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# CROSSTALK AND FIELD TO WIRE COUPLING PROBLEMS: THE SPICE SIMULATOR APPROACH

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## ABSTRACT

A new approach is presented for the simulation of crosstalk on a bundle of n conductors, and the effect of its excitation by an external field. This approach is based on a multiconductor line model in which the line is divided longitudinally into segments, on which the amplitude and phase of the incident field are assumed constant. We establish an equivalent circuit for the line, suitable for a SPICE simulation software. This equivalent circuit includes n-1 two-conductors transmission lines, each of them provided with two field equivalent sources, the amplitude of which is determined by an other equivalent circuit.

The determination of the parameters (i.e. the component values) of the equivalent circuits is obtained with a dedicated software, but the simulation itself is done by a conventional SPICE simulator. We demonstrate practical examples of calculation and their comparison with experiment is shown and discussed.

## 1. INTRODUCTION

PAUL already proposed a paper [1] on a SPICE model dedicated to the simulation of crosstalk in Multiconductor Transmission Lines (MTL). His results were limited to lossless lines, and he gave an example based on a symetrical 3-conductor MTL with linear terminations. A similar work has been repeated in a recent paper by the authors [2], yielding some results for non-linear sources and loads. Other authors are also currently interested in this kind of approach [3]. This communication extends our previous work to the simulation of field to wire coupling.

The simulation of the action of external fields on a bundle of conductors is obviously very important for EMC engineers, but it must be understood that a calculation yielding accurate values of voltages and currents coupled into the conductors is un-realistic. This is primarily because entering all parameters relevant for the computation of fields in the vicinity of the conductors is not feasible: too many data are needed and these data are not usually available. An other limitation arises from the complexity of calculation involving all those parameters.

Our approach for the computation of the action of an incident field is based on a MTL model in which the line is divided longitudinally into segments, on which the amplitude and phase of the incident field are assumed constant. The segments need not to be of equal length, nor to have a maximum size with respect to wavelength. Eventually, 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 the computation of the incident field itself, because a simple behaviour has to be assumed, for instance a plane wave, or the far field of a dipole.

An important part of our calculations is done with a Spice simulation software, and this implies that we found appropriate equivalent circuits. One of the advantages of the models described in this paper is that it is now possible to solve (with an acceptable accuracy) many complex EMC problems involving a bundle of wires and its nonlinear terminations, with popular hardware and software.

## 2. COMPUTATION OF CROSSTALK IN LOSSLESS MTL

In the following, MTL will stand for a very general multiconductor transmission line, that is to say a set of n parallel conductors (n ≥ 2). A MTL of length l is shown on figure 1 , and its n conductors have been numbered from 0 to n-1, the zeroth conductor being a ground plane in this example.

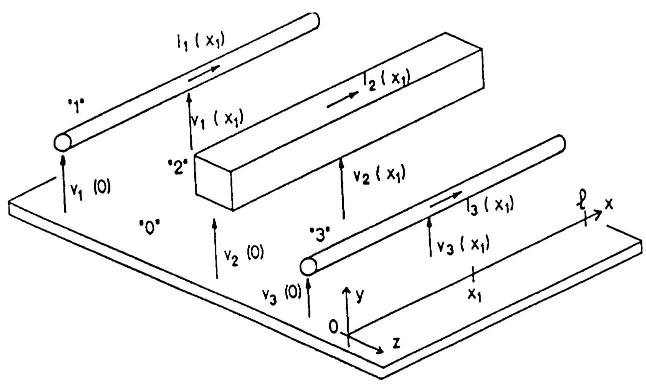


Figure 1 : A multiconductor transmission line (MTL). The numbers between quotes are the number of the conductors.

The zeroth conductor is always the voltage reference. It is not necessary that this conductor be a ground plane, however, when such a ground plane is present it is natural to use it as the conductor number zero. We will assume that we know the geometrical and electrical properties of the MTL. The electrical characteristic 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:

$$\frac{d\,[I]}{d\,x} = [C]\,\frac{d\,[V]}{d\,t} \tag{1}$$

and

$$\frac{d\,[V]}{d\,x} = [L]\,\frac{d\,[I]}{d\,t} \tag{2}$$

where

x is the abscisse along the MTL (see figure 1) ;
[I] is the column matrix of the n-1 currents $i_1,\ldots,i_{n-1}$ ;
[V] is the column matrix of the n-1 voltages $v_1,\ldots,v_{n-1}$ .

The numerical values of the coefficients of those two matrices can be obtained either by direct measurements, or by calculation, for example similar to the one used in [4].

Following a well known procedure [2], [5], [6] and thanks to the symmetry properties of [L] and [C], we can now introduce two new real matrices [T] and [S] which diagonalize respectively the matrices [C] [L] and [L] [C] as a unique real diagonal matrix D. The coefficients of D being all positive we may write them as $(1/c_i)^2$, and we obtain:

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

with

$$[T]^{-1}\,[C]\,[L]\,[T] = [S]^{-1}\,[L]\,[C]\,[S] = D \tag{4}$$

and

$$[S]=[C]^{-1}\,[T] \tag{5}$$

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. At this point, two matrices play an important role: the characteristic impedance matrix $[Z_o]$ and the matrix of modal characteristic impedances $[z_o]$, as defined by

$$[Z_o]=[C]^{-1}\,[T]\,\mathrm{diag}(1/c_1,\ldots,1/c_{n-1})\,[T]^{-1} \tag{6}$$

and

$$[z_o] = [S]^{-1}\,[Z_o]\,[T] = \mathrm{diag}(z_{o\,1},\ldots,z_{o\,n-1}) \tag{7}$$

The matrix $[z_o]$ is a diagonal matrix and its coefficients $z_{o\,1},\ldots,z_{o\,n-1}$ are the characteristic impedances of each mode of the MTL. The matrix $Z_o$ has the particular property of being the impedance of a n-pole which, when connected to an end of the MTL, would absorb any incoming signal without reflexion.

This short theoretical sketch contains more than what is needed for the computation of the crosstalk of a MTL: we can go in 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 $\underline{i}_1,\ldots,\underline{i}_{n-1}$ are the currents in each mode and if $\underline{v}_1,\ldots,\underline{v}_{n-1}$ are the voltages in each mode, the transformation into the modal domain is expressed by:

$$[\underline{V}] = [S]^{-1}\,[V] \tag{8}$$

and

$$[\underline{I}] = [T]^{-1}\,[I] \tag{9}$$

where

$[\underline{I}]$ is the column matrix of the n-1 modal currents $\underline{i}_1,\ldots,\underline{i}_{n-1}$;
$[\underline{V}]$ is the column matrix of the n-1 modal voltages $\underline{v}_1,\ldots,\underline{v}_{n-1}$.

## 3. A SPICE MODEL FOR CROSSTALK IN LOSSLESS MTL

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

This can be shown on the simple example of the line whose cross section is shown on figure 2 where dimensions are in mm.

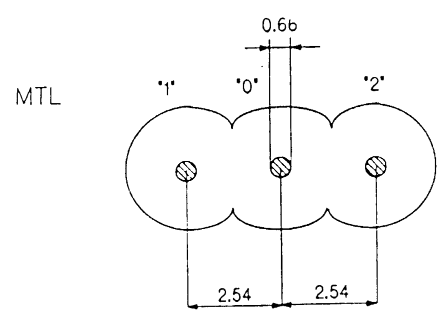


Figure 2 : Cross section of the symetrical 3-conductors MTL.

From these dimensions and the dielectric constant of the insulator, the different matrices take the following values:

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

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

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

$$[Z_o] = \begin{bmatrix} 133 & 38{,}6 \\ 38{,}6 & 133 \end{bmatrix} \Omega$$

Because of the symmetry of this MTL (which causes the particular simplicity of [T] and [S]), the modes have special names: the mode 1 is called the common mode, and the mode 2 is called the differential mode. We have:

$c_1 = 1.58 \quad 10^8$ m/s $\qquad c_2 = 1.73 \quad 10^8$ m/s

$z_{0\,1} = 172\ \Omega \qquad z_{0\,2} = 94{,}7\ \Omega$

The resulting equivalent circuit is shown on figure 3. When the MTL is terminated according to figure 4 (the improperly called "matched termination"), the transmission and crosstalk versus frequency characteristic of the line can be plotted, as shown on figure 5.

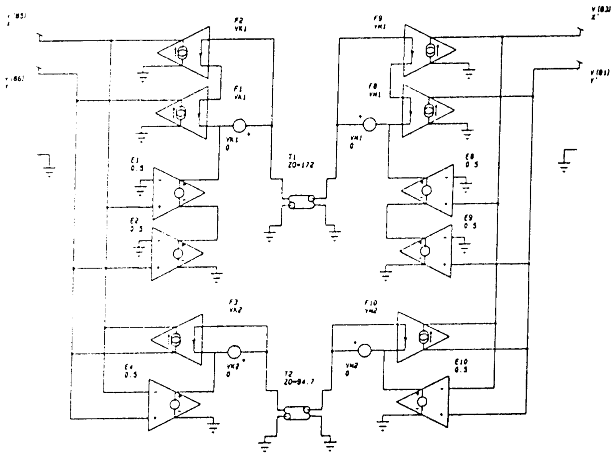

Figure 3: equivalent circuit for the 3-conductors MTL

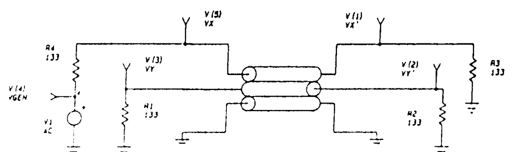

Figure 4: "matched" termination of the 3-conductors MTL

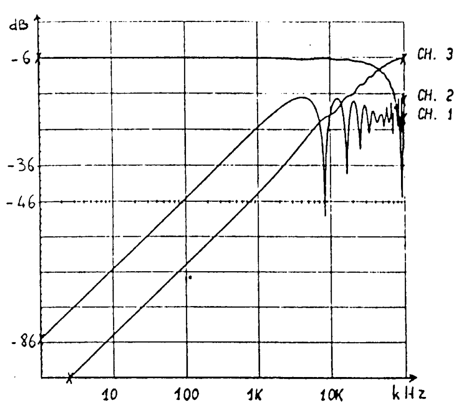


Figure 5: characteristics of the 3 conductors MTL: CH. 1 (upper curve at low frequencies) is the transmission; CH. 2 is the near end crosstalk; CH. 3 is the far end crosstalk.

The above results are quite easily obtained without Spice simulator. However, our Spice simulator approach makes the study of a complex schematic such as the one shown on figure 6 possible, where the upper circuit (digital) is the source of disturbances and the lower circuit (linear) the susceptor. Figure 7 shows the result of our transient simulation, when the pulsed source has a fast (1 ns) rise and fall time.

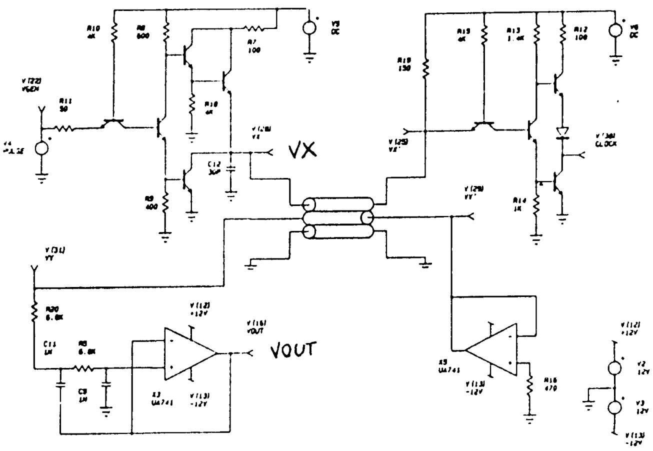


Figure 6: The 3 conductors MTL implemented in a mixed digital/linear circuit.

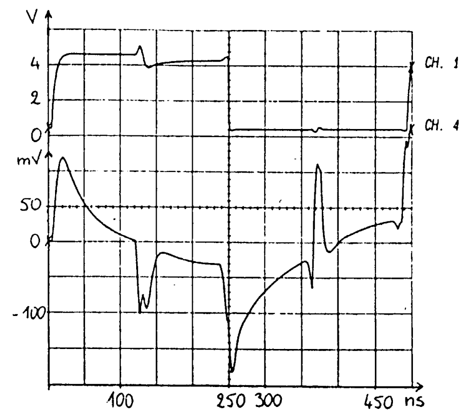


Figure 7: Noise produced at the output of the linear circuit: CH. 1 is the voltage VX produced by the driver stage on the upper conductor of the MTL; CH. 4 is the noise voltage VOUT at the output of the 10 kHz low-pass filter circuit.

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

## 4. FIELD TO WIRE COUPLING TRANSMISSION LINE MODELS

An equivalent circuit for the coupling of incident field to a MTL is shown on figure 8 in the case of a 4 conductor MTL. In this kind of equivalent circuits [5], [6] one introduces two Field-Equivalent-Sources (FES) at each end of each conductor. The amplitude of the FES for the i-th conductor are given by:

$$I_{k\,i} = \int_0^1 \left( - \frac{e_i(x;t-x/c)}{Z_o} + j_i(x;t-x/c) \right) dx \tag{10}$$

$$I_{m\,i} = \int_0^1 \left( \frac{e_i(x;t-(1-x)/c)}{Z_o} + j_i(x;t-(1-x)/c) \right) dx \tag{11}$$

where $e_i(x,\tau)$ 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 $j_i(x;\tau)$ is the displacement current injected into the conductor i on a length dx.

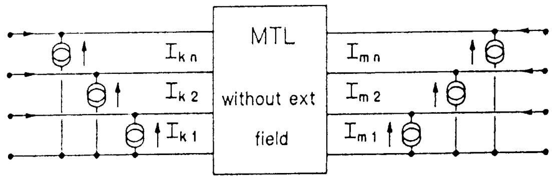


Figure 8: The 6 FES of a 4 conductor MTL submitted to an external field.

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:

$$e_i(x;\tau) \approx e_i(x_o;\tau) \qquad (12)$$

$$j_i(x;\tau) \approx j_i(x_o;\tau) \qquad (13)$$

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

The procedure will be shown in the following example of a single conductor MTL where the notations of figure 1 will be used. We will consider the external field produced by a broadside vertical antenna, in a set-up shown on figure 9 a).

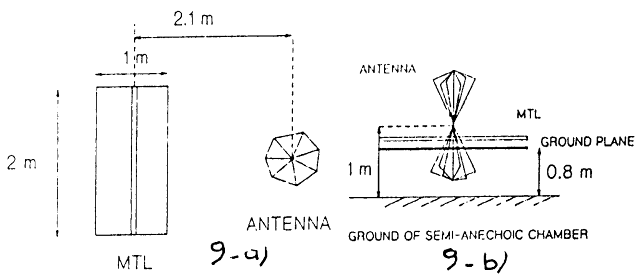


Figure 9: A single conductor MTL submitted to an external field.

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:

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

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

where r is the distance to the antenna, $r_o$ the distance from the antenna to the reference point, $\Delta t$ the delay and $\theta$ 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:

$$e = h\,\frac{\partial B_z}{\partial t} \quad \text{and} \quad j = h\,C\,\frac{\partial E_y}{\partial t} \qquad (16)$$

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

$$I_k = \frac{h\,r_o}{Zo\,r}\,(1-\sin\theta)\,\left[E_o(t-\Delta t)-E_o\left(t-\Delta t-\frac{\delta}{c}\right)\right] \qquad (17)$$

and

$$I_m = \frac{h\,r_o}{Zo\,r}\,(1+\sin\theta)\,\left[E_o(t-\Delta t)-E_o\left(t-\Delta t-\frac{\delta}{c}\right)\right] \qquad (18)$$

where $\delta$ is the length of the given segment.

We divided the 2 m long MTL of figure 9 into 3 segments of equal length, which was thought to be sufficient for frequencies up to 500 MHz. The single conductor of diameter 16 mm was 85 mm above ground plane, and the characteristic impedance was therefore Zo = 183 Ω. The equivalent circuit for the line illuminated by the field is given on figure 10, with two terminating resistances of 50 Ω and 47 Ω. One volt produced across V1 means 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 figure 10.

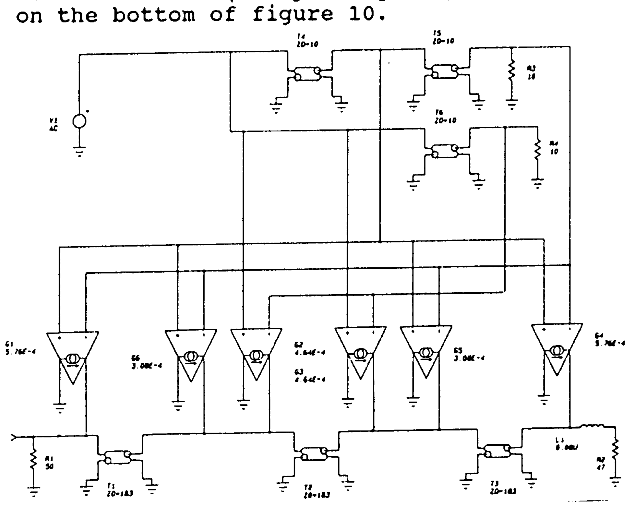

Figure 10: Equivalent circuit for field to MTL coupling.

Figure 11 shows the effective length of the MTL versus frequency up to 300 MHz, this effective length being defined as the voltage VMES divided by the field at the reference point, which was the middle of the MTL. Figure 12 is a time domain simulation where a biexponential field was assumed. In order to produce this waveform, a RLC generator was introduced in the circuit of figure 10.

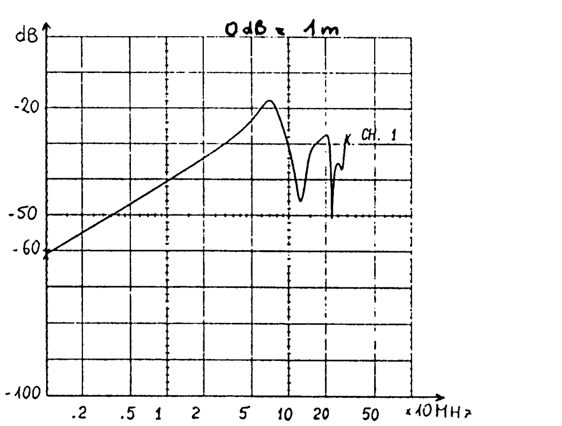


Figure 11: Effective length versus frequency, in dB relative to 1 m.

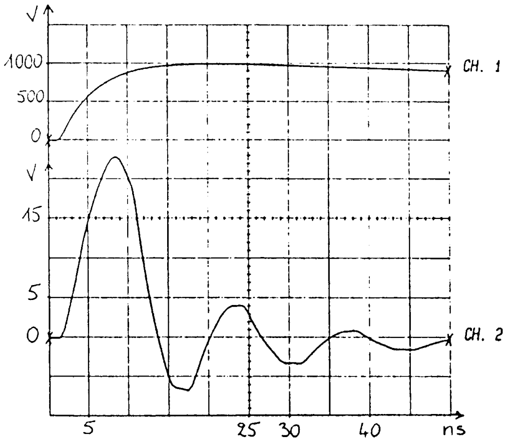


Figure 12: Time domain response. CH. 1 is the incident field (1kV peak, 10 ns rise time). CH. 2 is the voltage VMES, which peaks at 23 V.

## 5. EXPERIMENTAL VALIDATION

Our experimental set-up consisted of a semi-anechoïc chamber in which the 2 meters long MTL was installed. The ground conductor of the MTL was 80 cm above the conducting ground of the semi-anechoïc 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 as from 200 MHz. This homogeneity was not controlled. Measurements of the voltages appearing on the MTL were made with a Hewlett Packard 8590A spectrum analyser 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, 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.

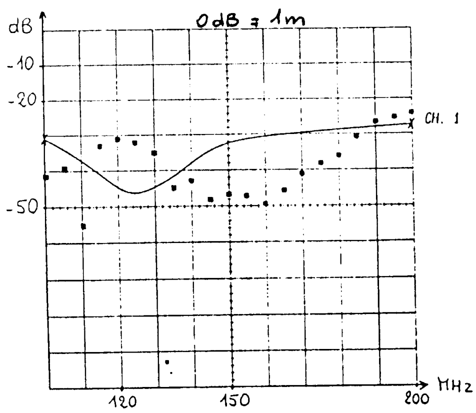


Figure 13: Effective length of the MTL terminated with 47 Ω.

Figure 13 shows the calculated and experimental results for the effective length of the line terminated with a 47 Ω resistor according to figure 10, in the 100 MHz to 200 MHz frequency range. We included a 0.06 μH inductance in our model, which represents the resistor leads inductance: this parameter was shown to be not very critical up to 200 MHz. Figure 14 shows the same data when the 47 Ω resistor is replaced by a 15 pF capacitor. Figure 15 gives the effective length when the 47 Ω resistor is replaced by a 1N4148 diode (our calculation assumed a linear 0 bias characteristic for the diode). In the experiments leading to figure 13 to 15, the field amplitude ranged between 128 dB(μV/m) and 138 dB(μV/m).

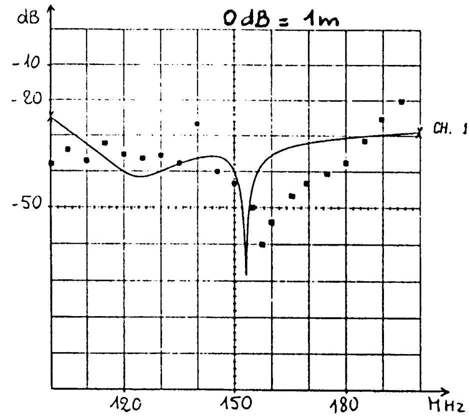


Figure 14: Effective length of the MTL terminated with 15 pF.

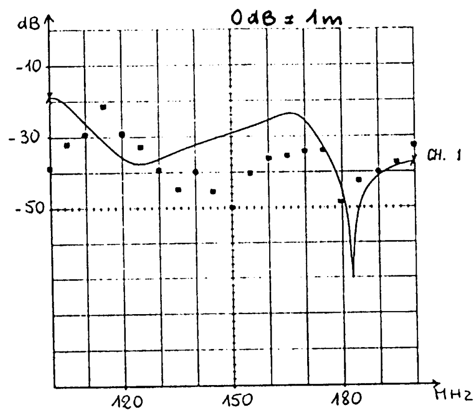


Figure 15: Effective length of the MTL terminated with a diode.

Agreement between our calculation and experiment is always within 20 dB, most of the time within 10 dB. Discrepencies can be explained by field calibration accuracy (± 3 dB), field inhomogeneity and standing waves (± 10 dB), and resonances caused by the finite size of the ground plane (± 15 dB).

Figures 16 and 17 show the computed time-domain voltages obtained with a causal 180 MHz sinusoïdal field of 76 V/m RMS, the MTL being terminated with the 1N4148 diode. This simulation uses a very accurate non-linear Spice model for the diode. Figure 17 includes a Fourier analysis giving for the amplitude of the fundamental frequency a value of 1.95 V RMS. The experimental value at this (high) field strength is 1.0 V RMS.

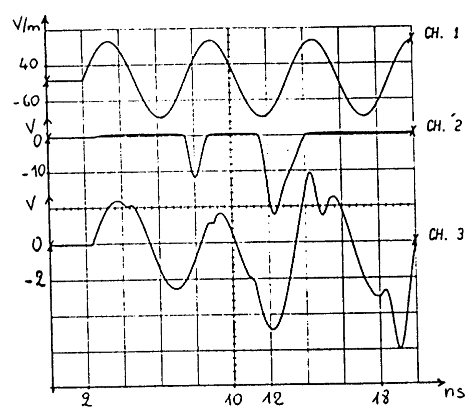


Figure 16: Response of the diode terminated MTL to a high amplitude field. CH.1 is the incident field (100 V/m per div.). CH.2 is the voltage across the diode (10 V per div.). CH.3 is the output voltage VMES (2 V per div.).

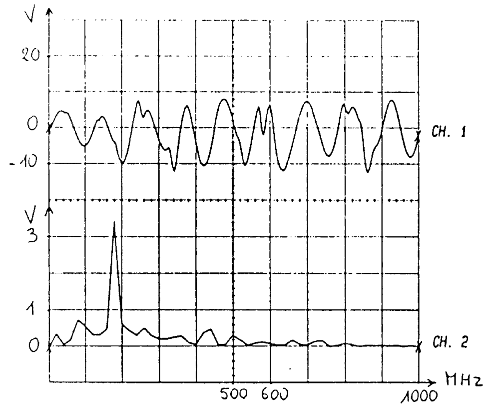


Figure 17: Response of the diode terminated MTL to a high amplitude field. CH. 1 is the output voltage VMES (5V per div.). CH. 2 is the FFT of this signal (1 V per div.).

## 6. SIMULATIONS

The determination of the Spice parameters to be used in our simulation was made with dedicated softwares. The schematics and simulation were accomplished with the Intusoft's ICAP/3 package, which contains a Spice 2G.6 compatible simulator. This simulator was used on a 80386 based PC with 80387 coprocessor, running at 25 MHz. Run time for the various simulation are given below:

| | | |
|---|---|---|
| figure 5 | 250 frequency values | 13 s |
| figure 7 | 500 time values | 395 s |
| figure 11 | 100 frequency values | 9 s |
| figure 12 | 100 time values | 617 s |
| fig.13,14,15 | 200 frequency values | ≈ 6 s |
| figure 16 | 200 time values | 163 s |
| figure 17 | 500 time values | 525 s |

Note: the simulation of figure 7 takes 3356 s on a 80286 based PC with 80287 coprocessor running at 8 MHz the ICAP/2 software of Intusoft!

Not all Spice simulators are equal in our kind of simulation: we found out that some other well-known softwares failed dramatically in time domain calculation involving non-linear devices. For instance, PAUL [1] using PSPICE would have been unable to produce simulations similar to the one shown on figures 6 and 7.

## 7. CONCLUSION

It is useful to keep in mind the major shortcomings of our transmission line models:

- our models are meaningless when the transverse separation of the conductors are not (much) smaller than the half wavelength at the relevant frequencies,

- the models do not account for true common mode currents (i.e. antenna mode currents), an instance of which are the currents which would appear on a rod in free space,

- we 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,

- most bundles of wires do not run parallel to each other and to a ground plane, and the simulation of non cylindrical structures gives some inaccurate results, especially for the influence of length and frequency on couplings in very long bundles where the relative position of wires is random,

- our assumptions concerning the amplitude variation of the incident field are not very accurate in most 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 current can usually be neglected.

Let's now review some of the advantages of our new approach. First of all, it allows calculation in the time domain as well as the frequency domain. Non-linearities, protective devices and even complex circuitry can be included at both ends of the line, and disturbing voltages can be studied at any node of a susceptor network. The possibility of including arbitrary sources of radiated disturbances is also very powerful; it allows for instance the simulation of aperture coupling in an enclosure.

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

## 8. BIBLIOGRAPHY

The reference list below only contains a very limited sample of useful references on crosstalk and field to wire coupling with MTL. This list is limited to contribution directly relevant to our Spice simulation of these phenomenons. More relevant references can be found in [3] and [4].

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