SCATTERING ANALYSIS OF SIGNAL DEGRADATION AND INTERFERENCES ON LONG AND LOSSY INTERCONNECTS

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Abstract
We present a time domain scattering formulation for low-loss nonlinearly loaded multiconductor transmission lines, which is suitable for an efficient and accurate evaluation of crosstalk and field coupling. A simulation of the effects of interferences on a long interconnect is given.

0-7803-1427-1/93/$3.00 ©1993 IEEE

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Owing to the recent tendency towards decreasing rise times of the signal waveforms and the higher frequency carriers, an increasingly larger part of interconnects, at any level of integration, require transmission line models to account for the signal degradation caused by both the propagation in the distributed structure and the effects of the interfering fields.

The field coupling to the line is described by equivalent voltage and current generators located along the line [1, 2], and, in the case of linear terminations, the analysis is effectively carried out in the frequency domain by a line characterization involving a proper set of network parameters. In the widely diffused case of nonlinear terminations, however, the analysis must be performed directly in the time domain and, unfortunately, the transient characteristics of the transmission line and the time evolution of the equivalent generators of the electromagnetic interference are usually particularly difficult to compute and handle.

A scattering parameter formulation of the transient problem was proved to be an efficient and fast method for the time-domain analysis of low-loss nonlinearly loaded multiconductor transmission lines (MTLs) [3]. In this paper, we extend the formulation to the study of MTLs subject to EMIs, and we show that the effects of the external field can be handled by simply adding to the line ends proper source terms, which do not affect the structure of the solution algorithm, and thereby preserve all the advantages of the matched scattering formulation. The example discussed in this paper aims at showing that our formulation represents a very efficient and accurate method for studying the effects of interferences (due both to crosstalk and field coupling) on "long" interconnects that can be modeled as transmission lines.

SCATTERING TRANSIENT ANALYSIS OF MTLs

Our formulation of the scattering transient analysis of MTLs is based on the structure illustrated in Fig. 1, where the voltage sources at the line ends ($e_{pq}, p, q = 1, 2$) account for the system excitation, and the functions $\mathbf{g}_p$ describe the (possibly nonlinear) loads. The N-conductor line is characterized by its impedance matrix $\boldsymbol{Z}^{\Box}{}_0(\omega)$ and the mode propagation constants $\{K_i(\omega)\}$, $i = 1, ..., N$.

The scattering parameter formulation expresses the unknown voltages and currents $\boldsymbol{v}_1$, $\boldsymbol{v}_2$, $\boldsymbol{i}_1$, $\boldsymbol{i}_2$ at both line ends, for each of the N conductors, in terms of the voltage wave vectors $\boldsymbol{a}_p$ and $\boldsymbol{b}_p$, *i.e.*,

$$\boldsymbol{v}_p = \boldsymbol{a}_p + \boldsymbol{b}_p \,, \qquad \boldsymbol{i}_p = \boldsymbol{y}^{\Box}{}_r * (\boldsymbol{a}_p - \boldsymbol{b}_p), \qquad p = 1,\ 2, \tag{1}$$

where $\boldsymbol{y}^{\Box}{}_r(t)$ is the transient expression of the reference admittance, and $*$ denotes the convolution operator. The transient equations for the voltage waves are obtained from the line scattering equations and from the load characteristics. In the special case of a matched characterization (*i.e.*, $\boldsymbol{y}^{\Box}{}_r = \boldsymbol{y}^{\Box}{}_o \triangleq \mathcal{F}^{-1}\{(\boldsymbol{Z}^{\Box}{}_o(\omega))^{-1}\}$) of the line, the scattering equations simplify and the transient equations take the form [3]

$$\begin{aligned} \boldsymbol{b}_1 &= \boldsymbol{s}^{\Box}{}_{21} * \boldsymbol{a}_2 \,, \\ \boldsymbol{b}_2 &= \boldsymbol{s}^{\Box}{}_{21} * \boldsymbol{a}_1 \,, \end{aligned} \tag{2}$$

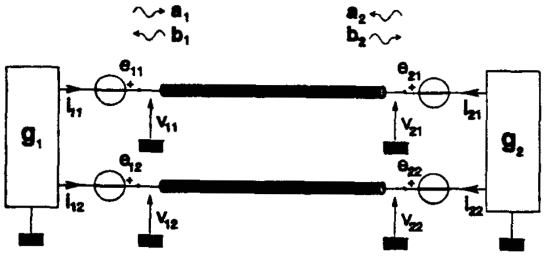

Figure 1: *Nonlinearly loaded transmission line: the relevant quantities for our formulation are indicated here and explained in the text. Although the scheme is limited to $N = 2$, the formulation is valid for any number of conductors.*

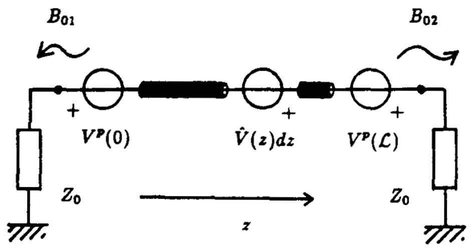

Figure 2: *The effect of radiated EMI on a transmission line translates into distributed and terminal lumped voltage sources on the line. The symbols are explained in the text. This conceptual scheme is valid also for MTLs.*

$$- \boldsymbol{y}^{\square}{}_{o} * (\boldsymbol{a}_p - \boldsymbol{b}_p) = \boldsymbol{g}_p(\boldsymbol{a}_p + \boldsymbol{b}_p - \boldsymbol{e}_p), \quad p = 1,\ 2, \tag{3}$$

where $\boldsymbol{s}^{\square}{}_{21}$ is the $N \times N$ transient scattering matrix of the matched line and $\boldsymbol{e}_p$ is the vector of the voltage sources connected to the p-th line end.

EFFECT OF THE EXTERNAL INTERFERENCES

To describe the radiated EMI effects, we employ a set of distribuited voltage generators along each conductor of the line and two lumped voltage generators at the ends, as illustrated in Fig. 2, [2]. The contributions of these generators are included in the above formulation in a straightforward manner, *i.e.*, by adding to the left hand side of (2) a source vector $\boldsymbol{b}_{0p}$, whose elements are the voltage waves caused by the interference generators, when both line ends are matched. This result is derived here in the frequency domain, since it involves only the line which is the linear part of the system. For simplicity, the derivation is limited to a single conductor transmission line: however, the extension to MTLs is straightforward. With reference to Fig. 2, and for a plane wave external interference, the distributed voltage source can be expressed as $\hat{V}(z) = \hat{E}(\omega)\exp\{-j\beta z\}$, where $\hat{E}$ depend on the interference spectrum, and β is the wavenumber along the transmission line direction. Complete expressions for $\hat{E}(\omega)$ and for the terminal lumped voltage sources $V^p(0)$, $V^p(\mathcal{L})$ are not given here for lack of space, but can be found in [2]. The frequency domain expressions of the voltage waves that account for the EMI generators are

$$\begin{aligned} B_{01} &= \frac{V^p(0)}{2} - S_{21}\frac{V^p(\mathcal{L})}{2} - \frac{\hat{E}}{2}(1 - e^{-(\gamma+j\beta)\mathcal{L}})/(\gamma + j\beta) \\ B_{02} &= -S_{21}\frac{V^p(0)}{2} + \frac{V^p(\mathcal{L})}{2} + \frac{\hat{E}}{2}(e^{-j\beta\mathcal{L}} - e^{-\gamma\mathcal{L}})/(\gamma - j\beta)\ , \end{aligned} \tag{4}$$

where the contribution of the distributed generators is integrated along the line, $S_{21} = \exp\{-\gamma\mathcal{L}\}$ is the transmission scattering parameter of the matched line, and γ is the line propagation function. It is convenient, after the inclusion of the voltage waves B_{0p} in the scattering equations, rearrange them in the form

$$\begin{aligned} B_{01} &= S_{21}(A_2 + \bar{A}_{02}) + \bar{B}_{01} \\ B_{02} &= S_{21}(A_1 + \bar{A}_{01}) + \bar{B}_{02}\ , \end{aligned} \tag{5}$$

where

$$\begin{aligned} \bar{A}_{01} &= -\tfrac{1}{2}\Big(V^p(0) + \hat{E}/(\gamma - j\beta)\Big)\ , & \bar{B}_{01} &= \tfrac{1}{2}\Big(V^p(0) - \hat{E}/(\gamma + j\beta)\Big)\ , \\ \bar{A}_{02} &= \tfrac{1}{2}\Big(- V^p(\mathcal{L}) + \hat{E}e^{-j\beta\mathcal{L}}/(\gamma + j\beta)\Big)\ , & \bar{B}_{02} &= \tfrac{1}{2}\Big(V^p(\mathcal{L}) + \hat{E}e^{-j\beta\mathcal{L}}/(\gamma - j\beta)\Big)\ . \end{aligned} \tag{6}$$

With this new set of source terms, all propagation effects included in the expressions of the voltage wave equivalents are extracted and accounted for by the transfer function S_{21} appearing in (5). The advantage of this new form of the scattering equations is that the terms $\tilde{A}_{0p}$ and $\tilde{B}_{0p}$ are simple transformations of the impinging signal and can be easily inverse transformed in the time domain. Additionally, multiple reflections are absent in the time behavior of the terms b_{0p}: this nice property is a consequence of the matched scattering characterization of the line, adopted in this formulation. In fact, if $\boldsymbol{y}^{\square}{}_{r} \neq \boldsymbol{y}^{\square}{}_{o}$ were used, not only the scattering parameter matrices in (2) whould have been more complex, but also the B_{0p} terms whould have had a different form from (4), and contain multiple reflections. For this reason, we decided not to adopt the formulation in terms of Thevenin equivalents of the EMI generators (see [2]) which, however, is valid and convenient when the entire system response is sought in the frequency domain.

Results

This approach to the EMI evaluation of MTLs was extensively validated by comparison with the result of a standard frequency domain analysis applied to structures with linear loads [2]. An example of the EMI evaluation for a realistic structure is given in Fig. 3, where the end voltage v_{22} caused by a double exponential pulse impinging on a two-conductor symmetric line driven and loaded by inverter circuits is shown. The transmission line is 0.5 m-long, and is composed of two parallel circular conductors with a diameter of 0.05 mm, separated by 0.2 mm, and at a distance of 0.5 mm above the ground plane; line losses include both the conductor DC resistance and the skin effect.

The inverters are assumed to be in the "high" logical state and are modeled as in [4]. The results of Fig. 3 show the time shape and levels of the noise induced into the electronic system: the abrupt variations of the voltage curve can be justified in terms of the reflections at the mismatched ends of the line. The efficiency of our formulation, which is implemented in an interpreted programming language, is proved by the fact that the results are produced in a matter of minutes on a personal computer.

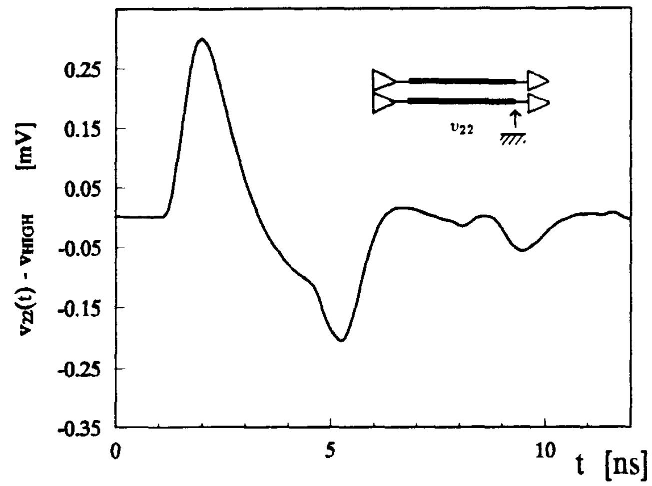

Figure 3: *End voltage v_{22} due to a plane wave interference on the transmission line loaded by inverters.*

References

[1] C. R. Paul, "Frequency response of multiconductor transmission lines illuminated by an electromagnetic field," *IEEE Trans. on EMC,* vol. EMC-18, pp. 183-190, Nov. 1976.

[2] F. Canavero, V. Daniele and R. Graglia, "Electromagnetic pulse interaction with multiconductor transmission lines," *Electromagnetics*, vol. 8, pp.293-310, Aug. 1988.

[3] I. Maio, S. Pignari e F. G. Canavero, "Efficient transient analysis of nonlinearly loaded low-loss interconnects," *Int. J. Analog Integrated Circuits and Signal Processing,* in press, June 1993.

[4] J. E. Schutt-Aine and R. Mittra, "Scattering parameters transient analysis of transmission lines loaded with nonlinear terminations," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-36, pp. 529-536, Mar. 1988.