High-Frequency Full-Wave Analysis of Interconnects with Inhomogeneous Dielectrics through an Enhanced Transmission Line Model

A. G. Chiariello[1], A. Maffucci[2], G. Miano[1], F. Villone[2], and W. Zamboni[2]

[1] Ass. EURATOM/ENEA/CREATE, DIEL
Università degli Studi di Napoli "Federico II", I-80125 Napoli, Italy
{miano, a.chiariello}@unina.it

[2] Ass. EURATOM/ENEA/CREATE, DAEIMI
Università degli Studi di Cassino, I-03043 Cassino (FR), Italy
{maffucci, villone, zamboni}@unicas.it

Abstract – The paper deals with the inclusion of inhomogeneous dielectrics in a full-wave transmission line model for high-frequency analysis of interconnects. This "enhanced" transmission line model is derived from a full-wave integral formulation of the electromagnetic problem, and the inclusion of dielectrics is performed by an accurate semi-analytical evaluation of the Green functions for layered planar structures. The resulting model has a computational cost typical of a TL model but is able to perform a full-wave analysis in frequency ranges where the standard TL model may no longer be used. Moreover, as shown in the proposed examples, the model gives the possibility to investigate separately several phenomena affecting the high-frequency behavior of interconnects, like losses in dielectrics, unwanted radiation and excitation of parasitic modes.

***Keywords*:** High-speed interconnects, transmission line model, full-wave analysis, Green functions, and parasitic modes.

I. INTRODUCTION

Electrical interconnects in high-speed circuits are usually modeled by means of the popular *transmission line model*, which assumes a propagation of quasi-TEM mode type. This model has been thoroughly studied in the past and it has been proven to be described in a simple and accurate way, the effects of interconnects on the signal (delays, mismatching, crosstalk, …) [1]. However, in many cases of practical interest due to the ever-increasing operating frequencies and to size decreasing, the quasi-TEM hypothesis of propagation no longer holds. In such cases high-frequency effects arise, such as radiation, mode conversion and dispersion, which are crucial to correctly estimate the system performance. These effects are not included in the *standard transmission line* model (*STL*) and would require, in principle, a full-wave analysis. This kind of analysis has two disadvantages: a high computational cost and a poor qualitative insight on the solution. It is indeed difficult to distinguish between the different phenomena quoted above. To solve both problems, several efforts have been made to obtain *generalized* transmission line models able to overcome the validity limits of the *STL* model while retaining the same simplicity and a low computational cost (e.g., [2-5]).

The Authors have recently proposed an *enhanced transmission line* (*ETL*) model which is able to describe in the frequency domain interconnects for which the characteristic transverse dimension is comparable to the characteristic wavelength of the carried signals. This has been done for two-conductor interconnects in [6-8] and for a multiconductor interconnect in [9-10]. The model has been successfully used to foresee effects like radiation in the transverse plane, dispersion due to the finite length of the interconnect, differential to common mode conversion in asymmetric interconnects and high-frequency crosstalk. However in all these papers the embedding dielectric has been assumed uniform. In this paper the multiconductor *ETL* model is extended to inhomogeneous dielectrics, so that the analysis of the above mentioned high-frequency effect may be performed for interconnects of practical interest such as the microstrips.

Section II is devoted to the problem formulation in presence of inhomogeneous dielectrics. The starting point is the integral formulation of the electromagnetic problem based on the vector and scalar potentials satisfying the Lorenz gauge. The formulation involves the Green functions for the considered structure: a general case is considered, where an expression of the Green functions for dielectric layers with different permittivity and a ground plane is used. These Green functions are evaluated semi-analytically: the principal part describing the propagation of signals is extracted analytically, whereas the remainder describing parasitic modes due to the non-ideal behavior of the interconnect is represented through equivalent low-order systems identified by a vector fitting procedure [11].

In Section III the *ETL* model is derived with suitable approximations from this integral formulation. The *ETL* model has the same mathematical structure as the *STL* model, the only difference being in the relations between the per-unit-length (p.u.l.) magnetic flux and the current and between the voltage and the *p.u.l.* charge, respectively. In the *STL* model they are local, whereas in the *ETL* they involve spatial convolutions.

In Section IV first the *ETL* model predictions are successfully compared to the full-wave solutions obtained by two different 3D numerical codes. The case study highlights the inaccuracy of the *STL* model in high-frequency ranges. Then a deep investigation of the high-frequency solution is performed, by analyzing the effects of different phenomena like frequency-dependent dielectric losses and unwanted radiation in the transverse plane. A second case-study refers to a coupled microstrip, and is analyzed in order to evaluate the high-frequency crosstalk noise.

II. INTEGRAL FORMULATION AND THE INCLUSION OF INHOMOGENEOUS DIELECTRICS

Let us consider the interconnect of Fig. 1, of total length l, made by two signal conductors on a dielectric layer of thickness h and a ground plane. Let us assume the conductors to be ideal and the dielectric permittivity to be $\varepsilon_r \varepsilon_0$ in the layer and ε_0 outside it (the magnetic permeability is everywhere $\mu = \mu_0$). Let us denote with Σ_1 and Σ_2 the signal conductor surfaces.

In frequency domain we can express the fields in terms of the electrical scalar and magnetic vector potentials φ and $\mathbf{A}$ (*Lorenz* gauge),

$$\mathbf{E} = -j\omega\mathbf{A} - \nabla\varphi, \quad \mathbf{B} = \nabla \times \mathbf{A}\,. \tag{1}$$

It is convenient to express the vector and scalar potentials in terms of the current density $\mathbf{J}_s(\mathbf{r}_\perp, x)$ and charge density $\sigma(\mathbf{r}_\perp, x)$, through the integrals,

$$\varphi(\mathbf{r}) = \frac{1}{\varepsilon_0}\iint_S G_\varphi(\mathbf{r},\mathbf{r}')\sigma(\mathbf{r}')dS, \tag{2}$$

$$\mathbf{A}(\mathbf{r}) = \mu_0 \iint_S G_A(\mathbf{r},\mathbf{r}')\cdot\mathbf{J}_s(\mathbf{r}')dS \tag{3}$$

which involve the Green functions for the considered multilayered structure (including the ground plane).

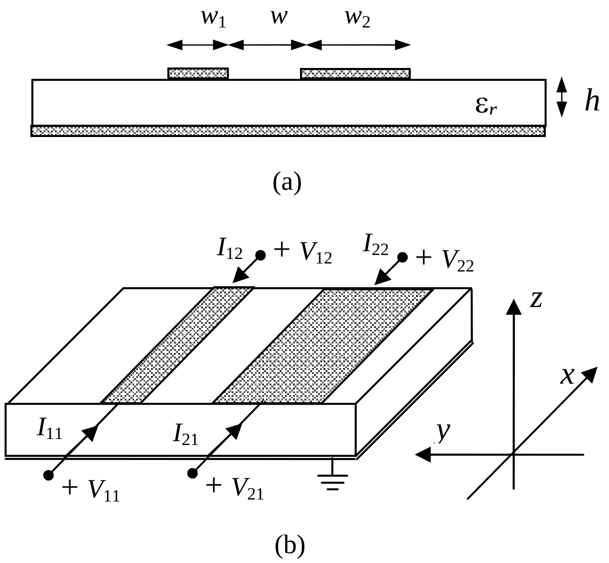


Fig. 1. The considered interconnect: (a) cross-section view; (b) adopted references for terminal voltages and currents .

The layers properties change along $\hat{z}$ (see Fig. 1), hence $G_A(r)$ has the structure [12],

$$G_A = \begin{bmatrix} G_{xx} & 0 & G_{xz} \\ 0 & G_{yy} & G_{yz} \\ G_{zx} & G_{zy} & G_{zz} \end{bmatrix}. \tag{4}$$

In many practical applications the thickness of conductors is small compared to their width w. If we consider zero-thickness for the signal conductors and assume the current density $\mathbf{J}_s$ directed along $\hat{x}$, we have the simple expression $G_A = G_{xx}$. We consider perfect conductors hence the sources are located on the surface $S = \Sigma_1 \bigcup \Sigma_2$ of the two conductors.

As for the dielectric, we can introduce frequency-dependent losses through a simple Debye model, assuming (e.g., [13]),

$$\varepsilon_r(\omega) = \varepsilon_\infty + \frac{\varepsilon_{DC} - \varepsilon_\infty}{1 + i\omega\tau}, \tag{5}$$

where ε_∞, ε_{DC} and τ are constant values associated to the particular dielectric chosen.

For the considered structure the Green functions may be evaluated in closed form in the spectral domain: let $\tilde{G}_{xx}(k_\rho)$ and $\tilde{G}_\varphi(k_\rho)$ be their transforms in such a domain, where k_ρ is the spectral domain variable. The spatial domain functions are obtained by evaluating the Sommerfeld integrals (e.g., [14]),

$$G_{xx}(r)=\frac{1}{4\pi}\int_{-\infty}^{+\infty}\tilde{G}_{xx}(k_\rho)H_0^{(2)}(k_\rho r)k_\rho dk_\rho , \qquad (6)$$

$$G_{\varphi}(r)=\frac{1}{4\pi}\int_{-\infty}^{+\infty}\tilde{G}_{\varphi}(k_\rho)H_0^{(2)}(k_\rho r)k_\rho dk_\rho , \qquad (7)$$

where $H_0^{(2)}$ is the Hankel function. The cost for computing such integrals is extremely high because of the slow decay of the integrands. A way to overcome this problem is to extract analytically the terms which are dominant in the low frequency range, referred to as the *quasi-static terms* [15],

$$G_{xx}^0(r)=\frac{e^{-ik_0\sqrt{x^2+y^2}}}{4\pi\sqrt{x^2+y^2}}-\frac{e^{-ik_0\sqrt{x^2+y^2+(2h)^2}}}{4\pi\sqrt{x^2+y^2+(2h)^2}}, \qquad (8)$$

$$G_{\varphi}^0(r)=(1+K)\frac{e^{-ik_0\sqrt{x^2+y^2}}}{4\pi\sqrt{x^2+y^2}}+ \\ +(K^2-1)\sum_{n=1}^{\infty}K^{n-1}\frac{e^{-ik_0\sqrt{x^2+y^2+(2nh)^2}}}{4\pi\sqrt{x^2+y^2+(2nh)^2}}, \qquad (9)$$

in which $K=(1-\varepsilon_r)/(1+\varepsilon_r)$ and $k_0=\omega\sqrt{\varepsilon_0\mu_0}$ is the vacuum space wavenumber. Once these terms have been extracted, the remainders (*dynamic terms*) may be evaluated in an efficient way by approximating the corresponding expressions in the spectral domain, for instance by using a vector fitting technique [11],

$$\tilde{G}_{Adyn}(k_\rho)=\sum_{n=1}^{N}\frac{b_n}{k_\rho^2-\alpha_n^2},\ \tilde{G}_{\varphi dyn}(k_\rho)=\sum_{n=1}^{N}\frac{d_n}{k_\rho^2-\beta_n^2}, \qquad (10)$$

so that the final expressions are given by

$$G_A^{xx}(r)=G_{A0}(r)-\frac{j}{4}\sum_{n=1}^{N}b_nH_0^{(2)}(\alpha_n r), \qquad (11)$$

$$G_{\varphi}(r)=G_{\varphi 0}(r)-\frac{j}{4}\sum_{n=1}^{N}d_nH_0^{(2)}(\beta_n r)\ . \qquad (12)$$

The quasi-static terms are associated to the fundamental mode, are the only terms left when $f\to 0$ and dominate the local range interactions. The dynamic terms are associated to parasitic waves (surface waves, leaky waves), vanish as $f\to 0$ and dominate the long-range interactions. Figure 2 gives an example of scalar potential Green function G_φ for a single microstrip with $\varepsilon_r=4.9$ and $h=0.7mm$. The quasi-static term dominates the near-field region, whereas for increasing distances the dynamic terms become the principal ones. Unless very high frequencies are considered, in practical interconnects the quasi-static terms are dominant, hence the approximation of the remainder is usually satisfactorily pursued by a low-order model. A reliable criterion [16] states that the Green functions are accurately represented by the quasi-static terms when $k_0h\sqrt{\varepsilon_r-1}<0.1$.

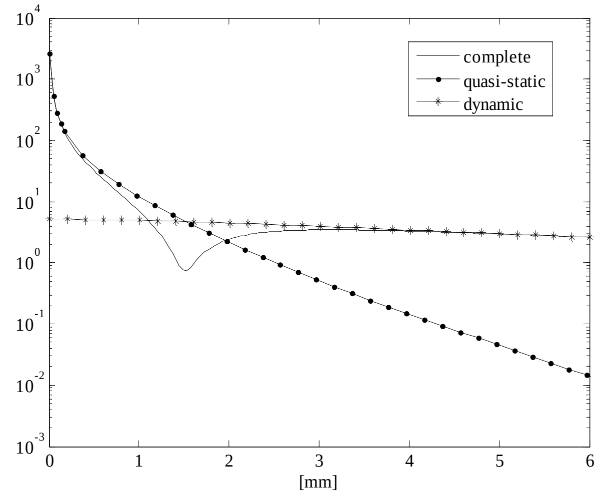


Fig. 2. Typical high-frequency behaviour of the scalar potential Green function: contributions of the quasi-static and dynamic terms.

III. THE ENHANCED TRANSMISSION LINE MODEL

In order to derive a transmission line model, let us impose the *charge conservation law*,

$$\nabla^{(s)}\cdot\mathbf{J}_s=-j\omega\sigma, \qquad (13)$$

where $\nabla^{(s)}$ is the *surface divergence operator*. In addition we must impose the *PEC boundary condition*,

$$\left(-j\omega\mathbf{A}-\nabla\varphi\right)\big|_S\times\hat{\mathbf{n}}=\mathbf{0}. \qquad (14)$$

Let us assume the dependence of the sources to be of separable type,

$$\sigma_1(s_1,x)=F_1(s_1)Q_1(x),\ \sigma_2(s_2,x)=-F_2(s_2)Q_2(x), \qquad (15)$$

$$J_{s1}(s_1,x)=F_1(s_1)I_1(x),\ J_{s2}(s_2,x)=-F_2(s_2)I_2(x) \qquad (16)$$

where $Q_{1,2}(x)$ and $I_{1,2}(x)$ are, respectively, the p.u.l. charges and the currents on the two signal conductors, $s_{1,2}$ are the curvilinear abscissas along the conductor contours Γ_1 and Γ_2, whereas the shape functions $F_{1,2}(s)$ describe the transverse distributions of the sources. Let a_n be a "characteristic dimension" of the cross-section of the *n*-th conductor, *i.e.* a characteristic distance between two points on the conductor contour. For instance a_n would be equal to the diameter for a circular cross-section. For this case we may assume $a_1 = w_1$ and $a_2 = w_2$ (Fig. 1). Next, let us introduce a "characteristic dimension" h_c of the transverse section. For a single trace microstrip we can assume $h_c = h$, where h is the dielectric thickness, whereas for the general case of coupled microstrips as in Fig.1 we may assume h_c as the mean value between *h* and the distance between the two traces *w*. Now, assuming $k_0 a_n << 1$ for any n and $h_c k_0 < 5$ it is possible to evaluate $F_{1,2}(s)$ once for all by solving a quasi-static problem in the transverse plane and to approximate at any abscissa *x* the values of $A(s_{1,2}, x)$ and $\varphi(s_{1,2}, x)$ on the surfaces Σ_1 and Σ_2 with their average values $< A_{1,2}(x) >$ and $< \varphi_{1,2}(x) >$ [7]. In all these conditions it is easy to derive from equations (13) and (14) the following *governing equations*,

$$\frac{d\mathbf{I}(x)}{dx} = -i\omega\mathbf{Q}(x), \quad \frac{d\mathbf{V}(x)}{dx} = -i\omega\mathbf{\Phi}(x), \tag{17}$$

$\mathbf{\Phi}(x)$ being the p.u.l. magnetic flux vector. In the same conditions, from equations (2) and (3) we derive the *constitutive relations*,

$$\mathbf{\Phi}(x) = \mu_0 \int_0^l H_I(x - x')\mathbf{I}(x')dx', \tag{18}$$

$$\mathbf{V}(x) = \frac{1}{\varepsilon_0} \int_0^l H_V(x - x')\mathbf{Q}(x')dx' \tag{19}$$

where the entries of the kernel H_I are given by,

$$H_I^{ik}(\zeta) = \frac{1}{c_i} \oint_{\Gamma_i} ds_i \oint_{\Gamma_k} G_{xx}(s_i, s_k'; \zeta) F_i(s_k') ds_k', \tag{20}$$

whereas H_V has the same expression involving G_φ. The kernels *H*(*x*) are computed numerically, paying attention to the logarithmic singularity of their diagonal terms, arising from the quasi-static parts of the Green functions [8-10].

The *ETL* model is given by equations (17) to (19) and is a generalization of the *STL* model, which is obtained when relations (18) and (19) are of *local type*. This happens for interconnects *electrically small* in the transverse plane ($h_C k_0 << 1$) and of infinite length along *x*. In this case the kernels in equations (18) and (19) tend to spatial Dirac pulses,

$$H_I(x) \to [\int_0^{+l} H_I(x')dx']\delta(x) = H_{I0}\delta(x), \tag{21}$$

$$H_V(x) \to [\int_0^{+l} H_V(x')dx']\delta(x) = H_{V0}\delta(x). \tag{22}$$

Hence equations (18) and (19) reduce to

$$\mathbf{\Phi}(x) = \mu_0 H_{I0}\mathbf{I}(x) = L\mathbf{I}(x), \tag{23}$$

$$\mathbf{V}(x) = \frac{1}{\varepsilon_0} H_{V0}\mathbf{Q}(x) = C^{-1}\mathbf{Q}(x). \tag{24}$$

Note that for homogeneous dielectrics it is $H_I = \varepsilon_r H_V$, hence $H_{I0} = \varepsilon_r H_{V0}$, and equations (23) and (24) yield the classical result $LC = \varepsilon_r \varepsilon_0 \mu_0 I$. By combining equations (23), (24), and (17) we obtain the *standard transmission line model*,

$$\frac{d\mathbf{I}(x)}{dx} = -i\omega C\mathbf{V}(x), \quad \frac{d\mathbf{V}(x)}{dx} = -i\omega L\mathbf{I}(x) \tag{25}$$

In conclusion, the *ETL* model (17) to (19) generalizes the *STL* one (*i.e.*, equation (25)), removing the assumptions that the transverse characteristic dimension of the interconnect is electrically small and that the interconnect is infinite. The *ETL* model is valid in the following limits: (i) the characteristic dimensions of the terminal devices are small compared to the interconnect length; (ii) the characteristic transverse dimension *a* of the conductors is electrically small, $k_0 a_n << 1$; and (iii) $h_c k_0 < 5$, where h_c is a characteristic dimension in the transverse plane.

IV. NUMERICAL RESULTS

The first case considered refers to a PCB microstrip, with the geometry of Fig. 1, assuming a single signal conductor above a ground plane and a length of 36 mm. The signal conductor has zero thickness, width $w_1 = 1.8$ mm and lies on a FR-4 dielectric layer of thickness $h = 1.016$ mm, dielectric constant $\varepsilon_r = 4.9$ and magnetic permeability $\mu = \mu_0$. The conductors and dielectric are assumed ideal.

The *ETL* model solution is compared to the *STL* one and to two 3D full-wave solutions; one provided by the commercial FEM code *HFSS* [17] and the other by the tool *SURFCODE*, which is based on the *Electric Field Integral Equation* formulation [18]. Assuming for this case $h_c = h$, since $\varepsilon_{reff} \approx 3.65$ we have $k_0 h_c \approx 0.1$ at 1.4 GHz, hence we expect the *STL*, *ETL* and full-wave solutions to agree up to frequencies around 1 GHz. This is clearly put on evidence in Fig. 3, where the input impedance of the line left open at the far end is plotted from DC to 0.8 GHz. For higher frequencies, the *STL* solution becomes inadequate, whereas the *ETL* one is still able to reproduce the full-wave solution, as shown in Fig. 4.

Since the conductors and the dielectric are assumed to be ideal, the finite amplitude of the peak is only due to the lossy effects related to the presence of unwanted parasitic modes (surface waves, leaky waves). In this condition a small but non-negligible amount of power is associated with radiation in the transverse plane. The real power absorbed by the interconnect fed at one end by a sinusoidal current of *rms* value I_0 and left open at the other end is given by,

$$P_{in}(\omega) = real\{Z_{in}(\omega)\} I_0^2 / 2. \quad (26)$$

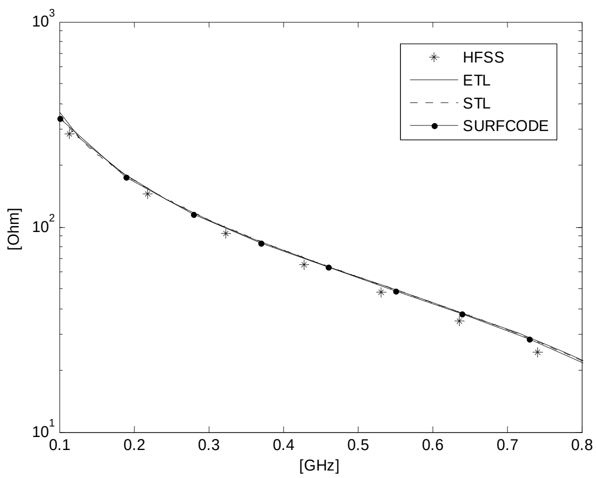


Fig. 3. Low frequency behaviour of the absolute value of the input impedance, Case 1.

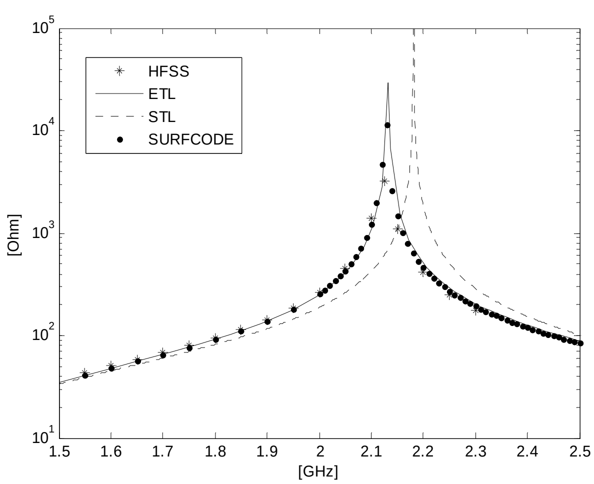


Fig. 4. High frequency behaviour of the absolute value of the input impedance, Case 1.

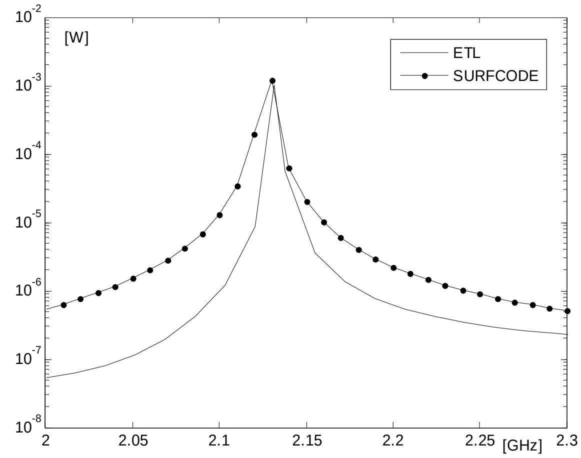


Fig. 5. Absorbed power for ideal dielectric, Case 1.

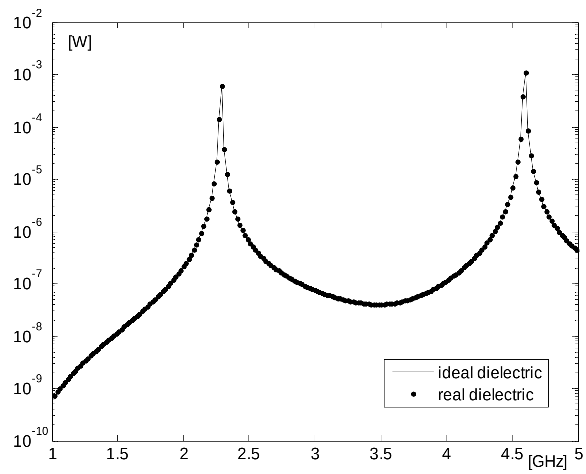


Fig. 6. ETL solution for absorbed power with ideal and real dielectric, Case 1.

In Fig. 5 it is shown the absorbed real power computed with $I_0 = 1mA$. The *ETL* solution is in good agreement with the full-wave one around the peak, whereas there is a deviation in the other ranges (however, the values of power are very low). Note that, since we are in the ideal case, the *STL* input impedance is strictly imaginary; hence the absorbed real power predicted by the *STL* model is always zero.

Let us now introduce a lossy dielectric described by the Debye model of equation (5), with $\varepsilon_\infty = 4.07, \varepsilon_{DC} = 4.178$, and $\tau = 1.15$ ps. Figure 6 shows the dissipated power computed in the same conditions described for Fig. 5, both considering a real and an ideal dielectric with $\varepsilon_r = \varepsilon_{DC} = 4.178$. It is clear that in this case the dielectric losses are negligible with respect to the losses associated to the other high-frequency phenomena.

The *ETL* model may be used to perform qualitative analysis on the solution. For instance, it is possible to distinguish between the high-frequency effects associated to the fundamental mode from those related to the excitation of parasitic modes. As shown in Section II, this could be easily done by switching on and off the contribution of the dynamic terms in the expression of the Green functions. Figure 7 shows, for instance, the mutual impedance of the above-considered line, computed by the *STL* model and by the *ETL* one, with or without the contribution of the dynamic terms. For this case, the quasi-static term is the only relevant, even for frequencies up to 7 GHz.

A second example is provided by a coupled microstrip made by two signal conductor and a ground plane. In this case (see Fig. 1) we have considered $w_1 = w_2 = w = 1.8\,\text{mm}$, $h = 1\,\text{mm}$, $\varepsilon_r = 4.17$ and a total length of 36 mm. The line behavior is investigated in the frequency range $(0-6)$ GHz, so extending to values of $k_0 h_c$ high enough to expect inaccurate results from the *STL* model. Here a crosstalk analysis has been performed, by assuming line 1 (see references in Fig. 1) to be fed at the near end by a unitary sinusoidal voltage source and open at the far end. The near and far ends of line 2 are both terminated on open circuits. Figure 8 shows the frequency behaviour of the near and far end crosstalk voltage defined as V_{21}/V_{11} and V_{22}/V_{11}, respectively. Note that in this case-study we have approximated the complete Green functions, considering only the contribution of the quasi-static term, which is again the dominant one.

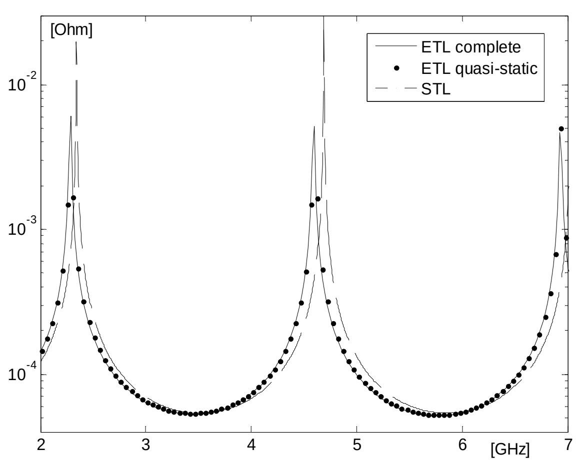
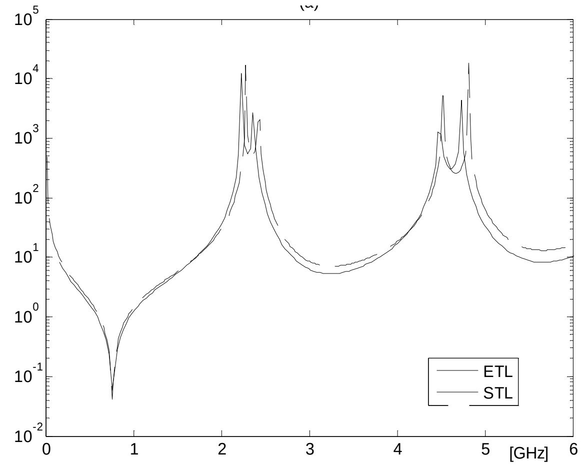
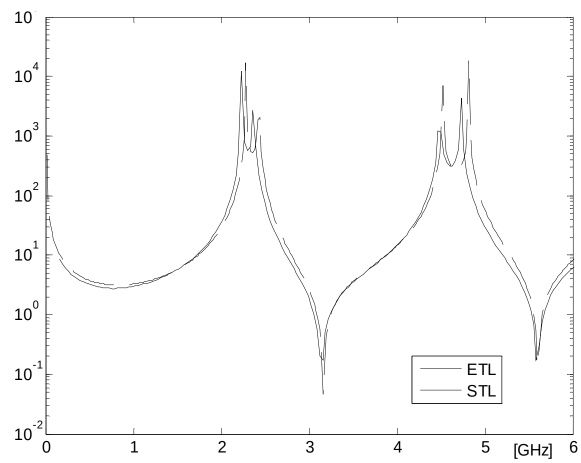


Fig. 7. Absolute value of the mutual impedance, Case 1.



(a)



(b)

Fig. 8. Near-end (a) and far-end (b) crosstalk voltage normalized to the input voltage, Case 2.

V. CONCLUSIONS

The extension to inhomogeneous dielectrics of a full-wave transmission line model is obtained by including in the integral formulation a semi-analytical expression of the Green function for planar layered interconnects. Case-studies show the reliability of the model, as compared to 3D full-wave numerical solutions. The model is able to foresee high-frequency effects like radiation and dispersion due to excitation of unwanted parasitic modes. The way used to include inhomogeneous dielectrics into such a formulation is promising, since it is possible to split the Green function in terms describing the signal propagation (evaluated analytically) and remainders associated to the unwanted parasitic modes. This allows an accurate evaluation of the influence of such unwanted modes, exploiting the possibility to switch on and off the corresponding terms in the Green function expression.

VI. ACKNOWLEDGEMENTS

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Andrea G. Chiariello received in 2004 the Laurea Degree in Electronic Engineering *summa cum laude* and from the 2004 he is a Ph.D. student in Electrical Engineering from the University of Naples "Federico II" , Italy.

In 2006 he was a visiting scholar at the electrical and computer engineering of the University of Illinois, Urbana Champain, USA. His research interests are the electromagnetic modeling of high-speed interconnects and the electromagnetic compatibility.

Antonio Maffucci received in 1996 the Laurea Degree in Electronic Engineering *summa cum laude* and in 2000 the Ph.D. degree in Electrical Engineering from the University of Naples "Federico II", Italy.

In 1997 he was with the nuclear fusion laboratory JET (Culham, U.K.). From 1998 to 2002 he was with the Department of Electrical Engineering at the University of Naples. From 2002 he is with the Elettrotecnica Group at the Faculty of Engineering of the University of Cassino, Italy, where he is currently an Associate Professor, teaching courses on circuit theory and electromagnetic compatibility.

His research areas include electromagnetic modeling of distributed systems, electromagnetic compatibility, computational electromagnetism, and nanotechnology.

He is author of more than 80 technical papers on international journals, conference proceedings and essays on books. He is also co-author of the book *Transmission lines and lumped circuits* (New York, Academic Press, 2001). He is Associate Editor of the *IEEE Transactions on Advanced Packaging*.

Giovanni Miano received the Laurea (*summa cum laude*) and Ph.D. degrees in electrical engineering from the University of Naples Federico II in 1983 and 1989, respectively.

From 1984 to 1985, he was with the PS Division, CERN, Geneva, Switzerland, where he was engaged in research on magnetic plasma lenses. From 1989 he Full Professor at the faculty of Engineering of the University of Naples Federico II, where he served as a Researcher until 1992 and as Associate Professor from 1992 to 2000. From 2005 he is the Director of the Dipartimento di Ingegneria Elettrica.

In 1996 he was a Visiting Scientist at the GSI Laboratories, Darmstadt, Germany, and in 1999 a Visiting Professor at the Department of Electrical Engineering, University of Maryland, College Park.

His research interests include ferromagnetic materials, nonlinear dielectrics, plasmas, electrodynamics of continuum media, nanotechnology and modeling of lumped and distributed circuits.

He is the author or co-author of more than 70 papers published in international journals, 60 papers published in international conference proceedings, two items in the *Wiley Encyclopedia of Electrical and Electronic Engineering* (New York: Wiley, 1999), and the monograph *Transmission Lines and Lumped Circuits* (New York: Academic, 2001). He is a Reviewer for various journals of the IEEE and past member of the Editorial Boards of the Conference on Electromagnetic Field Computation and the Conference on the Computation of Electromagnetic Field

Fabio Villone received the Laurea degree (summa cum laude) in Electronic Engineering in 1994 at the University of Naples Federico II, and the Ph.D. in Industrial Engineering in 1998 at the University of Cassino.

Currently, he is Full Professor at the Faculty of Engineering of the University of Cassino, where he formerly was Research Assistant (1997-2001) and Associate Professor (2001-2007). He teaches "Basic Electrical Engineering" and "Numerical models for electromagnetic fields and circuits".

In 1996 he was visiting scientist at CRPP-Lausanne (Switzerland), and in the last years at JET (Joint European Torus), close to Oxford (UK).

His scientific interests are in the field of computational electromagnetics, with particular reference to fusion plasma modelling and engineering, electromagnetic compatibility and eddy current non destructive testing.

He is co-author of more than 60 papers on international journals and essays on books, and more than 50 contributions to international conferences, among which several invited papers.

He was Scientific Coordinator of several experiments carried out at JET, and Principal Investigator of a national research project on fusion plasma modelling and control.

Walter Zamboni received the Laurea degree in Electrical Engineering (*summa cum laude*) in 2001 and the Ph. D. degree in Electrical Engineering in 2004 from the Faculty of Engineering, Università degli Studi di Napoli “Federico II”, Italy. In 2003 he has been Stager at CERN, Geneva, Switzerland, and Visiting Scientist at EPFL-CRPP Superconductivity Group, Villigen-PSI, Switzerland. From 2005 he is with the the Faculty of Engineering, Università degli Studi di Cassino, Italy, where he works as a research assistant. From 2006 he teaches Circuits Theory at the Università degli Studi del Sannio, Italy

His scientific interests are in the field of computational electromagnetics, with particular reference to interconnects and applied superconductivity.