Correspondence

Comments on "Capacitance Calculation for Cable Harnesses Using the Method of Moments"

B. N. Das and S. B. Chakrabarty

The interesting paper[1] on the evaluation of capacitance and EMI is very useful to the readers interested in theory and also to practicing engineers. It has been stated in the first paragraph of Section III B in the above paper[1], that no analytical expression for capacitance of a single isolated wire above a ground plane exits. It is shown in the present correspondence that this is not a proper statement. In this connection, attention is drawn to [1], which is in the same issue of EMC in which the above paper[1] has appeared.

For the particular case of insulated wire above a ground plane, the image of the insulated wire resulting from the presence of the ground plane together with the insulated wire itself leads to the configuration shown in [1, Fig. 1(a)]. Using the formulation presented in [1], it is found that (10) of the same article can be used for the evaluation of capacitance per unit length for the line parameters given in Figs. 5 and 6 of the above paper[1] and relative dielectric constant of 4. The numerical results on capacitance agree with those reported in the above paper in [1]. For cable harness of two insulated wires above a ground plane (configuration shown in Fig. 7 of the above paper [1]), the data on crosstalk evaluated using the formulas of [2], the circuit parameter values found from formulas of [1] have exact agreement with those reported in Fig. 8 of the above paper [1]. It is worthwhile to mention that the expression (10) of [1] has been derived using the analytical formulation based on conformal transformation.

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Comments on "A SPICE Model for Multiconductor Transmission Lines Excited by an Incident Electromagnetic Field"[1]

Frédéric Broydé, Evelyne Clavelier, and Lothar Hoeft

I. Introduction

In a recent paper,[1] C. Paul discusses the concept of using the SPICE circuit analysis code for simulating field-to-wire coupling. He may not be aware that similar work has already been undertaken [1], [2], [4]. Specifically, [1] appears to be the first published implementation of this idea. It described simulations of a near-field coupling problem, plus a comparison with experiments. In these simulations, the amplitude and delay of each segment of the incoming wave was modeled with an auxiliary circuit.

Since the time of the original publication many such simulations have been implemented by a team at Excem, both in the frequency and time domains. We have commonly used this approach to solve field-to-wire and crosstalk problems involving shielded cables. Up to now this approach has been limited to type 1 (transfer impedance) and type 2 (transfer admittance) couplings only. In addition, this approach requires that the per-unit-length transfer impedance of the shielded cable be synthesized using the lumped elements available in the SPICE code.

The present comments on Paul's mostly theoretical paper emphasize some practical aspects of the simulation of EMC problems using a SPICE simulation program when multiconductor transmission lines (MTL) models are implemented.

II. SPICE 2 versus SPICE 3 Simulation Programs

The equivalent circuit itself is a very important aspect of crosstalk and field-to-wire simulation using SPICE. As is well known to SPICE users, the lossless transmission line model in SPICE 2 simulation programs (e.g., the ubiquitous Berkeley SPICE 2G.6) suffered from severe limitations because of the large number of breakpoints generated at the beginning of transient simulations. In practice, this resulted in computational problems for simulations that were long with respect to the lines propagation time. The simulation often did not reach completion, and sometimes did not even start.

Today, most SPICE simulation codes use variations of SPICE 3 (e.g., the ICAP/4 package of Intusoft that we use, based on Berkeley SPICE 3F.2). In this case, the simulation code offers a completely new two-conductor lossless transmission line implementation (the so-called T-element), and a new two-conductor lossy transmission line model (the O-element) device. Though Paul's and our approaches are limited to lossless multiconductor transmission lines, it is often a good idea to formulate the equivalent circuit of the multiconductor

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[1] C. R. Paul, *IEEE Trans. Electromag. Compat.*, vol. 36, pp. 342–354, Nov. 1994.

$$\mathbf{L}(1\,\mathrm{MHz}) = \begin{pmatrix} 186.3 & 165.1 & 165.8 & 68.2 & 26.7 & 14.5 & 9.8 & 6.1 \\ 165.1 & 390.1 & 169.7 & 66.0 & 25.4 & 13.8 & 9.1 & 9.8 \\ 165.8 & 169.7 & 390.7 & 70.4 & 27.2 & 15.1 & 13.8 & 14.5 \\ 68.2 & 66.0 & 70.4 & 377.0 & 101.1 & 27.2 & 25.4 & 26.7 \\ 26.7 & 25.4 & 27.2 & 101.1 & 377.0 & 70.4 & 66.0 & 68.2 \\ 14.5 & 13.8 & 15.1 & 27.2 & 70.4 & 390.7 & 169.7 & 165.8 \\ 9.8 & 9.1 & 13.8 & 25.4 & 66.0 & 169.7 & 390.1 & 165.1 \\ 6.1 & 9.8 & 14.5 & 26.7 & 68.2 & 165.8 & 165.1 & 186.3 \end{pmatrix} \cdot \frac{\mathrm{nH}}{\mathrm{m}} \tag{1}$$

$$\mathbf{C} = \begin{pmatrix} 331.07 & -120.12 & -120.12 & -57.88 & -1.12 & 0 & 0 & -1.08 \\ -120.12 & 128.40 & -8.28 & 0 & 0 & 0 & 0 & 0 \\ -120.12 & -8.28 & -128.40 & 0 & 0 & 0 & 0 & 0 \\ -57.88 & 0 & 0 & 95.80 & -31.00 & 0 & 0 & -1.12 \\ -1.12 & 0 & 0 & -31.00 & 95.80 & 0 & 0 & -57.88 \\ 0 & 0 & 0 & 0 & 0 & 128.40 & -8.28 & -120.12 \\ 0 & 0 & 0 & 0 & 0 & -8.28 & 128.40 & -120.12 \\ -1.08 & 0 & 0 & -1.12 & -57.88 & -120.12 & -120.12 & 331.07 \end{pmatrix} \cdot \frac{\mathrm{pF}}{\mathrm{m}} \tag{2}$$

transmission line using the lossy transmission line model with a zero per-unit-length resistance and a zero per-unit-length conductance. This avoids many simulation problems and allows the model to be more easily optimized, thanks to the variety of parameters available for the O-elements (the TRYTOCOMPACT option, for instance, is very useful), and its superior implementation. In the place of the SPICE schematic of Fig. 6,[1] it would, therefore, be a good idea to eventually use O-elements.

III. Two Important Problems with Lossless SPICE Models

A lossless model of a cable is generally expected to overestimate the induced voltages and currents. This is usually acceptable to the user, because it leads to an additional margin in the design process. There are nevertheless two special cases where a lossless model usually gives rise to an overly optimistic assessment of an unwanted signal. The first case is related to the real part of the reference conductor impedance; the second case occurs if conductors of the MTL are terminated at both ends to the reference conductor with low impedances. For clarity, these two cases will be discussed using a crosstalk problem.

Fig. 1 shows a nine-conductor MTL with its terminations, the numbering index of the conductors appearing between quotes. Conductor 0 is the reference (ground plane) conductor and is everywhere the node 0 (ground) of the SPICE netlist. Conductor 1 is the shield of the shielded pair containing conductors 2 and 3, while conductor 8 is the shield of the shielded pair containing conductors 6 and 7. These two shielded pairs are identical. The other conductors are also identical to each other. The computed per-unit-length inductance matrix $\mathbf{L}$ of the transmission line at 1 MHz is shown in (1) at the top of the page.

The computed frequency independent per-unit-length capacitance matrix $\mathbf{C}$ of the transmission line is shown in (2) at the top of the page.

The per-unit-length inductance matrix of the transmission line is only mildly frequency dependent. The (more severely frequency dependent) per-unit-length impedance matrix of the transmission line is:

$$\mathbf{Z} = \mathbf{R} + j\omega\mathbf{L} \tag{3}$$

where $\mathbf{R}$ and $\mathbf{L}$ are real matrices. The imaginary part of the common impedances is obviously described by the off-diagonal terms of $\mathbf{L}$. At low enough frequency, these terms become negligible, whereas the $\mathbf{Z}$ matrix tends toward the dc per-unit-length resistance matrix

$$\mathbf{Z}(0\,\mathrm{Hz}) = \mathbf{R}(0\,\mathrm{Hz}) = \begin{pmatrix} 49 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 \\ 0.2 & 33 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 \\ 0.2 & 0.2 & 33 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 \\ 0.2 & 0.2 & 0.2 & 11 & 0.2 & 0.2 & 0.2 & 0.2 \\ 0.2 & 0.2 & 0.2 & 0.2 & 11 & 0.2 & 0.2 & 0.2 \\ 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 33 & 0.2 & 0.2 \\ 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 33 & 0.2 \\ 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 0.2 & 49 \end{pmatrix} \cdot \frac{\mathrm{m}\Omega}{\mathrm{m}} \tag{4}$$

where the value 0.2 mΩ/m is clearly the reference conductor per-unit length dc resistance. The lossless transmission line model neglects the effect of the real part of the impedance matrix and therefore, grossly underestimates the low frequency (typically below 10 kHz to 5 MHz depending on the MTL parameters) impedance matrix elements responsible for the crosstalk.

The first problem mentioned above can now be described as the consequence of neglecting the off-diagonal terms of the $\mathbf{R}$ matrix. This can be cured by adding a current controlled voltage source (an H-element according to the SPICE syntax), in series with each relevant conductor in the MTL SPICE model. It can be done at one or both ends of the conductor, in such a way that the total transimpedance (or transresistance in this case, e.g., the ratio of output voltage of the H-element to its input current) is equal to the off-diagonal term of $\mathbf{R}(0\ \mathrm{Hz})$. Let us call this "cosmetic measure no. 1."

The second problem mentioned above occurs when conductors other than the reference conductor are grounded or connected to ground with a low-impedance termination. For example, a common practice is to have many grounded-at-both-ends conductors on flat unshielded ribbon cables, for improved EMC characteristics. In the case of an MTL including one or several shields (like on Fig. 1), the shields are also normally grounded at both ends. Each conductor with low impedance to ground at both ends provides some shielding.

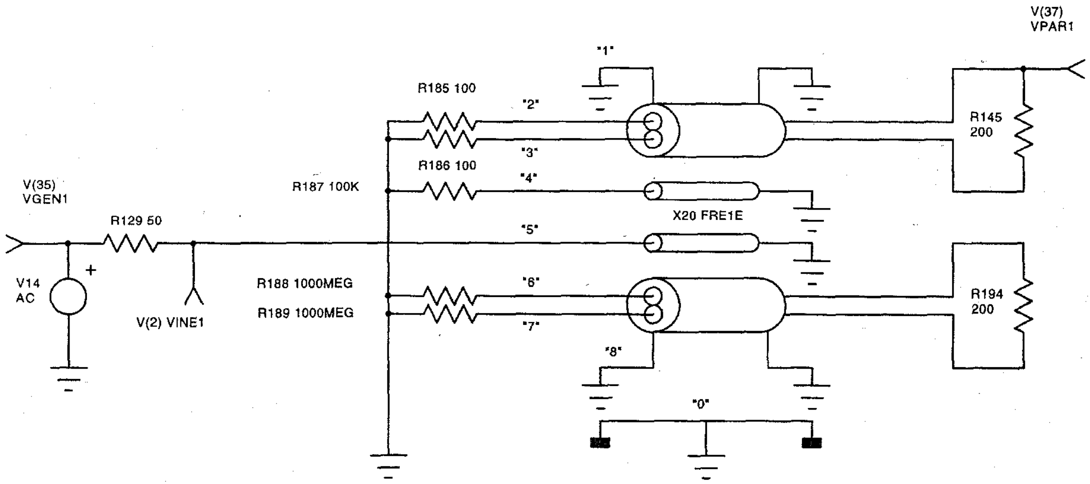

Fig. 1. A 9-conductor MTL and its terminations.

Neglecting losses overestimates their shielding effectiveness at low frequencies. This is because in the lossless model, their per-unit-length impedance decreases to zero at zero frequency, instead of being limited by their per-unit-length dc resistance. This per-unit-length resistance is equal to the corresponding diagonal term of $\mathbf{R}$(0 Hz) minus the per-unit-length dc resistance of the reference conductor. We therefore advocate the practice of adding in the MTL SPICE model, an appropriate lumped resistor at one or both ends of conductors potentially subject to this problem. Let us call this "cosmetic measure no. 2."

The two cosmetic measures proposed are equivalent to artificially adding the whole or part of the dc resistance matrix of the MTL to the lossless model. This is illustrated by the computation shown in Fig. 2, which plots the crosstalk transfer function for 30 m of the MTL characterized by (1) (the $\mathbf{L}$ matrix is assumed frequency independent) (2), and (4). This small signal (.AC) analysis was performed with the setup of Fig. 1, the crosstalk signal being measured at the far end of conductor 2. Curve 1 is the result for a lossless model of the MTL implemented using O-elements; curve 2 is the result with the diagonal term resistance added in series to the shield conductor (cosmetic measure no. 2); curve 3 is the result with an additional current controlled voltage source added in series with conductors 2 and 3 (cosmetic measures nos. 1 and 2). The results obtained with the models of curve 2 and 3 are clearly different up to 8 kHz and merge above that frequency, while the lossless model only starts to give reasonable results above 1 MHz, and agrees with the others above 3.5 MHz. The other terms of the dc resistance matrix were not included because they were not expected to be relevant to our problem and would have slowed down a later transient analysis. While performing similar calculations, we found that adding an H-element on conductor 1 does not change the coupling significantly.

Figs. 3 and 4 show voltages obtained with the same setup, except that the AC source is replaced by a circuit equivalent to the IEC combination wave generator [3], and the 50 Ω resistor by a 40 Ω one, as appropriate. Curve 1 of Fig. 3 was obtained with the lossless model. Curve 1 of Fig. 4 was obtained with cosmetic measure no. 2, while curve 2 of Fig. 4 was obtained with cosmetic measures numbers 1 and 2. The energy content of these three curves are, respectively, 2.4 $V^2\,\mu s$, 148 $V^2\,\mu s$, and 157 $V^2\,\mu s$. The improvements of the lossless model are obviously important for this time-domain problem because the coupled energy and maximum voltage are quite different. Note that the visible ringing at 1.5 MHz probably suffers from an

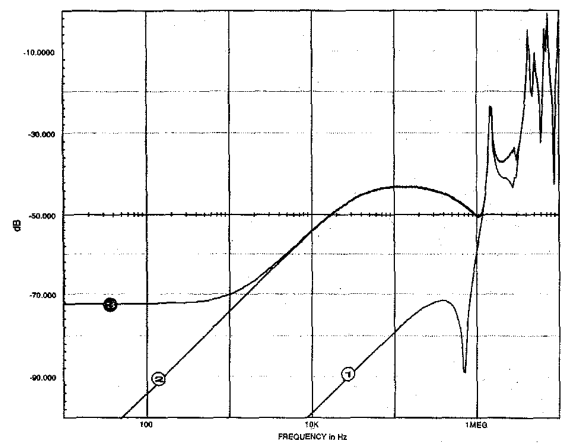

Fig. 2. Crosstalk transfer function versus frequency (in Hz) for the MTL of Fig. 1. Curve 1: Lossless MTL model. Curve 2: Same with cosmetic measures no. 2. Curve 3: Lossless MTL with cosmetic measures nos. 1 and 2.

unrealistically small damping factor. However, such artifacts, which are related to our starting from a lossless model, are of the kind EMC engineers can live with.

Another simulation, performed on a second nine-conductor MTL of 30 meters (this one has only one shielded pair and five identical conductors) gives the frequency domain results of Fig. 5. The pair (conductors 2 and 3) is terminated as on Fig. 1, the shield (conductor 1) is grounded at both ends, and one of the other conductors is excited by the AC source, similar to conductor 5 of Fig. 1. The four remaining conductors are grounded at the far end and left open-circuited at the near end, similar to conductor 4 of Fig. 1. The shield and reference conductor per-unit-length resistances are the same as with the first MTL. The legend of Fig. 5 is the same as that of Fig. 3. The differences between Figs. 2 and 5 are explained by the "shielding effectiveness" of conductor 8 of Fig. 1.

Note also that the "cosmetic measures" proposed above are only applicable when the added impedance and transimpedance remain much smaller than the corresponding terms of the characteristic impedance matrix of the MTL (assumed lossless). If this was not the case, they would generate a parasitic impedance mismatch in the problem and hurt the accuracy of high frequency coupling calculation.

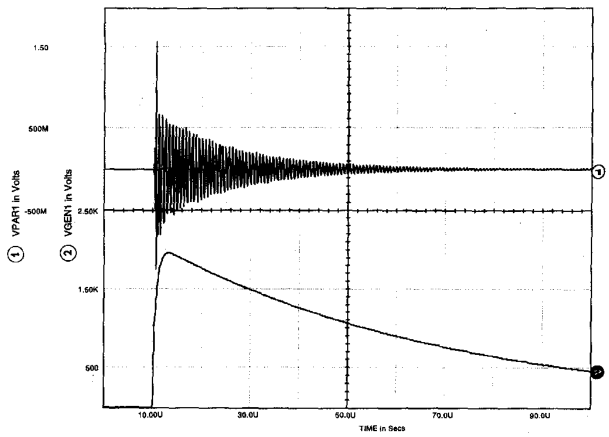

Fig. 3. Transient coupled voltage, lossless MTL model. Curve 1: Far-end signal on conductor 2. Curve 2: Voltage across the combination wave generator.

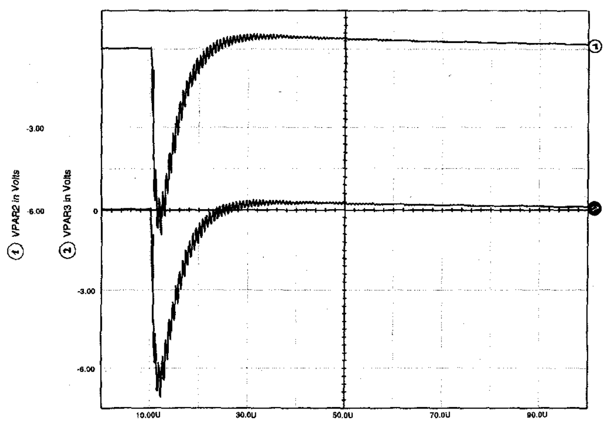

Fig. 4. Transient coupled voltage, modified lossless MTL models. Curve 1: Far-end signal on conductor 2 with cosmetic measures no. 2. Curve 2: Far-end signal on conductor 2 with cosmetic measures nos. 1 and 2.

Also, the shields in the example above are characterized by a per-unit-length transfer impedance with frequency independent resistive and inductive terms only: as is well known, this is only accurately valid for some types of screen, e.g., nonoptimized single-braided shields.

IV. Creation of MTL SPICE Models

Let us also mention that the creation of our lossless MTL models is not done manually. In practice, we enter the measured or computed **L** and **C** matrices in a Mathcad Plus 5.0 spreadsheet which computes the mode propagation velocities c_i, mode characteristic impedances z_i, and the usual $\mathbf{T}$ and $\mathbf{S}^{-1}$ matrices of the MTL [1], [2]. The resulting files are then treated by our in-house SpiceLine 1.2 software, which automatically generates the subcircuit equivalent to the MTL in an ICAP/4 compatible library. For instance, the model of the nine-conductor transmission line discussed above, implementing O-elements, is a 290 lines long subcircuit, blank lines not included!

Let us also mention that our practice of defining independently a matrix $\mathbf{T}$ and a matrix $\mathbf{S}$ for the modal transform

$$\begin{cases} \underline{\mathbf{V}} = \mathbf{S}^{-1}\mathbf{V} \\ \mathbf{I} = \mathbf{T}\underline{\mathbf{I}} \end{cases} \tag{5}$$

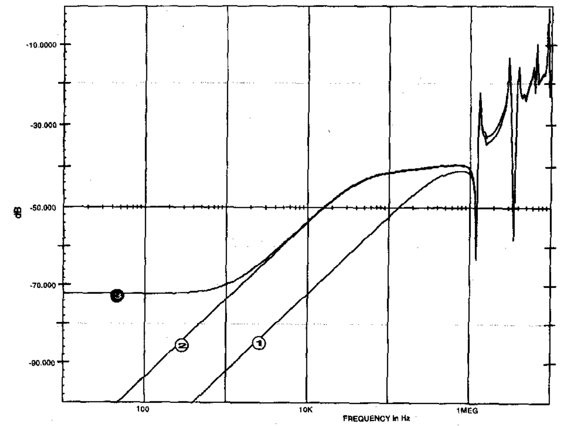

Fig. 5. Crosstalk insertion loss versus frequency (in Hz) for the second MTL. Curve 1: Lossless MTL model. Curve 2: Same with cosmetic measures no. 2. Curve 3: Lossless MTL with cosmetic measures nos. 1 and 2.

implemented in the MTL subcircuit, where the notations of [1] and [2] are used, can be used for arbitrarily selecting the mode characteristic impedance. We usually adopt the additional condition

$$\mathbf{S} = k\mathbf{C}^{-1}\mathbf{T} \tag{6}$$

where k is an arbitrary positive constant. This leads to the values

$$z_i = \frac{1}{kc_i} \tag{7}$$

for the mode characteristic impedance for mode i. Unlike what was said in [1] and [2], we usually adopt $k = 10^{-10} F/m$ instead of $k = 1 F/m$. This keeps the modal impedance within the physical bounds of a characteristic impedance, say between 10–1000 Ω, values for which SPICE may be expected to work properly. We do not know what Prof. Paul's rather complex normalization [cf., the above paper,[1] (13)–(19a)] procedure does in this respect.

V. Concluding Remarks

Previous reports shows that circuit analysis code simulations using SPICE-type softwares, combined with models of electronic circuits [1] or surge arrestors [4], can successfully analyze the transients that result from electromagnetic pulses, e.g., NEMP.

However, running such simulations and deriving useful information from them requires some care. Our team at Excem will be happy to support those who want to discuss any problems related to that kind of simulation project.

Author's Reply by C. R. Paul

The authors of the above comments state that they mentioned the idea of extending an earlier SPICE model for multiconductor transmission lines to the case of incident electromagnetic fields in two papers [1], [2] prior to my publication.[1] Their papers [1] and [2] are essentially identical, and so my comments are addressed to either. The authors state that [1] "appears to be the first published implementation of this idea." While this may be the case, the authors seem to imply that such mention precludes any future publications on the subject

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by others. It is not the mere mention of an idea that is important; the full and general implementation of the method as was provided in the above paper[1] provides the utility of the method. Reference [1] addresses a limited implementation, while my paper[1] gives a general implementation that does not suffer from the restrictions of their model. Specifically, their model in [1] requires the line to be broken into segments over which the phase and amplitude of the incident field is assumed constant. Apparently, this was done to simplify the required integrations of the incident field. But this represents an approximation in the model and is unnecessary. The general result for arbitrary angles of incident and polarization are easily derived and given in my paper.[1] The experiment they chose, broadside excitation of the line, happens to fit this restriction to some reasonable degree, but there are many more cases of incidence and polarization of the incident field where this restrictive assumption is either not valid or requires the line be broken into a large number of segments.

For example, consider an incident wave traveling along a line that is not electrically short. The phase of the incident field can change drastically along the line, requiring possibly many segments to represent this variation. The large experimental prediction errors of their model of some ± 20 dB are attributed by them in [1] and [2] to numerous possibilities in the measurement system, but may well be attributable in part to this approximation of requiring the line to be broken into electrically short segments. In fact, in [2] they state that "our assumptions concerning the amplitude variations of the incident field are not very accurate in many cases." In addition, the model of [1] uses a different structure than that of the above paper.[1] The model of [1] places current sources at the ends of the lines to model the effects of the incident field, whereas the model of the above paper[1] utilizes delay lines to represent this effect in the modal lines.

They also imply that [4] preceeded my publication.[1] Although [1] appeared in print in November of 1994, whereas [4] appeared in June of 1994, a close check of the above paper[1] shows that it was received by the editor in September of 1993, which is well before publication of [4]. Also, [4] derives the model specifically for an EMP waveform, whereas the above paper[1] provides for a general waveform.

Broydé *et al.* devote the remainder of their comments to pointing out that all such SPICE models are restricted to lossless lines, a fact that has been clearly stated.[1] They point out two well-known instances where lossless lines cannot be assumed: 1) common-impedance coupling wherein the circuits share a common return, thereby generating a coupled response from one circuit to the other at lower frequencies due to the voltage drop across the common impedance of the return conductor, and 2) lines that are terminated at their endpoints in very low impedances (low compared to the circuit characteristic impedances), such as a shield that is grounded at both ends. The first case, common-impedance coupling, is a well-known phenomenon throughout the EMC community [5], [6]. Whether it invalidates the lossless line assumption depends on how low a frequency one is interested in.

The second case, low impedance terminations, is also well known throughout the EMC community and is prevalent in the case of shielded lines where the shield endpoints are "grounded" to the reference conductor via short circuits [5], [6]. The shield inherently tends to remove capacitive coupling to the interior wires, whereas grounding the shield at both ends allows the return current to flow back along the shield rather than the return conductor hence tending to reduce inductive coupling [5], [6]. The frequency above which the shield reduces inductive coupling depends directly on the total impedance of the shield ground plane loop. From this, it should be abundantly clear that unless the terminations at the endpoints of a circuit are much larger than the "loop impedance" of the circuit consisting of the shield resistance and self inductance, then one cannot neglect the shield losses since they will dominate the loop impedance. To neglect conductor loss in such cases will clearly lead to erroneous results [5], [6].

Broydé *et al.* try to "patch this up" by putting half the total shield resistance at either end of the line in the SPICE model (which they refer to as a "cosmetic measure"). This is clearly an approximation to the fact that such losses are truly distributed. They recommend a similar "cosmetic measure" to try to augment the lossless SPICE model to predict common-impedance coupling. Showing computed results, as in their Fig. 2, which illustrates that losses cannot be neglected in either of these cases, does not prove that their method of trying to patch up the lossless SPICE model by lumping the total conductor resistances at either end of the lossless model is valid. Experimental results or some other exact calculation, such as in [7], would be required to demonstrate the adequacy of their "cosmetic measures."

The restriction of the SPICE models to lossless lines is necessitated in order that the diagonalization matrices be frequency independent. This restriction has been clearly stated.[1] Hence, the authors should not imply that this is a flaw in the lossless line SPICE model; one must adhere to the assumptions of any model. If the lossless line assumption is not valid, as in the above two obvious cases, one should not try to use it there.

And, finally, Broydé *et al.* state that they do not understand the "rather complex normalization procedure" of the above paper.[1] That procedure is not complex and simply produces modal characteristic impedances that are not extreme in value in order to minimize numerical errors.

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