A NEW PSEUDO-DIFFERENTIAL TRANSMISSION SCHEME FOR ON-CHIP AND ON-BOARD INTERCONNECTIONS

Frédéric Broydé *, Evelyne Clavelier *
(*) Excem, 12 chemin des Hauts de Clairefontaine - 78580 Maule, fredbroyde@excem.fr

Abstract. A new type of pseudo-differential link uses an interconnection comprising a return conductor wider than the transmission conductors, whereas conventional pseudo-differential links use a common conductor similar to the transmission conductors. This signaling scheme, referred to as the ZXnoise method, implements a particular type of termination and provides low reflections and a good protection against external crosstalk, thanks to the remarkable form of the characteristic impedance matrix of the interconnection. A simulation shows that the ZXnoise method is capable of high performances when a transmitting circuit producing a reduced common-mode current is used.

I. INTRODUCTION

In this paper, we consider analog or digital transmission through a multiconductor interconnection, for obtaining m channels, with $m \geq 2$. A differential link providing m channels uses an interconnection having $n = 2m$ transmission conductors. The authors have recently provided an analysis of crosstalk in balanced interconnections used for differential signaling [1]. A pseudo-differential link (PDL) providing m channels [2, § 4.2.3] uses an interconnection having $n = m$ transmission conductors (TCs) and a common conductor (CC) distinct from the reference conductor (ground). Compared to m single-ended links, a PDL features reduced external crosstalk (i.e. crosstalk with external conductors). A PDL is shown in Fig. 1, this link comprising an interconnection having $n = 4$ TCs. The transmitting circuit (TX circuit) receives at its input the signals of the 4 channels of the source, and its 5 output terminals are connected to the $n + 1 = 5$ conductors of the interconnection. The receiving circuit (RX circuit) has its 5 input terminals connected to the conductors of the interconnection, and its output terminals connected to the user.

In Fig. 1, there is no termination circuit, as is the case in many pseudo-differential signaling methods [2] [3] [4]. Consequently, substantial reflections of signals occur, and this implies limitations on the length $\mathcal{L}$ of the interconnection ($\mathcal{L}$ must be sufficiently small) and on the available bandwidth, since the limitation stated by Jarvis [5] applies ($\mathcal{L}$ should typically be less than one fourth of the distance traveled during the transition time).

Section II presents the principle of conventional PDLs, and the possible termination schemes. It explains a conflict between a good protection against external crosstalk and an effective reduction of reflections. Section III defines a new type of PDL, this scheme being referred to as the ZXnoise method. Section IV presents the theory of operation and shows that the ZXnoise method avoids said conflict. Section V briefly covers an example and provides simulation results for the transmission of signals, internal crosstalk and echo.

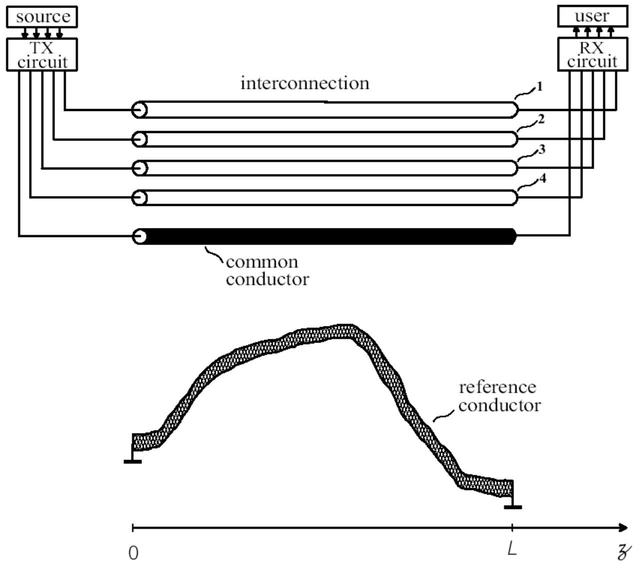

Fig. 1. A conventional pseudo-differential link (PDL) comprising an interconnection consisting of 4 transmission conductors (TCs) numbered from 1 to 4 and a common conductor (CC).

II. CONVENTIONAL PSEUDO-DIFFERENTIAL LINKS

In Fig. 1, since no termination is present, there is no need to maintain a constant characteristic impedance or characteristic impedance matrix along the interconnection, and there is consequently no constraint on the manner of routing the interconnection with respect to ground. Consequently, the reference conductor is shown as an irregular geometrical shape, such that the distance between the conductors of the interconnection and the reference conductor varies as a function of the abscissa z along the interconnection.

To obtain reduced internal crosstalk (i.e. crosstalk between the channels) and echo, other conventional PDLs use a termination, typically made of resistors [6], each resistor being connected between a TC and ground or a node intended to present a fixed voltage with respect to ground (for instance a power supply voltage), or connected between the CC and ground or a node intended to present a fixed voltage with respect to ground. This technique is for instance used in the pseudo-differential signaling scheme using integrated circuits of the Gunning Transceiver Logic (GTL) family [7] [8, pp. 2-3 to 2-17].

If a termination is used, the interconnection must be designed such that it is possible to model propagation in the interconnection using a uniform multiconductor transmission line (MTL) having $n + 2 = 6$ conductors, the MTL using as variables the n natural voltages (which are defined with respect to the reference conductor) and the n natural currents flowing on the TCs and on the CC [9] [10, ch. 6]. This result is typically obtained with a geometry such that the cross section of the interconnection and the reference conductor, in a plane orthogonal to the direction of propagation, does not change over the greatest part of the length of the interconnection, in the vicinity of the TCs.

The above-mentioned termination made of one resistor per TC should provide an impedance matrix not too different from the characteristic impedance matrix of said $(n+2)$-conductor MTL. This may happen only if the characteristic impedance matrix of this MTL is such that the absolute value of each of the diagonal components is much larger than the absolute value of each of the non-diagonal components, in a suitable frequency band. This implies that the TCs are in a way closer to the reference conductor than to the CC, as in the microstrip and stripline structures shown in Fig. 2.

The state of the art as regards fighting against external crosstalk requires that the routing of all TCs and of the CC must be matched [6], so that substantially equal noise voltages are obtained on all conductors of the interconnection. This seems to be achieved for the configurations shown in Fig. 2. Unfortunately, we see that, in Fig. 2, the CC (5) is close to a TC (4), but far from other TCs (1 and 2). The coupling parameters between an external conductor and a conductor of the interconnection will consequently be different for the different conductors, and it will not be possible to eliminate the external crosstalk.

We see that, according to these pseudo-differential transmission schemes, there is a discrepancy between an effective protection against external crosstalk which implies that the TCs are in a way closer to the CC than to the reference conductor, and reduced reflections which imply that the TCs are in a way closer to the reference conductor than to the CC. This is why, when m is large, several common conductors are needed, for instance one common conductor every fourth TC [3].

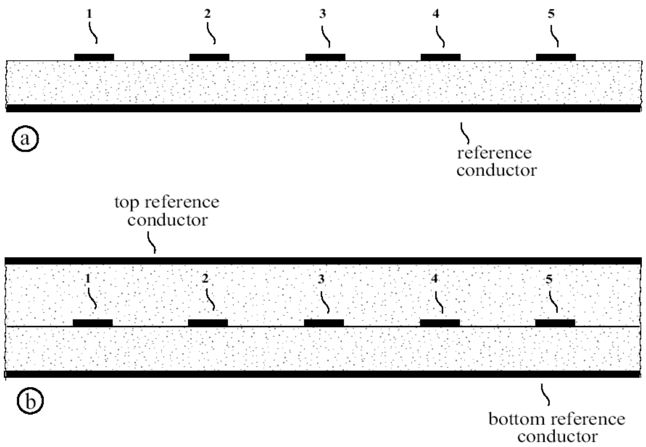

Fig. 2. Two possible cross-sections for the interconnection of a conventional PDL, where 1 to 4 are the TCs and 5 is the CC.

III. A New Type of Pseudo-Differential link

We now consider interconnections comprising $n \geq 2$ TCs and a return conductor (RC) distinct from the reference conductor, the interconnection being structurally combined with the reference conductor throughout the length of the interconnection. The new ZXnoise method is intended to provide, in a known frequency band, m transmission channels, with $n \geq m \geq 2$.

For any integer j such that $1 \leq j \leq n$, at a given abscissa z along the interconnection, let us use i_j to denote the natural current of index j, that is to say the current flowing on the TC number j, and let us use v_{Rj} to denote the natural voltage referenced to the RC of index j, that is to say the voltage between the TC number j and the RC. We may define the column-vector $\mathbf{I}_R$ of the natural currents $i_1,..., i_n$ and the column-vector $\mathbf{V}_R$ of the natural voltages $v_{R\ 1},..., v_{R\ n}$ referenced to the RC.

As a preamble for the implementation of the ZXnoise method, the designer proportions the interconnection in such a way that it may, in a part, denoted by B, of the known frequency band, taking into account the lumped impedances seen by the interconnection and caused by the circuits connected to the interconnection elsewhere than at the ends of the interconnection, be modeled with a sufficient accuracy as a MTL having $n + 1$ conductors [10, ch. 6], such that:

— the $(n+1)$-conductor MTL uses the natural voltages referenced to the RC and the natural currents as variables;

— the $(n+1)$-conductor MTL has uniform electrical characteristics over its length.

This implies that all conductors other than the conductors of the interconnection may be neglected when one models propagation in the interconnection and that, in particular, the reference conductor may be neglected when one models propagation in the interconnection. For instance, the coplanar-strips-over-return-conductor and coplanar-strips-inside-return-conductor structures shown in Fig. 3 are appropriate to obtain this result. It is consequently possible to define, for the $(n+1)$-conductor MTL, at any frequency f in B, a per-unit-length (p.u.l.) impedance matrix $\mathbf{Z}_R$ and a p.u.l. admittance matrix $\mathbf{Y}_R$, and the applicable telegrapher's equations are:

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

where $\mathbf{Z}_R$ and $\mathbf{Y}_R$ are matrices of size $n \times n$.

The ZXnoise method is further defined by the following additional steps:

■ determining, for the $(n+1)$-conductor MTL defined by (1), at any frequency in B, the characteristic impedance matrix

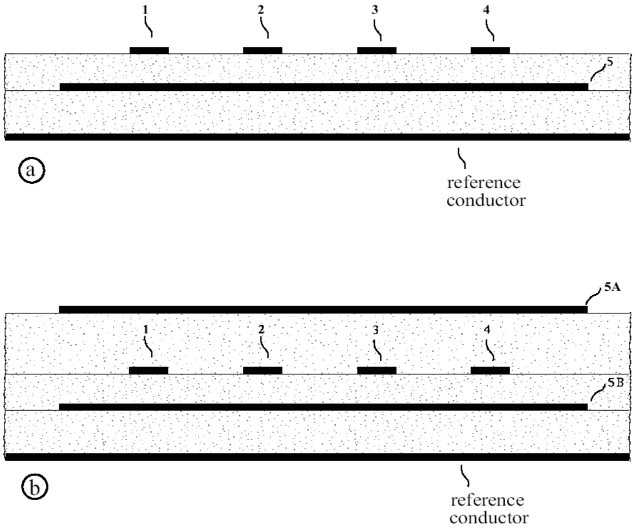

Fig. 3. Two possible cross-sections for the interconnection used in the ZXnoise method, where 1 to 4 are the TCs, where 5 is the RC in the coplanar-strips-over-return-conductor structure (*a*) and where the RC is made of the conductors 5A and 5B in the coplanar-strips-inside-return-conductor structure (*b*).

with respect to the RC, denoted by $\mathbf{Z}_{RC}$;

■ connecting, as shown in Fig. 4, the terminals of at least one termination circuit to the RC and the TCs, each termination circuit being, at any frequency in *B*, approximately characterized by an impedance matrix with respect to the RC, denoted by $\mathbf{Z}_{RL}$, this matrix $\mathbf{Z}_{RL}$ being of size $n \times n$, approximately equal to a diagonal matrix, and proportioned using $\mathbf{Z}_{RC}$;

■ using one TX circuit receiving the *m* signals to be sent, the output of said TX circuit being coupled to at least *m* TCs; and

■ using one RX circuit having its input coupled to the RC and the TCs, each "output signal of the RX circuit" being mainly determined by only one natural voltage referenced to the RC appearing at the input of the RX circuit.

Clearly, the key difference between the above definition and conventional PDLs resides in the properties of the interconnection and the terminations. They will be studied in Section IV.

Since $\mathbf{Z}_R$ and $\mathbf{Y}_R$ are independent from the abscissa *z*, (1) implies that the classical results about uniform MTLs may be transposed. In particular, the theory of pseudo-matched impedances [11] [12] gives the following results:

■ the matrix of the voltage reflection coefficients of the termination circuit with respect to the RC, denoted by $\mathbf{P}_R$, is given by

$$\mathbf{P}_R = \left(\mathbf{Z}_{RL} - \mathbf{Z}_{RC}\right)\left(\mathbf{Z}_{RL} + \mathbf{Z}_{RC}\right)^{-1} \qquad (2)$$

■ a floating termination circuit such that $\mathbf{Z}_{RL}$ is a diagonal matrix may be made of *n* impedors (i.e. linear two-terminal circuit elements), each impedor being connected between a TC and the RC, the *n* impedors being easily proportioned such that all components of the matrix $\mathbf{P}_R$ have an absolute value less than a sufficiently small arbitrary value;

■ it is even possible to determine a termination circuit presenting a diagonal matrix $\mathbf{Z}_{RL}$ minimizing the detrimental effects of reflections, by minimizing a suitable norm of $\mathbf{P}_R$.

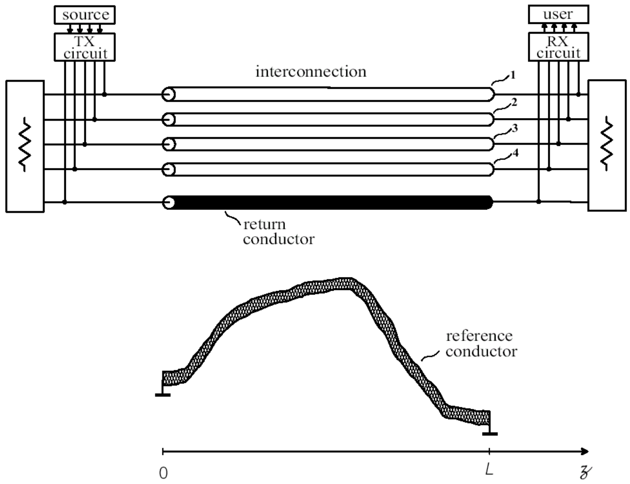

Fig. 4. A PDL implementing the ZXnoise method. At each end, the block containing the resistor symbol is a termination circuit. In some cases (e.g. unidirectional link), the termination circuit on the left is not present.

IV. Elementary Theory of the ZXnoise Method

A. Matrices with respect to ground

The $(n+1)$-conductor MTL defined by (1) uses natural voltages referenced to the RC and natural currents as variables. Of course, the interconnection used in the ZXnoise method may possibly also be modeled as a $(n+2)$-conductor MTL, this MTL using natural voltages referenced to ground and natural currents as variables. For such a model, it is necessary to consider, at a given abscissa *z* along the interconnection:

— for any *j* such that $1 \le j \le n$, the natural current i_j;

— the current flowing on the RC, denoted by i_{n+1};

— for any *j* such that $1 \le j \le n$, the voltage between the TC number *j* and the reference conductor, denoted by $v_{G\,j}$;

— the voltage between the RC and the reference conductor, denoted by $v_{G\,n+1}$.

We may then define the column-vector $\mathbf{I}_G$ of the currents $i_1,\ldots, i_{n+1}$ and the column-vector $\mathbf{V}_G$ of the natural voltages referenced to ground $v_{G\,1},\ldots, v_{G\,n+1}$. When it is possible to define, for the $(n+2)$-conductor MTL, at each abscissa *z* along the interconnection, at a frequency *f* in *B*, a per-unit-length (p.u.l.) impedance matrix $\mathbf{Z}_G$ and a p.u.l. admittance matrix $\mathbf{Y}_G$, the applicable telegrapher's equations are:

$$\begin{cases} \dfrac{d\,\mathbf{V}_G}{dz} = -\mathbf{Z}_G\,\mathbf{I}_G \\ \dfrac{d\,\mathbf{I}_G}{dz} = -\,\mathbf{Y}_G\,\mathbf{V}_G \end{cases} \qquad (3)$$

In (3), the matrices $\mathbf{Z}_G$ and $\mathbf{Y}_G$ are of size $(n+1) \times (n+1)$. We have said above that the interconnection is such that it may be modeled with a sufficient accuracy as a $(n+1)$-conductor MTL. Consequently, in (3), we may say that, to a sufficient accuracy:

— for any *j* such that $1 \le j \le n$, the voltage $v_{G\,j} - v_{G\,n+1}$ depends only on $i_1,\ldots, i_n$;

— the relationships between the voltages $v_{G\,j} - v_{G\,n+1}$ and $i_1,\ldots, i_n$ are determined by the matrices $\mathbf{Z}_R$ and $\mathbf{Y}_R$.

The matrix $\mathbf{Z}_G$ providing the spatial derivatives of the voltages $v_{G\,1}$, ..., $v_{G\,n+1}$ as a function of the currents i_1,..., i_{n+1}, the first condition implies that all entries of the last column of $\mathbf{Z}_G$ are equal. Because of reciprocity, all entries of the last row of $\mathbf{Z}_G$ are also equal. Using this result, the second condition implies that $\mathbf{Z}_G$ is, at a frequency in B, at a given z, approximately given by

$$\mathbf{Z}_G \approx \begin{pmatrix} \mathbf{Z}_R + Z_{RG}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix}\begin{pmatrix}1 & \cdots & 1\end{pmatrix} & Z_{RG}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \\ Z_{RG}\begin{pmatrix}1 & \cdots & 1\end{pmatrix} & Z_{RG} \end{pmatrix} \tag{4}$$

where Z_{RG} is a p.u.l. impedance. The matrix $\mathbf{Y}_G^{-1}$ providing the voltages $v_{G\,1}$, ..., $v_{G\,n+1}$ as a function of the spatial derivatives of the currents i_1,..., i_{n+1}, we can apply the reasoning used for $\mathbf{Z}_G$ to $\mathbf{Y}_G^{-1}$. We conclude that $\mathbf{Y}_G^{-1}$ is, at a frequency in B, at a given z, approximately given by

$$\mathbf{Y}_G^{-1} \approx \begin{pmatrix} \mathbf{Y}_R^{-1} + \dfrac{1}{Y_{RG}}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix}\begin{pmatrix}1 & \cdots & 1\end{pmatrix} & \dfrac{1}{Y_{RG}}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \\ \dfrac{1}{Y_{RG}}\begin{pmatrix}1 & \cdots & 1\end{pmatrix} & \dfrac{1}{Y_{RG}} \end{pmatrix} \tag{5}$$

where Y_{RG} is a p.u.l. admittance. We can easily check that

$$\mathbf{Y}_G \approx \begin{pmatrix} \mathbf{Y}_R & -\mathbf{Y}_R\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \\ -\begin{pmatrix}1 & \cdots & 1\end{pmatrix}\mathbf{Y}_R & Y_{RG} + \begin{pmatrix}1 & \cdots & 1\end{pmatrix}\mathbf{Y}_R\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \end{pmatrix} \tag{6}$$

In (4)-(6), Z_{RG} and Y_{RG} are z-dependent, in general. We observe that (4) and (5) imply that $v_{G\,n+1}$ is independent from any i_1,..., i_{n+1} such that $i_1 + ... + i_{n+1} = 0$. We also note that equations similar to (4) and (6) were used (without proof) in an analysis of shielded twisted-pair cables [1].

B. Characteristic impedance matrix with respect to ground

For the (n+2)-conductor MTL considered above, let us consider [12] a transition matrix from the "modal voltages referenced to ground" to the "natural voltages referenced to ground", denoted by $\mathbf{S}_G$. The column vectors of $\mathbf{S}_G$ are eigenvectors of $\mathbf{Z}_G\,\mathbf{Y}_G$. Assuming that (4) and (6) are exact, we find

$$\mathbf{Z}_G\mathbf{Y}_G = \begin{pmatrix} \mathbf{Z}_R\mathbf{Y}_R & Z_{RG}Y_{RG}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} - \mathbf{Z}_R\mathbf{Y}_R\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \\ \begin{pmatrix}0 & \cdots & 0\end{pmatrix} & Z_{RG}Y_{RG} \end{pmatrix} \tag{7}$$

Let us now look for the eigenvectors of $\mathbf{Z}_G\,\mathbf{Y}_G$. If $\mathbf{s}$ is an eigenvector of $\mathbf{Z}_R\,\mathbf{Y}_R$ associated with the eigenvalue γ^2, we find

$$\mathbf{Z}_G\mathbf{Y}_G\begin{pmatrix}\mathbf{s}\\ 0\end{pmatrix} = \begin{pmatrix}\mathbf{Z}_R\mathbf{Y}_R\mathbf{s}\\ 0\end{pmatrix} = \begin{pmatrix}\gamma^2\mathbf{s}\\ 0\end{pmatrix} = \gamma^2\begin{pmatrix}\mathbf{s}\\ 0\end{pmatrix} \tag{8}$$

and

$$\mathbf{Z}_G\mathbf{Y}_G\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} = Z_{RG}Y_{RG}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \tag{9}$$

Consequently, using ${}^t\mathbf{X}$ to denote the transpose of a vector or matrix $\mathbf{X}$, if $\mathbf{s}_1$,..., $\mathbf{s}_n$ is a basis of eigenvectors of $\mathbf{Z}_R\,\mathbf{Y}_R$, we have established that ${}^t({}^t\mathbf{s}_1, 0)$,..., ${}^t({}^t\mathbf{s}_n, 0)$, ${}^t(1,...,1)$ is a basis of eigenvectors of $\mathbf{Z}_G\,\mathbf{Y}_G$. In other words, a possible choice for $\mathbf{S}_G$ is

$$\mathbf{S}_G = \begin{pmatrix} \mathbf{S}_R & \begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \\ \begin{pmatrix}0 & \cdots & 0\end{pmatrix} & 1 \end{pmatrix} \tag{10}$$

where $\mathbf{S}_R$ is a transition matrix from the "modal voltages referenced to the RC" to the "natural voltages referenced to the RC", for the (n+1)-conductor MTL characterized by $\mathbf{Z}_R$ and $\mathbf{Y}_R$. The column-vector ${}^t(1,...,1)$ of $\mathbf{S}_G$ corresponds to the common-mode voltage. Let us use

$$\boldsymbol{\Gamma}_R = \mathrm{diag}_n(\gamma_1, \ldots, \gamma_n) \tag{11}$$

to denote the diagonal matrix of order n of the propagation constants of the (n+1)-conductor MTL characterized by $\mathbf{Z}_R$ and $\mathbf{Y}_R$, and

$$\boldsymbol{\Gamma}_G = \mathrm{diag}_{n+1}\left(\gamma_1, \ldots, \gamma_n, \sqrt{Z_{RG}Y_{RG}}\right) \tag{12}$$

to denote the diagonal matrix of order n+1 of the propagation constants of the (n+2)-conductor MTL characterized by $\mathbf{Z}_G$ and $\mathbf{Y}_G$. The characteristic impedance matrix of the (n+1)-conductor MTL characterized by $\mathbf{Z}_R$ and $\mathbf{Y}_R$, denoted by $\mathbf{Z}_{RC}$, and the characteristic impedance matrix of the (n+2)-conductor MTL characterized by $\mathbf{Z}_G$ and $\mathbf{Y}_G$, denoted by $\mathbf{Z}_{GC}$, are given by [12]

$$\mathbf{Z}_{RC} = \mathbf{S}_R\,\boldsymbol{\Gamma}_R^{-1}\,\mathbf{S}_R^{-1}\mathbf{Z}_R = \mathbf{S}_R\,\boldsymbol{\Gamma}_R\,\mathbf{S}_R^{-1}\mathbf{Y}_R^{-1} \tag{13}$$

and

$$\mathbf{Z}_{GC} = \mathbf{S}_G\,\boldsymbol{\Gamma}_G^{-1}\,\mathbf{S}_G^{-1}\mathbf{Z}_G = \mathbf{S}_G\,\boldsymbol{\Gamma}_G\,\mathbf{S}_G^{-1}\mathbf{Y}_G^{-1} \tag{14}$$

We want to compute $\mathbf{Z}_{GC}$. Using (5), (10), (12), (13) and (14), we finally obtain

$$\mathbf{Z}_{GC} = \begin{pmatrix} \mathbf{Z}_{RC} + \sqrt{\dfrac{Z_{RG}}{Y_{RG}}}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix}\begin{pmatrix}1 & \cdots & 1\end{pmatrix} & \sqrt{\dfrac{Z_{RG}}{Y_{RG}}}\begin{pmatrix}1\\ \vdots \\ 1\end{pmatrix} \\ \sqrt{\dfrac{Z_{RG}}{Y_{RG}}}\begin{pmatrix}1 & \cdots & 1\end{pmatrix} & \sqrt{\dfrac{Z_{RG}}{Y_{RG}}} \end{pmatrix} \tag{15}$$

In practice, (4) and (6) are not exact and the eigenvectors may be different from the one given by (8) and (9). However, $\mathbf{Z}_{GC}$ remains approximately given by (15).

C. Application of the theory

In order to ascertain that a structure is suitable for the ZXnoise method, the designer may determine the matrices $\mathbf{Z}_G$ and $\mathbf{Y}_G$ of the (n+2)-conductor MTL, as a function of frequency and z (from measurements and/or computations [9, ch. 3] [13]) and check that, at any frequency in B:

— the last row and column of $\mathbf{Z}_G$ and $\mathbf{Y}_G$ comply with (4) and (6) to a sufficient accuracy (taking into account the lumped impedances seen by the interconnection and caused by the circuits connected to the interconnection elsewhere than at the ends of the interconnection);

— the matrices $\mathbf{Z}_R$ and $\mathbf{Y}_R$ obtained from (4) and (6) are independent of z, with a sufficient accuracy; and

— one obtains a matrix $\mathbf{Z}_{RC}$ approximating a wanted matrix.

In the special case where Z_{RG} and Y_{RG} do not vary with z, the matrix of the voltage reflection coefficients of a termination with respect to the reference conductor, denoted by $\mathbf{P}_G$, is given by

$$\mathbf{P}_G = \left(\mathbf{Z}_{GL} - \mathbf{Z}_{GC}\right)\left(\mathbf{Z}_{GL} + \mathbf{Z}_{GC}\right)^{-1} \tag{16}$$

where $\mathbf{Z}_{GL}$ is the impedance matrix of the termination with respect to ground. The floating termination circuit defined in Section III does poorly in this respect, since its impedance matrix with respect to ground is not even defined. Consequently, resonances will occur for the common mode, and they may clearly damage the protection against external crosstalk. Resonances may be avoided if we use a termination consisting of a floating termination circuit defined in Section III and a damping circuit connected between the RC and the reference conductor. The lowest resonance are, according to (15), obtained if the impedance of the damping circuit is $(Z_{RG}/Y_{RG})^{0.5}$.

In the general case where Z_{RG} and Y_{RG} vary with z, $(Z_{RG}/Y_{RG})^{0.5}$ is not a constant. However, it in practice only incurs small relative variations, so that damping circuits presenting an impedance of the order of this value will provide adequate damping.

V. EXAMPLE

We now consider an example of PDL implementing the ZXnoise method, corresponding to the block-diagram of Fig. 4, without the termination circuit on the left. The discussion is based on a SPICE simulation, according to the schematic diagram shown in Fig. 5.

The 150-mm-long interconnection is a coplanar-strips-over-return-conductor structure (drawing *a* of Fig. 3) built in a printed circuit board, the TCs being traces of width equal to about 203 µm, the center-line to center-line spacing being 889 µm. It can be modeled, with a sufficient accuracy, as a (n+1)-conductor MTL, such that

$$\mathbf{Z}_R \approx j\omega \mathbf{L}_R \quad \text{with}$$

$$\mathbf{L}_R \approx \begin{pmatrix} 334 & 14 & 3 & 2 \\ 14 & 334 & 14 & 3 \\ 3 & 14 & 334 & 14 \\ 2 & 3 & 14 & 334 \end{pmatrix} \text{nH / m} \tag{17}$$

and

$$\mathbf{Y}_R \approx j\omega \mathbf{C}_R \quad \text{with}$$

$$\mathbf{C}_R \approx \begin{pmatrix} 107 & 0 & 0 & 0 \\ 0 & 107 & 0 & 0 \\ 0 & 0 & 107 & 0 \\ 0 & 0 & 0 & 107 \end{pmatrix} \text{pF/ m} \tag{18}$$

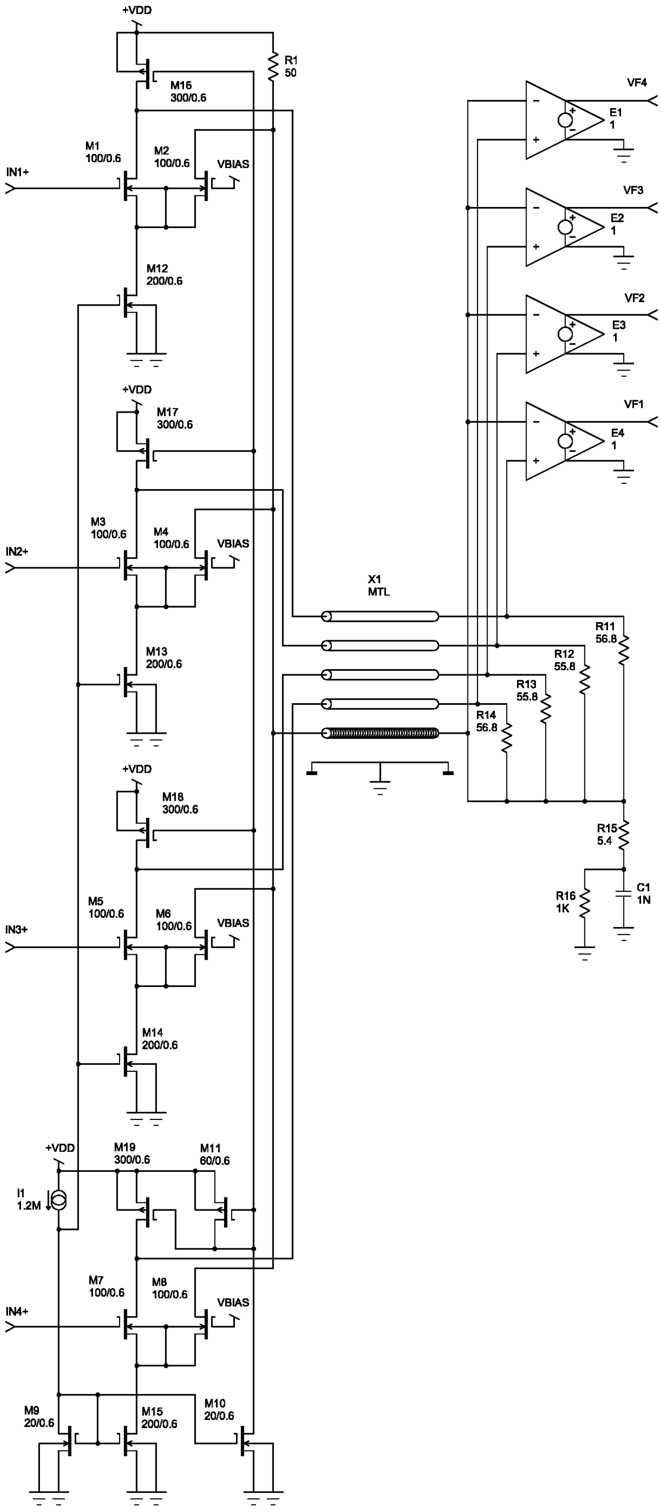

Fig. 5. Schematic diagram for the SPICE simulation of the 4-channel PDL.

The sub-circuit for the interconnection (X1 in Fig. 6) was created with SpiceLine [14], based on (17) and (18).

Because of the thickness of the conductors (34.8 µm), (17) is applicable to frequencies above 3 MHz, for which losses may be neglected in the computation of the characteristic impedance matrix $\mathbf{Z}_{RC}$ with respect to the RC. In this structure, the mutual capacitance between the TCs are very small and may be neglected in the computation of $\mathbf{Z}_{RC}$, but not the mutual inductances (this is typical of low-coupling microstrip structure). The matrices $\mathbf{Z}_R$ and $\mathbf{Y}_R$ are independent of z, and we find that $\mathbf{Z}_{RC}$ is given by

$$\mathbf{Z}_{RC} \approx \begin{pmatrix} 55.9 & 1.2 & 0.2 & 0.2 \\ 1.2 & 55.8 & 1.2 & 0.2 \\ 0.2 & 1.2 & 55.8 & 1.2 \\ 0.2 & 0.2 & 1.2 & 55.9 \end{pmatrix} \Omega \qquad (19)$$

In order to compute the diagonal matrix $\mathbf{Z}_{RL}$ defined above, the designer chooses to minimize [12] the matrix norm $\| \mathbf{P}_R \|_\infty$ of $\mathbf{P}_R$ given by (2), this matrix norm being equal to the largest sum of the absolute values of the entries of a row. In this manner, the designer obtains

$$\mathbf{Z}_{RL} = \mathrm{diag}_4(56.8, 55.8, 55.8, 56.8)\ \Omega \qquad (20)$$

for which $\| \mathbf{P}_R \|_\infty \approx 0.023$. The corresponding termination circuit is used in Fig. 5 (resistors R11 to R14).

In this design, the ratio Z_{RG}/Y_{RG} does not vary much as a function of z, and at frequencies above 3 MHz, we may consider that $(Z_{RG}/Y_{RG})^{0.5} \approx 5.4\ \Omega$. A damping circuit providing this impedance at high frequencies is included in Fig. 5 (R15, R16 and C1), at the far end. At the frequency of the $\lambda/4$ resonance of the RC, equal to 236 MHz, the reactance of C1 is smaller than the resistance of R15. At the near end, R1 also provides some damping.

In Fig. 5, the TX circuit is a device-level circuit using the device parameters of a 0.5-µm CMOS process, whereas the RX is modeled as 4 ideal voltage controlled voltages sources (E1-E4). This TX circuit, which comprises a balancing circuit such that the variations of the common mode current are effectively reduced, will be studied in a forthcoming paper. The frequency domain simulation results shown in Fig. 6 correspond to voltage gains of the different outputs of the link, with respect to a voltage applied to the input IN1. Similar results are obtained when the voltage is applied to the input IN2. For binary digital signaling, internal crosstalk becomes severe above 1 GHz. This limits the applicability of this circuit to rise and fall times of about 400 ps.

An important point is that the SPICE sub-circuit modeling the interconnection describes a 6-conductor MTL. This sub-circuit is based on $\mathbf{Z}_G$ and $\mathbf{Y}_G$ given by (4), (6), (17), (18), $Z_{RG} = j\omega$ 38.27 nH/m and $Y_{RG} = j\omega$ 1308 pF/m.

VI. CONCLUSION

The ZXnoise method provides low reflections and reduced external crosstalk, using an interconnection having a wide return conductor (RC). This result can be obtained because the ZXnoise method uses a termination scheme capable of reducing reflections without harming the protection against external crosstalk. A good protection against external crosstalk makes low-swing transmission possible, and low-swing is the key to high-speed transmission.

The interconnection may be formed in a rigid or flexible printed circuit board, in the substrate of a multi-chip module (MCM) or hybrid circuit or inside a monolithic integrated circuit. In the ZXnoise method, the RC is used as a return path for the currents corresponding to the signals, and the position of the TCs with respect to the RC and the reference conductor is such that the RC in a way acts as an electromagnetic screen. In a conventional PDL, the function of the CC is quite different, hence the different designations.

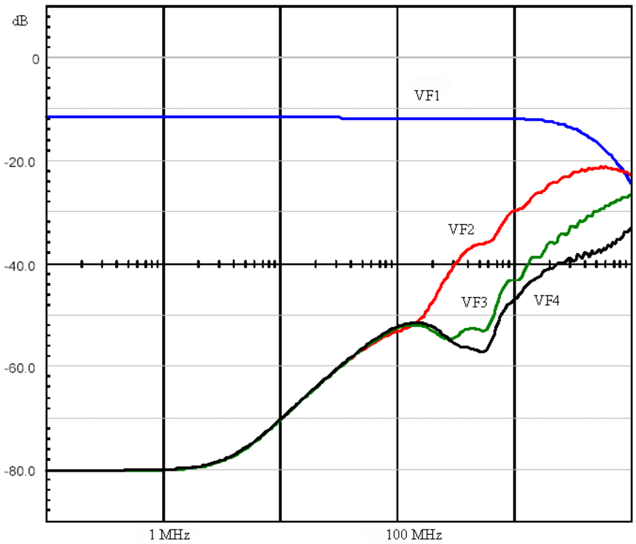

Fig. 6. Simulation results for the PDL discussed in Section V.

REFERENCES

[1] F. Broydé, E. Clavelier, "Crosstalk in balanced interconnections used for differential signal transmission", *IEEE Trans. Circuits Syst. I*, vol. 54, No. 7, pp. 1562-1572, July 2007.

[2] F. Yuan, *CMOS current-mode circuits for data communications*, New York, N.Y.:Springer, 2007.

[3] A. Carusone, K. Farzan, D.A. Johns, "Differential signaling with a reduced number of signal paths", *IEEE Trans. Circuits Syst. II*, vol. 48, No. 3, pp. 294-300, March 2001.

[4] S. Sidiropoulos, "Reducing coupled noise in pseudo-differential signaling", patent of the United States of America No. 7,099,395. Filed: November 7, 2000.

[5] D.B. Jarvis, "The Effect of Interconnections on High-Speed Logic Circuits", *IEEE Transactions on Electronic Computers*, Oct. 1963, pp. 476-487.

[6] T. Frodsham, "Multi-agent pseudo-differential signaling scheme", patent of the United States of America No. 6,195,395. Filed: March 18, 1998.

[7] W.F. Gunning, "Drivers and receivers for interfacing VLSI CMOS circuits to transmission lines", patent of the United States of America No. 5,023,488. Filed: March 30, 1990.

[8] *Design Considerations for Logic Products — Application Book — 1998*, Texas Instruments Deutschland, 1997.

[9] C.R. Paul *Analysis of Multiconductor Transmission Lines*, New York, N.Y.: John Wiley & Sons, 1994.

[10] F.M. Tesche, M.V. Ianoz, T. Karlsson, *EMC Analysis Methods and Computational Models*, New York, N.Y.: John Wiley & Sons, 1997.

[11] Broydé, "Clear as a Bell — Controlling Crosstalk in Uniform Interconnections ", *IEEE Circuits and Devices Magazine*, Vol. 20, No. 6, November/December 2004, pp. 29-37.

[12] F. Broydé, E. Clavelier, "A New Method for the Reduction of Crosstalk and Echo in Multiconductor Interconnections", *IEEE Trans. Circuits Syst. I*, vol. 52, No. 2, pp. 405-416, Feb. 2005, and "Corrections to «A New Method for the Reduction of Crosstalk and Echo in Multiconductor Interconnections»", *IEEE Trans. Circuits Syst. I*, vol. 53, No. 8, p. 1851, Aug. 2006.

[13] C. Wei, R.F. Harrington, J.R. Mautz, T.K. Sarkar, "Multiconductor transmission lines in multilayered dielectric media", *IEEE Trans. on Microwave Theory Tech*, vol. 32, No. 4, April 1984, pp. 439-450.

[14] *SpiceLine 2.23 with Telecom Line Predictor — User's Guide*, Excem document 00012107B, March 2000.