Proc. of the IEEE 18th Topical Meeting on Electrical Performance of Electronic Packaging and Systems, EPEPS 2009, Oct 19-21, 2009, pp. 5-8.

Designing a ZXnoise Pseudo-Differential Link

Frédéric Broydé
Excem,
12, chemin des Hauts de Clairefontaine, 78580 Maule, France
e-mail: fredbroyde@eurexcem.com
Tel.: + 33 1 34 75 13 65 Fax.: + 33 1 34 75 13 66

Bernard Démoulin
Université des Sciences et Technologies de Lille,
Bâtiment P3, 59655 Villeneuve d'Ascq Cedex, France
e-mail: bernard.demoulin@univ-lille1.fr
Tel.: + 33 3 20 43 48 56 Fax.: + 33 3 20 43 65 23

Abstract — An interconnection-ground structure used according to the ZXnoise method may be modeled using a $(n + 2)$-conductor multiconductor transmission line (MTL) model. However, a $(n + 1)$-conductor MTL model is used to design the link. We validate this design procedure.

I. Introduction

Pseudo-differential links (PDLs) providing several channels may be used to convey analog or digital signals between a transmitting circuit (TX circuit) and a receiving circuit (RX circuit). We shall refer to crosstalk between the different channels as *internal crosstalk* and to crosstalk with other circuits as *external crosstalk*. A multichannel PDL may provide a reduced external crosstalk compared to multiple single-ended links, using fewer conductors than multiple differential links. A low external crosstalk makes low-swing transmission possible, and low-swing is the key to high-speed transmission. A low external crosstalk is also necessary to implement multilevel digital signaling, which, for a given bit rate, uses a reduced bandwidth compared to binary signaling.

A wide variety of pseudo-differential transmission schemes use an interconnection having $n \geq 2$ transmission conductors (TCs) and one common conductor distinct from the reference conductor (ground):

■ some schemes use no termination circuit, so that their bandwidth is limited by reflections, for a given length;

■ some schemes use type 1 termination circuits, for which an impedance matrix with respect to ground is defined, this impedance matrix being diagonal, such termination circuits creating an undesirable coupling with other circuits using ground as return path;

■ some schemes use type 2 termination circuits or type 3 termination circuits, such termination circuits being characterized by an impedance matrix with respect to the return conductor, denoted by $\mathbf{Z}_{RL}$, $\mathbf{Z}_{RL}$ being a matrix of size $n \times n$, $\mathbf{Z}_{RL}$ being a diagonal matrix in the case of a type 2 termination circuit, or a non-diagonal matrix in the case of a type 3 termination circuit.

The combination of floating termination circuits (i.e., type 2 termination circuits or type 3 termination circuits) with appropriate interconnection-ground structures is referred to as *ZXnoise method* [1] [2]. In the ZXnoise method, the common conductor is usually referred to as *return conductor* because it is used as a return path for the return current produced by the current flowing in the TCs. In Section II, we present new theoretical results on propagation in an interconnection such as the one shown in Fig. 1, used for implementing the ZXnoise method. A design procedure is presented in Section III, and Section IV provides a design example.

II. Propagation in the Interconnection

Each interconnection-ground structure shown in Fig. 1 may be modeled as a $(n+2)$-conductor multiconductor transmission line (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 integer α such that $1 \leq \alpha \leq n$, the natural current i_α ;

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

■ for any integer α such that $1 \leq \alpha \leq n$, the voltage between the TC number α and the reference conductor, denoted by $v_{G\,\alpha}$;

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

We define the column-vector $\mathbf{I}_G$ of the currents $i_1,..., i_{n+1}$ and the column-vector $\mathbf{V}_G$ of the natural voltages referenced to ground $v_{G\,1},..., v_{G\,n+1}$. For the $(n + 2)$-conductor MTL, the telegrapher's equations are:

$$\begin{cases} \dfrac{d\,\mathbf{V}_G}{dz} = -\mathbf{Z}_G\,\mathbf{I}_G \\ \dfrac{d\,\mathbf{I}_G}{dz} = -\mathbf{Y}_G\,\mathbf{V}_G \end{cases} \qquad (1)$$

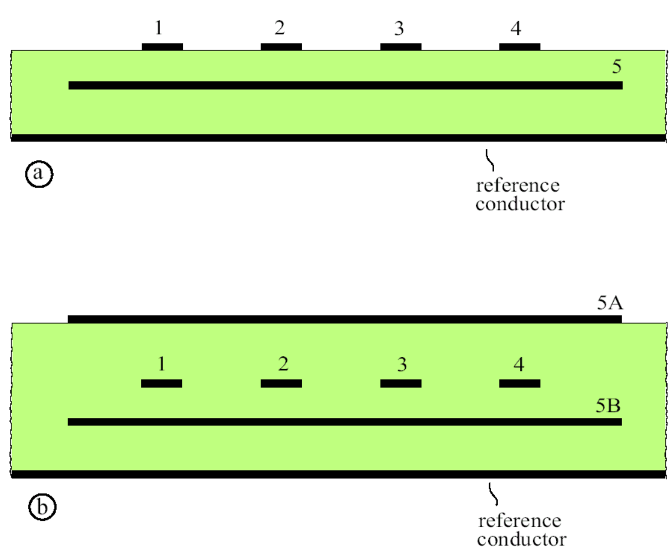

Fig. 1. Two possible cross-sections for an interconnection-ground structure used with floating termination circuits, where 1 to 4 are the TCs, where 5 is the return conductor in *a* and where the return conductor in *b* is made of 5A and 5B.

where $\mathbf{Z}_G$ and $\mathbf{Y}_G$ are the per-unit-length (p.u.l.) impedance matrix with respect to ground, and the p.u.l. admittance matrix with respect to ground, respectively. $\mathbf{Z}_G$ and $\mathbf{Y}_G$ are symmetric matrices of size $(n + 1) \times (n + 1)$.

These equations and appropriate boundary conditions provide a full description of all propagation phenomena in the interconnection-ground structure, including signal propagation, echo and internal crosstalk. They are also suitable for describing external crosstalk phenomena which do not involve a capacitive or inductive coupling with the conductors of the interconnection, for instance the very important problem of simultaneous switching output (SSO) noise produced within IC packages when the interconnection is built in the substrate of a multi-chip module (MCM) or in a printed circuit board (PCB). However these equations are not very convenient for the investigation of signal propagation since a receiving circuit (RX circuit) used in a PDL senses the natural voltages referenced to the return conductor, denoted by $v_{R\,1},..., v_{R\,n}$, where $v_{R\,\alpha}$ denotes the voltage between the TC number α and the return conductor. 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 referenced to the return conductor $v_{R\,1},..., v_{R\,n}$, such that $v_{R\,\alpha} = v_{G\,\alpha} - v_{G\,n+1}$ for $1 \le \alpha \le n$.

The equations involving $d\mathbf{V}_G/dz$ and $d\mathbf{V}_R/dz$ have been compared in the case $n = 1$, in a problem where the interconnection was a coaxial cable [3]. In order to write the telegrapher's equations applicable to $\mathbf{I}_R$ and $\mathbf{V}_R$ in the case $n \ge 2$ relevant to PDLs, we must use 2 additional variables and 6 new p.u.l. quantities. The additional variables are the common-mode current $i_{MC} = i_1 + ... + i_{n+1}$ and the common-mode voltage $v_{MC} = v_{G\,n+1}$. The new p.u.l. quantities are [4]:

■ the p.u.l. impedance matrix with respect to the return conductor, a symmetric matrix of size $n \times n$ denoted by $\mathbf{Z}_R$, the p.u.l. transfer impedance vector, of size $n \times 1$ and denoted by $\mathbf{Z}_E$, and the external impedance, denoted by Z_{EE}, defined by

$$Z_{G\alpha\beta} = Z_{R\alpha\beta} + Z_{EE} - Z_{E\alpha} - Z_{E\beta} \quad \text{and} \quad Z_{G\,n+1\,\alpha} = Z_{G\alpha\,n+1} = Z_{EE} - Z_{E\alpha} \quad \text{and} \quad Z_{G\,n+1\,n+1} = Z_{EE} \tag{2}$$

■ the p.u.l. admittance matrix with respect to the return conductor, a symmetric matrix of size $n \times n$ denoted by $\mathbf{Y}_R$, the p.u.l. transfer admittance vector, of size $n \times 1$ and denoted by $\mathbf{Y}_E$, and the external admittance, denoted by Y_{EE}, defined by

$$Y_{G\alpha\beta} = Y_{R\alpha\beta} \quad \text{and} \quad Y_{G\,n+1\,\alpha} = Y_{G\alpha\,n+1} = Y_{E\alpha} - \sum_{\beta=1}^{n} Y_{R\alpha\beta} \quad \text{and} \quad Y_{G\,n+1\,n+1} = Y_{EE} + \sum_{\alpha=1}^{n}\sum_{\beta=1}^{n} Y_{R\alpha\beta} - 2\sum_{\alpha=1}^{n} Y_{E\alpha} \tag{3}$$

where α and β are integers such that $1 \le \alpha \le n$ and $1 \le \beta \le n$ and where indices have been used to denote the entries of matrices and vectors. Using ${}^t\mathbf{X}$ to denote the transpose of $\mathbf{X}$, it can been shown that (1) is exactly equivalent to

$$\begin{cases} \dfrac{d\,\mathbf{V}_R}{dz} = -\mathbf{Z}_R\,\mathbf{I}_R + i_{MC}\,\mathbf{Z}_E \\ \dfrac{d\,\mathbf{I}_R}{dz} = -\mathbf{Y}_R\,\mathbf{V}_R - v_{MC}\,\mathbf{Y}_E \end{cases} \quad (4a) \qquad \text{and} \qquad \begin{cases} \dfrac{d\,v_{MC}}{dz} = {}^t\mathbf{Z}_E\,\mathbf{I}_R - i_{MC}\,Z_{EE} \\ \dfrac{d\,i_{MC}}{dz} = -{}^t\mathbf{Y}_E\,\mathbf{V}_R - v_{MC}\,Y_{EE} \end{cases} \quad (4b)$$

The equations (4a) and (4b) are not based on any approximation and are therefore valid for any pseudo-differential transmission scheme. Also, they clarify and improve the existing theory for the ZXnoise method.

In the case where the return conductor behaves as an ideal screen, the terms containing the p.u.l. transfer impedance vector $\mathbf{Z}_E$ or the p.u.l. transfer admittance vector $\mathbf{Y}_E$ vanish in (4a) and (4b), so that

$$\begin{cases} \dfrac{d\,\mathbf{V}_R}{dz} = -\mathbf{Z}_R\,\mathbf{I}_R \\ \dfrac{d\,\mathbf{I}_R}{dz} = -\mathbf{Y}_R\,\mathbf{V}_R \end{cases} \quad (5a) \qquad \text{and} \qquad \begin{cases} \dfrac{d\,v_{MC}}{dz} = -i_{MC}\,Z_{EE} \\ \dfrac{d\,i_{MC}}{dz} = -v_{MC}\,Y_{EE} \end{cases} \quad (5b)$$

Thus, the reference conductor being not regarded as a part of the interconnection, (5a) means that the interconnection may be modeled as a $(n + 1)$-conductor MTL using the natural voltages referenced to the return conductor and the natural currents $i_1,..., i_n$ as natural electrical variables. According to (5b), the propagation in the return-conductor-and-ground circuit may be modeled as a 2-conductor MTL using the common-mode voltage and the common-mode current as natural electrical variables. The equation (5a) is the basis of the published design procedure for ZXnoise PDLs [1] [2]. The equation (5b) describes only the propagation of noise produced by external sources since, in a PDL, the return-conductor-and-ground circuit is not used for signals.

In the case where the return conductor behaves as a good screen, the terms containing $\mathbf{Z}_E$ or $\mathbf{Y}_E$ in (4a) and (4b) may be regarded as small disturbances and treated using perturbation theory. In the framework of first order perturbation theory, the propagation of signal is still determined by (5a) and the propagation of noise produced by external sources by (5b), these results being used in (4a) to obtain the external crosstalk received by the PDL and in (4b) to obtain the external crosstalk produced by the PDL.

If the electric and magnetic fields of the signals are mainly confined between the TCs and the return conductor, we can say that the return conductor behaves as a good screen for signals. Thus, (5a) may be used to design the PDL so as to obtain a suitable propagation of signals. In particular, a termination circuit designed in this manner is a floating termination circuit, since it creates a boundary condition for the natural voltages referenced to the return conductor and the natural currents $i_1,..., i_n$. At a later stage, an analysis of the PDL may be directly based on (1), or on (4a) and (4b).

III. DESIGN FOR THE ZXNOISE METHOD

Based on (5a) alone, we shall use $\mathbf{T}_R$ and $\mathbf{S}_R$ to denote two regular matrices such that [5]:

$$\begin{cases} \mathbf{T}_R^{-1}\mathbf{Y}_R\mathbf{Z}_R\mathbf{T}_R = \Gamma_R^2 \\ \mathbf{S}_R^{-1}\mathbf{Z}_R\mathbf{Y}_R\mathbf{S}_R = \Gamma_R^2 \end{cases} \quad \text{where} \quad \Gamma_R = \operatorname{diag}_n(\gamma_{R1},\ldots,\gamma_{Rn}) \tag{6}$$

is the diagonal matrix of order n of the propagation constants for the different propagation modes of the $(n+1)$-conductor MTL, for waves propagating toward the far-end (that is to say in the direction of increasing z). The squares of the propagation constants are the eigenvalues of $\mathbf{Y}_R\mathbf{Z}_R$, which are also the eigenvalues of $\mathbf{Z}_R\mathbf{Y}_R$.

Any matrices $\mathbf{T}_R$ and $\mathbf{S}_R$ satisfying (6) define a "modal transform" for the natural currents and for the natural voltages referenced to the return conductor, and the results of this transform are called modal currents and modal voltages, respectively. If we use $\mathbf{I}_{RM}$ to denote the column-vector of the n modal currents $i_{RM\,1},\ldots, i_{RM\,n}$ and $\mathbf{V}_{RM}$ to denote the column-vector of the n modal voltages $v_{RM\,1},\ldots, v_{RM\,n}$, we get:

$$\begin{cases} \mathbf{V}_R = \mathbf{S}_R\mathbf{V}_{RM} \\ \mathbf{I}_R = \mathbf{T}_R\,\mathbf{I}_{RM} \end{cases} \tag{7}$$

The characteristic impedance matrix of the $(n+1)$-conductor MTL, denoted by $\mathbf{Z}_{RC}$, and the matrix of the voltage reflection coefficients of a floating termination circuit with respect to the return conductor, denoted by $\mathbf{P}_R$, are given by:

$$\begin{aligned} \mathbf{Z}_{RC} &= \mathbf{S}_R\Gamma_R^{-1}\mathbf{S}_R^{-1}\mathbf{Z}_R = \mathbf{S}_R\Gamma_R\mathbf{S}_R^{-1}\mathbf{Y}_R^{-1} \\ &= \mathbf{Y}_R^{-1}\mathbf{T}_R\Gamma_R\mathbf{T}_R^{-1} = \mathbf{Z}_R\mathbf{T}_R\Gamma_R^{-1}\mathbf{T}_R^{-1} \end{aligned} \tag{8a}$$

and

$$\mathbf{P}_R = (\mathbf{Z}_{RL} - \mathbf{Z}_{RC})(\mathbf{Z}_{RL} + \mathbf{Z}_{RC})^{-1} \tag{8b}$$

where $\mathbf{Z}_{RL}$ is the impedance matrix of the floating termination circuit with respect to the return conductor. $\mathbf{Z}_{RC}$ is a matrix of size $n \times n$ referred to as the "characteristic impedance matrix with respect to the return conductor".

Let us consider a given interconnection-ground structure meeting our requirements regarding external crosstalk, in which the propagation of signals may be modeled using (5a). If the desired reduction of reflection, expressed as a maximum value for a norm of $\mathbf{P}_R$, can be obtained with a diagonal matrix $\mathbf{Z}_{RL}$, and if the resulting internal crosstalk is acceptable, then a PDL implementing a type 2 termination circuit can be designed. Such a PDL uses one TC for each channel. In the opposite case, since $\mathbf{P}_R = \mathbf{0}$ for $\mathbf{Z}_{RL} = \mathbf{Z}_{RC}$, we may design a type 3 termination circuit producing very low echo. It will also provide a low internal crosstalk if we use one modal current or one modal voltage defined by (7) for each channel, so as to transpose to PDLs the ZXtalk method [5] or the special ZXtalk method for completely degenerate interconnection (CDI) [6]. In general, analog and/or digital processing are needed in the TX circuits and in the RX circuits, to perform linear combinations defined by $\mathbf{S}_R$ or $\mathbf{T}_R$. However, in the special case where the propagation constants of the $(n+1)$-conductor MTL model may be regarded as equal, linear combinations are not needed in the TX circuits and/or in the RX circuits, since $\mathbf{S}_R$ or $\mathbf{T}_R$ may be chosen equal to the identity matrix. Nearly equal propagation constants may for instance be obtained using the structure shown in Fig. 1*b*.

IV. EXAMPLE

We now discuss the design of a unidirectional PDL providing $m = 4$ channels, between two chips of a MCM. The PDL uses the ZXnoise method, according to the voltage-driven common conductor architecture. We assume that the PDL is equivalent to the schematic diagram shown in Fig. 2. At the near-end, the return conductor is connected to a node intended to present a fixed voltage e_{CC} and a low impedance Z_{CC} with respect to a ground plane of the substrate of the MCM. The TX circuit, being connected to this node, produces accurate open-circuit output voltages with respect to it. This node is also subject to a noise voltage e_N caused by other circuits within the chip of the TX circuit (e.g., SSO noise). At the far-end, the return conductor is connected to the termination circuit and to a damping resistor R_D. We assume an ideal RX circuit that is undisturbed by noise source inside its chip.

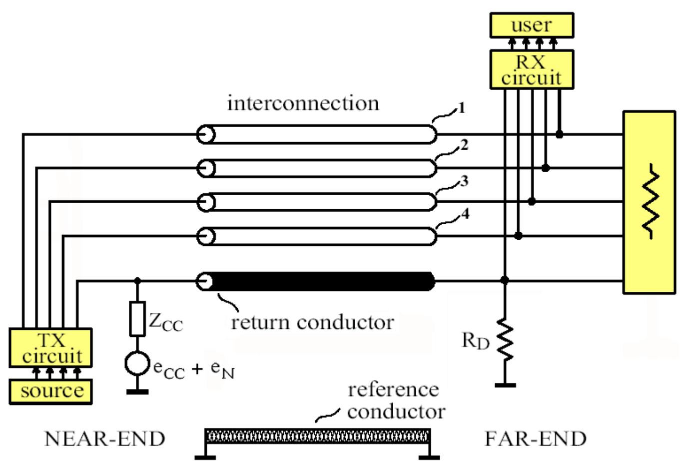

Fig. 2. A PDL with voltage-driven common conductor. The block containing the resistor symbol is a termination circuit.

The PDL uses a 30-mm-long interconnection having the cross-section shown in Fig. 1*a*. The parameters defined in Fig. 2 have the following nominal values: conductor spacing $s \approx 1.2\,w$, distance between conducting layers $h \approx H \approx 0.64\,w$ and a return conductor overhang $v \approx 1.2\,w$. The values of $\mathbf{Z}_G$ (at 50 MHz) and $\mathbf{Y}_G$ measured on a scaled-up model are given by $\mathbf{Z}_G = j\omega\,\mathbf{L}_G$ and $\mathbf{Y}_R = j\omega\,\mathbf{C}_R$, with

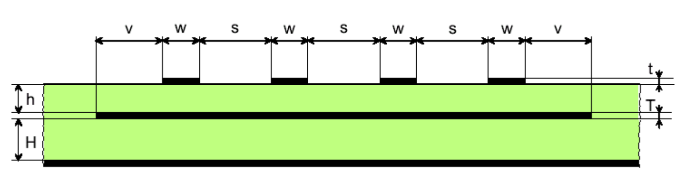

Fig. 3. Cross-section of an interconnection-ground structure built in a printed circuit board.

$$
\mathbf{L}_G \approx \begin{pmatrix} 407 & 92 & 69 & 61 & 66 \\ 92 & 414 & 94 & 69 & 70 \\ 69 & 94 & 410 & 93 & 69 \\ 61 & 69 & 93 & 404 & 66 \\ 66 & 70 & 69 & 66 & 85 \end{pmatrix} \text{ nH/ m}
$$

and

$$
\mathbf{C}_G \approx \begin{pmatrix} 116.6 & -3.8 & -0.4 & -0.2 & -108.4 \\ -3.8 & 115.9 & -3.7 & -0.4 & -106.4 \\ -0.4 & -3.7 & 116.9 & -3.9 & -107.6 \\ -0.2 & -0.4 & -3.9 & 117.3 & -109.4 \\ -108.4 & -106.4 & -107.6 & -109.4 & 1098 \end{pmatrix} \text{ pF/ m}
$$

For the p.u.l. impedance parameters of the (n+1)-conductor MTL model, using (2), we obtain $\mathbf{Z}_R = j\omega\, \mathbf{L}_R$ and $\mathbf{Z}_E = j\omega\, \mathbf{L}_E$ with

$$
\mathbf{L}_R \approx \begin{pmatrix} 359 & 41 & 17 & 13 \\ 41 & 359 & 39 & 17 \\ 17 & 39 & 356 & 42 \\ 13 & 17 & 42 & 356 \end{pmatrix} \text{ nH/ m}\,, \quad \mathbf{L}_E \approx \begin{pmatrix} 18 \\ 15 \\ 15 \\ 18 \end{pmatrix} \text{ nH/ m}
$$

and $Z_{EE} = j\omega\, L_{EE}$ with $L_{EE} \approx 85$ nH/m.

For the p.u.l. admittance parameters of the (n+1)-conductor MTL model, using (3), we obtain $\mathbf{Y}_R = j\omega\, \mathbf{C}_R$ and $\mathbf{Y}_E = j\omega\, \mathbf{C}_E$ with

$$
\mathbf{C}_R \approx \begin{pmatrix} 116.6 & -3.8 & -0.4 & -0.2 \\ -3.8 & 115.9 & -3.7 & -0.4 \\ -0.4 & -3.7 & 116.9 & -3.9 \\ -0.2 & -0.4 & -3.9 & 117.3 \end{pmatrix} \text{ pF/ m}\,, \quad \mathbf{C}_E \approx \begin{pmatrix} 3.8 \\ 1.6 \\ 1.4 \\ 3.5 \end{pmatrix} \text{ pF/ m}
$$

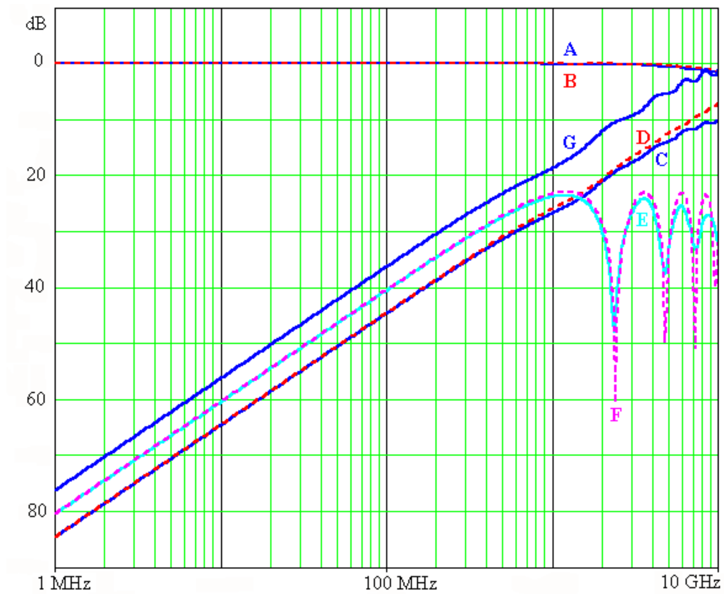

Fig. 4. Some frequency domain simulation results. Attenuation of transmitted signal when the TC number 1 is excited: curve A (blue, solid) according to (1) and curve B (red, dash) according to (5a). Far-end crosstalk loss on the TC number 2 when the TC number 1 is excited: curve C (blue, solid) according to (1) and curve D (red, dash) according to (5a). Near-end crosstalk loss on the TC number 2 when the TC number 1 is excited: curve E (cyan, solid) according to (1) and curve F (magenta, dash) according to (5a). Far-end external crosstalk loss on the TC number 1 when all conductors are excited at the near-end: curve G (blue, solid) according to (1).

and $Y_{EE} = j\omega\, C_{EE}$ with $C_{EE} \approx 1520$ pF/m. As expected, the entries of $\mathbf{L}_E$ and $\mathbf{C}_E$ are much smaller than the diagonal entries of $\mathbf{L}_R$ and $\mathbf{C}_R$, respectively, so that we may plan to use $\mathbf{L}_R$ and $\mathbf{C}_R$ for designing a ZXnoise PDL, according to Section III.

For a type 2 termination circuit, the minimum value of the matrix norm $\| \mathbf{P}_R \|_\infty$ is 0.083, which may be obtained with a termination circuit comprising two 57.7 Ω resistors and two 55.0 Ω resistors. However, our simulations use a termination circuit comprising one 50.0 Ω resistor connected between each TC and the return conductor, for which $\| \mathbf{P}_R \|_\infty = 0.131$, and a damping resistor of $R_D = 10$ Ω. Using the approximate (n + 1)-conductor MTL model corresponding to (5a), we obtain the curves B, D and F of Fig. 4 when a signal is sent through the channel 1. Using the exact (n + 2)-conductor MTL model for the same configuration, we obtain the curves A, C and E of Fig. 4. The curve G of Fig. 4 represents the rejection of external crosstalk, which can only be computed using the (n + 2)-conductor MTL model.

V. Conclusion

Up to 3 GHz, the rejection of external crosstalk exceeds 10 dB and the agreement between the (n + 1)-conductor MTL model and the (n + 2)-conductor MTL model is excellent. This validates (2), (3) and (4), and also the use of the design procedure outlined in Section III, which produces floating termination circuits because it is based on the (n + 1)-conductor MTL model. We have also computed the rejection of external crosstalk, which compares the performance of the PDL to that of multiple single-ended links. Our simple PDL design provides an effective reduction of external crosstalk, for instance 36 dB at 100 MHz and 19 dB at 1 GHz.

References

[1] F. Broydé, and E. Clavelier, “Pseudo-differential links using a wide return conductor and a floating termination circuit”, *Proc. of the 2008 IEEE International Midwest Symposium on Circuits and Systems (MWSCAS)*, August 10-13, 2008, pp. 586-589.

[2] F. Broydé, E. Clavelier, “A new pseudo-differential transmission scheme for on-chip and on-board interconnections”, *Proc. of the CEM 08 Int. Symp. on Electromagnetic Compatibility*, Paris, France, May 2008, Available: http://www.eurexcem.com/zxnoisedef.htm.

[3] B. Démoulin, A.P.J. van Deursen, “Deux approches pour établir le lien entre la notion usuelle d’impédance de transfert et le formalisme des lignes couplées”, *Proc. of the 10e Colloque International & Exposition sur la Compatibilité Électromagnétique, CEM 2000*, Clermont-Ferrand, France, March 2000, pp. 98-103.

[4] F. Broydé, and E. Clavelier, “Modeling the interconnection of a pseudo-differential link using a wide return conductor”, *Proc. of the 13th IEEE Workshop on Signal Propagation on Interconnects, SPI 2009*, May 12-15, 2009.

[5] F. Broydé, and E. Clavelier, “A New Method for the Reduction of Crosstalk and Echo in Multiconductor Interconnections”, *IEEE Trans. Circuits Syst. I: Regular Papers*, 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: Regular Papers*, vol. 53, No. 8, p. 1851, Aug. 2006.

[6] F. Broydé, and E. Clavelier, “Echo-Free and Crosstalk-Free Transmission in Particular Interconnections”, *IEEE Microwave and Wireless Components Letters*, Vol. 19, No. 4, April 2009, pp. 209-211.