, the output signals are given by

$$\mathbf{X}_0 = \mathrm{diag}_n\left(\beta_1,\ldots,\beta_n\right)\mathbf{S}^{-1}\mathbf{V}. \tag{23}$$

A receiving circuit connected in parallel with the interconnection and showing a high parallel impedance may directly acquire the natural voltages on the different conductors, and then perform a processing specified by (23). Alternatively, a receiving circuit may be connected in series with the interconnection and show a low series impedance. In this case, it would acquire the natural current on each conductor. If $\mathbf{I}$ is the column vector of these natural currents, we get

$$\mathbf{X}_0 = \pm\mathrm{diag}_n\left(\beta_1,\ldots,\beta_n\right)\mathbf{S}^{-1}\mathbf{Z}_C\mathbf{I} \tag{24}$$

where the sign depends on the direction of arrival.

C. If the Modal Electrical Variables are Currents

If we now use modal currents as modal electrical variables, we get, for a transmitting circuit

$$\mathbf{I}_M = \mathrm{diag}_n\left(\lambda_1,\ldots,\lambda_n\right)\mathbf{X}_I \tag{25}$$

where $\mathbf{I}_M$ is the column vector of the modal currents produced by the transmitting circuit and where $\mathrm{diag}_n(\lambda_1,\ldots,\lambda_n)$ is the diagonal matrix of the nonzero proportionality coefficients λ_i. The dimensions of each of these coefficients depend on the dimensions of the input signals. Using (4), we see that the transmitting circuit must produce, on each conductor, at its point of connection to the interconnection, the natural currents of the column vector $\mathbf{I}$ given by

$$\mathbf{I} = \mathbf{T}\,\mathrm{diag}_n\left(\lambda_1,\ldots,\lambda_n\right)\mathbf{X}_I. \tag{26}$$

Equation (26) could be used to specify a transmitting circuit connected in parallel with the interconnection, and presenting a high parallel impedance. In this case, the column vector $\mathbf{I}_T$ of the current injected by the output terminals of the transmitting circuit must be

$$\mathbf{I}_T = \pm a\mathbf{T}\,\mathrm{diag}_n\left(\lambda_1,\ldots,\lambda_n\right)\mathbf{X}_I \tag{27}$$

in which the sign depends on the side of the interconnection with respect to the point of connection, where the value given by (26) is expected. This is because in the case $a = 2$, $\mathbf{I}_T$ gives rise to two current waves of opposite polarity propagating in opposite directions, absorbed at each end of the interconnection.

If, on the contrary, a transmitting circuit is connected in series with the conductors of the interconnection and presents a low series impedance, the designer might prefer to consider its output as voltage sources. If $\mathbf{V}_T$ is the column vector of the voltages across each pair of output terminals of the transmitting circuit, we have

$$\mathbf{V}_T = a\mathbf{Z}_C\mathbf{T}\,\mathrm{diag}_n\left(\lambda_1,\ldots,\lambda_n\right)\mathbf{X}_I. \tag{28}$$

Moreover, since a receiving circuit produces at its output, for each channel, a signal proportional to the modal current corresponding to this channel, we get

$$\mathbf{X}_0 = \mathrm{diag}_n\left(\mu_1,\ldots,\mu_n\right)\mathbf{I}_M \tag{29}$$

where $\mathbf{I}_M$ is the vector of the modal currents received by the receiving circuit, and where $\mathrm{diag}_n(\mu_1,\ldots,\mu_n)$ is the diagonal matrix of the nonzero proportionality coefficients μ_i. The dimensions of these coefficients depend on the dimensions of the output signals. We see that the receiving circuit must read the set of conductors, to obtain the modal currents by applying (4). Therefore, if, at the connection point of the receiving circuit to the interconnection, the column vector of the natural currents is $\mathbf{I}$, the output signals are given by

$$\mathbf{X}_0 = \mathrm{diag}_n\left(\mu_1,\ldots,\mu_n\right)\mathbf{T}^{-1}\mathbf{I}. \tag{30}$$

A receiving circuit connected in series with the conductors of the interconnection and presenting a low series impedance may directly acquire the natural currents on the different conductors, and then perform a processing specified by (30). Alternatively, a receiving circuit may be connected in parallel with the conductors of the interconnection and present a high parallel impedance. In this case, it would acquire the natural voltage on each conductor. If we note $\mathbf{V}$ the column vector of these natural voltages, we get

$$\mathbf{X}_0 = \pm\mathrm{diag}_n\left(\mu_1,\ldots,\mu_n\right)\mathbf{T}^{-1}\mathbf{Z}_C^{-1}\mathbf{V} \tag{31}$$

where the sign depends on the direction of arrival.

D. Propagation of Signals

Given that, according to the new method, the waves propagate on the interconnection as they would in a uniform MTL, without significant reflection at the ends, it is possible, using (18) and (22) or (25) and (29), to clarify how the transmission of signals takes place. Between a transmitting circuit and a receiving circuit whose connection points to the interconnection show a difference of curvilinear abscissa ΔL, for any integer i between 1 and n, we obtain

$$x_{Oi} = \alpha_i\beta_i e^{-\gamma_i|\Delta L|}x_{Ii} \tag{32}$$

if modal voltages have been used, or

$$x_{Oi} = \lambda_i\mu_i e^{-\gamma_i|\Delta L|}x_{Ii} \tag{33}$$

if modal currents have been used.

VI. Relation With Associated Eigenvectors

Let us consider (21) where the coefficients α_i are arbitrary, and (27) where the coefficients λ_i are given by

$$\lambda_i = \pm j\omega c_K\frac{\alpha_i}{\gamma_i} \tag{34}$$

where c_K is an arbitrary nonzero scalar, which may depend on the frequency, and which has the dimensions of p.u.l. capacitance. We note that (34) is a reasonable choice, since the comparison of (21) and (27) shows that α_i/λ_i must have the dimensions of resistance. Since (21) and (27) are valid for any value of $\mathbf{X}_I$, we get

$$\mathbf{T} = \pm\mathbf{Z}_C^{-1}\mathbf{S}\,\mathrm{diag}_n\left(\frac{\alpha_1}{\lambda_1}, \ldots, \frac{\alpha_n}{\lambda_n}\right) = \frac{\mathbf{1}}{j\omega c_K}\mathbf{Z}_C^{-1}\mathbf{S}\mathbf{\Gamma}. \quad (35)$$

Then, using (5), we get

$$\mathbf{S} = j\omega c_K \mathbf{Y}^{-1}\mathbf{T}. \quad (36)$$

This result is identical to (9). It shows that associated eigenvectors are somehow embedded in the use of modal electrical variable for transmission, even though associated eigenvectors are not mentioned or implied in the definition of the new method. There is another way of looking at this. Once a matrix $\mathbf{S}$ or $\mathbf{T}$ is determined according to the second step of the new method, an associated transition matrix can be computed with (36). At this point, (10) is applicable. Therefore, in the definition of the new method

- it is physically equivalent that a transmitting circuit generates "modal voltages, each being proportional to one of the input signals", or "modal currents, each being proportional to one of the input signals;"
- it is physically equivalent that a receiving circuit delivers n output signals, "each being proportional to one of the modal voltages", or "each being proportional to one of the modal currents."

As a result, the use of either currents or voltages as electrical variables is without physical effect. From the standpoint of design however, it could be more convenient to use currents or voltages, depending on the type of device selected to implement the method, as explained in the Section V.

VII. First Example (Three Conductors)

In this first example, we consider a 30-cm-long interconnection on a PCB, having two transmission conductors, in a frequency range where it can be considered as lossless. It is made of two parallel traces over a ground plane, the traces being 0.127 mm in width and 35.6 μm in thickness. The substrate is 1.19-mm thick, with $\varepsilon_r = 4.7$, and the line spacing is 0.254 mm. The p.u.l. parameters of this interconnection and time-domain simulation results can be found in [15] and [16]. We compute

$$\mathbf{Z}_C = \begin{pmatrix} 147.187 & 60.1923 \\ 60.1923 & 147.187 \end{pmatrix}\Omega \quad (37)$$

$$\mathbf{S} = \begin{pmatrix} 1.0912 & 2.4616 \\ -1.0912 & 2.4616 \end{pmatrix}$$

$$\mathbf{T} = \begin{pmatrix} 0.707\,11 & 0.707\,11 \\ -0.707\,11 & 0.707\,11 \end{pmatrix}. \quad (38)$$

In (38), the matrices $\mathbf{S}$ and $\mathbf{T}$ are associated, for a value of the arbitrary p.u.l. capacitance c_K defined by (9) equal to 10^{-10} F/m, in line with the usual practice of the Authors [9]. The propagation velocities of the two modes are about 0.1774 and 0.1679 m/ns. We want to compare the crosstalk and echo obtained with optimal pseudo-matched impedances, to the one obtained with the new method.

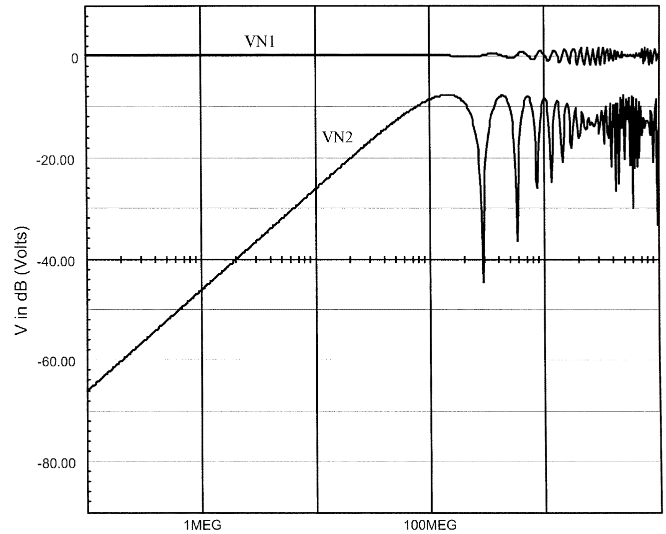

Fig. 3. Near-end signal VN1 and near-end crosstalk signal VN2.

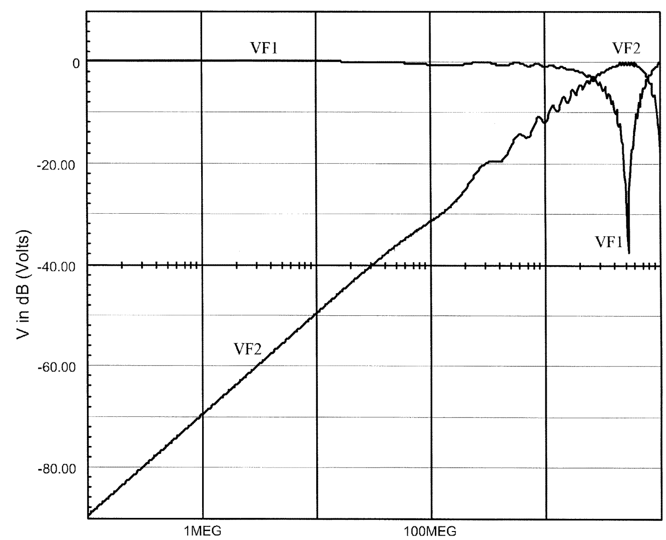

Fig. 4. Far-end signal VF1 and far-end crosstalk signal VF2.

According to Section II, we can use for the pseudo-matched impedances either the value 147.2 Ω of the diagonal elements of $\mathbf{Z}_C$, or the value 134.3 Ω, for which $||\mathbf{P}||_1$ and $||\mathbf{P}||_\infty$ are minimized and $||\mathbf{P}||_1 = ||\mathbf{P}||_\infty \approx 0.21$. We note that, following Jarvis [17], Paul [15] used pseudo-matched impedances of $(L_{11}/C_{11})^{0.5} = 135.8\ \Omega$. Using a resistor of 134.3 Ω between each end of each transmission conductor and ground, we have computed with SPICE the near-end and the far-end crosstalk signals in the frequency domain from 100 kHz to 10 GHz, shown in Figs. 3 and 4.

The schematic for the SPICE simulation of a circuit implementing the new method with this interconnection is shown in Fig. 5. At the ends of the MTL, we find identical circuits comprising each a transmitting circuit, a receiving circuit and a termination circuit. Consequently, bidirectional transmission is possible.

In Fig. 5, the ends of the interconnection are each connected to a termination circuit made of three resistors R401, R402 and R403. The value of R401 and R402 is 87 Ω and the value of

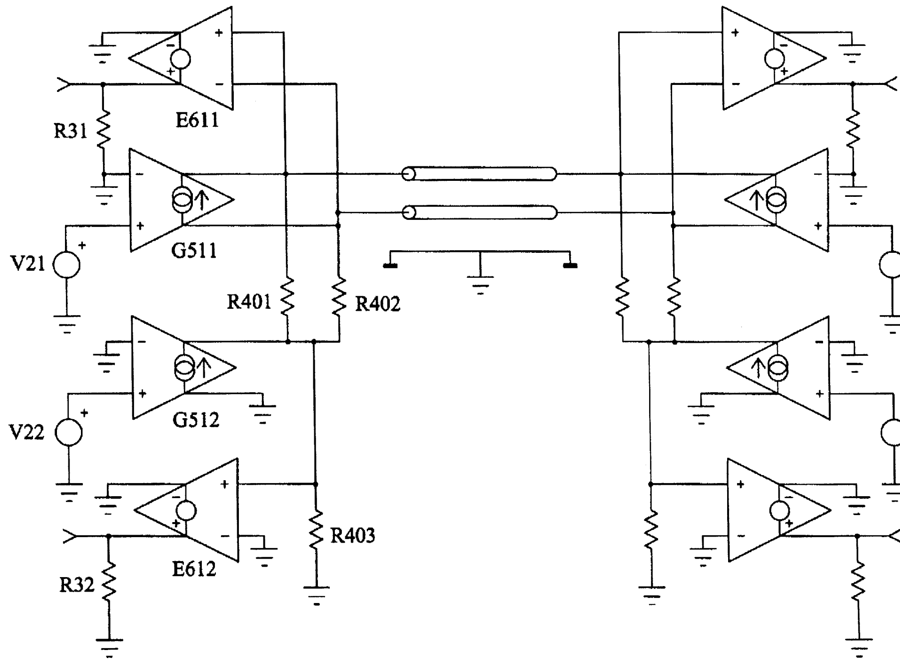

Fig. 5. First theoretical circuit for the implementation of the new method.

R403 is 60.2 Ω, because these values produce an impedance matrix very close to $\mathbf{Z}_C$ given by (37). Each transmitting circuit has parts in common with a termination circuit and comprises only two voltage-controlled current sources (VCCS) as its own circuit elements, labeled G511 and G512 on the left. The transmitting circuits receive at their input the signals of the two channels of the sources, each represented by two voltage sources, labeled V21 and V22 on the left. These transmitting circuits produce modal voltages such that each of them is proportional to the signal of one of the voltage sources. It is therefore necessary that only one of the two sources may be active at a given time, unless an active hybrid circuit is added to allow full duplex operation. The two receiving circuits have parts in common with each of the termination circuits and only comprise two voltage-controlled voltage sources (VCVS) as their own circuit elements, labeled E611 and E612 on the left hand side. These receiving circuits produce, on their two output channels connected to the users each represented by two resistors, labeled R31 and R32 on the left, two signals being each proportional to a modal voltage. The signals of the two channels of an active source are sent to the two channels of the two users.

In order to show some crosstalk (otherwise displaying dBs is a problem), we created some imbalance in the circuit by using a value of 87.1 Ω for R402. The SPICE simulations for instance produced the near-end voltage values shown in Fig. 6, when only V21 is active. The computed crosstalk coupling factor is independent of frequency and near −80 dB, and the voltages are in phase with V21. Time-domain waveforms across R31 and R32 are therefore identical to the one produced by V21, attenuated by the values shown in Fig. 6. The result obtained for the other transmission channel (i.e., when only V22 is active), and at the far-end are also frequency-independent crosstalk coupling factors, and the highest crosstalk coupling factors are about −70 dB. The phase difference with V21 being

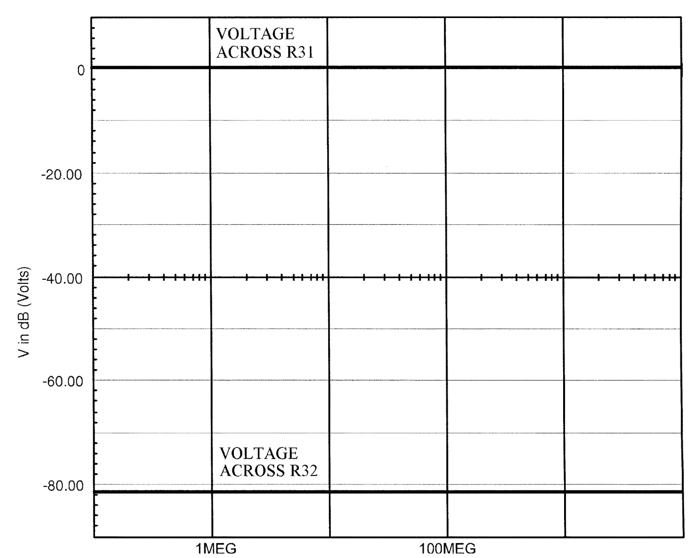

Fig. 6. Near-end signal across R31 and near-end crosstalk signal across R32.

always linear with respect to frequency (resulting only from the propagation time), time-domain waveforms are not distorted.

These results and their comparison with Figs. 3 and 4, show that crosstalk has been effectively reduced. Both crosstalk coupling factors are independent of the length of the interconnection, and may therefore be compared with the time-domain simulation provided by Paul, shown in [15, Fig. 6]. He calls “matched termination” a termination just made of grounded 135.8 Ω resistors, observing that grounded resistors cannot be considered as truly matched terminations. He refers to his book [1], in which he explains in § 5.2.6.1 that truly matched terminations unavoidably produce a large crosstalk. The theoretical interest of the ZXtalk method is clearer now: it proves that, contrary to prior beliefs, we can practically eliminate crosstalk

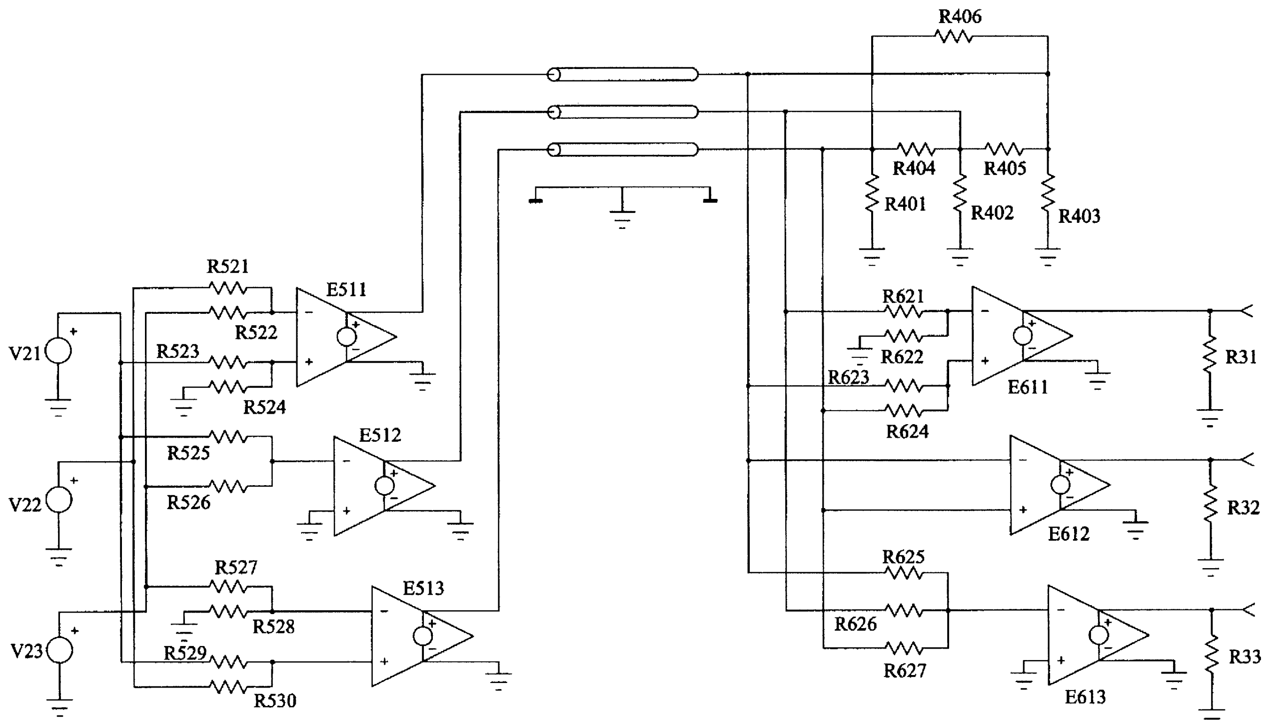

Fig. 7. Second theoretical circuit for the implementation of the new method.

(and echo) using truly matched terminations such as the one used in Fig. 5, provided they are implemented in combination with suitable transmitting circuits and receiving circuits.

VIII. Second Example (4 Conductors)

In this second example, we consider a 30-cm-long interconnection for which losses may be neglected at the frequencies used for transmission, having three transmission conductors and a reference conductor. This interconnection is defined by [16, (10) and (11)], but more information can be found in [18], including plots of time-domain signals obtained when it is used for single-ended transmission.

A SPICE simulation circuit for implementing the new method with this interconnection is shown in Fig. 7. This implementation is intended for unidirectional transmission only. In Fig. 7, only one end of the interconnection is connected to a termination circuit made of six resistors R401 to R406, with $\mathrm{R401} = \mathrm{R403} = 58.7\ \Omega$, $\mathrm{R402} = 69.2\ \Omega$, $\mathrm{R404} = \mathrm{R405} = 289.5\ \Omega$ and $\mathrm{R406} = 2781\ \Omega$. These values produce an impedance matrix very close to $\mathbf{Z}_C$. However, removing R406 would probably provide a good enough approximation in many cases [19]. The transmitting circuit comprises three VCVS labeled E511, E512, and E513 and ten resistors R521 to R530. It receives at its input the signal of the three channels of the source, represented by the voltage sources V21, V22, and V23. The receiving circuit comprises three VCVS labeled E611, E612, and E613 and seven resistors R621 to R627, which must not prevent the interconnection from seeing a matched termination. The proportioning of the unspecified circuit elements for this circuit is not difficult and may be based on (20) and (23). With a suitable proportioning, the transmitting circuit produces three modal voltages, each being proportional to the signal on one of the input channels, and the receiving circuit produces on the three output channels connected to the user represented by the resistors R31, R32, and R33, three signals being each proportional to a different modal voltage.

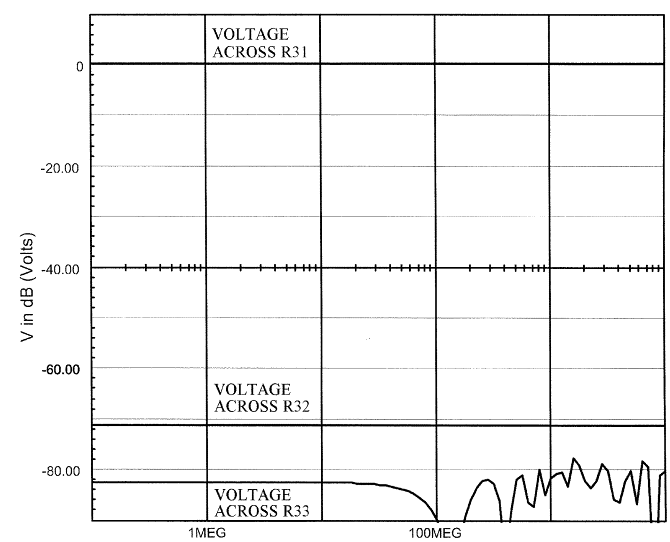

Fig. 8. Far-end signal across R31 and far-end crosstalk signals across R32 and R33.

A simulation result for a signal produced by V21 is shown in Fig. 8, which illustrates that crosstalk is very low, from 100 kHz to 10 GHz. A netlist for this simulation is given in [16]. The simulation results for the two other channels are similar. Transient simulation results (not shown) tell us that a 1-V step produced by V21 will produce a 1-V step across R31 and a crosstalk below 280 μV peak-to-peak across R32 and R33, in line with the coupling below -70 dB shown on Fig. 8. Anyhow, such numerical

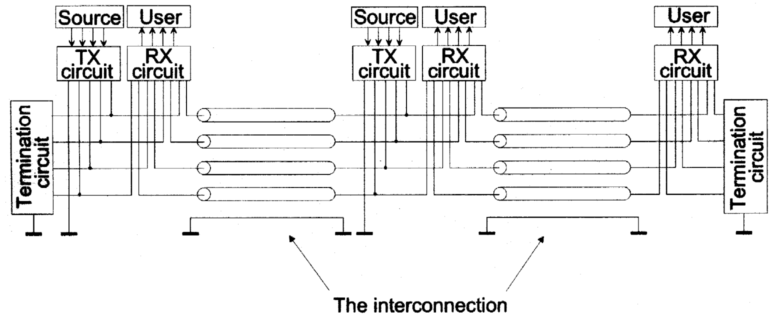

Fig. 9. Setup in which no nontrivial linear combination is needed.

results are not the consequence of a physical limitation of the new method, since theory says that crosstalk and echo are canceled in an ideal implementation. They are merely the effect of the number of digit entered for the part values in the simulation, and of the computational accuracy.

IX. Completely Degenerate MTL

When the propagation constants of the different propagation modes may be considered as equal, the MTL is said to be completely degenerate. In this case, (2) shows that the transition matrices from modal electrical variables to natural electrical variables ($\mathbf{S}$ or $\mathbf{T}$) used in the new method may be chosen equal to the identity matrix of order n. As a consequence, the linear combinations to be performed in the transmitting circuits and/or in the receiving circuits may become trivial [20], a linear combination being called "trivial" when it is merely the product of only one signal by a coefficient.

If we wish to build a device for implementing the ZXtalk method in which neither the transmitting circuits nor the receiving circuits perform nontrivial linear combinations, we should:

- when designing the transmitting circuits, use (20) if the electrical variables used in the new method are voltages, or use (27) if these electrical variables are currents;
- when designing the receiving circuits, use (23) if the electrical variables used in the new method are voltages, or use (30) if these electrical variables are currents.

As an example, we have represented in Fig. 9 an implementation of the new method, in which the electrical variables are currents. This implementation is similar to the one shown in Fig. 2, but it uses a completely degenerate interconnection, and the receiving circuits are connected in series with the interconnection. They must present a low series impedances to the interconnection. The transmitting circuits operate according to (27) and the receiving circuits operate according to (30). As a consequence, only trivial linear combinations are required. This feature is of course desirable, since in this case the number of parts is proportional to n, which allows operation at higher frequencies.

X. Comparison With Other Modal Transmission Schemes

The idea of using propagation modes for transmitting in n channels using n transmission conductors is not new. This concept of modal transmission was already suggested in a footnote of an IEEE paper [21] published in 1954, and later in a document [22] published in 1990. In fact both references use the same idea: "there is no crosstalk between modes as there is between nonmodal propagation" and they do not specify the terminations. Because of the latter, this idea is incorrect, since crosstalk is a phenomenon which occurs on interconnections of finite length, which are connected to some sort of termination at each end, and for which mismatch is a major potential cause of crosstalk.

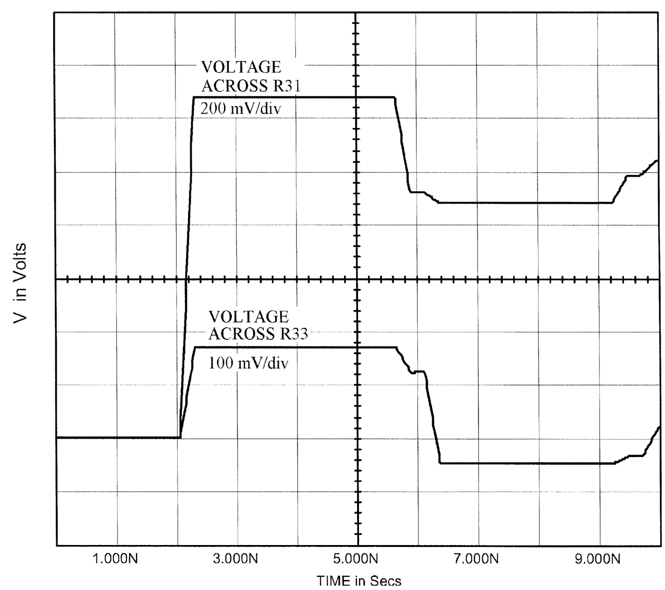

Fig. 10. Far-end signal across R31 and far-end crosstalk across R33, obtained with modal transmission, but without matched terminations.

Let us for instance consider the device shown in Fig. 7, but instead of using a matched termination, let us use a 40-Ω resistor for R401, a 100-Ω resistor for R402, a 50-Ω resistor for R403, and let us remove R406, R404, and R405. The transmitting circuit still produces modal voltages proportional to the input signals, and the receiving circuit still produces output voltages proportional to the modal voltages. A 1-V step is produced by the source V21, with a (0% to100%) rise time of 250 ps. The results of the SPICE time-domain simulation are shown in Fig. 10. The signal transmitted to R31 (test point VF1) is heavily distorted by echo, and the crosstalk signal reaching R33 has a peak value of about 210 mV. Comparing this value to the 280 μV mentioned at the end of Section VIII demonstrates that using modal transmission without matched terminations (as defined in Section II) does not remove crosstalk.

Conversely, using matched termination without modal transmission does not remove crosstalk, on the contrary, as said at the end of Section VII. This is why this scheme is not implemented in equipments. The new method provides a drastic reduction of crosstalk and echo because it appropriately combines the use of modal variables *and* the use of termination having an impedance matrix equal to $\mathbf{Z}_C$, in spite of the fact that, used separately, these requirements do not effectively reduce crosstalk.

The assumption "in general, n conductors and ground have n orthogonal modes" is another foundation of [22]. This is a widespread misconception, since the eigenvectors of a MTL are not necessarily orthogonal, even if the MTL is lossless [5]. For instance, the interconnection of Section VIII has nonorthogonal eigenvectors in the matrix $\mathbf{S}$, and nonorthogonal eigenvectors in the matrix $\mathbf{T}$ [16], [18]. Nothing can be changed about it, since they correspond to nondegenerate eigenvalues.

XI. Conclusion

Thinking of propagation in multiconductor interconnection in terms of modal electrical variables is an obvious choice when *associated* eigenvectors are used and the corresponding total decoupling of the telegrapher's equations occurs, as explained at the end of Section II. We combined the approach of using one propagation mode for each channel with the ideal absorption provided by matched terminations, to obtain transmission without crosstalk and without echo, and with reduced distortion caused by propagation. An interesting result is that associated eigenvectors show up in the design equation, even though they were not required in the definition of the new method!

The losses of the interconnection and the frequency dependence of its parameters are taken into account in the analysis and in (1)–(36). The validity of the concept is therefore proved for real interconnections. However, for clarity and brevity, we have neglected losses in the examples.

The new ZXtalk method requires that the characteristics of the interconnection be taken into account in the design of the transmitting circuits, receiving circuits and termination circuits, but an important feature is that the values of circuit elements do not depend on the length of the interconnection, as far as the compensation of the losses and phase and/or amplitude distortion between transmitting circuits and receiving circuits (equalization) is not required. The linear combination performed in the transmitting circuits and in the receiving circuits can be built with a hardware implementing analog signal processing and/or digital signal processing.

In this paper, we have only presented theoretical examples of the new method, using the ideal circuit elements of SPICE, and lossless MTL SPICE models. The effectiveness of the new method has been verified with an experimental setup [16], with which we could compare the crosstalk obtained with optimal pseudo-matched terminations used in a conventional circuit, to the crosstalk obtained with an implementation of the ZXtalk method, using the same interconnection. There is also a need to investigate the requirements and limitations of the ZXtalk method. For instance, the influence of inhomogeneity and losses in the interconnection, and of the tolerances of circuit elements should be studied, as well as the dissipation in matched terminations. However, this question is complex because several implementations of the new method have been described, and because different types of applications (e.g., analog RF signals, modulated digital signals, baseband digital signals) have different requirements.

The new method is in principle applicable to all level of analog and digital interconnection, ranging from inter-unit wiring to on-chip interconnects. The variation of the new method implementing an interconnection having substantially identical propagation constants seems to be the best approach to obtain the largest bandwidth with the largest number of conductors.

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