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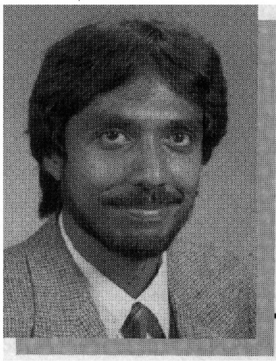
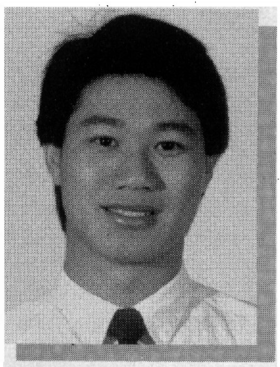
Simulation & Modeling





Resvé Saleh and Andrew Yang, Editors

Simulating Crosstalk and Field to Wire Coupling with a Spice Simulator

There is much interest in using the SPICE simulator for various calculations related to the electro-magnetic compatibility (EMC) of interconnections. EMC is a discipline fully recognized by the IEEE, with its own IEEE society, IEEE transactions publication, and its own symposia around the world.

EMC problems arise in virtually every phase of the electronic design process, from integrated circuit design to system design. This article focuses on the simulation of the EMC of interconnections. This topic can be further divided into three separate areas: crosstalk, emission (also called wire to field coupling) and cable pickup (also called field-to-wire coupling). We will limit the discussion to crosstalk and field to wire coupling.

A SPICE model [1] dedicated to the simulation of crosstalk in Multiconductor Transmission Lines (MTL) was limited to lossless lines, and an example based on a symmetrical 3-conductor MTL with linear terminations was provided. Similar work has been repeated in a recent paper [2], yielding some results for non-linear sources and loads. Other authors are also currently interested in this approach [3]. In this article, we present results using symmetrical and asymmetrical transmission lines.

The simulation of the action of external fields on a bundle of conductor is obviously very important for EMC engineers, but a calculation yielding accurate values of the voltages and currents coupled into the conductors is unrealistic. This limitation occurs primarily because entering all of the parameters that are relevant to the computation of fields in the vicinity of several conductors is not feasible; the data required is extensive, and usually is not available. Another limitation arises from the complexity of the calculations involving all of the parameters.

Our approach for computing the action of an incident field is based on a MTL model in which the line is divided longitudinally into segments. The amplitude and phase of the incident field are assumed constant [4]. The segments need not be of equal length or have a maximum size with respect to wavelength. Ultimately, in the case of a plane wave impinging on the line with normal incidence, the line can be considered as one single segment. The weakest point in this approach is computing the incident field itself; accuracy is limited because simple radiation geometries must be assumed, such as a plane wave or the far field of a dipole.

An important part of the calculation is performed with the IsSpice analog circuit simulation program. This implies that appropriate equivalent circuits have been found. One of the advantages of the models described in this article is that they allow the solving (with acceptable accuracy) of many complex EMC problems with readily available hardware and software.

Computing Crosstalk in Lossless MTLs

First, let's consider a very general multiconductor transmission line (MTL), that is, a set of n parallel conductors with $n \geq 2$ (Fig. 1). The MTL, of length l, has n conductors numbered from 0 to n-1, and the n = 0 (zero) conductor represents a ground plane.

The zero conductor is always the voltage reference point; however, it is not mandatory that this conductor be a ground plane. When such a ground plane is present, it is natural to use it as conductor number zero. It is assumed that we know the geometrical and electrical properties of the MTL. The electrical characteristics of our MTL, which is assumed lossless, can be described with two square matrices of order n - 1, which are the per-unit-length inductance matrix [L] and the per-unit-length capacitance matrix [C], as defined by:

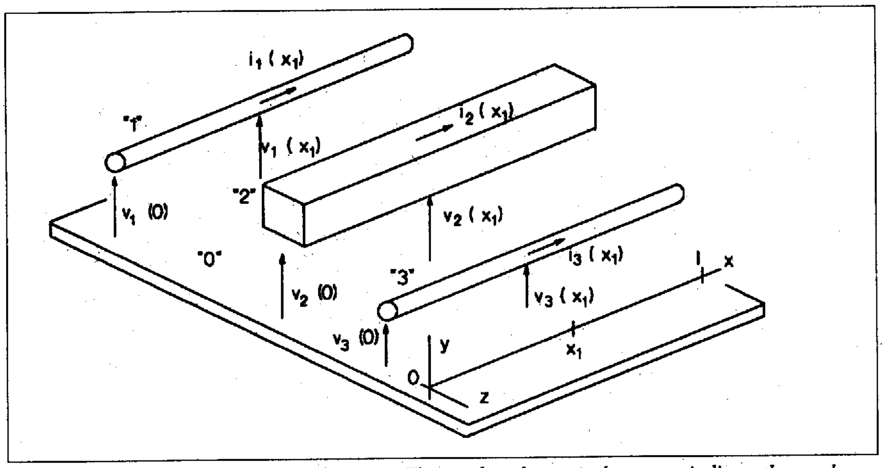
$$\frac{d\,[I]}{dx} = [C]\frac{d\,[V]}{dt} \qquad (1)$$

and

$$\frac{d\,[V]}{dx} = [L]\,\frac{d\,[I]}{dt} \qquad (2)$$

where x is the abscissa along the MTL, [I] is the column matrix of the n - 1 currents $i_1,...,i_{n-1}$, and [V] is the column matrix of the n - 1 voltages $v_1,...,v_{n-1}$. The numerical values of the coefficients of these two matrices can be obtained either by direct measurements or by calculation [5].

Following a well known procedure ([2], [6], [7]), and owing to the symmetrical properties of [L] and [C], we can now introduce two new real matrices, [T] and [S]. [T]



1. A multiconductor transmission line (MTL). The numbers between the quotes indicate the number of the conductor.

and [S] diagonalize, respectively, the matrices [C] [L] and [L] [C] as a unique real diagonal matrix, D. Because of the coefficients of D are positive, we may write them as $(1/c_i)^2$, and therefore obtain:

$$D = diag\ \{(1/c_1)^2,\ldots,(1/c_{n-1})^2\} \quad (3)$$

with

$$[T]^{-1}[C][L][T]=[S]^{-1}[L][C][S]=D \quad (4)$$

Once [T] has been computed, a particular solution for [S] is given by:

$$[S] = [C]^{-1}[T] \quad (5)$$

However, equation (4) has multiple solutions for [S] and [T]; for instance, any column vector those matrices can be multiplied by a non-zero real constant.

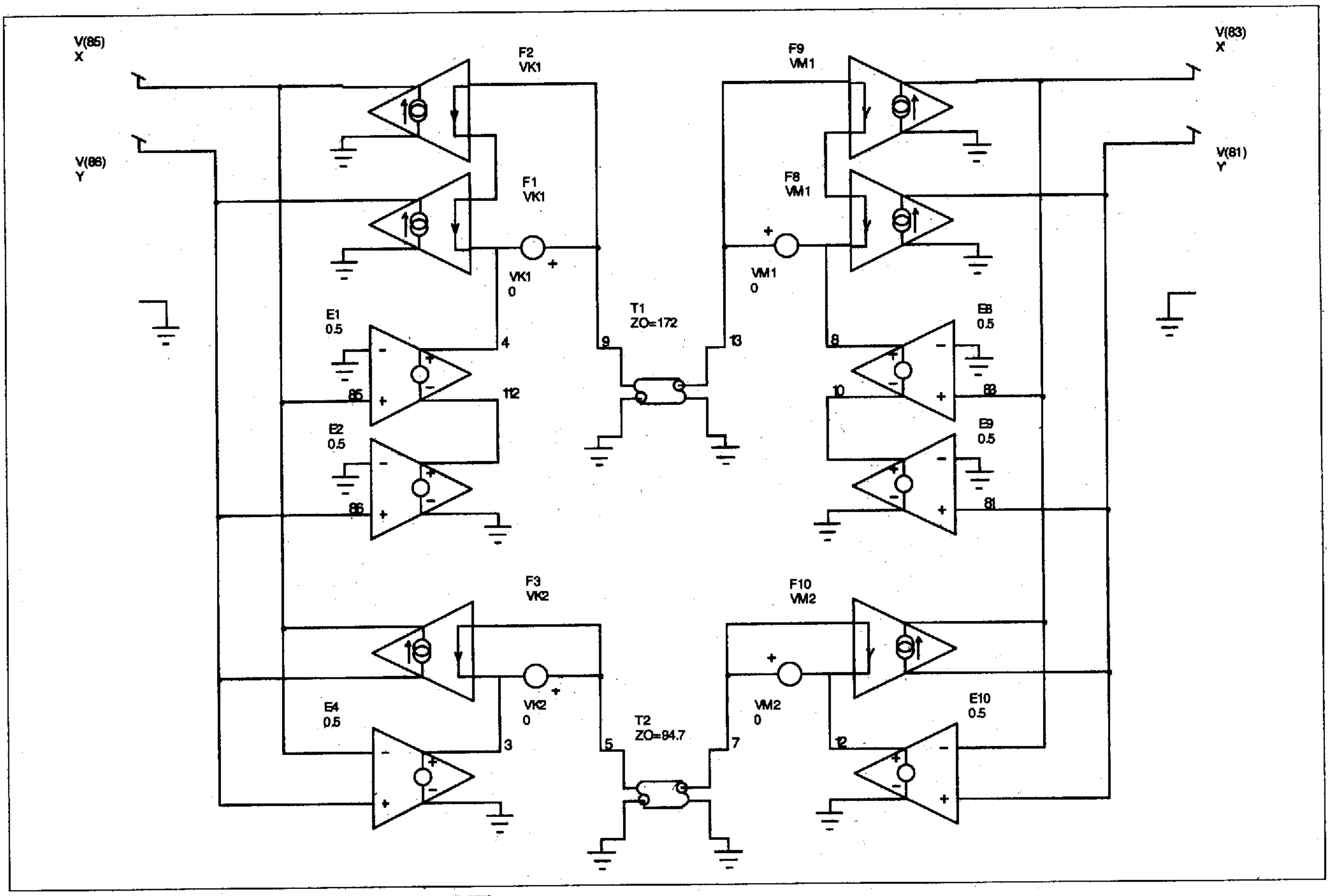
The coefficients $c_1,\ldots,c_{n-1}$ are the velocities of propagation for each of the n - 1 modes of propagation along the MTL. Two matrices play an important role; the characteristic impedance matrix $[Z_o]$, and the matrix of modal characteristics impedance, [zo], as defined by

$$[Zo]=[C]^{-1}\ [T]\ diag(1/c_n\ldots1/c_{n-1})\ [T]^{-1} (6)$$

and

$$[zo]=[S]^{-1}\ [Zo][T]=diag(zo_1,\ldots zo_{n-1}) \ (7)$$

The matrix [zo] is a diagonal matrix whose coefficients, $zo_1,\ldots,zo_{n-1}$, are the characteristic impedances of each mode of the MTL. The matrix [Zo] has the particular property of being the impedance of an n-pole which, when connected to an end of the MTL, would absorb any incoming signal without reflection.



2. Equivalent circuit for the 3-conductor symmetric MTL.

This short theoretical sketch contains all that is needed for the computation of the crosstalk of a MTL; we can now go into the modal domain, where we know the velocities of propagation and characteristic impedances of each mode, and we can return from the modal domain. More precisely, if $i_1,...,i_{n-1}$ are the currents in each mode and if $v_1,...,v_{n-1}$ are the voltages in each mode, the transformation into the modal domain is expressed by:

$$\underline{[V]} = [S]^{-1}[V] \qquad (8)$$

and

$$\underline{[I]} = [T]^{-1}[I] \qquad (9)$$

where [I] is the column matrix of the n - 1 modal currents $i_1,...,i_{n-1}$; and [V] is the column matrix of the n - 1 modal voltages $v_1,...,v_{n-1}$.

It must be emphasized that in a practical calculation, we will use a particular choice of matrix [T] and [S]. This choice will affect the modal domain voltage and current, and also the modal characteristic impedance. Obviously, the propagation velocities of the modes and the matrix [Zo] are not dependent upon the choice of [T] and [S].

Crosstalk Model: Symmetrical Case

In order to produce a SPICE model for crosstalk in a lossless MTL, we only need to translate the last two equations into an equivalent circuit, and treat the propagation in each mode with the two conductor transmission line model provided in SPICE.

This technique can be shown for the simple (and classical) example of a three conductor symmetrical transmission line. For example, we assume that the three conductors are equi-distant and the zero conductor is in the middle. For the parameter values, which we assumed in [2] and [4], the inductance and capacitance matrices are given by:

$$[L] = \begin{bmatrix} 0.816 & 0.270 \\ 0.270 & 0.816 \end{bmatrix} \mu H/m$$

$$[C] = \begin{bmatrix} 48.90 & -12.1 \\ -12.1 & 48.90 \end{bmatrix} pF/m$$

Or choice of modal transformation is:

$$[T] = \begin{bmatrix} 1 & -1 \\ 1 & 1 \end{bmatrix} = [S]$$

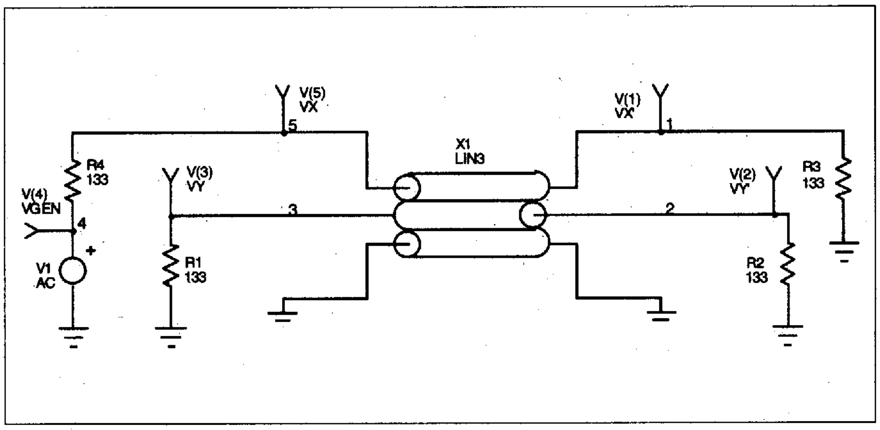
Because of the symmetry of this MTL, [T] and [S] take on the simple form shown above. Also, for this particular choice of [T] and [S], the modes traditionally have special names; mode 1 is called the common mode, and mode 2 is called the differential mode. Following that, we have: $c_1 = 1.58 \times 10^8$ m/s, $c_2 = 1.73 \times 10^8$ m/s; $zo_1 = 172$ ohms, $zo_2 = 94.7$ ohms. The characteristic impedance matrix is:

$$[Zo] = \begin{bmatrix} 133 & 38.6 \\ 38.6 & 133 \end{bmatrix} \Omega$$

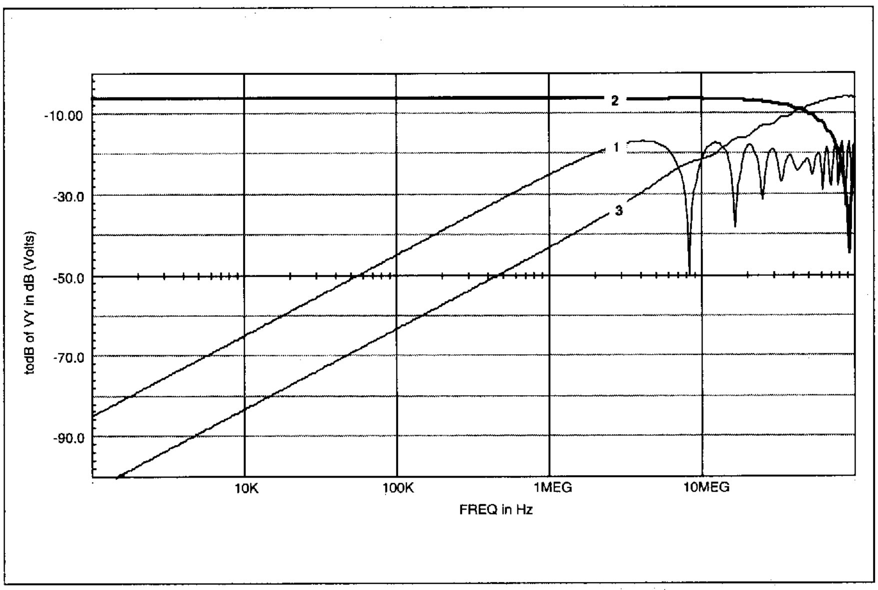
The resulting equivalent circuit for the MTL is shown in Fig. 2 (contact author Charles Hymowitz for the corresponding SPICE netlist, as well as for the netlists for the circuits in Figs. 3, 5, 7, and 10). Equation 8 is translated into three voltage dependent voltage sources on each side of the line, which impose the modal domain voltages. Eq. 9 is first inverted and then translated into three current dependent current sources. These convert the currents of the modal domain into the currents on the conductors.

When the MTL is terminated according to Fig. 3 (the improperly named "matched" termination), the transmission and crosstalk versus frequency characteristic of the line can be plotted (Fig. 4).

The previous results can be obtained without a SPICE simulation. However, the SPICE simulator approach makes the study of a complex circuit (Fig. 5) possible. In this case, the upper circuit portion (digital) is the



3. "Matched" termination of the 3-conductor symmetric MTL.



4. Characteristics of the 3 conductor MTL as plotted by IntuScope: Waveform 2 (upper curve at low frequencies) is the transmission (VX'); Waveform 1 is the near end crosstalk (VY); Waveform 3 is the far end crosstalk (VY').

source of the disturbances and the lower circuit (linear) is the susceptor. Figure 6 shows the result of a transient simulation run using IsSPICE. The pulsed source has a fast (1 ns) rise and fall time.

Crosstalk Model: Asymmetrical Case

The method described in the previous section can be applied to the case of asymmetrical lines. Of special interest is coaxial cable, considered next. Crosstalk through shielded cable is usually described with a transfer impedance and a trans fer admittance [8].

Transfer impedance relates the voltage appearing inside the cable to the common mode current on the cable. It can be divided into three terms: a diffusion term, an aperture coupling term, and a porpoising coupling term [9]. For braided shields, such as the one found on RG-58 coaxial cable, the diffusion term is negligible, and the transfer impedance is mostly a resistive term (about 10 m*/m) in series with an inductive term (about 1 nH/m).

Transfer admittance relates the current appearing inside the cable to the voltage between the cable shield and an external return path. It is a capacitive coupling term, which should be specified as a through elastance [10], in order to be a property of the cable only. That is, it is independent of the measurement set-up. Typical values for the through elastance for RG-58 coaxial cable are on the order of 4 x 10^7 m/F. By definition, if one multiplies this quantity by the product of the internal line capacitance of the cable and the line capacitance of the outer circuit, one obtains the mutual coupling (C_{12}) between the inner conductor and the external return conductor.

Now, let's consider a SPICE model for a shielded cable with the following parameters:

Characteristic impedance: 50 Ω
Propagation velocity inside the cable: 2 x 10^8 m/s
Inductive term of the transfer impedance: 1.6 nH/m
Through elastance = C_{12}/C_1C_2 = 2.74 x 10^6 m/F

We consider that this cable is over a ground plane, the characteristic impedance of the shield/ground plane line is 183 Ω, and the cable's propagation velocity is nearly 3 x 10^8 m/s. This structure is a three conductor transmission line with the shield as conductor 0, the inner conductor as conductor 1, and the ground plane as conductor 2. The resulting inductance and capacitance matrices are:

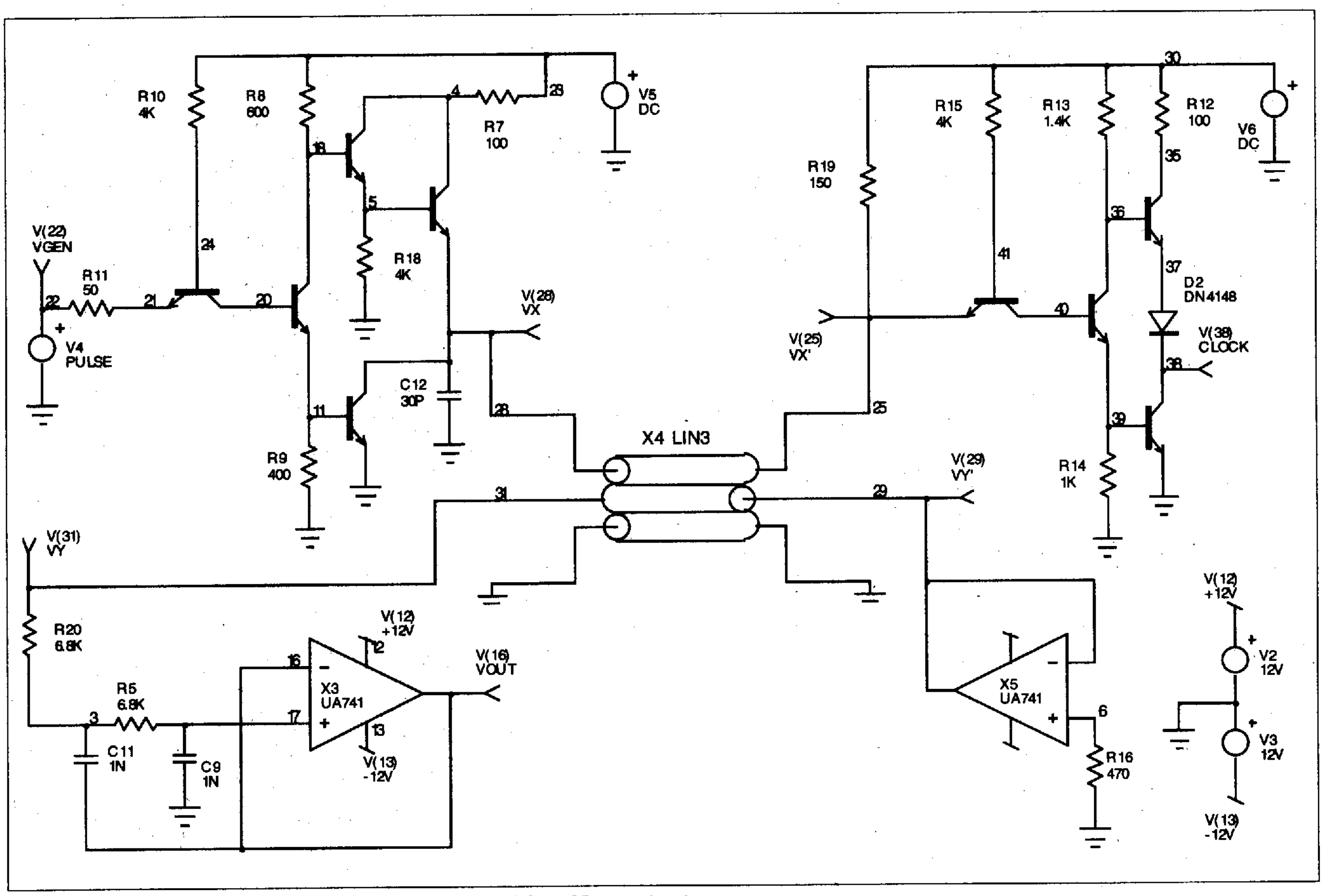
$$[L] = \begin{bmatrix} 0.250 & 1.60E{-3} \\ 1.60E{-3} & 0.610 \end{bmatrix} \mu H/m$$

$$[C] = \begin{bmatrix} 100 & 4.99E{-3} \\ 4.99E{-3} & 18.2 \end{bmatrix} pF/m$$

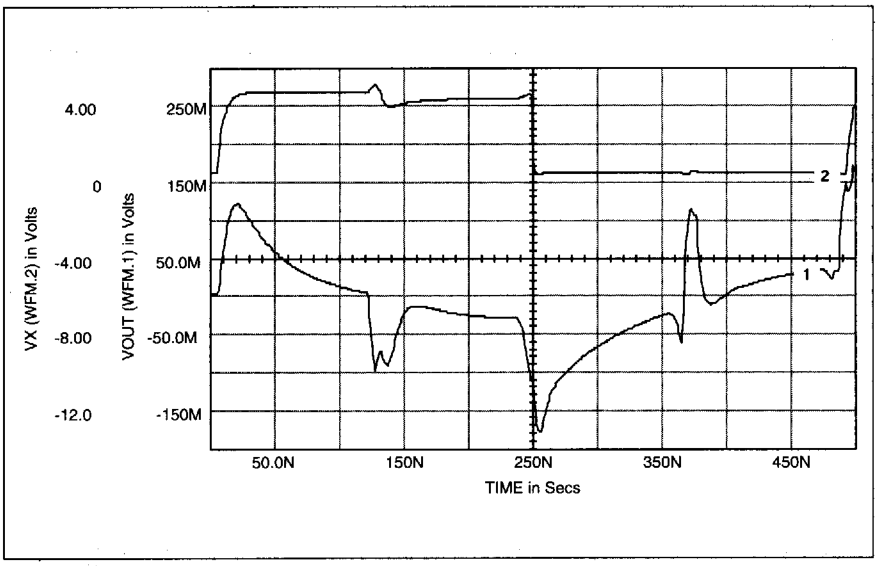
Our choice of modal transformation is:

$$[T] = \begin{bmatrix} 1.000 & 457.0 \\ -85.19 & 1.000 \end{bmatrix}$$

$$[S] = \begin{bmatrix} 1.000 & 85.19 \\ -457.0 & 1.000 \end{bmatrix}$$



5. The 3 conductor symmetric MTL implemented in a mixed digital/linear circuit.



6. Noise produced at the output of the linear circuit: Waveform 2 is the voltage, VX, produced by the driver stage on the upper conductor of the MTL; Waveform 1 is the noise voltage, VOUT, at the output of the 10kHz low-pass filter circuit.

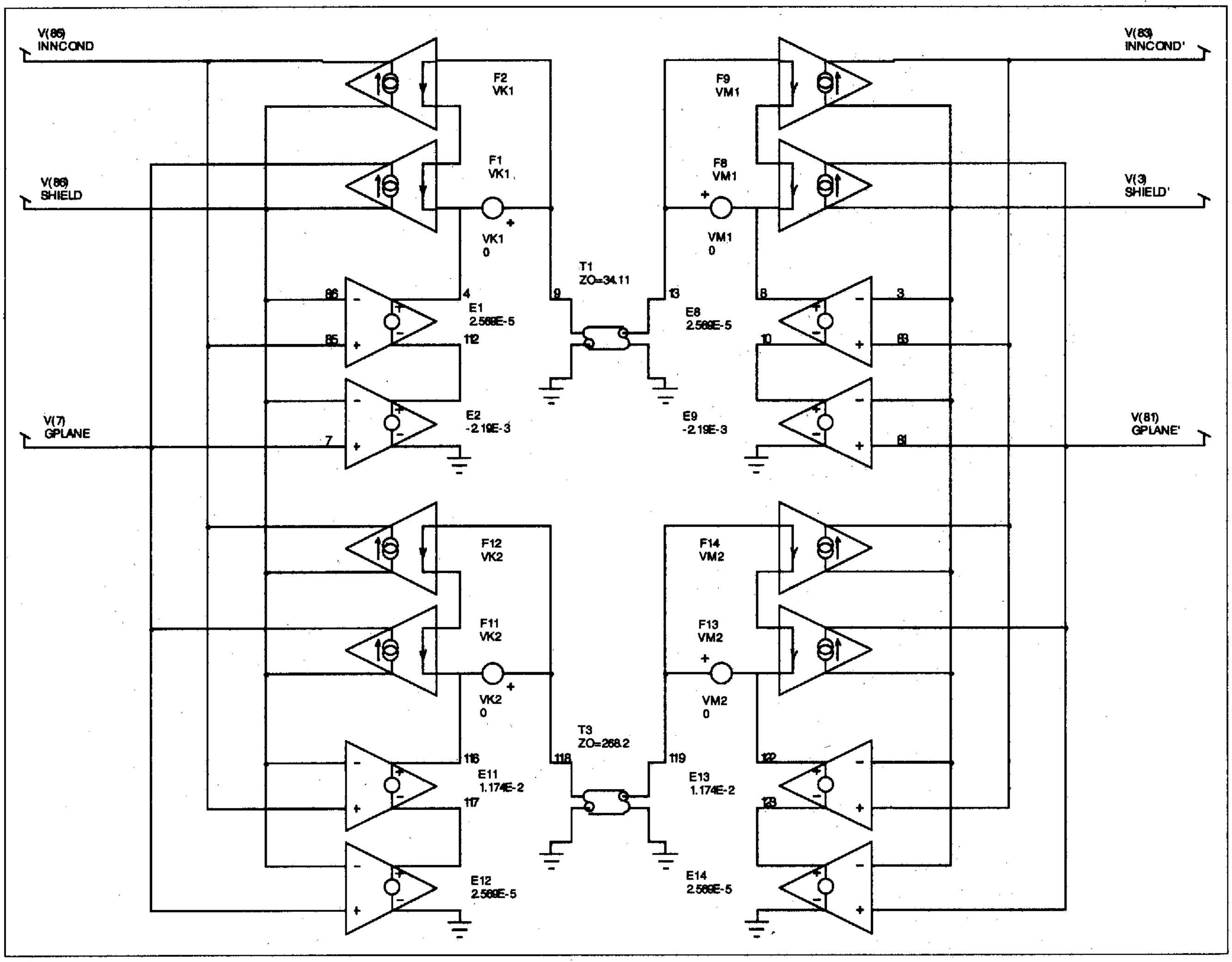
$$[S]^{-1} = \begin{bmatrix} 2.569E{-5} & -2.188E{-3} \\ 1.174E{-2} & 2.569E{-5} \end{bmatrix}$$

The propagation velocities and modal impedances are found as $c_1 = 3.00 \times 10^8$ m/s, $c_2 = 2.00 \times 10^8$ m/s, $zo_1 = 34.11\ \Omega$, and $zo_2 = 268.2\ \Omega$.

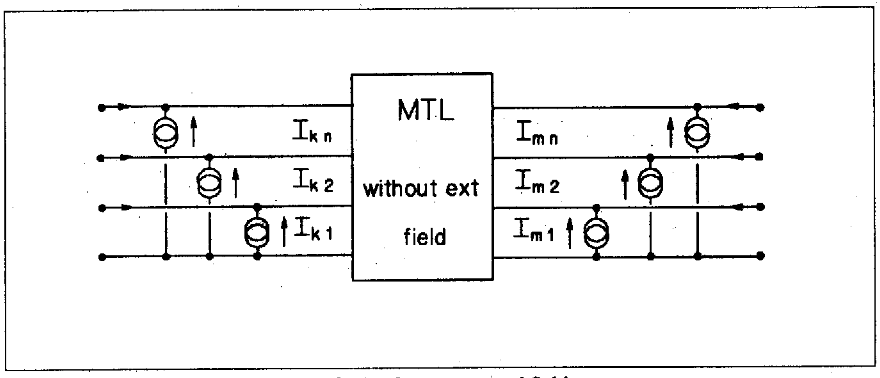
The propagation velocities are not affected by the coupling because the coupling is very weak. One should keep in mind that the modal impedances are somewhat arbitrary. The characteristic impedance matrix is:

$$[Zo] = \begin{bmatrix} 50.0 & 0.1865 \\ 0.1865 & 183 \end{bmatrix} \Omega$$

The resulting equivalent circuit for the MTL is shown in Fig. 7. The explanation is the same as for Fig. 2, except that due to the lack of symmetry of the matrix equations, we have four voltage dependent voltage sources and four current de-



7. Equivalent circuit for the asymmetric MTL.



8. The 6 FES of a 4 conductor MTL submitted to an external field.

pendent current sources on each side of the line.

SPICE Model for Crosstalk in Lossy MTL?

The possibility of having a simple time domain equivalent circuit in the case of a lossless MTL stems from the fact that the three matrices, Zo, T and S, are all real, and frequency independent. Unfortunately, this is not the case when the line becomes lossy. This is why the above approach is limited to lossless lines.

The first consequences of neglecting losses is that we grossly underestimate the coupling at lower frequencies. This is because the resistive part of the transfer impedance is lost. This part is simply the resistance of the zero conductor in the previous examples. In the case of the coaxial cable example, it means that we will underestimate the coupling below 1 MHz.

The second consequence is the lack of higher frequency losses, as encountered in two conductor lossy transmission lines. Other more complex consequences have been discussed in detail [6].

Field-to-Wire Coupling Transmission Line Models

An equivalent circuit for the coupling of an incident field to a 4 conductor MTL is shown in Figure 8. In this equivalent circuit, one introduces two field-equivalent-sources (FES) at each end of each conductor [5] [6]. The amplitudes of the FES for the i-th conductor are given by:

$$I_{ki} = \int_0^1 \{\frac{-e_i(x\,;t-\frac{x}{c})}{Zo} + j_i(x\,;t-\frac{x}{c})\}\mathrm{d}x \tag{10}$$

$$I_{mi} = \int_0^1 \{\frac{e_i(x;\,t-\frac{1-x}{c})}{Zo} + j_i(x;\,t-\frac{1-x}{c})\}\mathrm{d}x \tag{11}$$

where $e_i(x, \gamma)dx$ is the voltage induced by the varying magnetic field in the loop of length dx formed by the conductor i and the 0V conductor, and where $j_i(x;\gamma)$ is the displacement current injected into the conductor i on a length dx. These integrals simply state that the current due to external fields is the summation of the action of those fields on elementary wire segments along the wire, with an appropriate delay for each segment.

Our approach consists in dividing the line into segments on which the amplitude and phase of the incoming wave are assumed nearly constant. On each of these segments, we can therefore write:

$$e_{i(x;\gamma)} \approx e_{i(x_o;\gamma)} \tag{12}$$

$$j_{i(x;\gamma)} \approx j_{i(x_o;\gamma)} \tag{13}$$

where x_o is the abscissa of the middle of the segment. It is then possible to obtain a simple equivalent circuit for the integrals of Eq. 10 and 11 on each segment.

The procedure is now applied to a two conductor MTL, using the notations of Fig. 1. We will consider the external field produced by a broadside vertical antenna (Fig 9).

Assuming far-field conditions, we can compute the vertical electric field $E_y(t)$ and the transverse magnetic field $B_z(t)$ for a given field $E_o(t)$ produced by the antenna at a reference point on the ground plane as:

$$E_y(t) = E_o\,(t-\Delta t)\frac{r_o}{r} \tag{14}$$

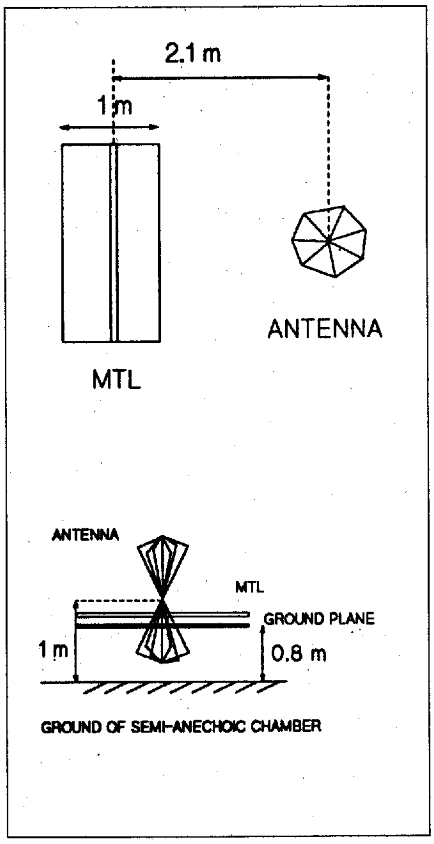
$$B_z(t) = \frac{E_o(t-\Delta t)}{c}\,\frac{r_o}{r}\sin\theta \tag{15}$$

where r is the distance to the antenna, r_o is the distance from the antenna to the reference point, Δt is the delay and θ is the appropriate angle. Dropping the unnecessary indices, and introducing the height, h, of the conductor above ground and its capacitance, C, to ground, we have:

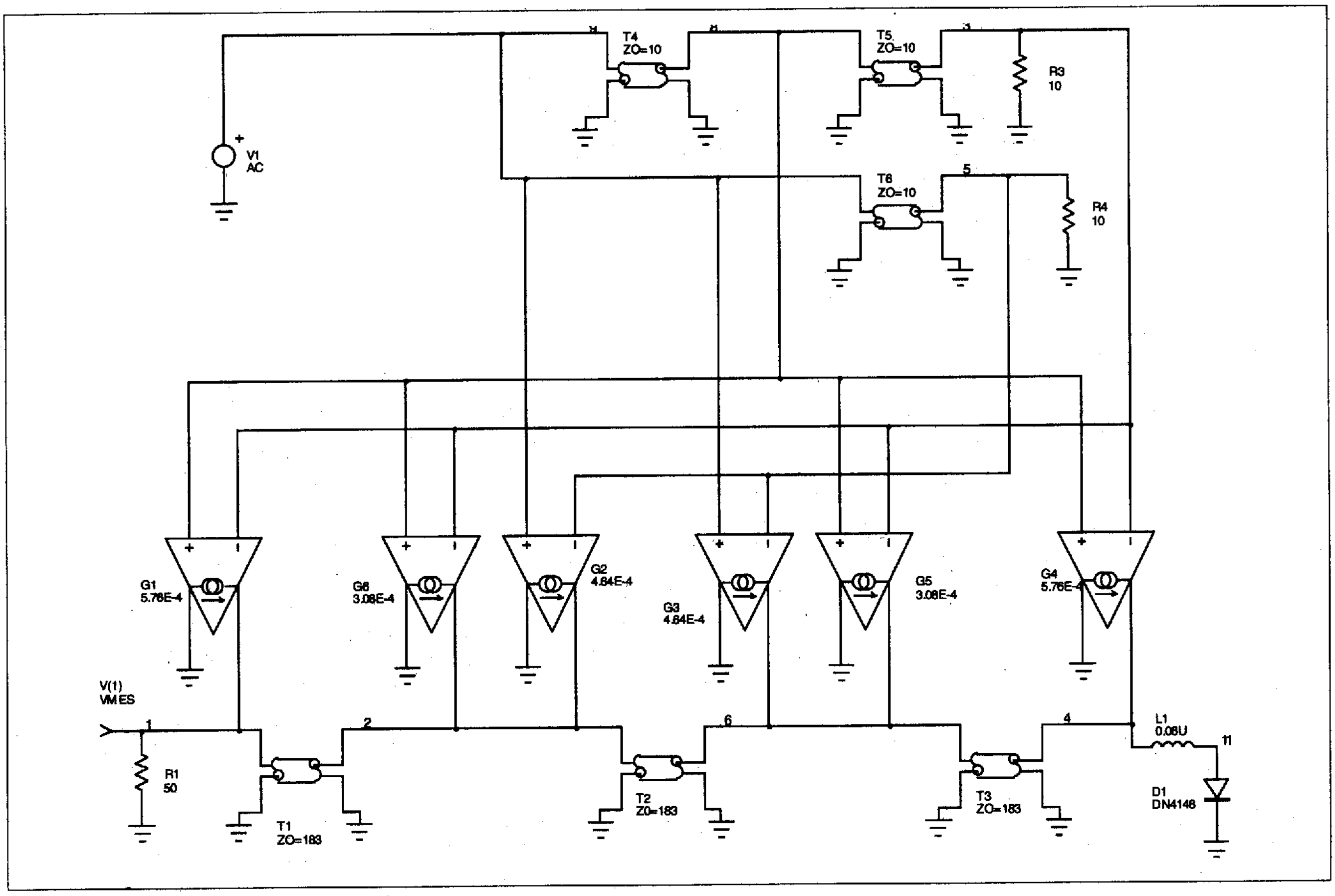
$$e = h\frac{\partial B_z}{\partial t};\; j = hC\frac{\partial E_y}{\partial Lt}$$

For a given segment of the MTL, the values of FES given by Eqs. 10 and 11 may now be written as:

$$I_K = \frac{h\,r_o}{Z_o\,r}(L-\sin\theta)\,[E_o(t-\Delta t) - E_o(t-\Delta t-\frac{\delta}{c})] \tag{17}$$



9. A single conductor MTL submitted to an external field.



10. Equivalent circuit for field to MTL coupling.

and

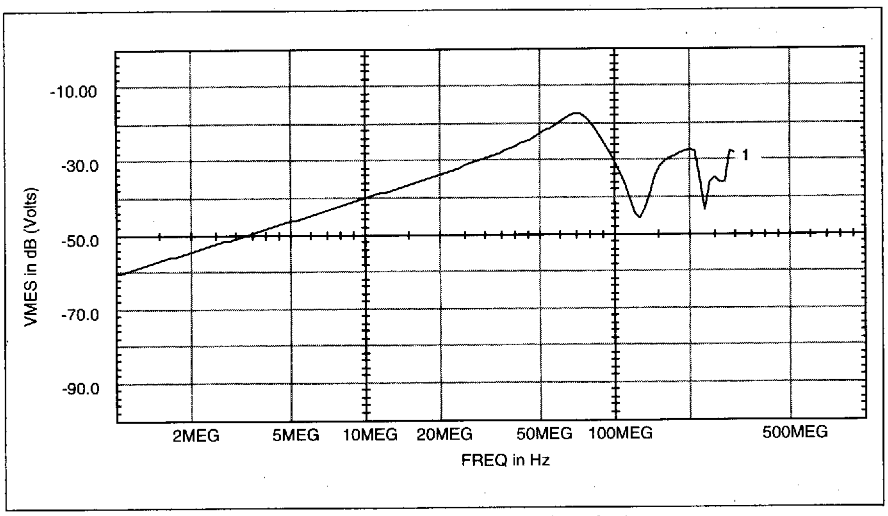
$$I_m = \frac{h\,r_o}{Z_o\,r}\,(L + \sin\theta)\,[E_o(t-\Delta t) - E_o(t-\Delta t-\frac{\delta}{c})] \qquad (18)$$

where δ is the length of the given segment.

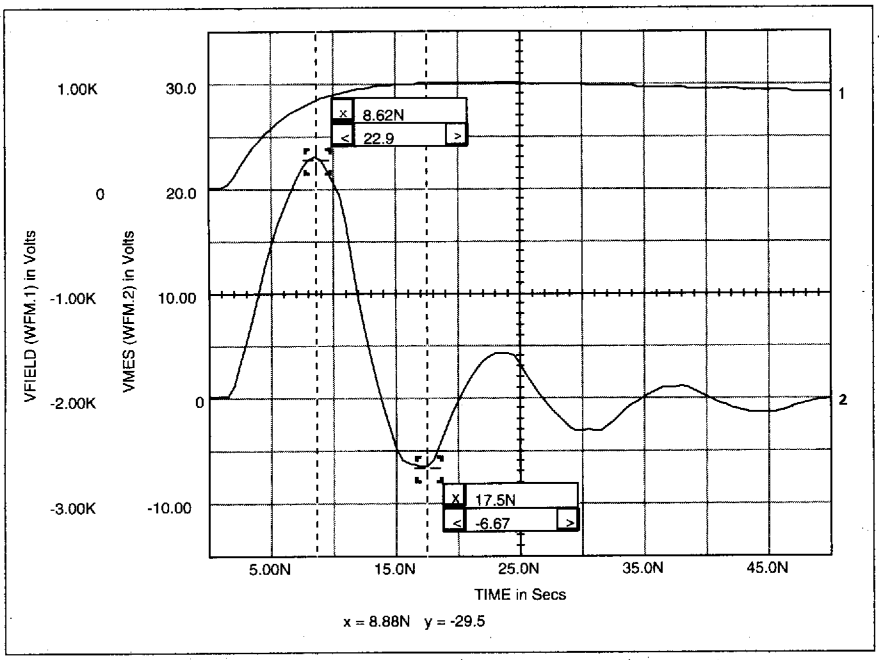
We divided the 2 m long MTL in Fig. 9 into 3 segments of equal length, which was thought to be sufficient for frequencies up to 500 MHz. Obviously, those segments are not electrically short at 500 MHz, but this is not important because our only assumption is that the phase and amplitude of the incident field must be nearly constant. The particular geometry chosen (almost a broadside incidence) for the antenna and the MTL allow for this assumption.

The single conductor of diameter 16 mm was 85 mm above the ground plane making the characteristic impedance Zo = 183 Ω. The equivalent circuit for the line illuminated by the field, given by Fig. 10, contains two terminating resistances of 50 Ω and 47 Ω. One volt produced across V1 translates to a field of 1 volt/meter at the center of the MTL. The 6 FES (two per segment) can be seen on the bottom of Fig. 10.

Figure 11 shows the effective length of the MTL versus frequency up to 300 MHz. This effective length is defined as the voltage, VMES, divided by the field at the reference point, which is the middle of the



11. Effective length versus frequency, in dB relative to 1 m. (0 dB = 1 m)



12. Time domain response. Waveform 1 (VField) is the incident field (1 kV peak, 10 ns rise time). Waveform 2 is the voltage VMES, which peaks at 22.9 V.

MTL. Figure 12 shows the time domain simulation where a bi-exponential field was assumed. In order to produce this waveform, a RLC generator was introduced in the circuit shown in Fig. 10.

It is valid to approximate the signal inside a coaxial cable due to external fields by first considering the two conductor problem of the shield and ground plane in the incident field; then, to use the three conductor approach for internal coupling along with the currents computed in the first step, but with the external field removed. This approach could be followed with a simulation using the models in this section.

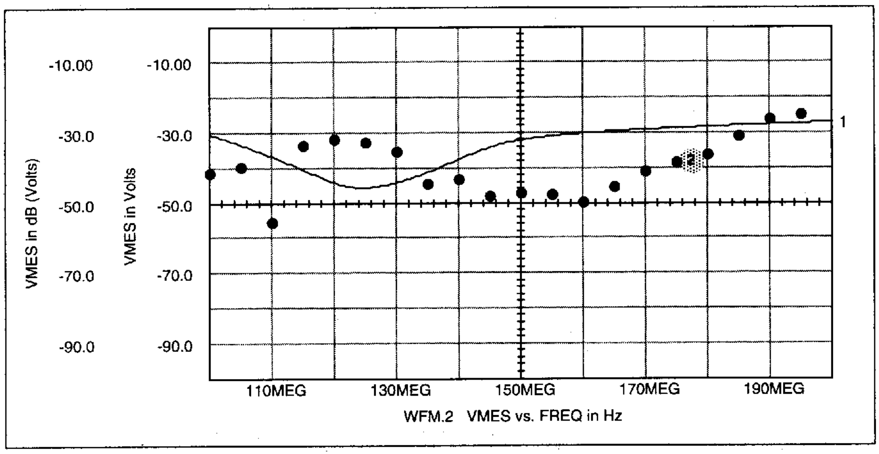
Experimental Validation

Our experimental set-up consisted of a semi-anechoic chamber in which the 2 meter long MTL was installed. The ground conductor of the MTL was 80 cm above the conducting ground of the semi-anechoic chamber. The source antenna was a conventional EMCO 3109 biconical antenna. From the thickness of the tapered absorbers, which was 60 cm, we expected fields with acceptable homogeneity at up to 200 MHz. This homogeneity was not controlled. Measurements of the voltages appearing on the MTL were made with a Hewlett-Packard 8590A spectrum analyzer connected at one end of the line (hence the 50 Ω termination). The field amplitude was calibrated with an electrically short antenna installed 85 mm above the ground plane and 25 cm away from the center of the MTL in the direction of the biconical antenna. This monitoring antenna was removed for the final measurements.

Figure 13 shows the calculated and experimental results for the effective length of the line terminated with a 47 Ω resistor according to Fig. 10 (100 MHz to 200 MHz). We included a 0.06 µH inductance in our model, which represents the resistor's lead inductance. This parameter, however, is not very critical below 200 MHz.

Agreement between the IsSPICE simulation and the experimental data is usually within 10 dB and always within 20 dB. Discrepancies can be explained by field calibration and measurement accuracy (±3 dB), and to other phenomena that we did not try to model, such as field non-homogeneity, standing waves (±10 dB), and resonances caused by the finite size of the ground plane (±15 dB). Also, Eqs. 14 and 15 yield only a very crude estimate of the near field of the biconical antenna.

Figure 14 shows the computed time-domain voltages obtained with a causal 180 MHz sinusoidal field of 76 V/m RMS and termination which uses the 1N4148 diode. This simulation uses a very accurate nonlinear SPICE model for the diode.



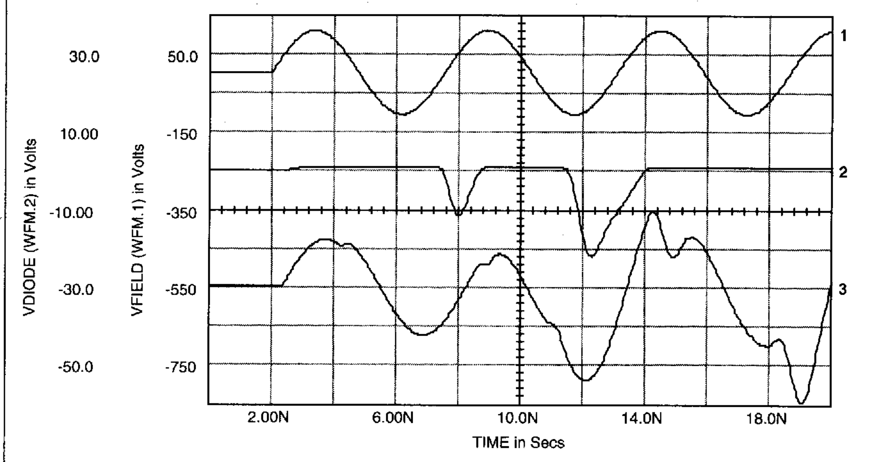
13. Effective length of the MTL terminated with 47 Ω.

Simulations

The determination of the SPICE parameters to be used in our simulation was made with dedicated software. The schematics and simulations were accomplished with the Intusoft ICAP/3 package, which contains SPICENET (a schematic entry program for SPICE), IsSPICE (a 32-bit version of SPICE), and IntuScope (a SPICE graphics post processor). IsSPICE was run on a 386 based PC with an 80387 coprocessor, running at 25 MHz. Run times for the various simulations are given below:

Fig 4	250 frequency values	13 s
Fig 6	500 time values	395 s
Fig 11	126 frequency values	9 s
Fig 12	100 time values	617 s
Fig 13	200 frequency values	6 s
Fig 14	200 time values	163 s

Note the simulation time for Figure 7 was 3356 s on a 286/8 PC with an 80287 co-



14. Response of the diode terminated MTL to a high amplitude field. Waveform 1 (VField) is the incident field (100 V/m per division). Waveform 2 (VDiode) is the voltage across the diode (10 V per division). Waveform 3 (VMes) in the output voltage VMES (2 V per division).

processor running a 16-bit version of SPICE. The simulation time on a 486/33 PC was 44.76 s indicating the tremendous advances in productivity resulting from current PC technology and 32-bit versions of SPICE.

Not all SPICE simulators perform equally well in this kind of simulation. We found out that some other well-known software programs failed dramatically in the time domain calculation involving non-linear devices.

Conclusion

Calculations dedicated to EMC, similar to the ones proposed in this article, were formerly available only on mainframe computers. An important consequence of our work is that it demonstrates the feasibility of these complex calculations on personal computers, and therefore, with very low hardware and software costs.

It is useful to keep in mind some of the shortcomings of the transmission line models. The models

- Are only useful when the transverse separation of the conductors is smaller than half a wavelength at the relevant frequencies
- Do not account for true common mode currents, i.e., antenna mode currents which would appear on a rod in free space
- Neglect the effect of losses in the lines, and therefore, the consequences of common resistance coupling, the increase of this coupling by skin effect, and also the proximity effects between nearby conductors
- Do not take into account the fact that most bundles of wire do not run parallel to each other and to a ground plane. Additionally, the simulation of non-cylindrical structures is not accounted for. These features can cause the results to be inaccurate due to the influence of length and frequency on couplings in very long bundles of wire where the relative position of the wires are random

In addition, our assumptions concerning the amplitude variations of the incident field are not very accurate in many cases. Despite these difficulties, our approach is very useful as long as the upper frequency limit of the model is not reached, because true common mode currents can usually be neglected.

In review, the new modeling techniques allow calculation in the time domain, as well as the frequency domain. Nonlinearities, protective devices, and even complex circuitry can be included at both ends of the transmission line. Disturbing voltages can also be studied at any node in a susceptor network. The possibility of including arbitrary sources of radiated disturbances is also very powerful because it could allow, for instance, the simulation of aperture coupling in an enclosure.

CD

—Frédéric Broyde, Evelyne Clavelier and Charles Hymowitz

Frederic Broyde is chairman of EXCEM, a company based in Rueil-Malmaison, France, where he also manages research projects related to EMC.
Evelyne Clavelier is general manager of EXCEM. She is responsible for supervising scientific computation related to research and engineering projects.
Charles Hymowitz is vice president of INTUSOFT, makers of circuit simulation tools. He is responsible for writing the Intusoft Newsletter.

References

1. C.R. Paul, "A Simple SPICE Model for Coupled Transmission LInes," Proc. of the 1988 IEEE Symposium on EMC," Seattle, p. 327-333.
2. F. Broyde, E. Clavelier, "Le Simulateur IsSpice pour evaluer "l'immunite d'une liaison," Electronique industrielle et Toute l'electronique, n°174, 18 juin 1990, p. 39-42.
3. G.L. Matthaei, H.C.-H. Shu, S.I. Long, "Simplified Calculation of Wave-Coupling Between Lines in High-Speed Integrated Circuits," IEEE Trans. on Circuits and Systems, vol. 37, no. 10, Oct 1990, p. 1201-1208.
4. F. Broyde, E. Clavelier et al, "Crosstalk and field to wire coupling problems: The SPICE simulator approach," Proc. of the 9th Int'l. Zurich Symposium on EMC, March 1991, p. 23-28.
5. C.R. Paul, "Computation of the Transmission Line Inductance and Capacitance Matrices from the Generalized Capacitance Matrix," IEEE Trans. on EMC, vol. EMC-18, no. 4, Nov. 1976, p. 175-183.
6. O. Pardo-Gibson, Ph. Auriol, "Etude et Simulation Temporelle de la susceptibilite Electromagnetique de Systemes Interconnects," these presentee a l'Ecole Centrale de Lyon, Sept. 1987.
7. F. Broyde, "Compatibilite Electromagnetique des Liaisons et Interconnexions," support de stage (stage n° 3) de la societe EXCEM, Nov. 1989.
8. E.F. Vance, "Coupling to shielded cables," R.E. Krieger, 1978.
9. L.O. Hoeft, J.F. Hofstra, "Experimental evidence of porpoising coupling and optimization in braided cables," Proc. of the 7th International Zurich symposium on EMC, March 1989, p. 505-509.
10. E.P. Fowler, L.K. Halme, "State of the art in screening measurements," Proc. of the 9th International Zurich symposium on EMC, March 1989, p. 151-158.
11. L.G. Meares, C.E. Hymowitz, "Simulating With SPICE," Intusoft, 1988.