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A Modal Transmission Technique Providing a Large Reduction of Crosstalk and Echo

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Abstract — **This paper is about a method for the reduction of crosstalk and echo in multiconductor interconnections, using special active receiving circuits and transmitting circuits. We first review the three basic concepts needed to explain this method: the different possible modal transforms and the properties of eigenvectors, the termination of each conductor with a grounded impedor having a pseudo-matched impedance, the termination of the interconnection with a (truly) matched termination which eliminates reflections. The new method is based on the use of such matched terminations and of one modal variable for each transmission channel. We then compare this new method with other concepts of modal transmission.**

I. Introduction

In this paper, we consider a generic multiconductor interconnection, which may be an on-chip interconnect, an on-board interconnection made of printed circuit board (PCB) traces, a board-to-board interconnection using for instance a ribbon cable or a backplane, or an inter-unit interconnection. The interconnection may be used for sending any type of signal: analog signal, baseband digital signal, modulated digital signal. The interconnection is used to obtain n transmission channels.

We want that the signal be not degraded by echo and crosstalk. In this paper, we 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 *reflection*, because reflections occurring for instance at an end of the interconnection may be responsible for echo and/or crosstalk.

Designers usually choose [1] between *single-ended transmission*, for which n transmission conductors plus a reference conductor are needed, and *differential transmission*, for which $2n$ transmission conductors are needed. Differential transmission is known to provide a better integrity of signal, especially when the line transmitters, the line receivers and the interconnection are well balanced.

A third type of transmission uses n transmission conductors plus a reference conductor and allocates a different propagation mode to each channel. This paper will define and discuss such *modal transmission* schemes, which are intended to provide very low echo and crosstalk.

II. Notations and Definitions

Let us consider an interconnection having uniform electrical properties over its length, comprising n transmission conductors and a reference conductor. We are therefore dealing with a $(n+1)$ conductors interconnection.

Using the model [2, ch. 6] of multiconductor transmission lines (MTL), echo and crosstalk can be computed from the knowledge of the devices connected to the interconnection and of the per-unit-length (p.u.l.) impedance matrix $\mathbf{Z} = \mathbf{R} + j\omega\mathbf{L}$ and p.u.l. admittance matrix $\mathbf{Y} = \mathbf{G} + j\omega\mathbf{C}$ of the interconnection, where ω is the angular frequency. Using a suitable definition of *natural currents* (the currents on the transmission conductors) and of *natural voltages* (the voltages between each transmission conductor and the reference conductor), the phenomena occurring on the MTL are described by the telegrapher's equations:

$$\begin{cases} \dfrac{d\mathbf{V}}{dz} = -\mathbf{Z}\,\mathbf{I} \\ \dfrac{d\mathbf{I}}{dz} = -\mathbf{Y}\,\mathbf{V} \end{cases} \tag{1}$$

where $\mathbf{I}$ is the column-vector of the natural currents $i_1,..., i_n$, where $\mathbf{V}$ is the column-vector of the natural voltages $v_1,..., v_n$, and where z is the abscissa along the interconnection.

This equation can easily be solved using a diagonalization of the matrices $\mathbf{ZY}$ and $\mathbf{YZ}$. The eigenvectors define the propagation modes, and the eigenvalues correspond to their propagation constants. We shall use $\mathbf{T}$ and $\mathbf{S}$ to denote two transition 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. The matrices $\mathbf{Z}$ and $\mathbf{Y}$ being symmetrical matrices, we observe that, if we first compute, with a diagonalization of the matrix $\mathbf{YZ}$, a matrix $\mathbf{T}$ satisfying the first line of (2), then

$$\mathbf{S} = {}^t\mathbf{T}^{-1} \tag{4}$$

is a solution of the second line of (2). Most authors [2, § 6.2.6.3] use (4) for solving (2), and therefore obtain bi-orthonormal eigenvectors [3]. However, (4) is not a necessary condition, and another possible choice to obtain a solution **S** of the second line of (2) is

$$\mathbf{S} = j\omega\, c_K\ \mathbf{Y}^{-1}\mathbf{T} \tag{5}$$

where c_K is an arbitrary non-zero scalar, which may depend on the frequency, and which has the dimensions of a p.u.l. capacitance. This choice is not as usual as (4), but it has for instance been used in [4] [5] and [6]. In order to indicate that a matrix **S** and a matrix **T** are defined by (2), (3) and (5) we shall say that they are *associated*, and that the eigenvectors contained in **S** and **T** (i.e. their column-vectors) are associated.

III. Matched Terminations and Pseudo-Matched Impedances

The characteristic impedance matrix $\mathbf{Z}_C$ of the MTL is defined [7] by

$$\mathbf{Z}_C = \mathbf{S}\,\Gamma^{-1}\,\mathbf{S}^{-1}\mathbf{Z} = \mathbf{S}\,\Gamma\,\mathbf{S}^{-1}\mathbf{Y}^{-1}$$
$$= \mathbf{Y}^{-1}\mathbf{T}\Gamma\mathbf{T}^{-1} = \mathbf{Z}\mathbf{T}\Gamma^{-1}\mathbf{T}^{-1} \tag{6}$$

where

$$\Gamma = \mathrm{diag}_n(\gamma_1, \ldots, \gamma_n) \tag{7}$$

is the diagonal matrix of order n of the propagation constants γ_i. At an end of the MTL, no reflection occurs for incident waves if and only if this end is connected to a (n+1)-terminal linear termination showing an impedance matrix equal to $\mathbf{Z}_C$. In this case, the termination is said to be *matched* to the MTL. When the transmission conductors are coupled by mutual capacitance and/or inductance, $\mathbf{Z}_C$ is a full matrix. As a consequence, a matched termination typically requires a network of n (n + 1)/2 resistors [8] when the MTL is assumed to be lossless.

In order to obtain the integrity of signals, one of the first rule is to reduce reflections. To this end, designers never use matched terminations (according to the above definition), because such terminations create crosstalk. A simulation showing this phenomenon is presented in [9, § 3], but the corresponding explanation is very simple. If for instance the far-end of the interconnection is terminated with a matched termination, the impedance matrix observed at the near-end by looking into the interconnection is equal to $\mathbf{Z}_C$. Since $\mathbf{Z}_C$ is a full matrix, a current injected in a given transmission conductor at the near-end produces a crosstalk voltage traveling toward the near-end, because of the non-zero non-diagonal matrix elements. A similar phenomenon takes place at the far-end.

Instead of using a matched termination, a common practice implements, for each transmission conductor, an impedor (i.e. a two-terminal linear circuit element) inserted between the transmission conductor and ground, having an impedance chosen in such a way that it reduces reflections in single-ended transmission. The phenomenon described above does not take place because the corresponding impedance matrix is diagonal. Though the impedances of the grounded impedors do not match the MTL, they may be called *pseudo-matched impedances*.

It should be noted that some authors improperly refer to grounded impedors having a pseudo-matched impedance as "matched terminations". In this paper, the impedance matrix of a matched termination is $\mathbf{Z}_C$.

IV. Modal Voltages and Modal Currents

Matrices **T** and **S** solutions of (2) and (3) 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 note $\mathbf{I}_M$ the vector of the n modal currents $i_{M\,1}, \ldots, i_{M\,n}$ and $\mathbf{V}_M$ the vector of the n modal voltages $v_{M\,1}, \ldots, v_{M\,n}$, we get (by definition):

$$\begin{cases} \mathbf{V} = \mathbf{S}\mathbf{V}_M \\ \mathbf{I} = \mathbf{T}\,\mathbf{I}_M \end{cases} \tag{8}$$

Consequently, we shall call **S** the *transition matrix from modal voltages to natural voltages*, and we shall call **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. We can point out that (8) also implies that for any integer α between 1 and 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.

When associated eigenvectors defined by (2), (3) and (5) are used, for a wave propagating in a given direction and for any integer α such that $1 \leq \alpha \leq n$, we have :

$$v_{M\alpha} = \frac{\varepsilon}{j\omega\ c_K}\,\gamma_\alpha\, i_{M\alpha} \tag{9}$$

ε being 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 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 [9] choosing associated eigenvectors provides 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 2-conductor transmission lines.

Such equivalent circuits can be used to solve problems involving interconnections, and to create simple SPICE circuits when losses can be neglected [5]. This is how the simulations of this paper have been obtained.

V. Losses and Properties of Eigenvectors

In the theory presented so far, we have made no assumption regarding losses in the interconnection. We have only assumed that we could diagonalize **YZ** at any relevant angular frequency ω. It can be shown that this is always possible when the MTL is lossless [7], and in this case **S** and **T** can be chosen real and frequency-independent. Diagonalizing **YZ** is usually also possible when the MTL is lossy, though in this case it might not be possible to define frequency-independent matrices for **S** and **T**. The reason for this is that, for a lossy MTL:

- if the electromagnetic fields propagate in an homogeneous medium, then **YZ** can be diagonalized [7] and there is only one (completely degenerate) propagation constant;
- if the electromagnetic fields propagate in an

inhomogeneous medium, we can find n different (non degenerate) propagation constants, in which case **YZ** is diagonalizable;
■ in an inhomogeneous medium we can also find the case where the characteristic equation has multiple roots, typically because of an intentional symmetry, but we have never seen the case where this had led to a non-diagonalizable **YZ**.

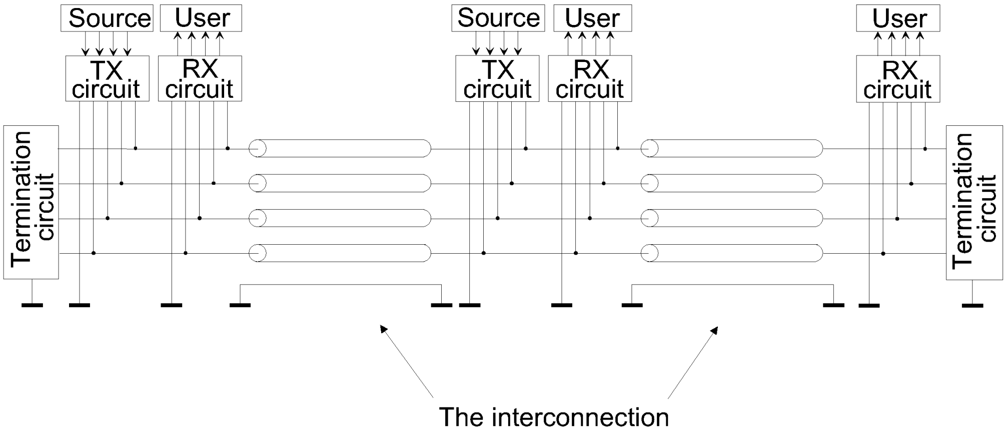

Fig. 1: the ZXtalk method applied to a bidirectional bus.

We also note that the possibility of obtaining a non-diagonalizable matrix **YZ** by chance, that is to say without intentional symmetry, corresponds to a singularity of no physical significance because of uncertainties. We will therefore exclude from our discussion the special circumstances in which **YZ** cannot be diagonalized. As a consequence, we consider that the modal currents and the modal voltages can always be defined by (2) and (3), regardless of losses.

In the following, we will not assume any particular relation between **S** and **T**, such as (4) or (5). It should be stressed that the eigenvectors of **S** are not necessarily orthogonal [3], nor the eigenvectors of **T**, this statements applying to the lossy case and to the lossless case. The reader should not be confused by unrelated properties like the orthogonality of the E and H modes of a lossless hollow waveguides, and their lack of orthogonality when the waveguide is lossy [10, § 5.2].

VI. A Method for Reducing Crosstalk and Echo

A new method for the reduction of crosstalk and echo called ZXtalk is implemented on the example shown in Fig. 1. It is applicable to interconnections with n transmission conductors which may be modeled as a uniform MTL with a sufficient accuracy, in a given frequency band. This method is mainly characterized by the following points [11]:
■ the interconnection is connected at at least one end to a termination circuit behaving as a matched termination;
■ one or several transmitting circuits ("TX circuit" in Fig. 1) combine the input signals generated by sources according to linear combinations defined by a transition matrix from modal electrical variables to natural electrical variables (i. e. **S** or **T**), the output of the transmitting circuit being connected to the n transmission conductors;
■ the n transmission conductors are connected to the input of at least one receiving circuit ("RX circuit" in Fig. 1), which combines 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, the receiving circuit providing at its output the signals for a destination.

The circuit of Fig. 1 implements a data bus architecture intended for bidirectional transmission, but the signals needed to control the active state of at most one transmitting circuit at a given time are not shown. We also note that the transmitting circuits and the receiving circuit being connected in parallel to the interconnection, they must show a high impedance to the interconnection, in order not to disturb the propagation and not to produce undesirable reflections.

The ZXtalk method uses a superposition of waves being each composed of a unique modal electrical variable corresponding to a single channel, because:
■ the wave of a modal electrical variable propagates along the MTL without being coupled to other modal electrical variables of a different index,
■ at an 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.

The signals of the n channels of a source connected to an active transmitting circuit are therefore sent to the n channels of the destinations, without noticeable crosstalk and echo.

There are many possible implementations for this method, which may use analog circuits and/or digital circuits [12]. We will not discuss real implementations in this paper. However, we will discuss a theoretical example of the ZXtalk using a 30 cm long lossless interconnection having three parallel transmission conductors, with:

$$\mathbf{L} = \begin{pmatrix} 313.9 & 67.5 & 22.2 \\ 67.5 & 319.3 & 67.5 \\ 22.2 & 67.5 & 313.9 \end{pmatrix} \text{nH/ m}$$

$$\mathbf{C} = \begin{pmatrix} 130.3 & -16.2 & -0.8 \\ -16.2 & 133.7 & -16.2 \\ -0.8 & -16.2 & 130.3 \end{pmatrix} \text{pF/ m}$$

We determined the matrices $\mathbf{Z}_C$, **S** and **T**, and the SPICE model of the MTL, using SpiceLine [6]. These matrices and results of time-domain simulations for this interconnection implemented in a standard single-ended transmission set-up, are provided in [13]. For instance, the results of a SPICE simulation corresponding to a 1 Volt step sent on conductor 2 is shown in Fig. 2. The transmitted signal VF2 is not significantly distorted, but the far-end crosstalk signal VF1 has a peak value of about 340 mV. This is a good example of a lossless MTL for which the eigenvectors of **S** are not orthogonal, nor the eigenvectors of **T**.

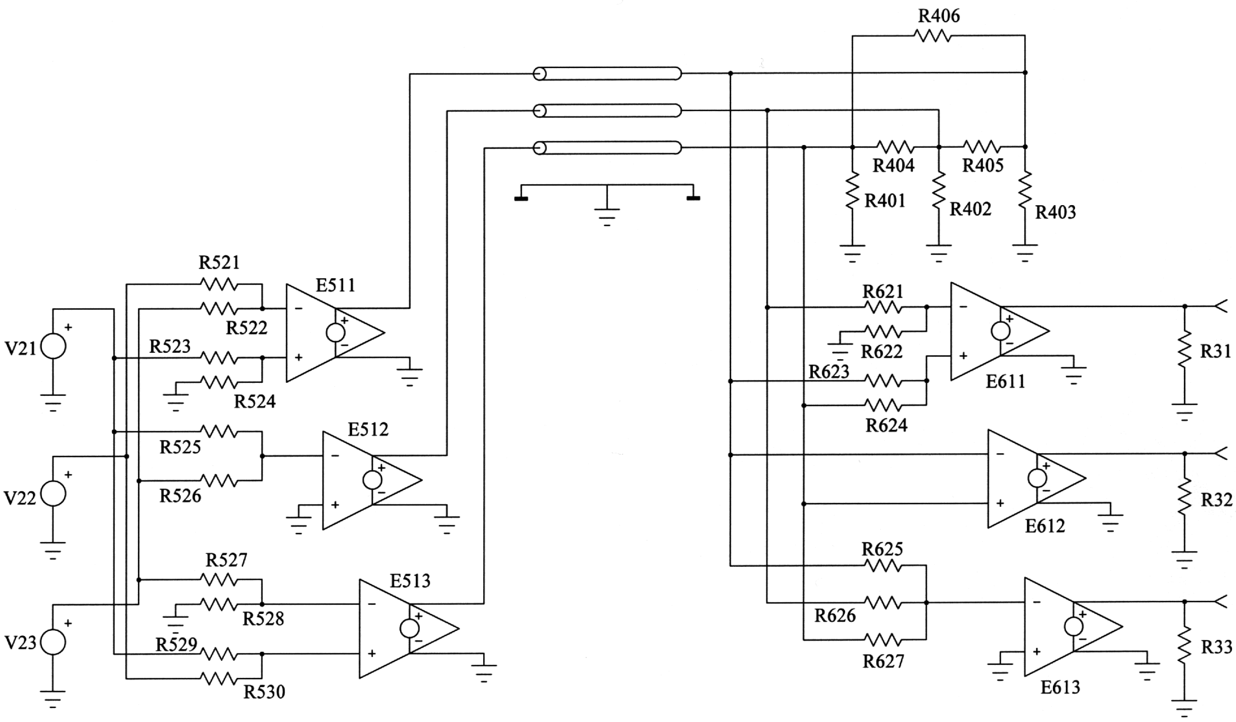

Fig. 3 : example of the ZXtalk using a 4-conductor interconnection for unidirectional transmission

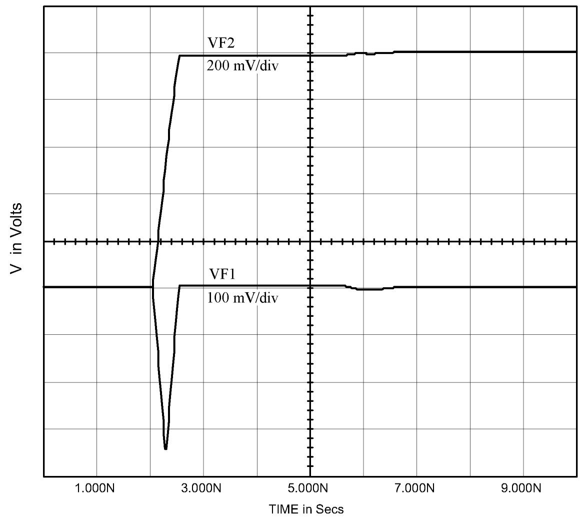

Fig 2: voltages at the far-end in a standard set-up.

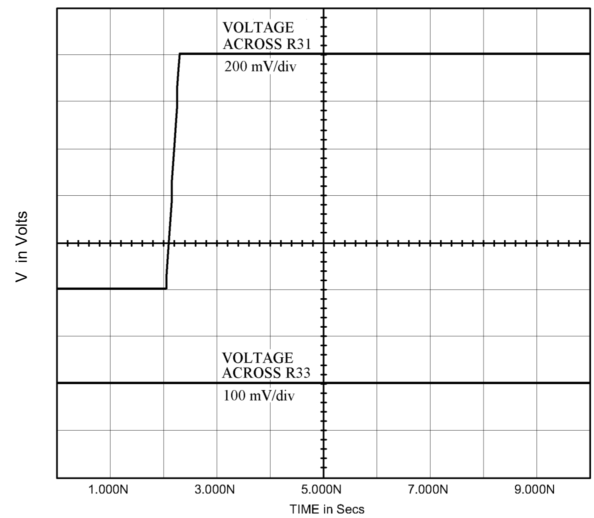

Fig 4: voltages at the far-end.

We show in Fig. 3 the schematic for the SPICE simulation of a theoretical example of the ZXtalk using the same 30 cm long lossless interconnection. In this simple example, the interconnection is intended for unidirectional transmissions. Only one end of the interconnection is connected to a termination circuit made of six resistors R401 to R406, their values being determined in such a way that the impedance matrix of the termination is close to the characteristic impedance matrix.

We find R401 = R403 = 58.7 Ω, R402 = 69.2 Ω, R404 = R405 = 289.5 Ω and R406 = 2781 Ω. The transmitting circuit comprises three voltage controlled voltage sources (VCVS) E511, E512 and E513, and 10 resistors R521 to R530. It receives at its input the signals of the three channels of the source represented by the voltage sources V21, V22 and V23. The receiving circuit comprises three VCVS E611, E612 and E613 and seven resistors R621 to R627. It delivers to the resistors R31, R32 and R33 the output signals of the three channels. With a suitable choice of part values and a (0% to100%) rise time of 250 ps, we compute waveforms such as the one shown in Fig. 4, which can be compared to Fig. 3. This shows that there is no crosstalk and no echo left.

The ZXtalk method is applicable to analog and digital signals. It may also be implemented in such a way that bidirectional transmission is obtained, in which case the near-end crosstalk and the far-end crosstalk vanish.

VII. Homogeneity and Losses in the ZXtalk

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 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.

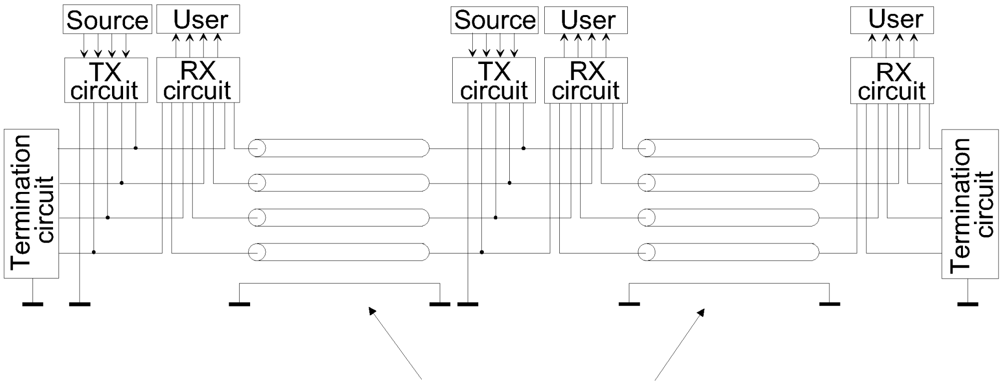

Fig. 5. A set-up in which no non-trivial linear combination is needed.

In Fig. 1 for instance, the transmitting circuits and the receiving circuits, being connected in parallel with the interconnection, must present a high impedance to the interconnection.

Considering the observations of § V, we see that the definition of the ZXtalk method is applicable to (ideal) lossless interconnections and to (real) lossy interconnections. Obviously, when losses are significant, implementations may become more complex. We note that, when computing the matrices $\mathbf{Z}_C$, **S** and **T** of the MTL, we may often consider that losses are negligible above some frequency, for instance above 1 MHz. In this case, $\mathbf{Z}_C$ is real and frequency-independent and the matrices (it can be realized with a network of resistors), and **S** and **T** chosen may be real and frequency-independent. 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.

VIII. Completely Degenerate Interconnection

When the propagation constants of the different propagation modes may be considered as equal, the MTL is said to be completely degenerate. This for instance occurs with an interconnection made of perfect conductors in an homogenous dielectric. When the MTL is completely degenerate, (2) shows that the transition matrices from modal electrical variables to natural electrical variables (**S** or **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 [14], 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 non-trivial linear combinations, we must apply an appropriate design structure.

As an example, we have represented in Fig. 5 an implementation of the new method, in which the electrical variables are currents. This implementation is similar to the one shown in Fig. 1, 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. We could prove that these circuits can be designed in such a way that 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.

IX. Other Modal Transmission Schemes

The idea of using propagation modes for obtaining n channels using n transmission conductors is not new. It was mentioned [15, footnote 8] [16] as a way of eliminating crosstalk, based on a principle stating that "there is no crosstalk between modes as there is between non-modal propagation". However, this idea is incorrect, since it assumes that crosstalk is only caused by coupling during propagation.

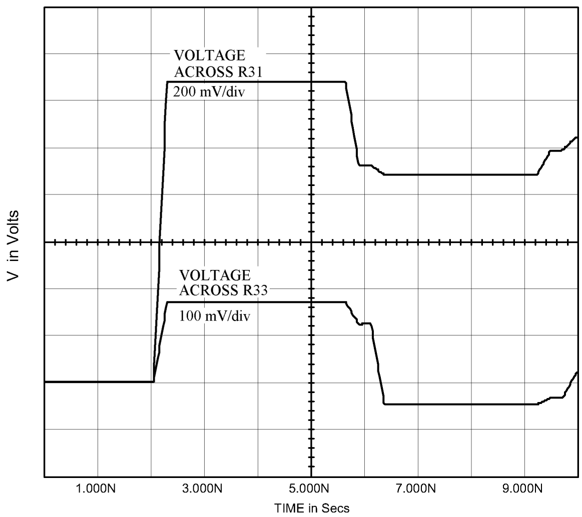

Fig. 6: voltages without matched terminations.

A correct analysis of crosstalk shows that it is caused by coupling during propagation, and by coupling occurring with reflections. As a consequence, any transmission scheme which does not specify appropriate termination is unable to eliminate crosstalk. Let us for instance consider the circuit shown in Fig. 2, 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. Computed waveforms for this case are shown in Fig. 4. Obviously, a comparison with Fig. 4 confirms that using modal transmission without matched terminations does not remove crosstalk.

Conversely, using matched termination without modal transmission does not remove crosstalk, on the contrary, as said in § III. 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 measures do not effectively reduce crosstalk.

The patent [17], like [16], is based on an hypothesis: "in general, n conductors and ground have n orthogonal modes". We have shown in § V and § VI that it is incorrect. Unlike [16], [17] does not use one of the available modes, the discarded mode being defined as the common-mode. Unfortunately this requirement is inappropriate since the common mode is usually not a propagation mode of an interconnection. Finally, another incorrect assumption of [17] is that short-circuiting some conductors at the end(s) of the interconnection reduces the size of the square matrices $\mathbf{Z}$ and $\mathbf{Y}$, as if these conductors were merged into a single one. Computations based on such errors cannot provide eigenvectors corresponding to the propagation modes of the MTL. As a result, crosstalk and echo will appear in transmission systems so designed.

X. Conclusion

We have defined a new modal transmission technique for reducing echo and crosstalk in multiconductor interconnections. The general equations needed for designing circuits implementing the ZXtalk technique will be published [18]. This reference also proves that there is an interesting relation between the new technique and the analysis of the MTL with associated eigenvectors defined by (5), though such a relation is not assumed in the definition of the ZXtalk, as explained in § V.

Real interconnections have losses and their $\mathbf{R}$, $\mathbf{L}$, $\mathbf{G}$ and $\mathbf{C}$ matrices are therefore frequency-dependent. For instance, considering "the inductance matrix" of a real interconnection is meaningless if the frequency is not specified. It is therefore essential that the ZXtalk method has been defined in such a way that losses are taken into account, though the interconnection used for the example was considered lossless for simplicity. We also note that, for interconnections having particular properties, for instance substantially equal propagation constants, the implementation of the ZXtalk method becomes simpler.

As shown in § IX, adequate operation of any modal transmission scheme requires a correct analysis of crosstalk and echo in uniform interconnections.

The ZXtalk method has been implemented in a laboratory environment [19] using wide-band analog circuits. It allowed to compare the signal integrity obtained using the ZXtalk method, to the one obtained with conventional line drivers and line receivers, and terminations made of grounded resistors having a pseudo-matched impedance.

The practical reduction of crosstalk depends on the type and length of the interconnection, on the bandwidth, and on the specific implementation. It is therefore difficult to provide a general rule of thumb for the achievable performances. However, our results show that this method can readily be implemented with PCB traces, backplanes, flex circuits and cables, for the purpose of using denser or longer interconnections, or a wider bandwidth.

Economical implementations of the ZXtalk on PCB require appropriate interface IC or interface circuits inside IC performing other functions. The ZXtalk could also be implemented with on-chip interconnects. However, specific features in design automation tools are instrumental for implementing the ZXtalk on PCB or on-chip.

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