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A New Interconnection Architecture for the Reduction of Crosstalk and Reflections

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Abstract — This paper is about a method for the reduction of crosstalk and reflections in multiconductor interconnections, using special receiving circuits and transmitting circuits. We first review the conventional techniques used for reducing crosstalk. When the interconnection can be modeled as a uniform transmission line, we compare the termination of each conductor with a grounded impedor having a pseudo-matched impedances, with (truly) matched terminations, which eliminate reflections. The new method is based on the use of such matched termination and of one modal variable for each transmission channel. The conversion from modal to natural variables and *vice versa* is performed by active circuits.

I. Introduction

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 [1, ch. 6] of multiconductor transmission lines (MTL), reflections and crosstalk may be computed using 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. 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 ω is the angular frequency, where $\mathbf{I}$ is the column-vector of the natural currents $i_1,\ldots, i_n$, where $\mathbf{V}$ is the column-vector of the natural voltages $v_1,\ldots, v_n$, 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. More precisely, 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 [1, § 6.2.6.3] use (4) for solving (2), and therefore obtain bi-orthonormal eigenvectors [2]. However, (4) is not a necessary condition, and another possible choice to obtain a solution $\mathbf{S}$ 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 [3] [4] and [5]. In order to indicate that a matrix $\mathbf{S}$ and a matrix $\mathbf{T}$ are defined by (2), (3) and (5) 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.

The particular properties of the solutions based on (5) are related to a new method for the reduction of crosstalk and reflections presented in this paper.

II. Classical techniques for crosstalk reduction

Let us assume that we want to send signals through n transmission channels. Six classical solutions are available for the reduction of crosstalk. A first solution consists in using balanced transmission lines, to which differential signal transmitters and differential signal receivers will be connected. This solution requires two transmission conductors for each transmission channel. It is used from d.c. to several hundreds of megahertz. When the interconnection is a cable, this solution uses twisted pairs, such as the one found in paired cables for telephone local loops or in cables for wide-band local area networks (e.g. category 5 UTP or STP cables). For interconnections on a printed circuit board (PCB), the use of differential transmission is now commonplace for the analog inputs and outputs of high performance of A/D converters and D/A converters, and for digital serial links in backplanes using serializer/deserializer (SerDes) chips.

A second solution consists in shielding: conductors connected to ground at both ends must in this case be used to separate (from the electromagnetic standpoint) the signals to be sent. A shield is needed for each transmission channel, and one therefore uses at least *2n* conductors. If the interconnection is a cable, it could for instance contain n coaxial pairs. This type of cable (multi-coax cable) is used in video applications and is somewhat expensive. This solution may be combined with the first one, giving rise to cables made of individually shielded twisted pairs, such as some cables for high speed data transmission. Adding conductor grounded at both ends to get shielding is not convenient for PCB designers, who must create a structure behaving more or less like a shield, using traces and eventually ground planes (they have to draw at least one ground trace along the transmission conductor(s) used for each channel, which takes up much board space and increases the number of vias).

A third solution consists in increasing the distance between the transmission conductors used for different channels. This approach is often not compatible with cost and size requirements.

A fourth solution consists in decreasing the distance between the transmission conductors and the ground conductor. For instance, on a multilayer PCB, one creates a ground plane on the layer just below and/or on the layer just above the traces used as transmission conductors. The combined use of the last two solutions (the third and the fourth) gives good results, but it is not welcome in current designs of interconnections implemented on PCB as well as on cables, because of cost and size considerations.

A fifth solution consists in reducing the upper limit of the frequency band used by the signals to be sent. This solution can of course not be used in situations where the bandwidth of these signals cannot be modified. A sixth solution consists in terminating the transmission conductors with pseudo-matched impedances, according to the definition of the next paragraph. The main effect of this solution is in fact a reduction of reflections at the ends of these conductors, the reduction of crosstalk being obtained indirectly, as a byproduct.

All these solutions have limitations: they either provide a small reduction of crosstalk, or they require a large transverse dimension for the interconnection, because of an increased spacing of the transmission conductors, or because they typically require twice as many conductors as transmission channels. We will now leave the first five solutions, in order to focus on approaches which are less demanding with respect to the cross section of the interconnection. We will now only consider interconnections with n transmission conductors and a reference conductor, providing n transmission channels.

III. Matched terminations and pseudo-matched impedances

The characteristic impedance matrix $\mathbf{Z}_C$ of the MTL is defined [6] by

$$\mathbf{Z}_C = \mathbf{S}\,\boldsymbol{\Gamma}^{-1}\,\mathbf{S}^{-1}\mathbf{Z} = \mathbf{S}\,\boldsymbol{\Gamma}\,\mathbf{S}^{-1}\mathbf{Y}^{-1} = \mathbf{Y}^{-1}\mathbf{T}\boldsymbol{\Gamma}\mathbf{T}^{-1} = \mathbf{Z}\mathbf{T}\boldsymbol{\Gamma}^{-1}\mathbf{T}^{-1} \quad (6)$$

where

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

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. Such a termination typically requires a network of $n\,(n+1)/2$ resistors [7] 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 [8, § 3]. Instead, 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. Though these impedances 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 $\mathbf{Z}_C$.

One way of computing optimal pseudo-matched impedances is to minimize a suitable norm of the matrix $\mathbf{P}$ of the voltage reflection coefficients ρ_{ij} defined by

$$\mathbf{P} = (\mathbf{Z}_L - \mathbf{Z}_C)(\mathbf{Z}_L + \mathbf{Z}_C)^{-1} \quad (8)$$

with a well-chosen value of the diagonal impedance matrix $\mathbf{Z}_L$ of the termination.

We can for instance choose to minimize the maximum absolute row sum norm $\|\mathbf{P}\|_\infty$ of $\mathbf{P}$. This norm [9, p. 1148] is defined by

$$\|\mathbf{P}\|_\infty = \max_i \sum_{j=1}^{n} |\rho_{ij}| \qquad (9)$$

Since it is the natural norm induced by the L_∞-norm for vectors, defined by

$$\|\mathbf{V}\|_\infty = \max_i |v_i| \qquad (10)$$

this norm $\|\mathbf{P}\|_\infty$ is such that the reflected wave $\mathbf{V}_-$ caused by a non-zero incident wave $\mathbf{V}_+$ satisfies:

$$\frac{\|\mathbf{V}_-\|_\infty}{\|\mathbf{V}_+\|_\infty} \le \max \frac{\|\mathbf{V}_-\|_\infty}{\|\mathbf{V}_+\|_\infty} = \|\mathbf{P}\|_\infty \qquad (11)$$

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

In order to illustrate these concepts, a few simulations will show signals obtained with pseudo-matched impedances. We will use an interconnection having three parallel transmission conductors and a reference conductor, which was described by Nickel *et al* [10]. We will neglect losses. For this particular interconnection, the matrices $\mathbf{L}$ and $\mathbf{C}$ are

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

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

We determined the matrices $\mathbf{Z}_C$, $\mathbf{S}$ and $\mathbf{T}$, and the SPICE model of the MTL, using the SpiceLine software [5]. We obtained:

$$\mathbf{Z}_C = \begin{pmatrix} 49.41 & 8.35 & 2.24 \\ 8.35 & 49.53 & 8.35 \\ 2.24 & 8.35 & 49.41 \end{pmatrix} \Omega$$

$$\mathbf{S} = \begin{pmatrix} 0.3101 & -0.5394 & -0.4793 \\ -0.4755 & 0 & -0.6232 \\ 0.3101 & 0.5394 & -0.4793 \end{pmatrix}$$

$$\mathbf{T} = \begin{pmatrix} 0.4786 & -0.7071 & 0.5198 \\ -0.7361 & 0 & 0.6780 \\ 0.4786 & 0.7071 & 0.5198 \end{pmatrix}$$

In the last two expressions, the matrices $\mathbf{S}$ and $\mathbf{T}$ are associated, for a value of the arbitrary p.u.l. capacitance c_K defined by (5) equal to 10^{-10} F/m. The propagation velocities of the three modes are about 0.1670 m/ns, 0.1617 m/ns and 0.1475 m/ns.

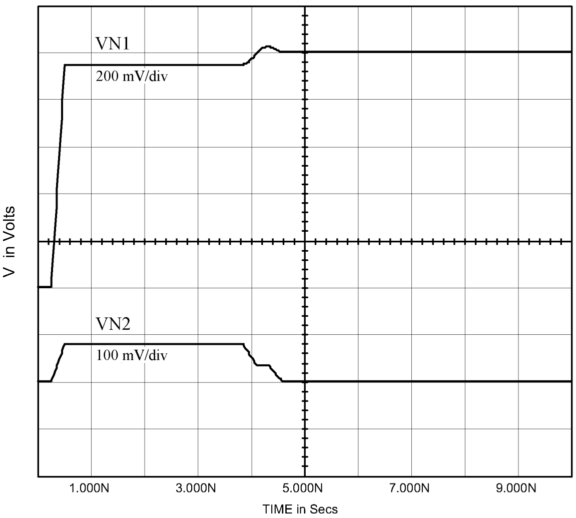

Fig 2: voltages at the near-end.

We have performed the SPICE simulation of this 30 cm long interconnection, for which the propagation times for the three propagation modes are about 1.80 ns, 1.86 ns and 2.03 ns. The simulated circuit is shown in Fig. 1, comprising at each end three resistors having the values of optimal pseudo-matched impedance for the L_∞-norm. The source produces a step with a (0% to100%) rise time of 250 ps. The results of the SPICE simulation when the current of the current source is injected on the conductor 1 are the following.

- The voltages at the near-end test points VN1 and VN2 are shown in Fig. 2. The signal at VN1 does not reach its final value immediately and has a spike after about 2 propagation times. The NEXT signal at VN2 reaches about 80 mV peak.
- The voltages at the far-end test points VF1 and VF2 are shown in Fig. 3. The signal at VF1 is not significantly distorted. The settling time is therefore short. The FEXT signal VF2 has a peak value of about 305 mV.

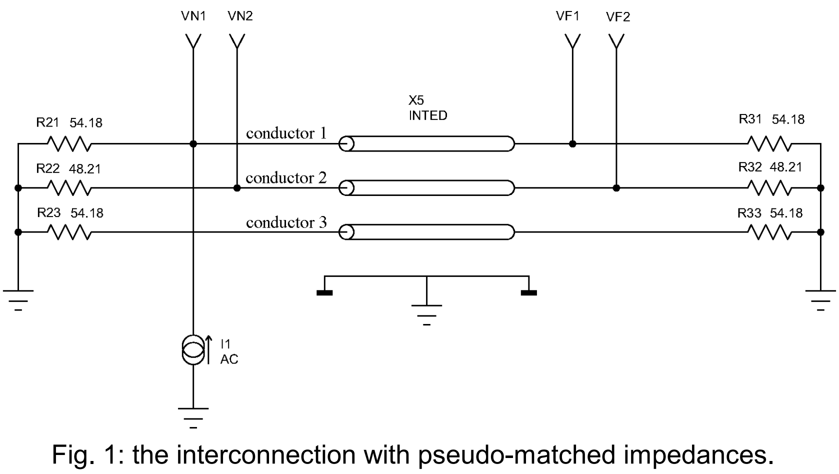

Fig. 1: the interconnection with pseudo-matched impedances.

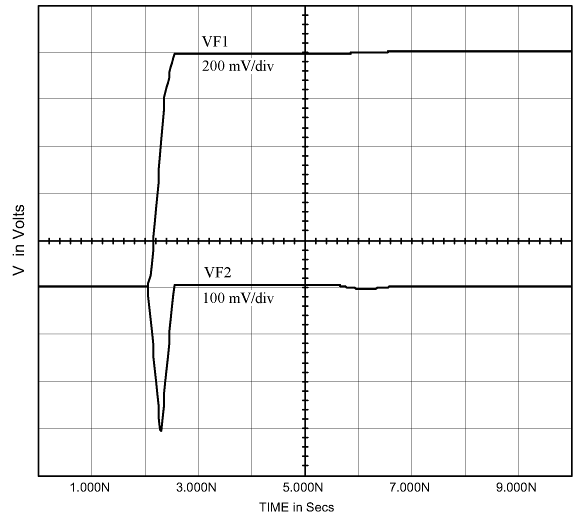

Fig 3: voltages at the far-end.

The results of the SPICE simulation when the current is injected on the conductor 2 instead of the conductor 1 are the following.

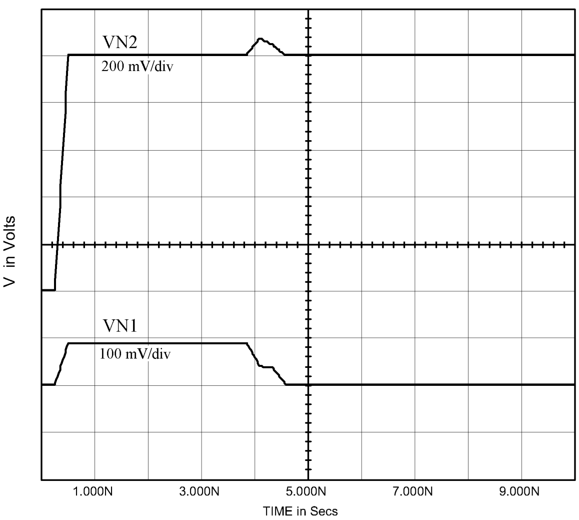

Fig 4: voltages at the near-end.

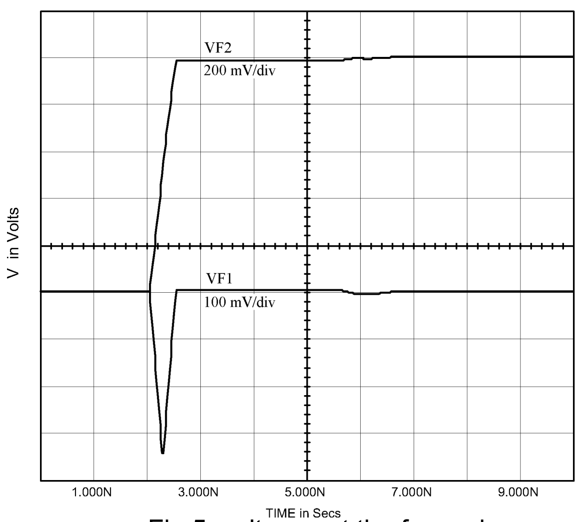

Fig 5: voltages at the far-end.

■ The voltage at the near-end test points VN1 and VN2 are shown in Fig. 4. For the signal at VN2, the final value is almost reached immediately, but there is a spike after about two propagation times. The NEXT signal at VN1 reaches about 90 mV peak.

■ The voltages at the far-end test points VF1 and VF2 are shown in Fig. 5. The transmitted signal at VF2 is not significantly distorted. The FEXT signal at VF1 has a peak value of about 340 mV.

We have just seen that the use of terminations made of pseudo-matched grounded impedors leaves reflections and crosstalk in single-ended transmission..

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 $i_{M\,1},..., i_{M\,n}$ and $\mathbf{V}_M$ the vector of the n modal voltages $v_{M\,1},..., v_{M\,n}$, we get (by definition):

$$\begin{cases} \mathbf{V} = \mathbf{S}\mathbf{V}_M \\ \mathbf{I} = \mathbf{T}\,\mathbf{I}_M \end{cases} \tag{12}$$

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 (12) 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 :

$$v_{M\alpha} = \frac{\varepsilon}{j\omega\, c_K}\,\gamma_\alpha\, i_{M\alpha} \tag{13}$$

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 [8] 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 circuit when losses can be neglected [4]. This is how the simulations of § III have been obtained.

V. A new method for reducing crosstalk and reflections

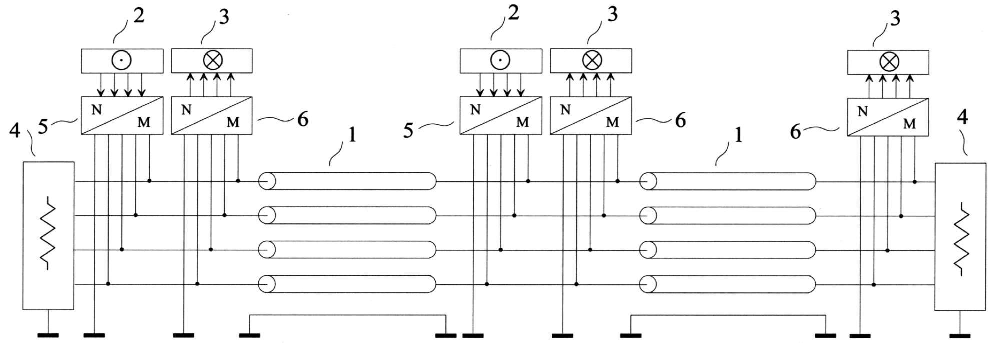

Fig. 6 : the ZXtalk method applied to a bidirectional bus.

We are now able to describe the basics of a new method which in theory allows to cancel crosstalk and reflections. This method, called ZXtalk is implemented on the example shown in Fig. 6. It is applicable to interconnections with n transmission conductors which may be modeled as a uniform MTL with a sufficient accuracy. This method is mainly characterized by the following points [11]:

- the interconnection (# 1 in Fig. 6) is connected at at least one end to a matched termination (# 4 in Fig. 6);
- one or several transmitting circuits (# 5 in Fig. 6) combine the input signals generated by sources (# 2 in Fig. 6) 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 of the interconnection;
- the n transmission conductors are connected to the input of at least one receiving circuit (# 6 in Fig. 6), 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 (# 3 in Fig. 6).

The circuit of Fig. 6 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

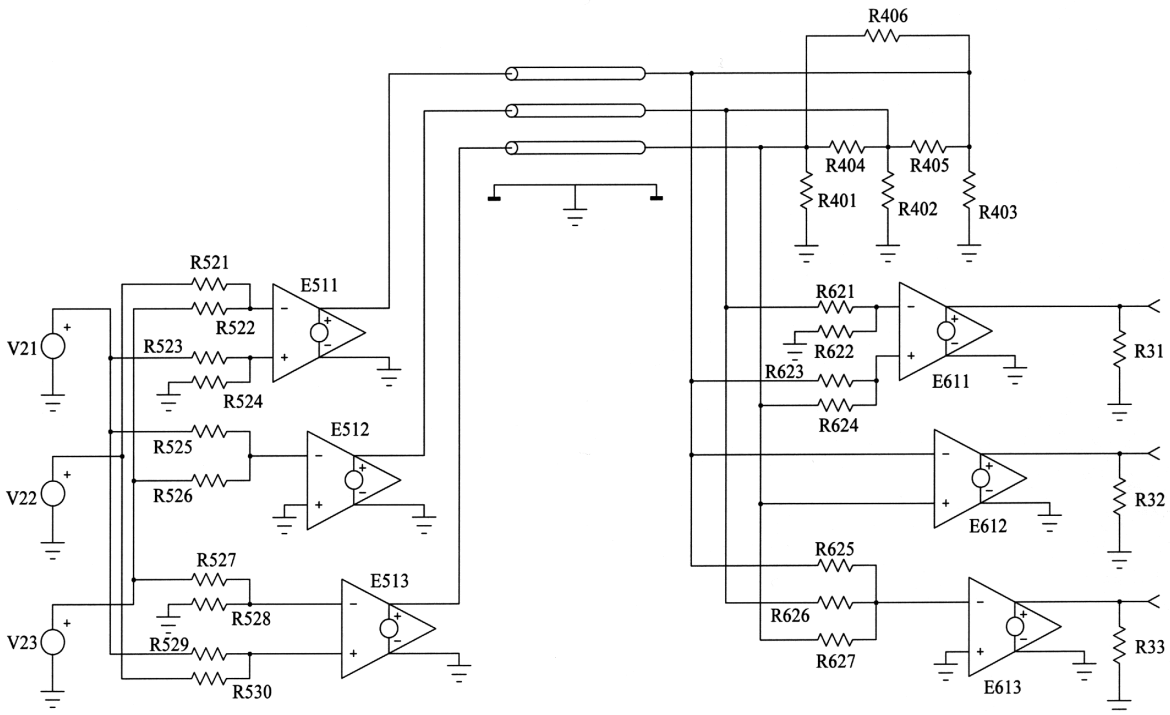

Fig. 7 : example of the ZXtalk using a 4-conductor interconnection for unidirectional transmission

crosstalk and reflections.

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 have shown in Fig. 7 the schematic for the SPICE simulation of a theoretical example of the ZXtalk using the 30 cm long interconnection studied in § III. 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. 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 we obtained simulated waveforms such as the one shown in Fig. 8 for a 250 ps rise time, which confirm that there is no crosstalk and no reflection left.

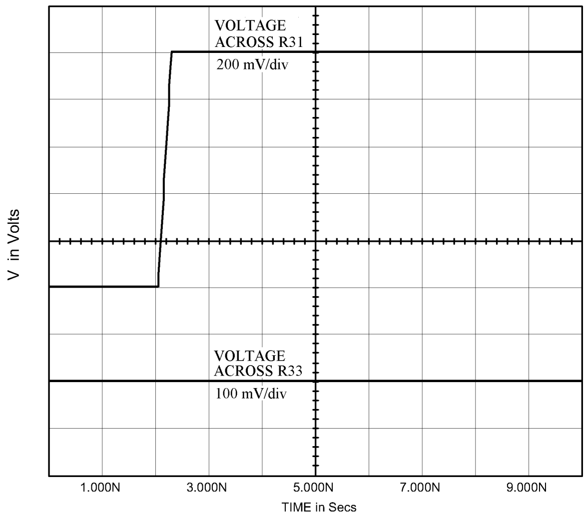

Fig 8: voltages at the far-end.

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. We may also note that, for interconnections having particular properties, the implementation of the ZXtalk method is simpler. This is for instance the case when all propagation constants are equal [13].

VI. Conclusion

The ZXtalk method has been implemented in a laboratory environment [14]. 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 impedors having a pseudo-matched impedance.

Such works allow to confirm that the ZXtalk method works as expected. However, the achievable 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. Anyway, our results show that this method can readily be implemented with PCB traces, backplanes, flex circuits and cables, for the purpose of using denser (hence cheaper), or longer interconnections, or a wider bandwidth.

Economical implementations of the ZXtalk require the use of appropriate interface IC or interface circuits inside IC performing other functions. Also, the ZXtalk could be implemented with on-chip interconnects.

Several organizations are currently joining their efforts to develop the ZXtalk under the Eureka banner.

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