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Some Internal Crosstalk Reduction Schemes

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***Abstract* — We present and discuss different schemes which can be defined to mitigate internal crosstalk in transposed and untransposed uniform interconnections, and in non-uniform interconnections. We emphasize the unique characteristics of modal transmission schemes.**

I. Introduction

This paper is about a multichannel point-to-point electrical link, shown in Fig. 1, built in a multi-chip module (MCM) or PCB, providing $m \geq 2$ channels from a near-end interface and termination device (NIT) to a far-end interface and termination device (FIT). The link is linear and comprises a multiconductor interconnection having $n \geq m$ transmission conductors (TCs) and a reference conductor or ground conductor (GC). The link may also provide one or more channels from the FIT to the NIT.

We assume a close spacing of the TCs, which causes a significant *TC-to-TC coupling*, which collectively designates mutual capacitance between the TCs and mutual impedance between loops each comprising one of the TCs and the GC. Thus, some analog and/or digital processing of signals present on two or more TCs is necessary to obtain each channel with a sufficiently low *internal crosstalk*, that is to say a sufficiently low interaction with the other channels of the link.

We want to study different schemes which may be defined to mitigate internal crosstalk in transposed and untransposed uniform interconnections, and in non-uniform interconnections. Section II discusses uniformity and transposition. The sections III and IV present the internal crosstalk mitigation problem, and discuss possible solutions. Section V covers modal transmission schemes in transposed and untransposed interconnections.

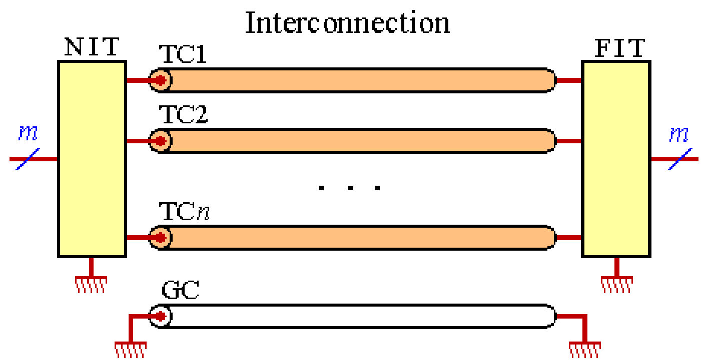

Fig. 1. A point-to-point link providing m channels, consisting of an interconnection, a near-end interface and termination device (NIT) and the far-end interface and termination device (FIT). The interconnection has n transmission conductors (TCs) and a reference conductor (GC).

II. Uniformity and Transposition

We use z to denote the curvilinear abscissa along the interconnection, which extends from $z = 0$ to $z = \mathcal{L}$. We shall only consider frequency domain variables. In the framework of multiconductor transmission line (MTL) theory, at a given z, the interconnection is characterized by a per-unit-length (p.u.l.) impedance matrix denoted by $\mathbf{Z}'$, and a p.u.l. admittance matrix, denoted by $\mathbf{Y}'$. From the standpoint of the computation of wave propagation in the interconnection, it is legitimate to use values of $\mathbf{Z}'$ and $\mathbf{Y}'$ which are averaged over a length sufficiently smaller than the shortest wavelength of interest, denoted by λ. From the standpoint of measuring techniques, a sufficient length of the interconnection, compared to this wavelength, is necessary to obtain accurate measurements of $\mathbf{Z}'$ and $\mathbf{Y}'$, which are consequently averaged over the length of the sample under test.

Traditionally, *uniform* means independent of z. However, in this paper, we shall define a uniform interconnection as being such that $\mathbf{Z}'$ and $\mathbf{Y}'$ are independent of z, after a suitable averaging over a reasonable length, e.g., $\lambda/100$, is taken into account.

At the beginning of the 20th century, engineers knew that an effective method of reducing crosstalk in telegraph and telephone transmission is the use of a separate circuit for each signal to be transmitted, the circuits being substantially perfectly balanced to each other by means of very frequent transposition of the TCs of each circuit [1]. In transposition, permutations of the positions of the TCs, at intervals along the interconnection, are used to simultaneously obtain a balanced interconnection, and a uniform interconnection in the bandwidth of interest.

A perfectly balanced interconnection comprises p pairs such that [2]: the TCs of the same pair have the same *averaged* p.u.l. impedance and p.u.l. admittance with respect to the reference conductor; and the excitation of any pair in differential mode induces no voltage and injects no current in any other conductor. Let us number the TCs of a p-pair interconnection in the following way: for $\alpha \in \{1,..., p\}$, the TC number $2\alpha - 1$ is the 1st wire of the α-th pair, and the TC number 2α is the 2nd wire of the α-th pair. Let us use $\mathbf{X}'$ to denote any one of the *averaged* natural matrices $\mathbf{Z}'$ or $\mathbf{Y}'$. $\mathbf{X}'$ is of size $2p \times 2p$ and the interconnection is perfectly balanced if and only if

$$\forall \alpha \in \{1,\ldots,p\} \quad \forall \beta \in \{1,\ldots,p\}$$

we have $$X'_{2\alpha-1\;2\alpha-1} = X'_{2\alpha\;2\alpha} \tag{1}$$

and $$(\alpha \neq \beta) \Rightarrow \left(X'_{2\alpha-1\;2\beta-1} = X'_{2\alpha-1\;2\beta} = X'_{2\alpha\;2\beta-1} = X'_{2\alpha\;2\beta}\right)$$

Particular properties are obtained with a super-balanced

Fig. 2. A 2-pair super-balanced interconnection built in a PCB. The two colors used for the traces correspond to different layers.

interconnection, defined as a perfectly balanced interconnection in which any pair can be exchanged with any other pair without changing $\mathbf{Z}'$ or $\mathbf{Y}'$. The interconnection is super balanced if and only if there exist X_A, X_B and X_M such that [2]:

$$\begin{aligned} &\forall\, \alpha \in \{1,\ldots,p\} \quad \forall\, \beta \in \{1,\ldots,p\} \\ &\qquad X_{2\alpha-1,\,2\alpha-1} = X_{2\alpha,\,2\alpha} = X_A \\ &\qquad X_{2\alpha-1,\,2\alpha} = X_{2\alpha,\,2\alpha-1} = X_B \\ &(\alpha \neq \beta) \Rightarrow \left(\begin{aligned} X_{2\alpha-1,\,2\beta-1} &= X_{2\alpha-1,\,2\beta} \\ &= X_{2\alpha,\,2\beta-1} = X_{2\alpha,\,2\beta} = X_M \end{aligned} \right) \end{aligned} \tag{2}$$

This corresponds to what Carson and Hoyt called "the ideal telephone transmission system" in 1927 [3, eq. (19)].

Transposition is a current technique used to obtain high-speed balanced interconnection, for instance in UTP category 5e twisted-pair cables for 1000BASE-T local area networks, which comprise 4 pairs and can be used up to 100 MHz over a distance of 100 m. It can also be used in a link built in a PCB or MCM [4] [5], for instance using a super-balanced interconnection such as the one shown in Fig. 2.

III. General Formulation of Internal Crosstalk Mitigation Schemes

Let us use $\mathbf{e}_S$ to denote the column vector of the open-circuit voltages applied by the NIT to the interconnection, for the m channels from the NIT to the FIT, and for a column vector of the input signals $\mathbf{x}_{IS}$ applied to these channels. $\mathbf{e}_S$ is of size $n \times 1$, and $\mathbf{x}_{IS}$ is of size $m \times 1$. A *premixing matrix* $\mathbf{A}_S$, of size $n \times m$ and rank m, defines linear combinations of signals performed in the transmitting circuit (TX-circuit) of the NIT, such that

$$\mathbf{e}_S = \mathbf{A}_S\ \mathbf{x}_{IS} \tag{3}$$

As said above, one or more channels from the FIT to the NIT may be present, in which case the link is bidirectional. Thus we need to consider a column vector of the open-circuit voltages applied by the FIT to the interconnection, denoted by $\mathbf{e}_L$, of size $n \times 1$, for a column vector of the input signals $\mathbf{x}_{IL}$, of size $q \times 1$, applied to these channels. A premixing matrix $\mathbf{A}_L$, of size $n \times q$ and rank q, defines linear combinations of signal performed in the TX-circuit of the FIT, such that

$$\mathbf{e}_L = \mathbf{A}_L\ \mathbf{x}_{IL} \tag{4}$$

The column vector of the voltages measured by the FIT is denoted by $\mathbf{v}_L$, and given by

$$\mathbf{v}_L = \mathbf{H}_{LS}\,\mathbf{e}_S + \mathbf{H}_{LL}\,\mathbf{e}_L + \mathbf{n}_L \tag{5}$$

where $\mathbf{H}_{LS}$ and $\mathbf{H}_{LL}$ are matrix transfer functions, both of size $n \times n$, and $\mathbf{n}$ a noise vector. The FIT determines the column vector of the output signals $\mathbf{x}_{OL}$ of the m channels from the NIT to the FIT, $\mathbf{x}_{OL}$ being of size $m \times 1$ and given by

$$\mathbf{x}_{OL} = \mathbf{B}_L\,(\mathbf{v}_L - \mathbf{D}_L\,\mathbf{e}_L) \tag{6}$$

where $\mathbf{B}_L$ is a *demixing matrix*, of size $m \times n$, which defines linear combination of signals intended to recover the wanted signal sent by the NIT, and where $\mathbf{D}_L$ is a *duplexing matrix*, of size $n \times n$, whose purpose is the reduction (ideally, the cancellation) of the contribution of $\mathbf{x}_{IL}$ to $\mathbf{x}_{OL}$ to allow simultaneous bidirectional (full duplex) transmission. In a context where digital signal processing would be used to implement said linear combinations, $\mathbf{A}_S$ and $\mathbf{A}_L$ could each be referred to as a *coding matrix* or a *precoding matrix*, and $\mathbf{B}_L$ as a *decoding matrix*. We have:

$$\mathbf{x}_{OL} = \mathbf{B}_L\left(\mathbf{H}_{LS}\mathbf{A}_S\mathbf{x}_{IS} + \left[\mathbf{H}_{LL} - \mathbf{D}_L\right]\mathbf{A}_L\mathbf{x}_{IL} + \mathbf{n}_L\right) \tag{7}$$

which describes transmission from the NIT to the FIT. All variables in (7) may be frequency dependent. Thus, the mixing and demixing matrices can in principle provide preemphasis and deemphasis, respectively to obtain a compensation of the frequency dependent losses in the matrix transfer functions.

If the link provides only unidirectional (simplex), or alternate bidirectional (half-duplex) transmission, we may consider that $\mathbf{x}_{IL} = \mathbf{0}$ in (7). Here, the problem of internal crosstalk reduction consists in finding an interconnection structure (which determines $\mathbf{H}_{LS}$), a premixing matrix $\mathbf{A}_S$ and a demixing matrix $\mathbf{B}_S$ such that $\mathbf{A}_S$ and $\mathbf{B}_S$ each corresponds to a causal impulse response matrix, and such that, in the relevant frequency range, for any $\alpha \in \{1,\ldots, m\}$

$$R'_\alpha\left(\mathbf{B}_L\mathbf{H}_{LS}\mathbf{A}_S\right) \le \varepsilon_\alpha \left|\left[\mathbf{B}_L\mathbf{H}_{LS}\mathbf{A}_S\right]_{\alpha\,\alpha}\right| \tag{8}$$

where ε_α is an arbitrary positive real which defines the maximum allowed signal to crosstalk ratio, where $[\mathbf{M}]_{\alpha\,\beta}$ is the entry of the row α and column β of a matrix $\mathbf{M}$, and where, if $\mathbf{M}$ is of size $m \times m$, $R'_\alpha(\mathbf{M})$ is the αth deleted absolute row sum of $\mathbf{M}$, given by

$$R'_\alpha(\mathbf{M}) = \sum_{\substack{\beta=1 \\ \beta\neq\alpha}}^{m} \left|\left[\mathbf{M}\right]_{\alpha\,\beta}\right| \tag{9}$$

We see that (8) means that $\mathbf{B}_L\ \mathbf{H}_{LS}\ \mathbf{A}_S$ is sufficiently close to a diagonal matrix, and implies a requirement on the location of the eigenvalues of $\mathbf{B}_L\ \mathbf{H}_{LS}\ \mathbf{A}_S$, because each $R'_\alpha(\ \mathbf{B}_L\ \mathbf{H}_{LS}\ \mathbf{A}_S\ )$ is the radius of a Geršgorin disk [6, § 6.1]. If crosstalk cancellation is required, then $\varepsilon_\alpha = 0$ for any $\alpha \in \{1,\ldots, m\}$, so that (8) means that $\mathbf{B}_L\ \mathbf{H}_{LS}\ \mathbf{A}_S$ is diagonal.

We are mostly interested in interconnections having a length $\mathcal{L}$ which is not very small compared to λ, because they exhibit a higher crosstalk when no crosstalk mitigation technique is used. In this case, $\mathbf{H}_{LS}$ corresponds to a causal time domain response which contains a propagation delay, so that $\mathbf{H}_{LS}^{-1}$ does not correspond to a causal time domain response. Thus, using $\mathbf{1}_n$ to denote the identity matrix of size $n \times n$, we see that, in the case $n = m$:

- $\mathbf{B}_L = \mathbf{1}_n$ and $\mathbf{A}_S = \mathbf{H}_{LS}^{-1}$ is not an acceptable solution of (8);
- $\mathbf{B}_L = \mathbf{H}_{LS}^{-1}$ and $\mathbf{A}_S = \mathbf{1}_n$ is not an acceptable solution of (8).

If the link provides full duplex transmission, internal crosstalk cancellation additionally requires that we find a duplexing matrix $\mathbf{D}_L$ corresponding to a causal impulse response matrix, such that, in the relevant frequency range, for any $\alpha \in \{1,..., m\}$

$$\sum_{\beta=1}^{m} \left| \left[(\mathbf{H}_{LL} - \mathbf{D}_L) \mathbf{A}_L \right]_{\alpha\,\beta} \right| \le \varepsilon_\alpha \left| \left[\mathbf{B}_L \mathbf{H}_{LS} \mathbf{A}_S \right]_{\alpha\,\alpha} \right| \tag{10}$$

We see that (10) means that $(\mathbf{H}_{LL} - \mathbf{D}_L)\,\mathbf{A}_L$ is sufficiently close to a null matrix. If crosstalk cancellation is required, (10) means that $(\mathbf{H}_{LL} - \mathbf{D}_L)\,\mathbf{A}_L$ is a null matrix. In the case $q = n$, crosstalk cancellation is obtained if and only if $\mathbf{D}_L = \mathbf{H}_{LL}$, which of course corresponds to a causal impulse response matrix.

IV. Case of a Uniform Interconnection

We now assume that the interconnection is uniform in the meaning of § II. It may be transposed or untransposed. Here, we can compute $\mathbf{H}_{LS}$ and $\mathbf{H}_{LL}$ explicitly. Using [7, eq. 72] we get

$$\mathbf{H}_{LS} = \frac{1}{2}(\mathbf{1}_n + \mathbf{P}_L) e^{-\mathcal{L}\mathbf{G}} \times \left(\mathbf{1}_n - \mathbf{P}_S\, e^{-\mathcal{L}\mathbf{G}}\, \mathbf{P}_L\, e^{-\mathcal{L}\mathbf{G}}\right)^{-1} (\mathbf{1}_n - \mathbf{P}_S) \tag{11}$$

and

$$\mathbf{H}_{LL} = \frac{1}{2}\left(\mathbf{1}_n + e^{-\mathcal{L}\mathbf{G}}\, \mathbf{P}_S\, e^{-\mathcal{L}\mathbf{G}}\right) \times \left(\mathbf{1}_n - \mathbf{P}_L\, e^{-\mathcal{L}\mathbf{G}}\, \mathbf{P}_S\, e^{-\mathcal{L}\mathbf{G}}\right)^{-1} (\mathbf{1}_n - \mathbf{P}_L) \tag{12}$$

where $\mathbf{G}$ is the lineic propagation matrix given by $\mathbf{G} = \sqrt{\mathbf{Z}'\mathbf{Y}'}$ and where $\mathbf{P}_S$ and $\mathbf{P}_L$ are the matrix of the voltage reflection coefficients at the near end and the matrix of the voltage reflection coefficients at the far end, respectively, given by

$$\mathbf{P}_S = (\mathbf{Z}_S - \mathbf{Z}_C)(\mathbf{Z}_S + \mathbf{Z}_C)^{-1} \tag{13}$$

and

$$\mathbf{P}_L = (\mathbf{Z}_L - \mathbf{Z}_C)(\mathbf{Z}_L + \mathbf{Z}_C)^{-1} \tag{14}$$

where $\mathbf{Z}_S$ and $\mathbf{Z}_L$ are the impedance matrices presented by the NIT and the FIT, respectively, to the interconnection, and where $\mathbf{Z}_C$ is the characteristic impedance matrix given by

$$\mathbf{Z}_C = \sqrt{\mathbf{Z}'\mathbf{Y}'}^{-1} \mathbf{Z}' \tag{15}$$

The multiple reflection terms, $(\mathbf{1}_n - \mathbf{P}_S\ e^{-\mathcal{L}\mathbf{G}}\,\mathbf{P}_L\ e^{-\mathcal{L}\mathbf{G}})^{-1}$ in (11) and $(\mathbf{1}_n - \mathbf{P}_L\ e^{-\mathcal{L}\mathbf{G}}\,\mathbf{P}_S\ e^{-\mathcal{L}\mathbf{G}})^{-1}$ in (12), may cause an impulse response which lasts for a long time, and consequently increase dramatically the cost of the signal processing circuit used to implement $\mathbf{A}_S$, $\mathbf{B}_L$ or $\mathbf{D}_L$. It is therefore always advisable to require that

$$|||\, \mathbf{P}_L \,|||_\infty \;\; |||\, \mathbf{P}_S \,|||_\infty \;\; |||\, \mathbf{e}^{-\mathcal{L}\mathbf{G}} \,|||_\infty << 1 \tag{16}$$

which is easily obtained if the NIR provides reflectionless matching, i.e. $\mathbf{P}_S = \mathbf{0}_{n\,n}$, or if the FIR provides reflectionless matching, i.e. $\mathbf{P}_L = \mathbf{0}_{n\,n}$, where $\mathbf{0}_{n\,n}$ is the null matrix of size $n \times n$.

V. Modal Signaling

Let us now consider a uniform interconnection such that $\mathbf{Z}'\,\mathbf{Y}'$ is diagonalizable. The transition matrix from modal voltages to natural voltages, denoted by $\mathbf{S}$, is a solution of

$$\mathbf{S}^{-1}\mathbf{Z}'\,\mathbf{Y}'\,\mathbf{S} = \Gamma^2 \tag{17}$$

where

$$\Gamma = \mathrm{diag}_n(\gamma_1, \ldots, \gamma_n) \tag{18}$$

is the diagonal matrix of order n of the propagation constants.

The definition of $\mathbf{S}$ involves multiple choices. Let assume that such choices lead us to obtain $\mathbf{S}$ as a function of frequency. Since $e^{-\mathcal{L}\mathbf{G}} = \mathbf{S}\ e^{-\mathcal{L}\Gamma}\ \mathbf{S}^{-1}$, we observe that, if $\mathbf{P}_S = \mathbf{0}_{n\,n}$ and $\mathbf{P}_L = \mathbf{0}_{n\,n}$, or $\mathbf{P}_S = -\mathbf{1}_n$ and $\mathbf{P}_L = \mathbf{0}_{n\,n}$, or $\mathbf{P}_S = \mathbf{0}_{n\,n}$ and $\mathbf{P}_L = \mathbf{1}_n$, then by (11) and (12) we may conclude that $\mathbf{S}^{-1}\,\mathbf{H}_{LS}\,\mathbf{S}$ and $\mathbf{S}^{-1}\,\mathbf{H}_{LL}\,\mathbf{S}$ are diagonal matrices. This indicates that we could consider using $\mathbf{A}_S = \mathbf{S}$ and $\mathbf{B}_L = \mathbf{S}^{-1}$ to solve (8). However, there is no guarantee that this solution is acceptable, because $\mathbf{S}$ and $\mathbf{S}^{-1}$ need not be frequency-domain descriptions of linear systems having a causal impulse response matrix. In fact, $\mathbf{S}$ and $\mathbf{S}^{-1}$ do not even need to be continuous functions of frequency.

This lead us to the general concept of modal signaling as defined in [8]. In a conventional modal signaling scheme:

- the interconnection model used for the initial design of the link is a uniform MTL model, referred to as the underlying MTL model, which need not be a perfectly accurate model, for which the p.u.l impedance matrix and the p.u.l impedance matrix will be denoted by $\mathbf{Z}'_U$ and $\mathbf{Y}'_U$, respectively;
- each of the m transmission channels from the NIT to the FIT is allocated to a modal electrical variable, that is a modal voltage or modal current of the underlying MTL model.

Assuming that $\mathbf{Z}'_U\,\mathbf{Y}'_U$ is diagonalizable, let us define, for the underlying MTL model, the transition matrix from modal voltages to natural voltages, denoted by $\mathbf{S}_U$ and the transition matrix from modal currents to natural currents, denoted by $\mathbf{T}_U$, as solutions of

$$\begin{cases} \mathbf{T}_U^{-1}\mathbf{Y}'_U\mathbf{Z}'_U\,\mathbf{T}_U = \Gamma_U^2 \\ \mathbf{S}_U^{-1}\mathbf{Z}'_U\,\mathbf{Y}'_U\,\mathbf{S}_U = \Gamma_U^2 \end{cases} \tag{19}$$

where

$$\Gamma_U = \mathrm{diag}_n(\gamma_{U1}, \ldots, \gamma_{Un}) \tag{20}$$

is the diagonal matrix of order n of the propagation constants, $\mathbf{S}_U$ and $\mathbf{T}_U$ being additionally required to meet the necessary and sufficient condition for total decoupling [7, § 7] [9, § III]

$$\mathbf{S}_U = j\omega\ \mathbf{Y}'^{-1}_U\,\mathbf{T}_U\,\mathbf{c}_{UK} \tag{21}$$

where $\mathbf{c}_{UK}$ is an arbitrary invertible diagonal matrix, possibly frequency-dependent, and having the dimensions of p.u.l. capacitance. The characteristic impedance matrix of the underlying MTL model, given by

$$\mathbf{Z}_{CU} = \sqrt{\mathbf{Z}'_U\mathbf{Y}'_U}^{-1}\mathbf{Z}'_U \tag{22}$$

may be used to define the values of $\mathbf{Z}_S$ and $\mathbf{Z}_L$. Total decoupling entails that, for any $i \in \{1,..., n\}$, the propagation of the i-th modal voltage corresponds to the propagation of the i-th modal current, (in line with [7, eq. (48)] or [9, eq. (24)]). We may therefore assume that each of the m transmission channels is allocated to a modal voltage of the underlying MTL model, so that, if $m = n$,

$$\mathbf{A}_S = (\mathbf{Z}_S + \mathbf{Z}_{CU})\mathbf{Z}_{CU}^{-1}\mathbf{S}_U\ \mathrm{diag}_n(\alpha_1, \ldots, \alpha_n) \tag{23}$$

and

$$\mathbf{B}_L = \mathrm{diag}_n(\beta_1, \ldots, \beta_n)\,\mathbf{S}_U^{-1}\mathbf{Z}_{CU}(\mathbf{Z}_S + \mathbf{Z}_{CU})^{-1} \tag{24}$$

where the α_i and the β_i are arbitrary nonzero possibly frequency dependent parameters. If $m < n$, $\mathbf{A}_S$ and $\mathbf{B}_L$ are submatrices of the matrices given by (23) and (24), respectively. A key aspect of modal signaling is that $\mathbf{A}_S$ and $\mathbf{B}_L$ are independent of $\mathcal{L}$, if the α_i and the β_i are chosen to be independent of $\mathcal{L}$. This simplifies the design of the NIT and FIT.

In a link using a transposed interconnection, the underlying MTL model typically assumes a symmetry which defines $\mathbf{S}_U$. For instance, if the underlying MTL model assumes a super-balanced interconnection, $\mathbf{S}_U$ may be chosen to be a frequency-independent real (FIR) matrix given by $\mathbf{S}_U = \mathbf{A}^{-1}$, where $\mathbf{A}$ is given by [2, eq. 9]. For instance, for the two-pair interconnection shown in Fig. 2, we may use

$$\mathbf{S}_U = \begin{pmatrix} 1/2 & 0 & 1/2 & 1 \\ -1/2 & 0 & 1/2 & 1 \\ 0 & 1/2 & -1/2 & 1 \\ 0 & -1/2 & -1/2 & 1 \end{pmatrix} \tag{25}$$

Many designs based on a transposed interconnection only use the first p modes, referred to as differential modes, which correspond to the first two-columns in (25). The other p modes are, however, also available for transmission.

In the case of an untransposed interconnection, the underlying MTL model is typically chosen to be an approximate model such that $\mathbf{S}_U$ and $\mathbf{Z}_{CU}$ are FIR matrices, for instance a lossless model, or preferably a model which can accurately take high-frequency losses into account, such as the "fourth MTL" defined in [9, § IV]. If $\mathbf{S}_U$, $\mathbf{Z}_{CU}$ and $\mathbf{Z}_S$ are FIR, and if the α_i and the β_i correspond to causal time domain responses, then $\mathbf{A}_S$ and $\mathbf{B}_L$ correspond to causal time domain responses. Thus, the α_i and the β_i can be chosen to provide preemphasis and deemphasis.

The ZXtalk method [7, § 14] refers to a special case of modal signaling using an untransposed interconnection, in which a non-diagonal $\mathbf{Z}_S$ and/or a non-diagonal $\mathbf{Z}_L$ are used to approximate $\mathbf{Z}_{CU}$, in order to satisfy (16), so as to comply with (8). As an example, we consider a 4-channel point-to-point link in which the 20 mm-long interconnection is the multiconductor microstrip used in [9, § V], which presents resistive and dielectric losses. Here, $m = n = 4$. The link uses the ZXtalk method for simplex transmission, and only the FIT comprises a termination circuit. $\mathbf{A}_S$ and $\mathbf{B}_L$ are real and frequency independent, so that there is no equalization. The termination circuit is made of $2\,n - 1 = 7$ resistors. The signals at the far-end, computed with the lossy MTL model described in [9, § V], are shown in Fig. 3. We see that crosstalk is not canceled, because of the approximations made in the synthesis of the circuits of the link. However, compared to the single-ended link considered in [9, § V], a reduction of internal crosstalk of about 40 dB has been obtained above 10 GHz, and also a reduction of echo and linear distortions in the channels.

In the general ZXtalk method described above, the number of terms of the linear combinations performed in the NIT and FIT increases as n^2. This becomes a problem for large values of n. A special ZXtalk method [7, § 15] uses an interconnection for which we can use $\mathbf{S}_U = \mathbf{1}_n$, so that this problem is not present.

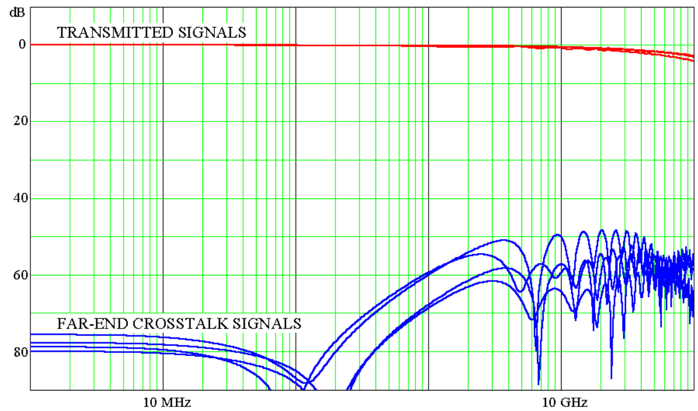

Fig. 3. Attenuations at the far-end when a signal is applied at the near-end of any channel in a ZXtalk link providing $m = 4$ channels. Sixteen curves are plotted, four for the transmitted signals and twelve for the far-end crosstalk signals.

VI. Conclusion

A link using an internal crosstalk reduction schemes must satisfy (8). We have stressed that, in addition, the premixing matrix $\mathbf{A}_S$ and a demixing matrix $\mathbf{B}_S$ complying with (8) must each corresponds to a causal time domain response. We have also noted that the cost of the analog or digital processing used to realize $\mathbf{A}_S$ and $\mathbf{B}_S$ is decreased when (16) is satisfied.

These considerations are used to support a view presented in [8]: modal signaling is a special case of noise subtraction where the signal processing requirements defined by (23) and (24) are light, and the ZXtalk technique is a special case of modal signaling in which the signal processing requirements are minimal because non-diagonal $\mathbf{Z}_S$ and/or $\mathbf{Z}_L$ are used to reduce reflections.

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