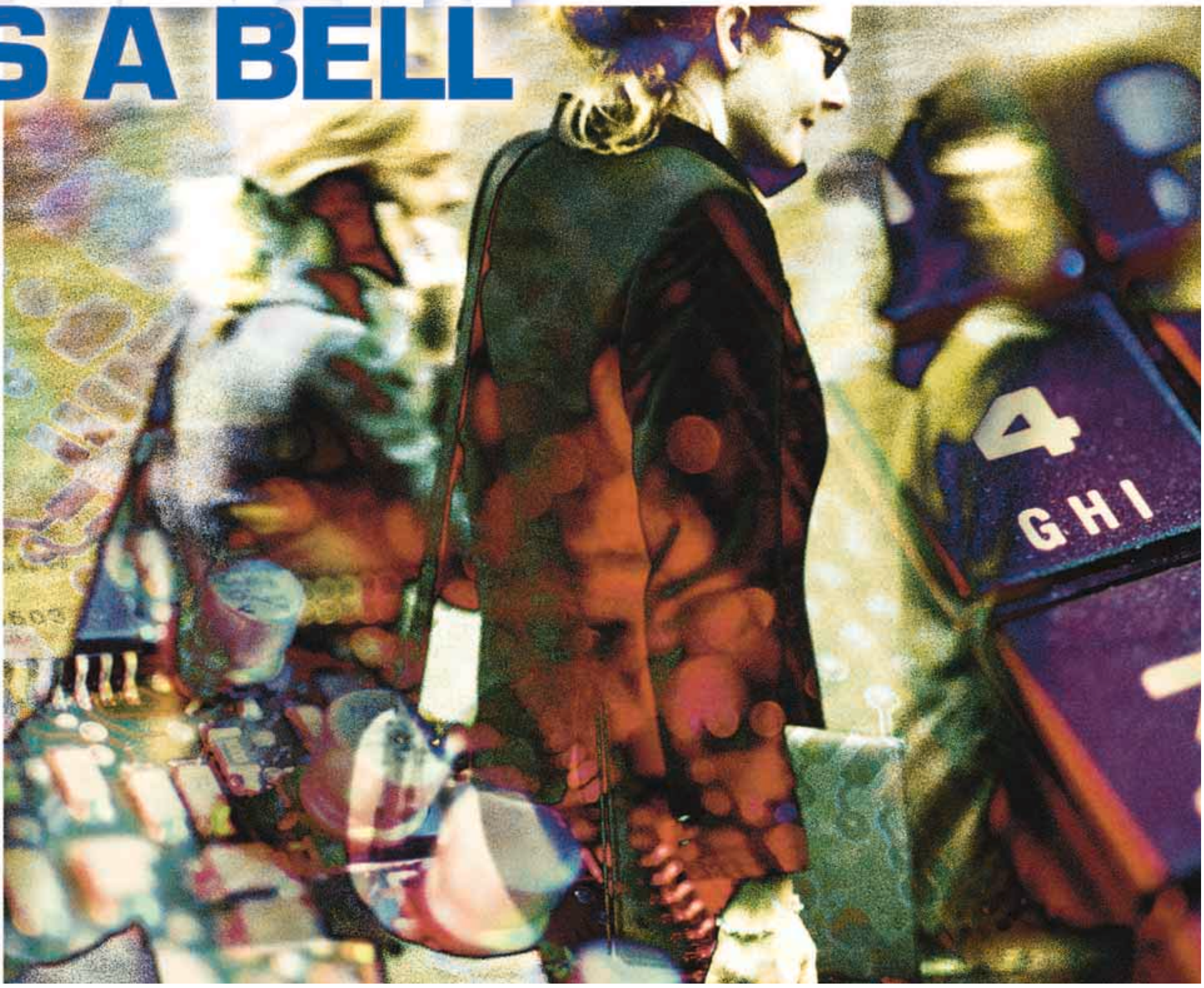
CLEAR AS A BELL



Controlling Crosstalk in Uniform Interconnections

Frédéric Broydé

Designers are aware of several approaches to reduce crosstalk, such as the use of shields or differential links, that require a particular structure for the interconnection. This article discusses known crosstalk mitigation methods and presents a different technique applicable in various situations, which provides a drastic reduction of crosstalk.

Electronic designers must often fight the phenomenon of crosstalk in interconnections: a signal intentionally injected on a conductor of the interconnection gives rise to an unwanted crosstalk signal, which may degrade the performances of a given system or make it inoperative. Since the old telegraph days, engineers have fought crosstalk, because it takes place in various types of interconnections such as telecommunication cables, local area networks, printed circuit boards, and on-chip interconnects. The increase of speed and bandwidth on the one hand and of circuit and interconnection density on the other hand worsen this phenomenon, so that crosstalk is sometimes the limiting factor for the use of the most advanced technologies. Crosstalk also depends on the length of the interconnection. As a consequence, controlling crosstalk is desirable

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CLASSICAL SOLUTIONS FOR THE REDUCTION OF CROSSTALK

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

The second solution consists of 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 (multicoax 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 STPs, such as some cables for high speed data transmission. Adding conductors 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, vias, 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).

The third solution is to increase the distance between the transmission conductors used for different channels. This approach is often not compatible with cost and size requirements.

The fourth solution is to decrease the distance between the transmission and ground conductors. 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.

The fifth solution consists in reducing the upper limit of the frequency band used by the signals to be send. This solution can, of course, not be used in situations where the bandwidth of these signals cannot be modified.

The sixth solution consists of 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 with limited performances.

All of 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 concentrate on approaches which are less demanding with respect to the cross section of the interconnection; in order to send n signals, we would like to only use n transmission conductors and a ground conductor, in the most compact way.

MATCHED TERMINATIONS AND PSEUDOMATCHED IMPEDANCES

Rigorously matching a multiconductor transmission line (MTL) at one end amounts to completely removing reflections at the other end for waves coming from the MTL. This implies that, at one end, a termination is connected, which presents an impedance matrix equal to the characteristic impedance matrix of the MTL (see "Characteristic Impedance Matrix and Transition Matrices of a Multiconductor Transmission Line"). The circuit of such a termination will typically be a network of $n(n+1)/2$ resistors in the case of sufficiently high frequencies, for which the characteristic impedance matrix may be regarded as real.

In order to obtain the integrity of signals, one of the first rules is to reduce reflections. To this end, designers never use matched terminations (according to the above definition) because such terminations create crosstalk. Instead, a common practice implements, for each transmission conductor, a two-terminal linear circuit element inserted between the transmission conductor and ground [1], 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*. We note that their value is not defined in a unique way in the literature.

In order to illustrate these concepts, a few simulations will show signals obtained with typical input and output impedances, and with pseudo-matched impedances. We have shown in Figure 1 a schematic including a 30-cm long MTL with two transmission conductors, connecting linear circuits simply represented by their Thevenin equivalent circuit. Referring to their position on the schematic, these conductors will be called "top" and "bottom." For the top transmission conductor, an output is connected on the left end of the MTL (internal impedance $R1 = 35\ \Omega$), and an input is connected to the right

Characteristic Impedance Matrix and Transition Matrices of a Multiconductor Transmission Line

The $n+1$ conductors of a multiconductor transmission line (MTL) are numbered from 0 to n, the conductor 0 being the reference conductor (for instance, a ground plane). When the MTL can be regarded as having electrical characteristics sufficiently uniform over its length (that is to say, independent of the curvilinear coordinate z along the MTL), it is characterized for the transmission of signals and for crosstalk by a per-unit-length (p.u.l.) inductance matrix $\boldsymbol{L}$, a p.u.l. resistance matrix $\boldsymbol{R}$, a p.u.l. capacitance matrix $\boldsymbol{C}$, and a p.u.l. conductance matrix $\boldsymbol{G}$, these four matrices being independent of z. When losses are neglected, one assumes that $\boldsymbol{R}=0$, $\boldsymbol{G}=0$, and that $\boldsymbol{L}$ and $\boldsymbol{C}$ are independent of frequency. For instance, for the MTL of Figure 1, neglecting losses, we have used the matrices provided by Paul [2] for a particular configuration of two PCB traces above a ground plane:

$$\boldsymbol{L}=\begin{pmatrix}0.8629 & 0.3725\\ 0.3725 & 0.8629\end{pmatrix}\mu\text{H/m} \tag{1}$$

$$\boldsymbol{C}=\begin{pmatrix}46.762 & -18.036\\ -18.036 & 46.762\end{pmatrix}\text{pF/m}. \tag{2}$$

Propagation and crosstalk on the MTL are governed by the telegrapher's equations:

$$\begin{cases}\dfrac{d\boldsymbol{V}}{dz}=-(\boldsymbol{R}+j\omega\boldsymbol{L})\boldsymbol{I}\\[2mm] \dfrac{d\boldsymbol{I}}{dz}=-(\boldsymbol{G}+j\omega\boldsymbol{C})\boldsymbol{V}\end{cases} \tag{3}$$

where ω is the radian frequency, $\boldsymbol{V}$ is the vector of the n natural voltages (a natural voltage is a voltage between a transmission conductor and the reference conductor) and $\boldsymbol{I}$ is the vector of the n natural currents (a natural current is a current on one of the transmission conductors).

We shall now use $\boldsymbol{Z}=\boldsymbol{R}+j\omega\boldsymbol{L}$ to denote the p.u.l. impedance matrix and $\boldsymbol{Y}=\boldsymbol{G}+j\omega\boldsymbol{C}$ to denote the per-unit length admittance matrix. A suitable diagonalization of the matrices $\boldsymbol{ZY}$ and $\boldsymbol{YZ}$ can be used to solve (3). The eigenvectors obtained in this manner define the propagation modes, and the eigenvalues correspond to the propagation constants. More precisely, we shall use $\boldsymbol{T}$ and $\boldsymbol{S}$ to denote two regular matrices, called transition matrices from modal electrical variables to natural electrical variables, such that:

$$\begin{cases}\boldsymbol{T}^{-1}\boldsymbol{YZT}=\boldsymbol{D}\\ \boldsymbol{S}^{-1}\boldsymbol{ZYS}=\boldsymbol{D}\end{cases} \tag{4}$$

where

$$\boldsymbol{D}=\text{diag}_n(\gamma_1^2,\ldots,\gamma_n^2) \tag{5}$$

is the diagonal matrix of order n of the eigenvalues. From (4) and (5), it is possible to define the characteristic impedance matrix $\boldsymbol{Z}_C$ of the MTL as:

$$\begin{aligned}\boldsymbol{Z}_C&=\boldsymbol{S}\Gamma^{-1}\boldsymbol{S}^{-1}\boldsymbol{Z}=\boldsymbol{S}\Gamma\boldsymbol{S}^{-1}\boldsymbol{Y}^{-1}\\ &=\boldsymbol{Y}^{-1}\boldsymbol{T}\Gamma\boldsymbol{T}^{-1}=\boldsymbol{Z}\boldsymbol{T}\Gamma^{-1}\boldsymbol{T}^{-1}\end{aligned} \tag{6}$$

where

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

is the diagonal matrix of order n of the propagation constants γ_i, which have the dimensions of the inverse of a length. The characteristic impedance matrix is automatically computed by the SpiceLine [4] software. For instance, for the MTL of Figure. 1, we obtain

$$\boldsymbol{S}=\begin{pmatrix}1.0912 & 2.4616\\ -1.0912 & 2.4612\end{pmatrix}\quad \boldsymbol{T}=\begin{pmatrix}0.70711 & 0.70711\\ -0.70711 & 0.70711\end{pmatrix} \tag{8}$$

$$\boldsymbol{Z}_C=\begin{pmatrix}147.187 & 60.1923\\ 60.1923 & 147.187\end{pmatrix}\Omega, \tag{9}$$

which explains the value 147.2 Ω used for the resistors R1–R4 when pseudo-matched impedances are considered.

For the MTL of Figure 9, we have used the following matrices [5]:

$$\boldsymbol{L}=\begin{pmatrix}0.3139 & 0.0675 & 0.0222\\ 0.0675 & 0.3139 & 0.0675\\ 0.0222 & 0.0675 & 0.3139\end{pmatrix}\mu\text{H/m} \tag{10}$$

$$\boldsymbol{C}=\begin{pmatrix}130.3 & -16.2 & -0.8\\ -16.2 & 133.7 & -16.2\\ -0.8 & -16.2 & 130.3\end{pmatrix}\text{pF/m} \tag{11}$$

and SpiceLine has computed

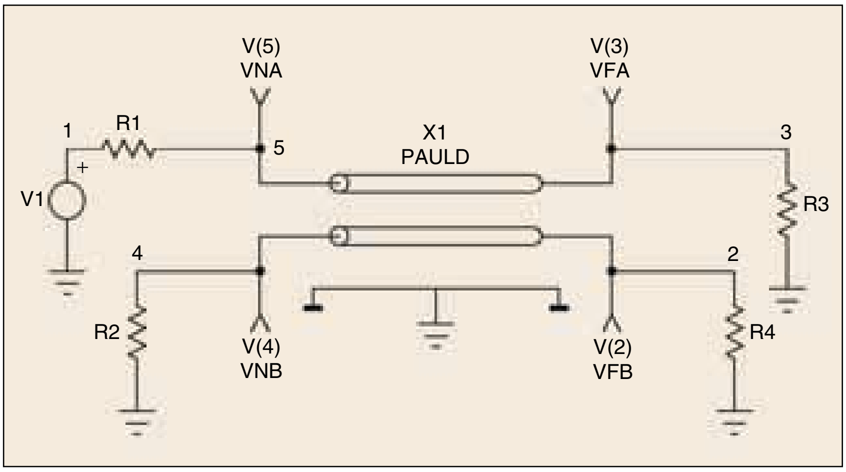
$$\boldsymbol{S}=\begin{pmatrix}0.3101 & -0.5394 & -0.4793\\ -0.4755 & 0 & -0.6232\\ 0.3101 & 0.5394 & -0.4793\end{pmatrix} \tag{12}$$

$$\boldsymbol{T}=\begin{pmatrix}0.4786 & -0.7071 & 0.5198\\ -0.7361 & 0 & 0.6780\\ 0.4786 & 0.7071 & 0.5198\end{pmatrix} \tag{13}$$

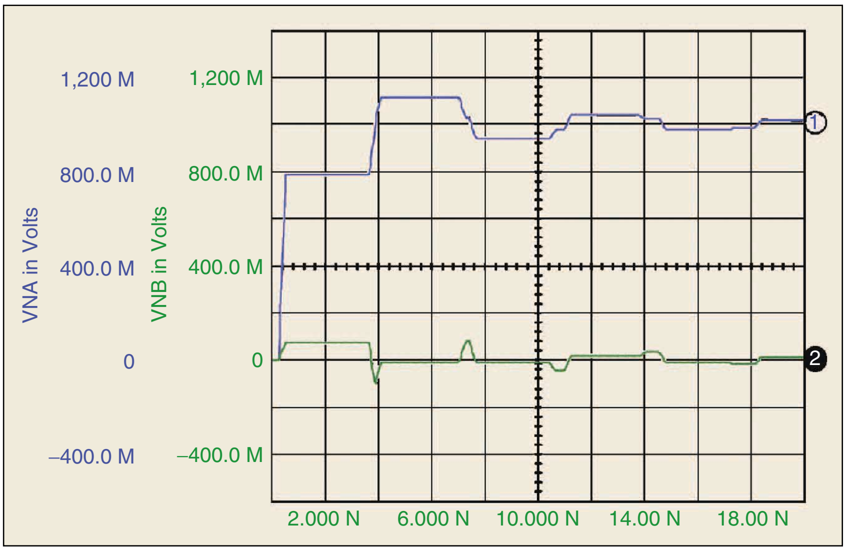
and

$$\boldsymbol{Z}_C=\begin{pmatrix}49.41 & 8.35 & 2.24\\ 8.35 & 49.35 & 8.35\\ 2.24 & 8.35 & 49.41\end{pmatrix}\Omega. \tag{14}$$

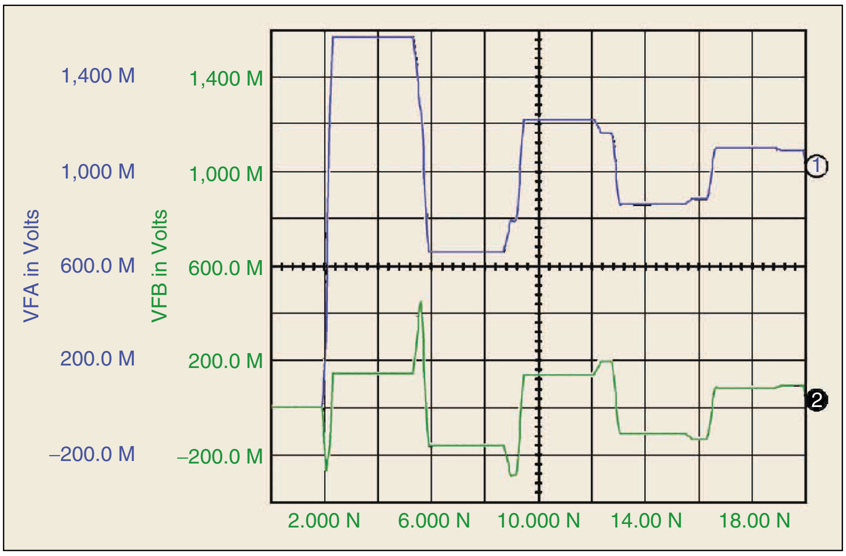
For this MTL, the pseudomatched impedances are close to 49.4 Ω . One can easily check that the resistances R401 = R403 = 58.7 Ω, R402 = 69.2 Ω, R404 = R405 = 289.5 Ω and R406 = 2781 Ω correspond to a termination having an impedance matrix very close to $\boldsymbol{Z}_C$.



1. An MTL with two transmission conductors and its terminations.



2. Signals measured at the near end for R1 = R2 = 35 Ω and R3 = R4 = 100 kΩ: the VNA signal (top) and the NEXT signal VNB (bottom).



3. Signals measured at the far end for R1 = R2 = 35 Ω and R3 = R4 = 100 kΩ: the transmitted signal VFA (top) and the FEXT signal VFB (bottom).

end of the MTL (very high input impedance R3 = 100 kΩ). Such low output impedances and high input impedances are typical of MOS digital circuits, but we neglect so many features of real circuits, for instance, nonlinearities, input capacitances, etc, that we prefer to say that we do not intend to represent any particular technology in the following simulations. For the bottom transmission conductor, two cases will be considered: the case of an output at the left end with an input at the right end (R2 = 35 Ω and R4 = 100 kΩ) and the case of an input at the left end with an output at the right end (R2 = 100 kΩ and R4 = 35 Ω). Since we consider linear circuits for which voltages can be scaled to any value, we have considered that the active output produces a step with an open-circuit amplitude of 1 V, corresponding to the value of the voltage source V1. The assumed rise time (0–100%) is 250 ps. The results of SPICE simulations performed according to a classical method [3] using transmission line models created by SpiceLine [4] are the following.

- For R2 = 35 Ω and R4 = 100 kΩ, the voltages at the test points VNA and VNB are shown in Figure 2. The signal VNA is degraded by a strong echo (that is to say, a reflection of the signal at the right end of the MTL), which will also give rise to an added crosstalk. The near-end crosstalk (NEXT) signal VNB remains below 100 mV because it is measured across the low impedance of an output.
- For R2 = 35 Ω and R4 = 100 kΩ, the voltages at test points VFA and VFB are shown in Figure 3. The transmitted signal VFA is strongly distorted by multiple reflections and presents a strong overshoot. The far-end crosstalk (FEXT) signal VFB is very strong (about 450 mV peak).
- For R2 = 100 kΩ and R4 = 35 Ω, the voltages at the test points VNA and VNB are shown in Figure 4. The signal VNA is degraded by a strong echo, as was the case in Figure 2. The NEXT sig-

nal VNB is extremely strong (about 800 mV peak-to-peak).

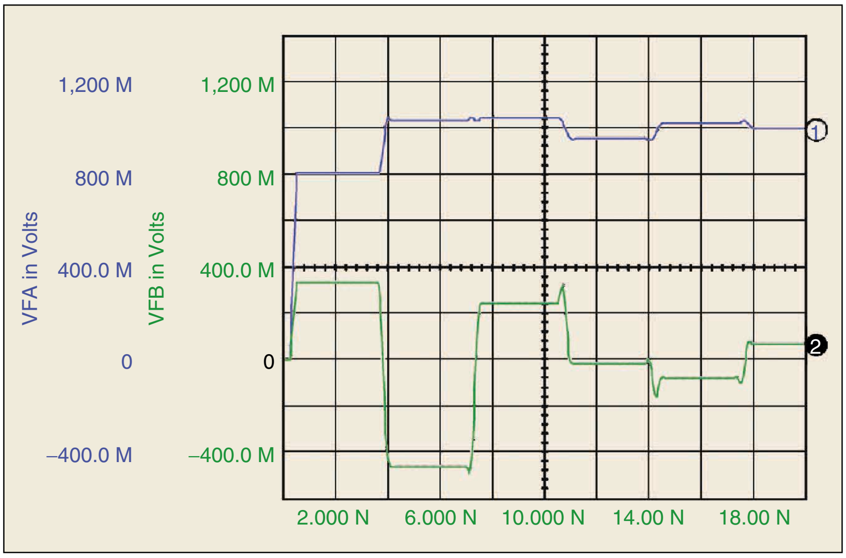
- For R2 = 100 kΩ and R4 = 35 Ω, the voltages at test points VFA and VFB are shown in Figure 5. The transmitted signal VFA is strongly distorted by multiple reflections and presents a strong overshoot, as was the case in Figure 3. The FEXT signal VFB remains below 140 mV because it is connected to the low impedance of an output.

The ZXtalk technique provides a large reduction of crosstalk and echo, improved signal integrity, and increased speed.

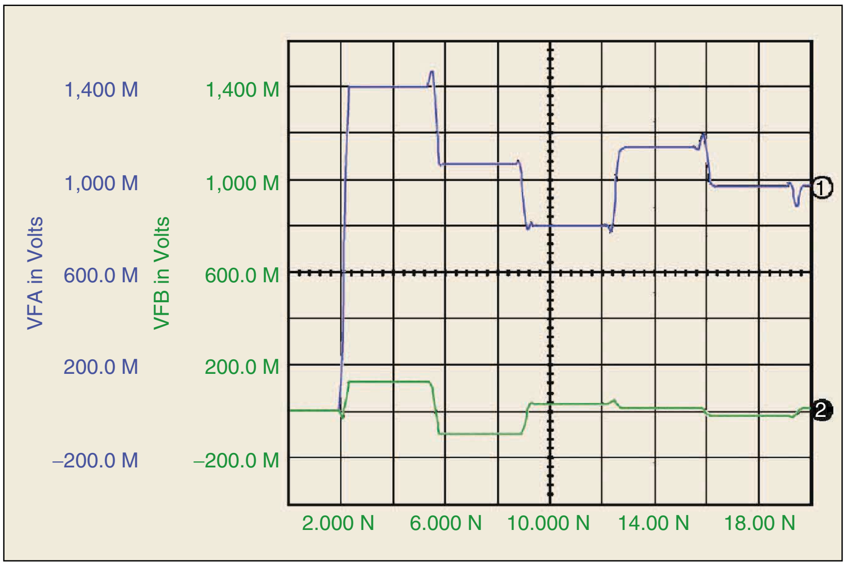
The echo and crosstalk received at inputs with the values considered above for R1–R4 are clearly not satisfactory for signal transmission. We will now investigate the case of the same interconnection used with pseudo-matched impedances connected to each end of each transmission conductor. We will use the values R1 = R2 = R3 = R4 = 147.2 Ω (see "Characteristic Impedance Matrix and Transition Matrices of a Multiconductor Transmission line"). A 2-V step is produced by the voltage source V1 in order to obtain a nominal 1-V step at VNA and VFA. The rise time is 250 ps, as previously noted. The results are the following:

- The voltages at the test points VNA and VNB are shown in Figure 6. The shape of the signal VNA shows that there is only a small remaining mismatch. As a result, the NEXT signal VNB is much smaller than the one obtained in Figure 4, but it is not negligible (about 220-mV peak).
- The voltages at the test points VFA and VFB are shown in Figure 7. The transmitted signal VFA is not severely distorted. The FEXT signal VFB is much smaller (about 180-mV peak) than the one shown in Figure 3.

However, for an interconnection connected to terminations having an impedance matrix corresponding to pseudo-matched impedances tied to ground, the FEXT signal is (within a known maximum value) practically proportional to the length of the interconnection and inversely proportional to the rise time of the step [6]. As a result, if slower transition [after all, the rise time of typical bipolar compementary metal-oxide semiconductor (BiCMOS) devices are longer than the one used in our simulations] and/or a shorter MTL had been considered, the FEXT signal would have been smaller than the one shown in Figure 7. In addition, we note that if the limitation stated by Jarvis [6] concerning the length of the interconnection (one fourth of the distance traveled during the transition time) had been observed, the FEXT and NEXT signal would have been smaller. However, for a 250-ps rise time and a propagation velocity of about 1.7×10^8 m/s applicable to this interconnection, its length would have been limited to about 1 cm!



4. Signals measured at the near end for R1 = R4 = 35 Ω and R2 = R3 = 100 kΩ: the VNA signal (top) and the NEXT signal VNB (bottom).



5. Signals measured at the far end for R1 = R4 = 35 Ω and R2 = R3 = 100 kΩ: the transmitted signal VFA (top) and the FEXT signal VFB (bottom).

A NEW TECHNIQUE TO CANCEL CROSSTALK

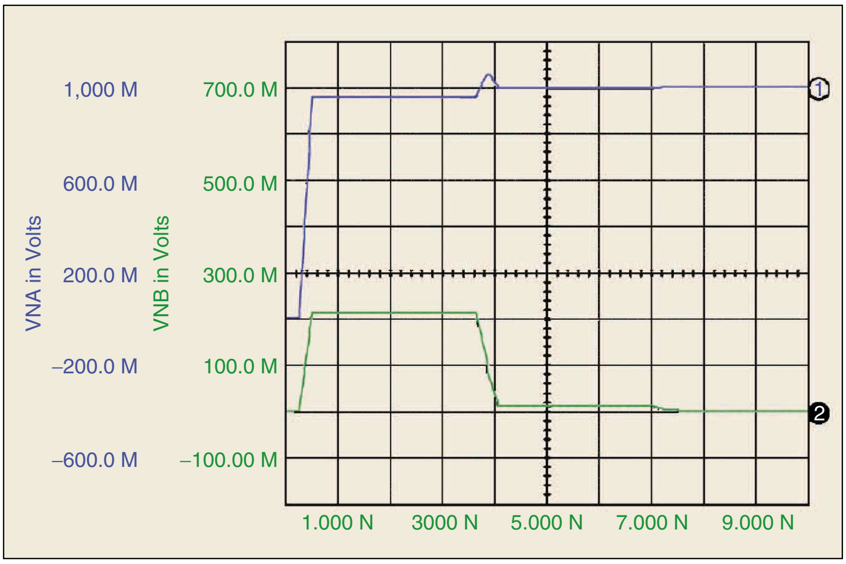
We have just seen that the use of pseudomatched impedances does not always solve the problem of crosstalk when an interconnection using the reference conductor as return path for all signals is too long or used with too fast signals. In the above examples, we note that even the use of pseudo-matched termination at both ends leaves NEXT and FEXT levels which are likely to be a problem (see the bottom plots of Figures 6 and 7).

We are now going to describe the basics of a new technique that, in theory, allows the cancellation of crosstalk and echo completely. This technique, called ZXtalk, is implemented in Figure 8. It is applicable to interconnections with n transmission conductors, which may be modeled as a uniform MTL with a sufficient accuracy. This technique is mainly characterized by the following points.

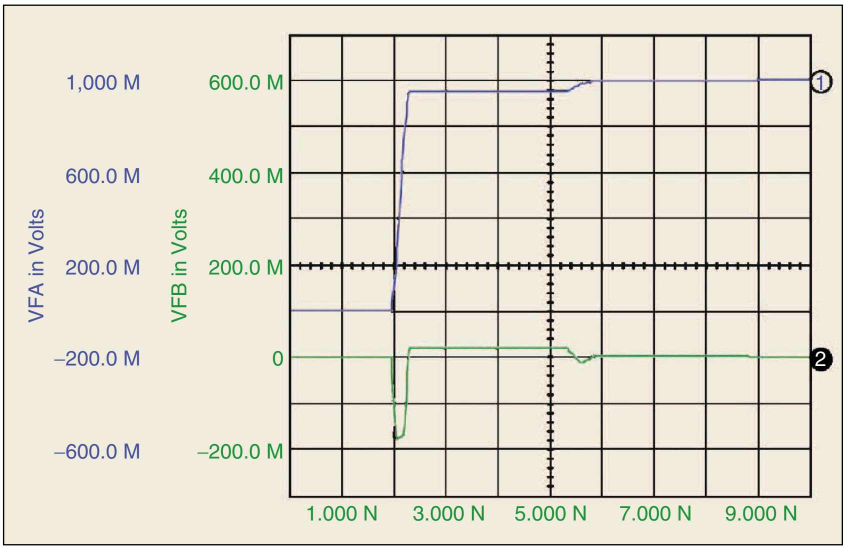
- The interconnection (#1 in Figure 8) is connected at at least one end to a matched termination (#4 in Figure 8); that is to say, a termination having an impedance matrix close to the characteristic impedance matrix of the MTL.
- One or several transmitting circuits (#5 in Figure 8) combine the input signals generated by sources (#2 in Figure 8) according to linear combinations defined by a transition matrix from modal electrical variables to natural electrical variables (see "Characteristic Impedance Matrix and Transition Matrices of a Multiconductor Transmission Line"), the output of such a 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 Figure 8) that each combines the signals present on these conductors according to linear combinations defined by the inverse of the transition matrix, each receiving circuit providing at its output the signals for a destination (#3 in Figure 8).

We note that, in the special case of Figure 8, we have a data bus architecture intended for bidirectional transmission and that 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 present a high impedance to the interconnection, in order not to disturb the propagation of waves along the interconnection and not to produce undesirable reflections. Using the concept of modal voltages and modal currents (not addressed in this article), one may show that the signals of the four channels of a source connected to an active transmitting circuit are sent to the four channels of the destinations without noticeable crosstalk.

There are many possible implementations for this technique, which may use analog and/or digital circuits. We will not discuss real implementations in this article. However, we have shown in Figure 9 the schematic for the SPICE simulation of a theoretical example of a 30 cm long interconnection with three transmission conductors. In this simple example, the interconnection is intended for unidi-

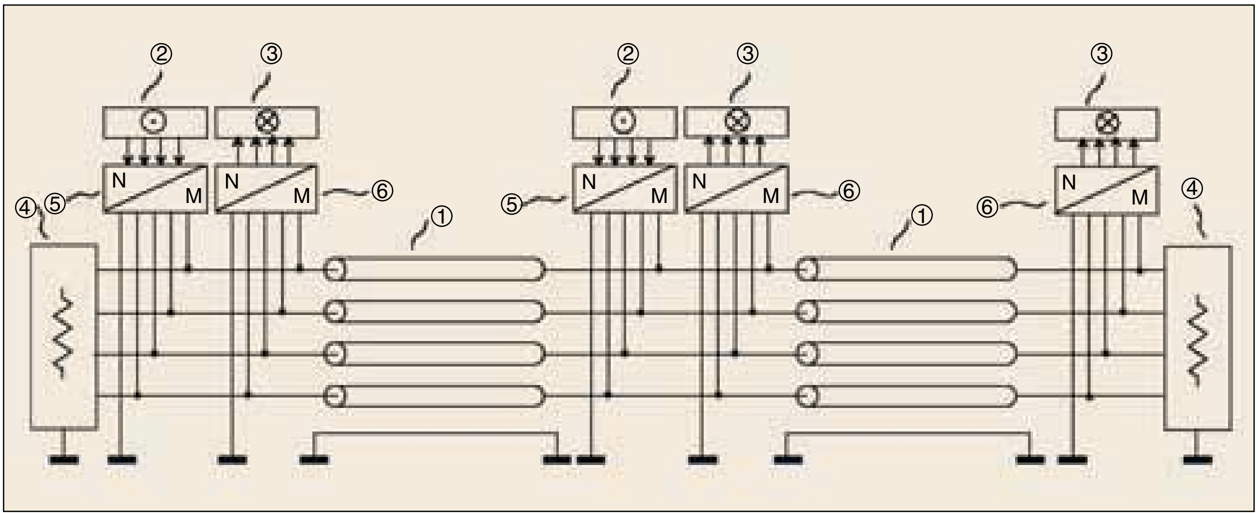


6. Signals measured at the near end with pseudomatched impedances R1 = R2 = R3 = R4 = 147.2 Ω: the VNA signal (top) and the NEXT signal VNB (bottom).

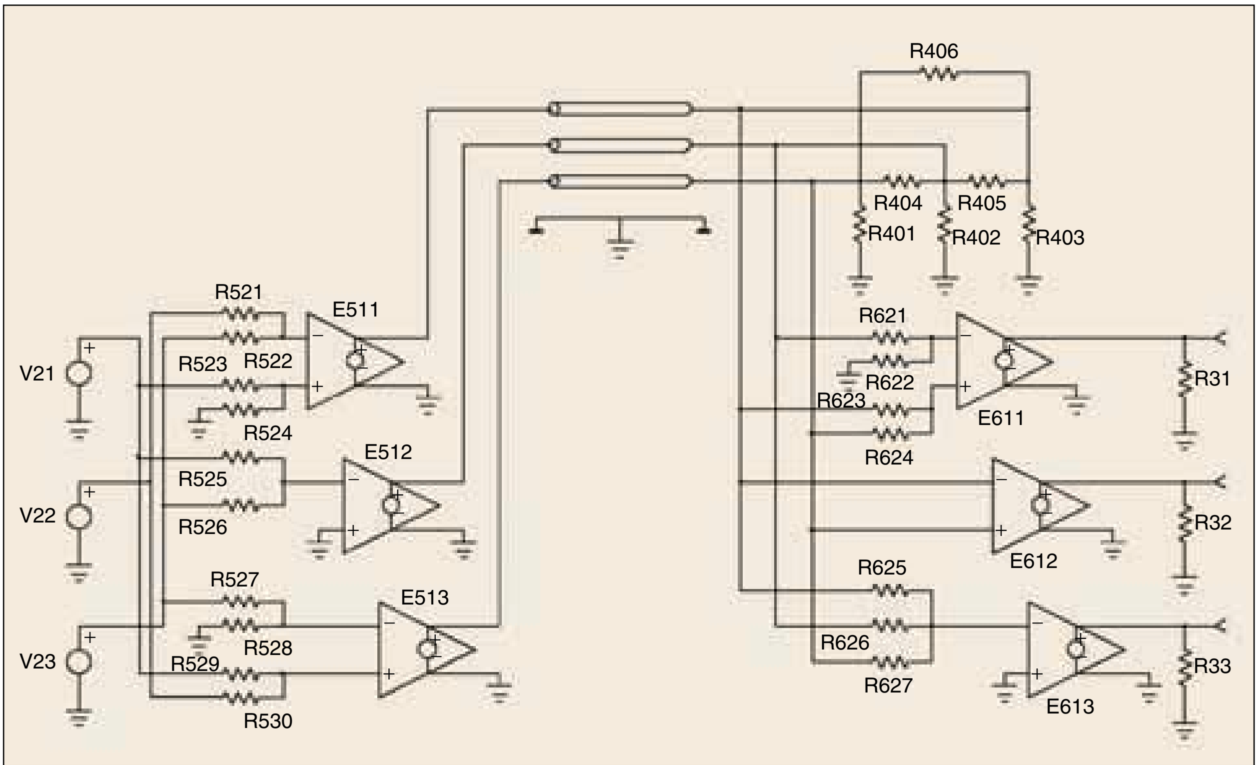


7. Signals measured at the far end with pseudo-matched impedances R1 = R2 = R3 = R4 = 147.2 Ω: the transmitted signal VFA (top) and the FEXT signal VFB (bottom).

rectional transmission. Only one end of the interconnection is connected to a termination circuit made of six resistors R401–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 E511, E512, and E513, and ten resistors R521–R530. This transmitting circuit 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 voltage-controlled voltage sources E611, E612, and E613 and seven resistors R621–R627. It delivers to the resistors R31, R32, and R33 the output signals of the three channels. All part values appear in the input file shown in "SPICE Input File for Simulating the ZXtalk Circuit" for the case in which only V21 is present. This input file also shows how the sub-circuit of the



8. The new method for crosstalk reduction applied to an interconnection with four transmission conductors ① including two sources of signals ②, three destinations ③, transmitting circuits ⑤, receiving circuits ⑥, and two termination circuits ④.



9. Theoretical circuit for the implementation of the new method to an interconnection with three transmission conductors, providing unidirectional transmission.

SPICE Input File for Simulating the ZXtalk Circuit

The subcircuit representing the interconnection is generated automatically by the SpiceLine software [4]. This subcircuit, in fact, accepts the length of the interconnection as parameter, but this vendor-specific feature has been removed from the listing below.

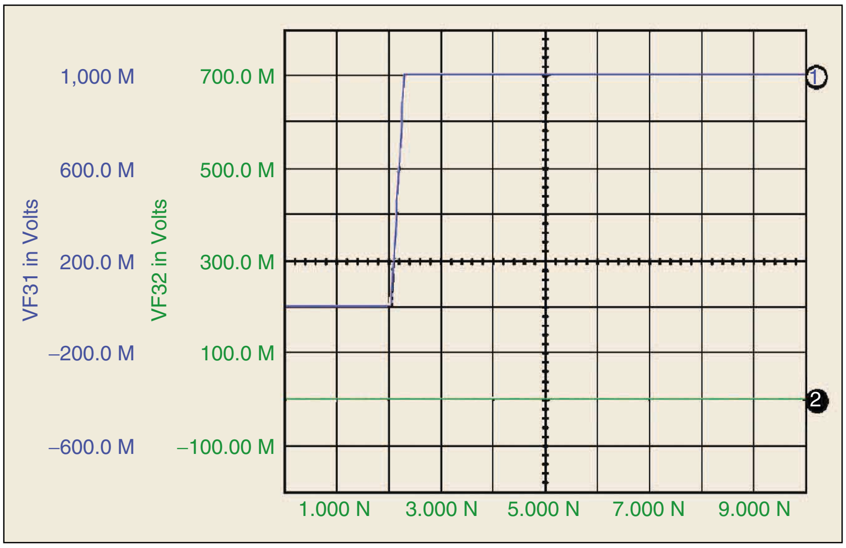
```
Simulation papier7 modifiee

.SUBCKT INTED#0 E1 E2 E3 S1 S2 S3 G
T1 K1 0 M1 0 Z0=59.874 TD=1.7962 N
T2 K2 0 M2 0 Z0=61.840 TD=1.8552 N
T3 K3 0 M3 0 Z0=67.818 TD=2.0345 N
VK1 KV1 K1 0
VM1 MV1 M1 0
VK2 KV2 K2 0
VM2 MV2 M2 0
VK3 KV3 K3 0
VM3 MV3 M3 0
BVK1 KV1 0 V= 7.39910187 E-01*(V(E1)-V(G))
+              + -1.13813380 E+00*(V(E2)-V(G))
+              + 7.39910187 E-01*(V(E3)-V(G))
BVK2 KV2 0 V= -9.27016990 E-01*(V(E1)-V(G))
+              + 1.10618426  E-15*(V(E2)-V(G))
+              + 9.27016990 E-01*(V(E3)-V(G))
BVK3 KV3 0 V= -5.64455475 E-01*(V(E1)-V(G))
+              + -7.36246518 E-01*(V(E2)-V(G))
+              + -5.64455475 E-01*(V(E3)-V(G))
BVM1 MV1 0 V= 7.39910187 E-01*(V(S1)-V(G))
+              + -1.13813380 E+00*(V(S2)-V(G))
+              + 7.39910187 E-01*(V(S3)-V(G))
BVM2 MV2 0 V= -9.27016990 E-01*(V(S1)-V(G))
+              + 1.10618426 E-15*(V(S2)-V(G))
+              + 9.27016990 E-01*(V(S3)-V(G))
BVM3 MV3 0 V= -5.64455475 E-01*(V(S1)-V(G))
+              + -7.36246518 E-01*(V(S2)-V(G))
+              + -5.64455475 E-01*(V(S3)-V(G))
BIE1 E1 G I= 4.78579809 E-01*(I(VK1))
+            + -7.07106781 E-01*(I(VK2))
+            + -5.19782227 E-01*(I(VK3))
BIE2 E2 G I= -7.36154014 E-01*(I(VK1))
+            + 9.99977570 E-16*(I(VK2))
+            + -6.77977045 E-01*(I(VK3))
BIE3 E3 G I= 4.78579809 E-01*(I(VK1))
+            + 7.07106781 E-01*(I(VK2))
+            + -5.19782227 E-01*(I(VK3))
BIS1 S1 G I= 4.78579809 E-01*(I(VM1))
+            + -7.07106781 E-01*(I(VM2))
+            + -5.19782227 E-01*(I(VM3))
BIS2 S2 G I= -7.36154014 E-01*(I(VM1))
+            + 9.99977570 E-16*(I(VM2))
+            + -6.77977045 E-01*(I(VM3))
BIS3 S3 G I= 4.78579809 E-01*(I(VM1))
+            + 7.07106781 E-01*(I(VM2))
+            + -5.19782227 E-01*(I(VM3))
.ENDS

.AC DEC 20 0.1MEGHZ 10000MEGHZ
.TRAN 0.02 N 10 N 0 N 0.02 N
.PRINT AC V(11) VP(11) V(19) VP(19)
.PRINT AC V(20) VP(20)
.PRINT TRAN V(11) V(19) V(20)
V21 9 0 AC 1 PULSE 0 1 0.25 N 0.25 N
R401 13 0 58.7
R402 17 0 69.2
R403 18 0 58.7
R404 13 17 289
R405 17 18 289
R406 13 18 2781
E511 5 0 3 6 1.019
E512 4 0 0 8 1.099
E513 1 0 10 14 0.849
R521 0 6 1889
R522 0 6 2125
R523 9 3 3285
R524 0 3 1438
R525 9 8 2311
R526 0 8 1763
R527 0 14 1772
R528 0 14 2295
R529 9 10 2739
R530 0 10 1575
X5 5 4 1 18 17 13 0 INTED#0
E611 11 0 12 7 1.48
E612 19 0 13 18 0.927
E613 20 0 0 16 1.865
R621 17 7 130 K
R622 0 7 433 K
R623 18 12 200 K
R624 13 12 200 K
R625 18 16 330.4 K
R626 17 16 253.3 K
R627 13 16 330.4 K
R31 11 0 10 K
R32 19 0 10 K
R33 20 0 10 K
.END
```

interconnection containing all interaction between the conductors looks like. With this input file, we obtain the simulated transmission and crosstalk characteristics shown in Figure 10, on which VF31 is a transmitted signal measured across R31, and VF32 is a FEXT signal measured across R32. There is no crosstalk and no echo.

The ZXtalk technique is, in fact, applicable to analog and digital signals. It may also be implemented in such a way that bidirectional transmission is obtained. In such a case, the near-end crosstalk and the far-end crosstalk vanish. For instance, simulations performed with a ZXtalk circuit using the interconnection of Figure 1 and the same rise times shows that bidirectional transmission without any crosstalk and without echo are obtained. It is also interesting to note that, for interconnections having particular properties, the circuits for implementing the ZXtalk technique are simpler. This is, for instance, the case when all propagation constants are equals.

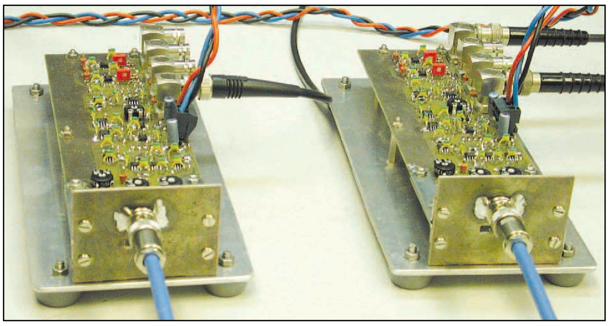


10. Signals measured at the far end of the circuit of Figure 9: the transmitted signal VF31 across R31 (top) and a FEXT signal VF32 across R32 (bottom).

THE ZXTALK TECHNIQUE IS VALIDATED

The ZXtalk crosstalk reduction technique has been implemented in a laboratory environment, for instance, using the experimental set-up shown in Figure 11, which comprises a cable for the interconnection (the thick blue cable) and two identical "cable connection boards," each hosting a transmitting circuit, a receiving circuit, and a termination circuit. The cable connection boards were intended to be connected to standard laboratory instruments for time-domain and frequency-domain measurements. They allowed the comparison of the signals obtained using the ZXtalk technique to the one obtained with conventional line drivers and line receivers and terminations made of pseudo-matched impedances connected to ground.

Such experimental results allow us to confirm that the ZXtalk technique works as expected: it provides a large reduction of crosstalk and echo, improved signal integrity, and increased speed. The achievable improvements depend 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. Our results show that this technique can readily be implemented with PCB traces, backplanes, flex circuits, and cable 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 integrated circuits (ICs) or interface circuits inside ICs performing other functions. Also, the ZXtalk could be implemented with on-chip interconnects.



11. Experimental setup used for the validation of the new method using two "cable connection boards," the cable connected to them (bottom) being the interconnection.

REFERENCES

[1] H. Johnson and M. Graham, *High-Speed Digital Design—A Handbook of Black Magic*. Upper Saddle River, NJ: Prentice Hall PTR, 1993.

[2] C.R. Paul, "Solution of the transmission-line equation under the weak-coupling assumption," *IEEE Trans. Electromagn. Compat.*, vol. 44, pp. 413–423, Aug. 2002.

[3] F. Broydé, E. Clavelier, and C. Hymowitz, "Simulating crosstalk and field to wire coupling with a SPICE simulator," *IEEE Circuits Devices Mag.*, vol. 8, pp. 8–16, Sept. 1992.

[4] *SpiceLine 2.23 with Telecom Line Predictor—User's Guide*, Document 00012107B, Excem Consultants, Maule, France, 2000 [Online]. Available: http://www.excem.fr/eurexcem.com/eurexdef.htm.

[5] J.G. Nickel, D. Trainor, and J.E. Schutt-Ainé, "Frequency-domain-coupled microstrip-line normal-mode parameter extraction from *S*-parameters," *IEEE Trans. Electromagn. Compat.*, vol. 43, pp. 495–503, Nov. 2001.

[6] D.B. Jarvis, "The effect of interconnections on high-speed logic circuits," *IEEE Trans. Electron. Comput.*, vol. 12, pp. 476–487, Oct. 1963.

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