Circuit Models for Interconnects Using 3D Computational Techniques

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Abstract **–** This paper presents a method to generate circuit models from 3D distributed structures. In the first step a broadband characterization of the device is obtained through a reduced order modeling technique. Then in the second step a rational approximation of the coefficients of the impedance matrix is derived using a root mean squared procedure. An equivalent circuit is then synthesized and allows a global circuit simulation of the whole structure. The proposed methodology can be used with a segmentation technique for the analysis of complex structures: a whole 3D structure can be subdivided into smaller parts. Each part is characterized by an equivalent circuit deduced from wideband analysis. The connection between the substructures makes available a global simulation of the whole system inside a circuit platform (SPICE for example). Numerical results are shown for different kinds of interconnects (tracks and cables).

Keywords: Electromagnetic compatibility, interconnecting wires, and equivalent circuits.

I. INTRODUCTION

With the increasing use of on-board electronic equipment, mastering Electromagnetic Compatibility (EMC) compliance at early design stage is becoming a crucial technical issue for the automotive industry. Computer simulation of the EMC properties of electronic devices is a promising way to make EMC design faster and cost-effective, since it can be applied to the virtual product before fabrication.

In complex equipments involving interconnecting wires and electronic components a complete understanding requires a global analysis studying both electromagnetic modelling (for distributed parts) and circuit simulation (for lumped components). For a reliable simulation at the sub-system or system level efficient techniques are needed to address a hybrid field-circuit analysis. For a time domain analysis a strong coupling between a field solver in the time domain and a circuit simulator leads to a heavy process updating at each time step both the field quantities and the circuit values [1].

An efficient solution for transient analysis is to extract lumped circuit parameters from broadband field computation. Circuit extraction is a well known procedure in case of high speed interconnects [2]. Equivalent circuits or circuit macromodels can be easily derived from a rational approximation of the frequency response of the structure [2, 3]. In case of conducting cables an adequate modelling technique is usually achieved in the frame of the transmission line theory. However such approach does not remain valid because of the increase of the frequency range involved in EMC analysis. In [4] a technique based on a full wave integral approach has been developed and applied to unshielded cables.

The main computational cost relevant to circuit extraction techniques is the broadband full wave analysis. In order to minimize this computational cost a Padé approximation procedure provides an efficient way [5]. However such an approach requires getting the solution for a set of frequencies distributed over the whole frequency band. The results may be sensitive to the choice of the sample frequencies. A much more powerful approach is reduced order modelling: instead of solving the field problem, the equations characterizing the device are first brought to the state space form of the linear system and the system is reduced by appropriate algebraic transformation.

In recent years many reduced-order modelling (ROM) techniques have been studied [6-9]. In [9] a Laguerre SVD (Singular Value Decomposition) algorithm was shown to provide an efficient ROM technique. The capabilities of the approach were demonstrated in the case of linear systems resulting from the telegrapher equations or from the PEEC (Partial Element Equivalent Circuit) method. In our work we show that such a technique can be efficiently used in connection with a 3D finite element approach. With such an approach the impedance of distributed structures can be easily obtained over a broadband. Then circuit models are deduced using a root mean square procedure and can be directly incorporated into a circuit simulator if required.

II. REDUCED-ORDER MODELING

Consider an electromagnetic problem involving one excitation port. From a 3D edge finite element approach the metrical system governing the vector of unknowns e(t) is given by [10],

$$\begin{cases} M_m \dfrac{d^2e}{dt^2} + M_d \dfrac{de}{dt} + M_s e = Lu(t) \\ y(t) = L^T e(t) \end{cases} \tag{1}$$

where $\mathbf{M_m}$ is the mass matrix, $\mathbf{M_d}$ is the damping matrix, $\mathbf{M_s}$ is the stiffness matrix, **L** is a selector matrix, u(t) is the excitation and y(t) is the output of interest.

In the Laplace domain with zero initial condition the transfer function H(s), Y(s) = H(s)U(s) is given by,

$$H(s) = L^T \left(M_s + sM_d + s^2 M_m\right)^{-1} L \tag{2}$$

the second-order system can be reformulated as a first-order linear system of order N,

$$\begin{aligned} C\frac{dx}{dt} + Gx(t) &= Bu(t) \\ y(t) &= B^T x(t) \end{aligned} \tag{3}$$

Consequently from equation (3) the transfer function can be written as,

$$H(s) = B^T (G + sC)^{-1} B. \tag{4}$$

The aim of model order reduction is to replace the mathematical model (equation (4)) by a model which is much smaller but keeps the same behaviour over a given frequency band. In other words the purpose is to find a system governed by a reduced state space form,

$$\begin{aligned} C_r \frac{dx}{dt} + G_r x(t) &= B_r u(t) \\ y(t) &= B_r{}^T x(t) \end{aligned} \tag{5}$$

where the dimension q of the reduced matrices $\mathbf{C_r}$, $\mathbf{G_r}$, and $\mathbf{B_r}$ verifies $\mathbf{q << N}$.

The new transfer function $\mathbf{H_r}$ is,

$$H_r(s) = B_r{}^T (G_r + sC_r)^{-1} B_r. \tag{6}$$

The ROM technique used in this work has been described in [9]. It is based on a system description in terms of orthonormal Laguerre functions. It uses the singular value decomposition and Arnoldi algorithm. To illustrate, the efficiency of this ROM technique is demonstrated in the case of the loop antenna shown in Fig. 1. The variation of the impedance computed with the reduced order method is compared to Fig. 2 with a standard 3D finite element method for two different orders **(q = 7, and q = 10)**. It is shown that for **q = 10** an excellent agreement is obtained over the entire frequency band.

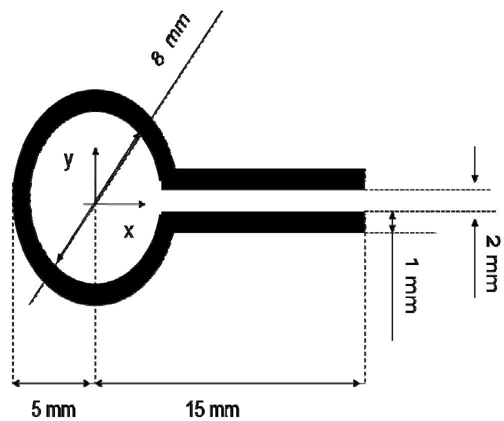


Fig. 1. Loop antenna.

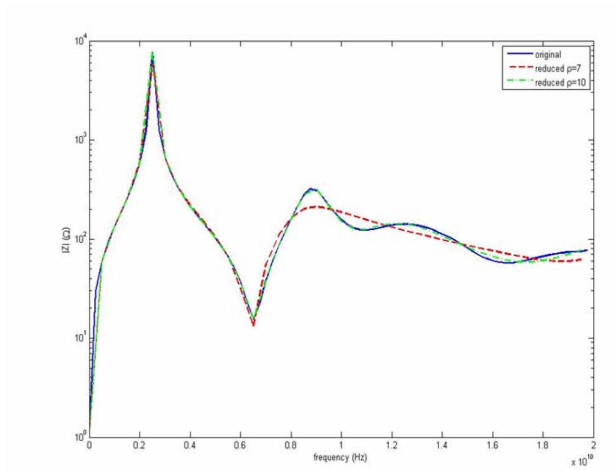


Fig. 2. Comparison between ROM technique and standard 3D finite element method.

III. CIRCUIT MODELS

Once the impedances of distributed structures are known over a given frequency band, an approximate expression based on rational functions can be determined. For a lossless transmission line of length l, the coefficients of the two port impedance matrix are known analytically [11],

$$Z_{11}(s) = Z_{22}(s) = Z_c \coth(\tau s)$$
$$Z_{12}(s) = Z_{21}(s) = Z_c \frac{1}{sh(\tau s)} \tag{7}$$

where $Z_c = \sqrt{\frac{L}{C}}$ is the characteristic impedance and $\tau = l\sqrt{LC}$.

These coefficients can be expressed as Fourier series,

$$Z_{11}(s) = \frac{Z_c}{\tau}\left(\frac{1}{s} + \sum_{n\geq 1} \frac{2s}{s^2 + n^2\omega_0^2}\right)$$
$$Z_{12}(s) = \frac{Z_c}{\tau}\left(\frac{1}{s} + \sum_{n\geq 1} \frac{(-1)^n 2s}{s^2 + n^2\omega_0^2}\right) \tag{8}$$

where $\omega_0 = \frac{\pi}{\tau}$.

Since the structures studied in this work behave like transmission lines, a rather natural rational approximation of this impedance uses second order rational functions with real coefficients. It is determined via a least mean squared procedure. The whole band is divided into $\mathbf{N_b}$ sub-intervals where $\mathbf{N_b}$ is the number of resonant frequencies. For example, the value of $\mathbf{Z_{11}}$ is searched as,

$$Z_{11}(s) \approx \sum_{k=1}^{N_b} Z_{ap}^k(s) = \sum_{k=1}^{N_b} \frac{b_o^k + b_1^k s}{a_o^k + a_1^k s + a_2^k s^2}. \tag{9}$$

Each second order rational function corresponds to an electrical circuit constituted with an inductance L with resistive loss (corresponding to a resistance R_1) in parallel with a capacitance C with a leakage conductance $G = 1/R_2$ shown in Fig. 3.

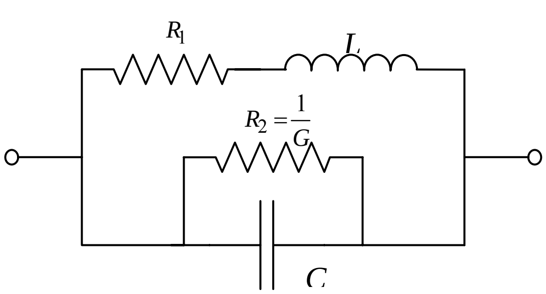


Fig. 3. Equivalent circuit for a second order rational function.

IV. CONDUCTING CABLES

Figure 4 shows a perfectly conducting cable above a ground plane. The load is 50 Ω. The ground plane is a perfect conductor. Figure 5 shows the 3D finite element computation over the whole band and the corresponding rational approximation. This approximated impedance is built using a sum of 4 second order rational functions. Each rational function corresponds to a resonance peak. The distances between the different peaks characterize the resonances of the transmission line. A reasonable agreement between the finite element based approach and the approximation is obtained.

$$Z_{11}(s) \approx \sum_{k=1}^{4} \frac{b_o^k + b_1^k s}{a_o^k + a_1^k s + a_2^k s^2} \tag{10}$$

where the coefficients are given in Table 1.

Table 1. Coefficients of the rational approximation.

	i	b_i^k	a_i^k
k = 1	0	-3.14	1
	1	-1.56e+02	-5.21e-02
	2		2.02
k = 2	0	-2.19	1
	1	4.33e+01	3.54e-02
	2		5.09e-01
k = 3	0	-3.52e-01	1
	1	-1.74e+01	-1.80e-02
	2		2.20e-01
k = 4	0	-3.75	1
	1	9.48	1.88e-02
	2		1.27e-01

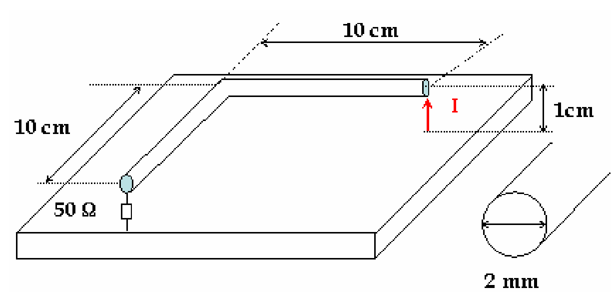
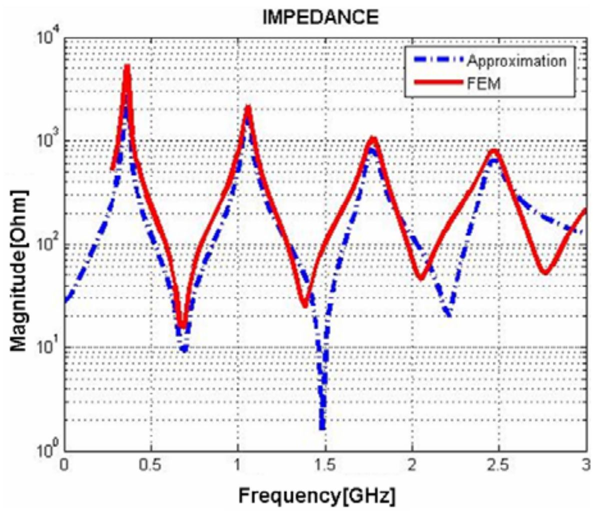
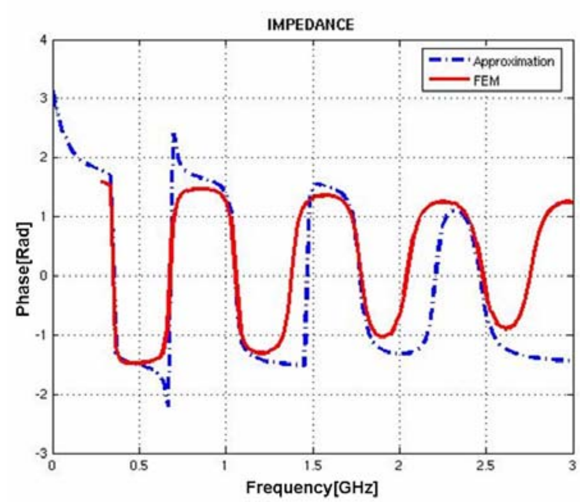


Fig. 4. Cable of diameter 1 mm above the ground plane.

Figure 6 shows a more complicated case: the cable is 50 cm long and the height is not constant. The corresponding numerical results for the full wave computation and the approximation are shown on Fig. 7. The coefficients are given in Table 2. The rational series can be directly incorporated into a circuit simulator like SPICE for example and can provide an efficient time domain simulation of the signal propagating along the transmission lines.



a) Magnitude of the impedance.



b) Phase of the impedance.

Fig. 5. Comparison between direct 3D FEM computations (solid line) and rational approximation (dashed line).

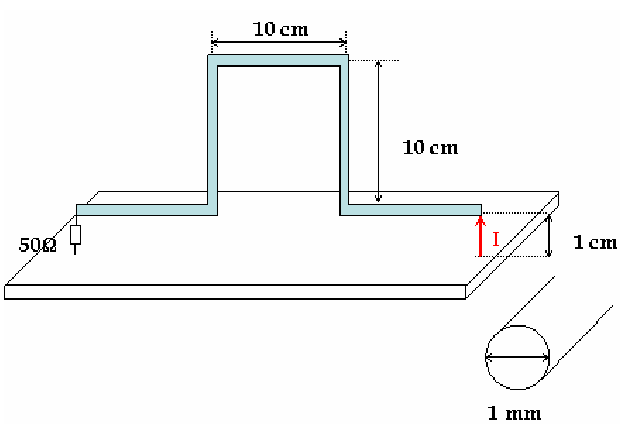
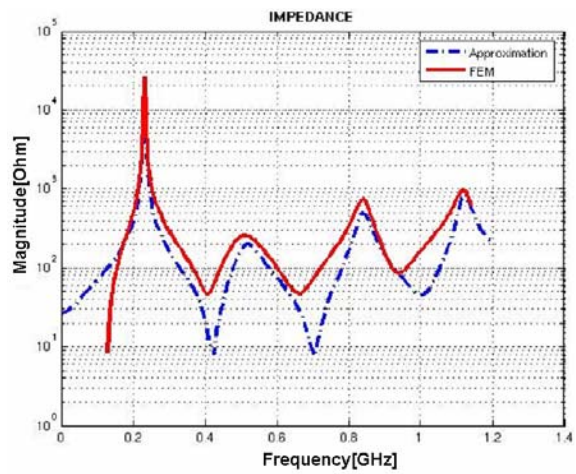


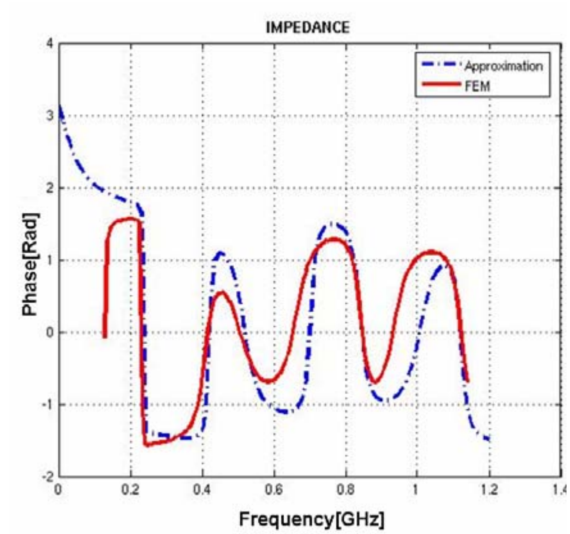
Fig. 6. Cable of diameter 1 mm above the ground plane.

Table 2. Coefficients of the rational approximation.

	i	b_i^k	a_i^k
	0	-1.89e+01	1
k = 1	1	4.31e+02	5.52e-03
	2		1.84e+01
	0	-5.00	1
k = 2	1	4.58e+01	2.32e-01
	2		3.75
	0	-5.38	1
k = 3	1	2.20e+01	4.62e-02
	2		1.41
	0	2.62	1
k = 4	1	1.79e+01	2.11e-02
	2		7.94e-01



a) Magnitude of the impedance.



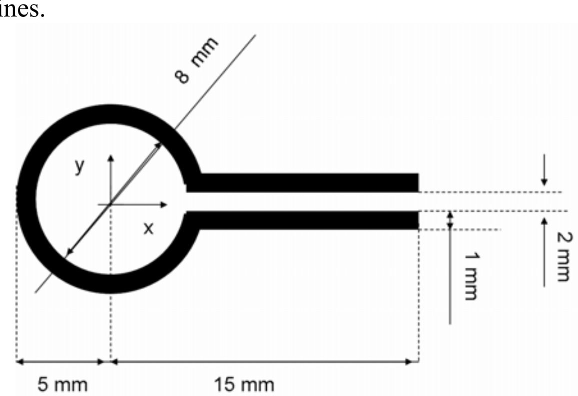
b) Phase of the impedance.

Fig. 7. Comparison between direct 3D FEM computation (solid line) and rational approximation (dashed line).

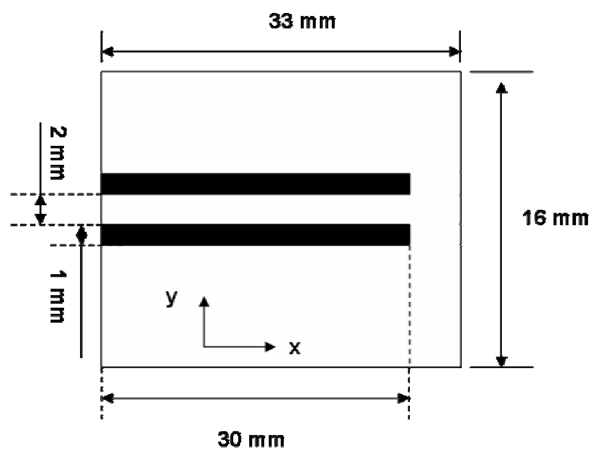
V. TRACKS ETCHED ON PRINTED CIRCUIT BOARDS

The loop antenna studied in Section I is associated with a section of a multi-conductor transmission line etched on the same substrate (Fig. 8). The characteristic impedance of the transmission line can be evaluated through a 2D cross-section analysis. So a global impedance of the whole structure (loop antenna + transmission line) can be determined at the input of the transmission line according to the rules of the transmission line theory. An approximate analytical expansion of including only two rational functions is also evaluated. A good agreement is obtained between the SPICE rational result and the two kinds of analytical solutions (Fig. 9). These solutions are compared to the whole 3D computation: it is worth noting that the SPICE result is closed to the global impedance deduced from the full wave 3D code ASERIS-BE (from EADS). This can be explained by the fact that the electromagnetic coupling between the loop antenna and the transmission line is weak in this case: the global behaviour of the structure roughly follows the theory of the transmission lines. This electromagnetic coupling is significant in the low frequency range: the global structure does not behave like a transmission line and a macromodelling SPICE simulation is no longer available in this range.

This methodology could provide an efficient way to simulate how conducted emissions can be induced along systems of transmissions lines by a perturbating electromagnetic field. The whole 3D transmission line system can be divided into sub-structures. Each part can be separately handled and the corresponding equivalent SPICE circuit can be deduced. As shown in the example above, in the high frequency case, all the parts can be connected together making available a global circuit model. So, provided that the spectrum of the illuminating exciting source is in a high frequency range, a SPICE simulation will allow giving how a conducted interference will be carried by the system of transmission lines.



a) Loop antenna.



b) Transmission line.

Fig. 8. Global 3D structure.

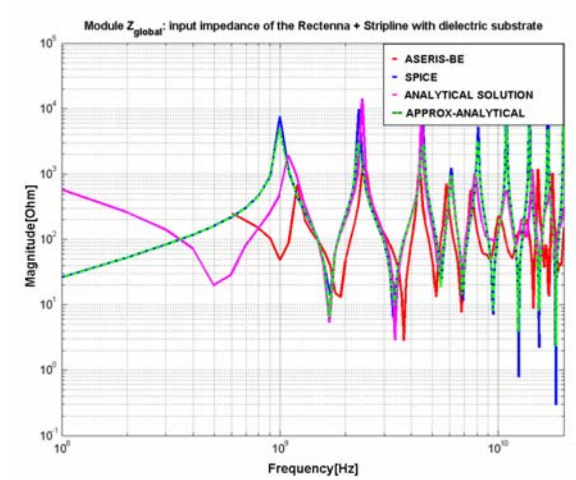


Fig. 9. Global impedance obtained with full wave computation, SPICE, and analytical expressions.

VI. CONCLUSION

A macromodelling approach was presented for EMC analysis of interconnected systems in the field of electromagnetic compatibility. In the methodology a broad-band characterization of distributed structures is performed with a finite element based approach. Then circuit models are deduced using a root mean square procedure. These circuit macromodels, SPICE compatible for example, provide a straightforward technique to simulate the propagation of parasitic signals along tracks and/or transmission lines. The procedure can be efficiently used in the simulation of time domain reflectometry for cable diagnosis.

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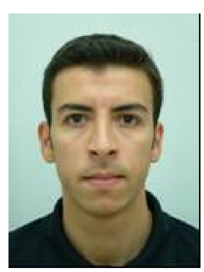

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