ON-CHIP INTERCONNECTS: NEW ACCURATE NOMINAL AND PARAMETRIZED MODELS

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Această lucrare descrie tehnici specializate de extragere a modelelor nominale şi parametrice pentru interconexiunile lungi descompuse în linii drepte, modelate ca linii de transmisie şi în componente de joncţiune, modelate ca dispozitive pasive. Un pas important în modelare este extragerea parametrilor lineici pentru linia de transmisie. Este prezentată o nouă abordare pentru calculul conductanţei şi capacităţii lineice. Noutatea studiului este parametrizarea atât în funcţie de dimensiunile geometrice cât şi de frecvenţă. Parametrizarea în funcţie de geometrie este bazată pe calculul sensitivităţilor de ordinul întâi din modele de câmp electromagnetic, în timp ce influenţa frecvenţei este aproximată prin polinoame raţionale obţinute prin fitting. Abordarea propusă este validată prin comparaţie cu experimentele.

This paper describes specialized techniques to extract nominal and parametric models for long interconnects decomposed in straight parts, modeled as transmission lines and in junction components, modeled as passive components. An important step in modeling is the extraction of the per unit length parameters for the transmission line. A new approach to compute line conductance and capacitance is presented. The novelty of the study is the parameterization with respect both to the geometric parameters and the frequency. The parameterization with respect to the geometry is based on the computation of first order sensitivities from electromagnetic field models, whereas the influence of the frequency is approximated by rational polynomials obtained by fitting. The approach proposed is validated by comparing with experiments.

Keywords: variability, litography, interconnects, transmission lines

1. Introduction

Much research is focusing on interconnects as their performances impact has become important due to the fact that million closely spaced interconnections in one or more levels connect various components on the integrated circuit [1]. In general, if on-chip interconnects are sorted with respect to their electric length, they may be categorized in three classes: short, medium and long. While the short

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interconnects have simple circuit models with lumped parameters, the extracted model of the interconnects longer than the wave length has to consider also the effect of the distributed parameters. Fortunately, the long interconnects have usually the same cross-sectional geometry along their extension. If not, they may be decomposed in straight parts connected by junction components (Fig.1). The former are represented as transmission lines (TLs) whereas the latter are modeled as common passive 3D components. For relatively low frequencies, adverse side-effects, such as parasitic effects, are neglectable. But at high frequencies, these effects must be included. So, standard computational modeling methods as FIT (Finite Integral Technique) or FEM (Finite Element Method) used to simulate electromagnetic field, are not sufficient. In order to cope with the new challenges approaches, very often extensions or modifications of classical methods are used.

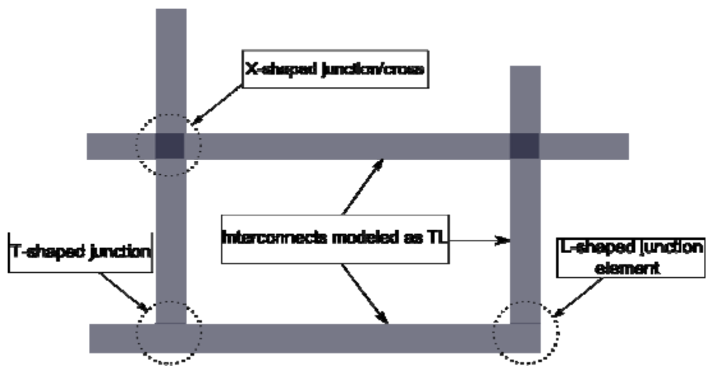


Fig.1 Decomposition of the interconnect net in 2D TLs and 3D junctions

Manufacturing variability in the fabrication process of the ICs is gaining more attention as technology dimensions become smaller and the operation frequencies continue to go up. Such variations, which are hard to predict and to control, may have an important effect on the functionality of the design or on the accuracy of the resulting device. The parameter variability can no longer be disregarded during modeling, simulation and verification of the device. Parasitic capacitances, resistances and inductances of the interconnects have become major factors in the evolution of very high speed IC technology. The subject of this paper is how nominal and parametric models for interconnects modeled at high frequencies can be extracted in a fast and robust manner. First, the extraction of the per unit length parameters for transmission lines is presented, then a new Modified Analytical-Numerical Two Fields Approach for admittance computation is discussed. In the third part we present how sensitivities are extracted and then the parametric models based on these sensitivities are considered. The authors investigate promising alternatives beside the classic models of first-order truncations of Taylor expansions. Two types of parametric models are developed: one for geometrical dimensions variations and the other for variations w.r.t frequency. The results presented validate all these approaches and conclusions are drawn at the end.

2. Extraction of line parameters

Models with various degrees of fineness can be established for TLs. The coarsest ones are circuit models with lumped parameters, such as the Π equivalent circuit for a single TL shown in Fig. 2. The values of the parameters can be roughly estimated either starting from the geometry data by field solution, or from measurements, if available. As expected, the characteristic of such a circuit is appropriate only at low frequencies, over a limited range, and for short lines.

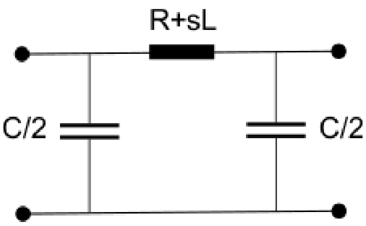


Fig. 2. The coarsest model for a single transmission line: a pi equivalent circuit

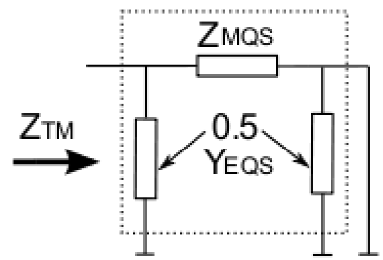


Fig. 3. The pi equivalent circuit for a simulated short line segment. Parameters are evaluated from field simulations.

At high frequencies, the distributed effects have to be considered as an important component of the model. Proper values for the line parameters can be obtained only by simulating the electromagnetic (EM) field. The extraction of line parameters is the main step in TLs modeling since the behavior of a line of a given length can be computed from them. For instance, for a multiconductor transmission line, from the line parameters matrices $\mathbf{R}$, $\mathbf{L}$, $\mathbf{C}$ and $\mathbf{G}$ the transfer matrix can be computed as $\mathbf{T} = \exp(\mathbf{D} + j\omega\mathbf{E})$, where $\mathbf{D} = \begin{bmatrix} \mathbf{0} & -\mathbf{R} \\ -\mathbf{G} & \mathbf{0} \end{bmatrix}$ and $\mathbf{E} = \begin{bmatrix} \mathbf{0} & -\mathbf{L} \\ -\mathbf{C} & \mathbf{0} \end{bmatrix}$.
From them, other parameters (impedance, admittance or scattering) can be computed as shown for instance in [2]. When considering *geometric data*, the simplest model may consider uniform fields in steady-state electric conduction (EC), electrostatics (ES) and magnetostatics (MS) to asses the line resistance, capacitance and inductance, respectively. Empirical formulas may also be found in the literature, such as the ones given in [3, 4] for the line capacitance. None of them take the frequency dependence into account. A first attempt to take into consideration the frequency effect is to compute the skin depth in the conductor and to use a better approximation for the resistance. In [5] we proposed a much more

accurate estimation based on the numerical modeling of the EM field. Two complementary problems are solved, one which describes the transversal behavior of the line from which the line admittance $\mathbf{Y}(\omega) = \mathbf{G}(\omega) + j\omega\mathbf{C}(\omega)$ is extracted and a second one which describes the longitudinal behavior of the line and from which the line impedance $\mathbf{Z}(\omega) = \mathbf{R}(\omega) + j\omega\mathbf{L}(\omega)$ is extracted. The first problem is dedicated to the computation of the transversal parameters and it uses a 2D transversal electro-quasi-static (EQS) field in dielectrics, considering the line wires as perfect conductor with given voltage. The second problem focuses on the longitudinal electric and the generated transversal magnetic field. Consequently, a short line-segment of length l is considered in which a full-wave (FW) but transversal magnetic (TM) field approximation is used. The transversal component is finally subtracted from the FW-TM simulation to obtain an accurate approximation of the line impedance, as given by

$$\mathbf{Z}_{MQS} = \left(\mathbf{Z}_{TM}^{-1} - \frac{1}{2}\mathbf{Y}_{EQS} \right)^{-1} \tag{1}$$

This subtraction is carried out according to a pi-like equivalent net for the simulated short segment (Fig. 3). Finally, the line parameters are:

$$\mathbf{G}(\omega) = \mathrm{Re}(\mathbf{Y}), \quad \mathbf{C}(\omega) = \mathrm{Im}(\mathbf{Y})/\omega, \quad \mathbf{R}(\omega) = \mathrm{Re}(\mathbf{Z})$$
$$\mathbf{L}(\omega) = \mathrm{Im}(\mathbf{Z})/\omega \tag{2}$$

where

$$\mathbf{Y} = \mathbf{Y}_{EQS}/l, \qquad \mathbf{Z} = \mathbf{Z}_{MQS}/l \tag{3}$$

The obtained values of the line parameters are frequency dependent.

If *measurements* are available, an estimation of the line parameters at low frequencies can be done by considering the simplest Π equivalent lumped circuit (Fig.2) and extrapolating experimental data towards zero frequency. A more accurate estimation can be done if TL theory is used. In the case of single TL, it can be easily derived that for every frequency the line parameters can be computed as $R = real(\gamma Z_c)$, $G = real(\gamma / Z_c)$, $L = imag(\gamma Z_c)/\omega$, $C = imag(\gamma / Z_c)/\omega$, where the complex propagation constant can be computed from the components of the impedance matrix as $\gamma = \arg\cosh(Z_{11}/Z_{12})/l$ and the complex characteristic impedance can be computed as $Z_c = Z_{12}\sinh(\gamma l)$. There are difficulties related to the fact that the $\arg\cosh$ function is multi-valued, but these can be overcome in a correction step, as described in [5]. The obtained values of the line parameters are frequency dependent as well.

3. Modified Analytical-Numerical Two Fields Approach

The starting point of this method is the standard two field problems approach presented above. Based on the results obtained, we have observed that the line capacitance is approximately constant w.r.t. frequency. Only at high frequencies, close to 60GHz its value slightly decreases. Analyzing the p.u.l. conductance graphs we have observed that its value is constant up to 10 GHz but then it grows rapidly. The entire course of the curve is a standard second degree curve. The modified analytical-numerical two field problems approach is based on these two observations. The main difference, comparing to the previous, standard method is that the admittance:

$$\mathbf{Y}_{lA-N}(\omega)=\mathbf{G}_{lA-N}(\omega)+j\omega\mathbf{C}_{lA-N}(\omega) \quad (4)$$

is computed not from 2D EQS simulations, but from analytical expressions, whereas the impedance is calculated as in the previous method. To designate the p.u.l. C value empirical expressions were used. Afterwards, the results from these formulas were compared with simulation data. For further applications, expression that fits the best the simulation data was used. In order to designate the p.u.l. conductance G, three methods were used and compared with the simulation results. The first method is based on the fact that the relation between frequency and p.u.l. G can be described as a second order polynomial. The quadratic polynomial coefficients are obtained through a fitting procedure. The polynomial has the following form:

$$G_{lA-N}(f)=a_1 f^2+a_2 f-a_3 \quad (5)$$

The second method uses the transfer function *H(s)* obtained from the Vector Fitting procedure. The real part of the obtained transfer function is p.u.l. conductance.

The third method requires as in the first case the calculation of the coefficients of the second order polynomial. They were obtained with Matlab *cftool* procedure.

The three methods were compared and the best results are obtained for the third method. So far, the p.u.l. G analysis was limited to constant values of geometric dimensions. The purpose is to develop a method with universal character and to include also the variations of the geometric dimensions. So, the line conductance will be then computed w.r.t. frequency and geometric variations:

$$G(f,\alpha)=(b_1\alpha+b_2)f^2+(b_3\alpha^2+b_4\alpha+b_5)f+(-b_6\alpha^2-b_7\alpha-b_8) \quad (6)$$

The coefficients have been computed using *cftool* from Matlab.

As an alternative to this analytical parametric formulation, we developed a method based on Taylor series expansion for the quantity that varies.

4. Computation of sensitivities for per unit length parameters

The process uncertainty usually directly affects the geometrical or electrical properties of the layout, and therefore, most of these variations can be represented as modifications of the values of the system matrices inside a state space descriptor:

$$\mathbf{C}(\alpha)\frac{d\mathbf{x}(\alpha)}{dt} + \mathbf{G}(\alpha)\mathbf{x}(\alpha) = \mathbf{Bu} \tag{7}$$

$$\mathbf{y}(\alpha) = \mathbf{Lx}(\alpha) \tag{8}$$

Parametric models are often obtained by truncating the Taylor series expansion for the quantity of interest. This requires the computation of the derivatives of the device characteristics with respect to the design parameters [6]. Let us assume that $y(\alpha_1, \alpha_2, \ldots, \alpha_n) = y(\alpha)$ is the device characteristic which depends on the design parameters $\alpha = [\alpha_1, \alpha_2, \ldots, \alpha_n]$. The quantity y may be, for instance the real or the imaginary part of the device admittance at a given frequency. In our case this quantity is any of the p.u.l. parameters. The parameter variability is thus completely described by the real function, y, defined over the design space S, a subset of $\Re^n$ The nominal design parameters correspond to the particular choice $\alpha_0 = [\alpha_{01}, \alpha_{02}, \ldots, \alpha_{0n}]$. First order truncation of the Taylor series is the affine function:

$$F(\alpha_1, \alpha_2, \ldots, \alpha_n) = F(\alpha_{01}, \alpha_{02}, \ldots, \alpha_{0n}) + S_{\alpha 1}(\alpha_1 - \alpha_{01}) + \ldots + S_{\alpha n}(\alpha_n - \alpha_{0n}) \tag{9}$$

where $S_{\alpha k} = \partial F / \partial \alpha_k$ are the first order sensitivities defined as partial derivatives of the device characteristic w.r.t. design parameters, computed for the nominal values of the parameters. This definition is available not only for the real part of the characteristic, but also when F is a complex number, a vector or a matrix.

The next level of the approximation in the modeling process is the computation of the first order sensitivities of the output quantity from the sensitivities of the state space matrices:

$$\frac{\partial \mathbf{y}}{\partial \alpha} = \mathbf{L}\frac{\partial \mathbf{x}}{\partial \alpha} \tag{10}$$

where

$$\frac{\partial \mathbf{x}}{\partial \alpha} = -(j\omega\mathbf{C} + \mathbf{G})^{-1}\left[\left(j\omega\frac{\partial \mathbf{C}}{\partial \alpha} + \frac{\partial \mathbf{G}}{\partial \alpha}\right)\mathbf{x}\right] \tag{11}$$

$$\mathbf{x} = \left(j\omega\mathbf{C} + \mathbf{G}\right)^{-1}\mathbf{B}u \tag{12}$$

Sensitivities of the p.u.l. parameters can be expressed as follows:

$$\frac{\partial \mathbf{G}_l}{\partial \alpha} = \frac{1}{l}\mathrm{Re}\frac{\partial \mathbf{Y}_{EQS}}{\partial \alpha} \tag{13}$$

$$\frac{\partial \mathbf{C}_l}{\partial \alpha} = \frac{1}{l}\mathrm{Im}\frac{\partial \mathbf{Y}_{EQS}}{\partial \alpha} \tag{14}$$

$$\frac{\partial \mathbf{R}_l}{\partial \alpha} = \frac{1}{l}\mathrm{Re}\frac{\partial \mathbf{Z}_{MQS}}{\partial \alpha} \tag{15}$$

$$\frac{\partial \mathbf{L}_l}{\partial \alpha} = \frac{1}{\omega l}\mathrm{Im}\frac{\partial \mathbf{Z}_{MQS}}{\partial \alpha} \tag{16}$$

5. Geometrical and frequency dependent parametric models

In this section we discuss the alternative models developed for geometrical variations and then we insert the frequency dependence in the parametric models. The advantage of these models is that they don't require additional iterations. We define two types of parametric models: additive (A) and rational (R) [7].

The *additive model* is a simply first order normalized standard version of the truncated Taylor expansion

$$\hat{y}_A(\alpha) = y_0\left(1 + \sum_{k=1}^{n} S_{\alpha_k}^{y}\,\delta\alpha_k\right), \tag{17}$$

where we denote by $\frac{\partial y}{\partial \alpha_k}(\alpha_0)\frac{\alpha_{0k}}{y_0} = S_{\alpha_k}^{y}$ the relative sensitivities with respect to each parameter and by $(\alpha_k - \alpha_{0k})/\alpha_{0k} = \delta\alpha_k$ the relative variations of the parameters. The *rational model* is the additive model for the reverse quantity $1/y$. It is obtained from the first order truncation of the Taylor Series expansion for the function $1/y$. In the general case, the rational model is:

$$\hat{y}_R(\alpha) = \frac{y_0}{1 + \sum_{k=1}^{n} S_{\alpha_k}^{1/y}\delta\alpha_k} \tag{18}$$

where it can be easily shown that $S_{\alpha}^{1/y} = -S_{\alpha}^{y}$.

The first idea to include the frequency dependence in these models is to consider the variation with respect to the frequency of the nominal values and of the sensitivities:

$$\hat{y}_A(s,\alpha) = y_0(s)\left(1 + \sum_{k=1}^{n} S_{\alpha_k}^{y}(s)\delta\alpha_k\right), \hat{y}_R(s,\alpha) = \frac{y_0(s)}{1 + \sum_{k=1}^{n} S_{\alpha_k}^{1/y}(s)\delta\alpha_k} \tag{19}$$

The implementation of formulas (19) in a computer code is straightforward. However, this approach is not appropriate if the final goal is to obtain a synthesized small circuit with parameterized values of components.

The alternative we propose to obtain a frequency dependent parametric model is to use a rational approximation in the frequency domain. We have shown in [8] that the most efficient method for the class of problems we address is the vector fitting method proposed in [9] and improved in [10,11], which finds the transfer function matching a given frequency characteristic. The resulting approximation has guaranteed stable poles and the passivity can be enforced in a post-processing step [11]. Thus, in the frequency domain, for the output quantity *y(s)*, this procedure finds the poles p_m (real or complex conjugate pairs), the residuals k_m and the constant terms k_∞ and k_0 of a rational approximation $\hat{y}(s)$ of the admittance:

$$y(s) \cong \hat{y}_{VFIT}(s) = \sum_{m=1}^{q} \frac{k_m}{s - p_m} + k_\infty + s k_0 \tag{20}$$

To keep the explanations simple, we assume that there is only one parameter that varies, i.e. the quantity α is a scalar. Assuming that keeping the order q is satisfactory for the whole range of the variation of this parameter, this means that (20) can be parameterized as:

$$y(s,\alpha) \cong \hat{y}_{VFIT}(s,\alpha) = \sum_{m=1}^{q} \frac{k_m(\alpha)}{s - p_m(\alpha)} + k_\infty(\alpha) + s k_0(\alpha) \tag{21}$$

Without loss of generality, we can assume that the additive model is more accurate than the rational one. If not, the reverse quantity is used, which is equivalent, for our class of problems, to change the excitation of terminals from voltage excited to current excited, and use an additive model for the impedance $z = y^{-1}$. The additive model (17) can be written as

$$y(s,\alpha) \cong y_A(s,\alpha) = y(s,\alpha_0) + \frac{\partial y}{\partial \alpha}(s,\alpha_0)(\alpha - \alpha_0) \tag{22}$$

where here y is a matrix function (e.g. for a single TL, it is a 2x2 matrix).

By combining (21) and (22) we obtain an approximate additive model based on VFIT:

$$y(s,\alpha) \cong \hat{y}_{A-VFIT}(s,\alpha) = \hat{y}_{VFIT}(s,\alpha_0) + \frac{\partial \hat{y}_{VFIT}}{\partial \alpha}(s,\alpha_0)(\alpha - \alpha_0) \tag{23}$$

From (21) it follows that the sensitivity of the VFIT approximation needed in (23) is

$$\frac{\partial \hat{y}_{VFIT}}{\partial \alpha} = \sum_{m=1}^{q} \left[\frac{\partial k_m / \partial \alpha}{s - p_m} + \frac{k_m}{(s - p_m)^2} \frac{\partial p_m}{\partial \alpha} \right] + \frac{\partial k_\infty}{\partial \alpha} + s \frac{\partial k_0}{\partial \alpha} \tag{24}$$

The sensitivity $\partial y / \partial \alpha$ are computed as described in section 4 for as many frequencies as required and thus the sensitivities of poles and residues in (24) can be computed solving the linear system (24) by least square approximation. Finally, by substituting (24) and (21) in (23), the final parameterized and frequency dependent model is obtained:

$$\hat{y}_{A-VFIT}(s,\alpha)=\sum_{m=1}^{q}\left[\frac{k_m+(\alpha-\alpha_0)\partial k_m/\partial\alpha}{s-p_m}\right]+(\alpha-\alpha_0)\sum_{m=1}^{q}\left[\frac{k_m}{(s-p_m)^2}\frac{\partial p_m}{\partial\alpha}\right]+$$
$$+\left[k_\infty+(\alpha-\alpha_0)\frac{\partial k_\infty}{\partial\alpha}\right]+s\left[k_0+(\alpha-\alpha_0)\frac{\partial k_0}{\partial\alpha}\right] \quad (25)$$

Expression (25) has the advantage that it has an explicit dependence with respect both to the frequency $s = j\omega$ and parameter α, is easy to implement and feasible to be synthesized as a second order net-list having components with dependent parameters.

6. Results

6.1. Results obtained for the nominal models

Two class of problems have been tested. First, a **transmission line** having the configuration shown in Fig. 4. The geometrical and electrical characteristics of the problem are h1= 1µm, h2 = 0.69 µm, h3 = 10 µm, a = 130.5 µm, p3 = 3 µm, p1 = 0, p2 = h2, xmax = 264 µm.

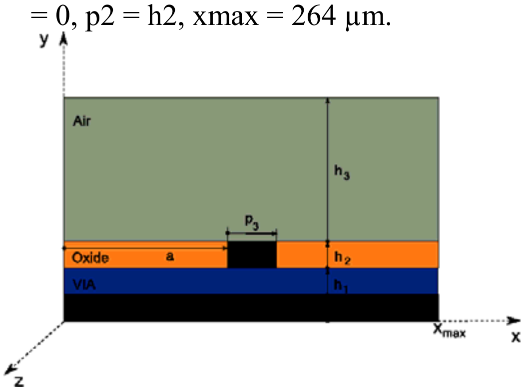


Fig. 4. Test problem .

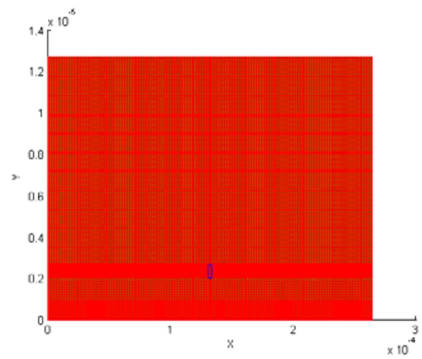
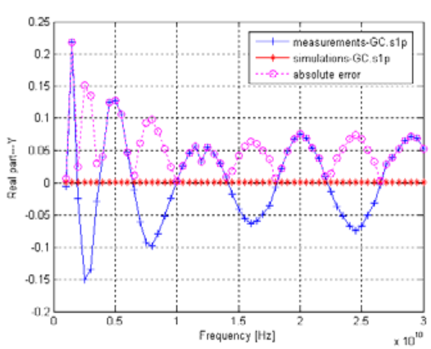
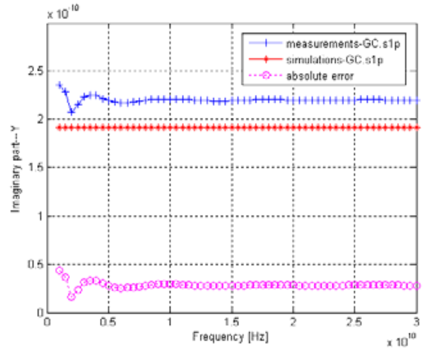


Fig.5 – 2D grid used in the modelling procedure

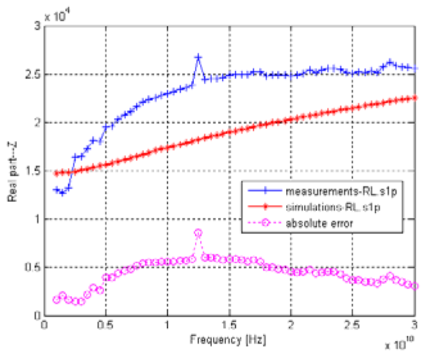
For this problem the p.u.l. parameters have been extracted first with the two field problems approach and then with the modified method. These methods have been implemented in Chamy tool developed in Matlab in the frame of the European Project Chameleon-RF [12]. The grid used has nx=261, ny=178, nz=2 nodes (Fig. 5). The number of degrees of freedom is 92030 for the 2D-EQS problem and 230132 for FW-TM problem. Fig. 6 shows the comparison of the line parameters extracted from measurements (blue) and those extracted from simulations (red). As expected, the longitudinal parameters depend on frequency.



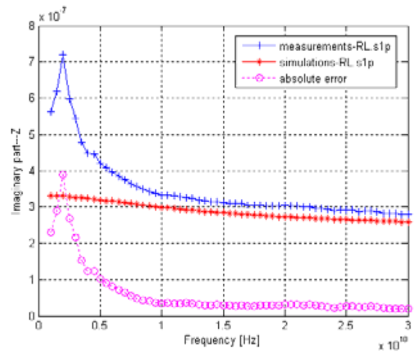
a) Line capacitance



b) Line conductance



c) Line resistance



d) Line inductance

Fig. 6. Comparison of line parameters extracted from measurements and simulations

Line conductance and line capacitance have also been extracted using the Modified Analytical-Numerical Two Fields Approach. Fig. 7 shows the comparison between different analytical approaches for p.u.l. capacitance extraction and the simulation results (obtained from the 2D-EQS field). The best analytical result is the one obtained for Meijs and Fokkema formula [3].

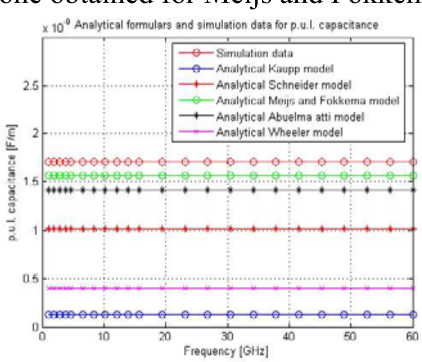


Fig. 7 – Comparison between different analytical approaches and the simulation for p.u.l. capacitance

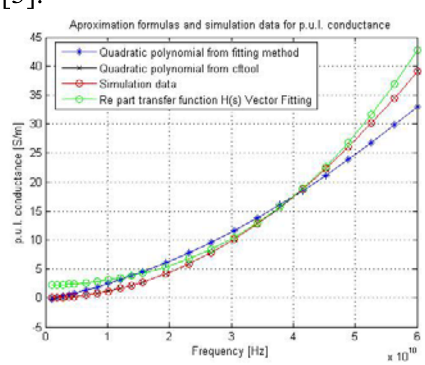
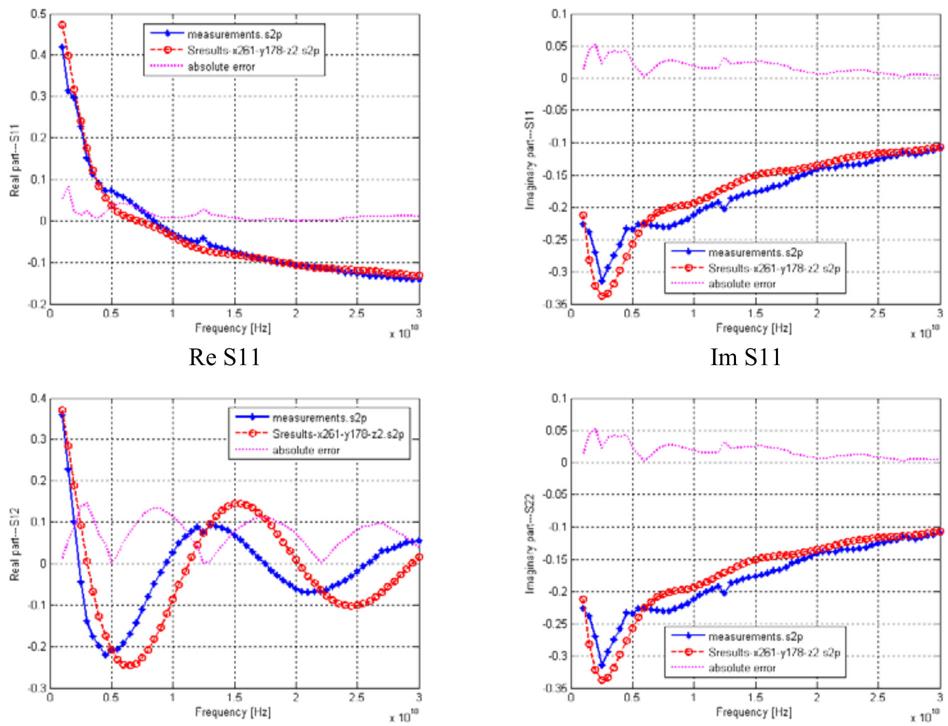


Fig. 8 – Comparison between the three approximation formulas for p.u.l. conductance calculation with simulation data

Fig. 8 presents the comparison between the three methods of computation the p.u.l. conductance and the results obtained for the simulations of the EQS field. Best results are obtained for the third method. The values of the coefficients obtained with Matlab cftool are: $a_1 = 1{,}083 \cdot 10^{-20}$, $a_2 = 2{,}314 \cdot 10^{-12}$, $a_3 = 0.01064$.

Comparison results for the scattering parameters between measurements (blue) and simulations (red) are shown in Fig. 9. The error between the measurements and the simulations is 17.92%. These results validate our procedure of extracting line parameters from two field problems.



Re S11 Im S11

Re S12 Im S12

Fig. 9 – S11 and S12 parameters

6.2. Results obtained for parametric models

For the transmission line (Fig. 4), both geometrical and frequency dependent parametric models have been developed. The first sets considered one parameter that varies, namely the height of the line, h_2. The nominal value chosen was $h_2 = 0.67\,\mu m$ and samples in the interval $[0.59, 0.79]\,\mu m$ were considered. The reference result of the p.u.l. resistance was obtained by doing “exact” simulation for the samples. These were compared with the approximate values obtained from models A and R (Fig. 10). In order to evaluate the appropriateness of these models for the analysis of technology variability we considered the parameter variations less than 15%, which is a typical limit for nowadays technologies. The errors of both additive and rational first order models are shown in Fig. 11.

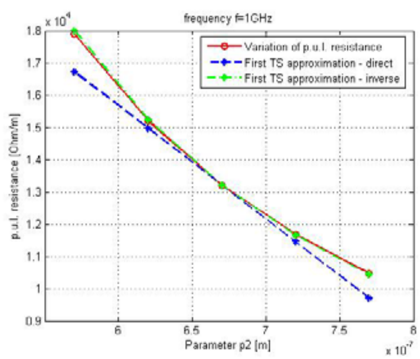


Fig. 10 Reconstruction of the p.u.l. C from Taylor Series first order expansion

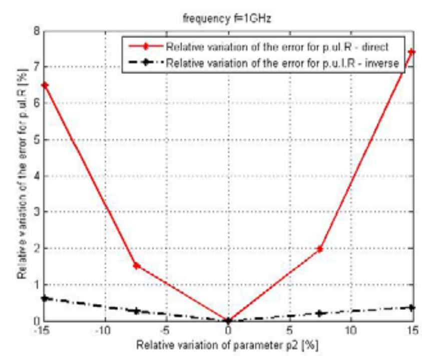


Fig. 11 Relative error w.r.t. the relative variation of parameter α_2.

The sensitivity of the admittance with respect to this parameter has been calculated according to section 4, using EM field solution. By applying Vector Fitting, a transfer function with 8 poles has been obtained. This conduced to an over determined system of size (236,26) which has been solved with an accuracy (relative residual) of 3.7 % (Fig. 12). Finally, the relative error of the A-VFIT model is 1.09 % compared to the relative error of the A model which is 0.95 % for a relative variation of the parameter of 10 % (in Fig. 13 the three curves are on top of each other).

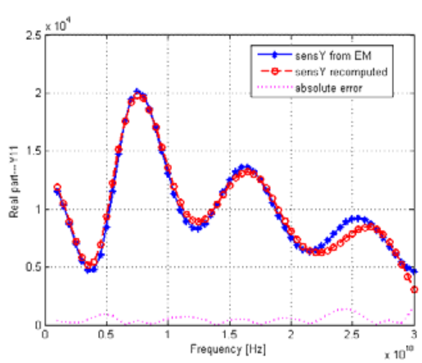


Fig. 12. Variation of the admittance sensitivity with respect to the frequency.

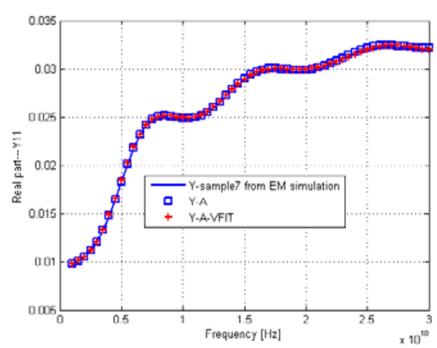


Fig. 13. Reference simulation vs. answer obtained from the frequency dependent

parametric model (13).

The second class of problems addresses the junction components of the interconnections modeled as 3D passive components, more precisely we analyze the parameterized **T-shape conductor** with the configuration shown in Fig 14. The aim is to model different relative positioning of contact shapes in the substrate. The sensitivities are computed and the reconstruction of the Taylor series expansion is shown. The geometrical parameter that varies is p1 (Fig. 16). The aim is to model different relative positioning of contact shapes in the substrate. The relative position, p1 is varying in a set of samples from 40µm to 60µm. Fig. 17 represents an accurate TS approximation for a relative variation of parameter of 20%.

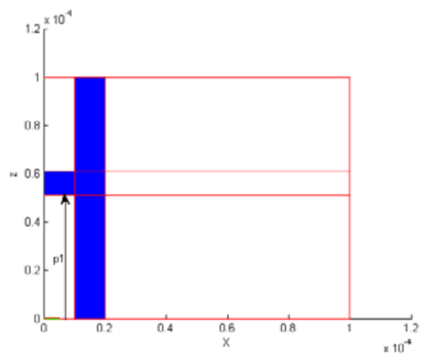


Fig. 16 Parameterised conductor

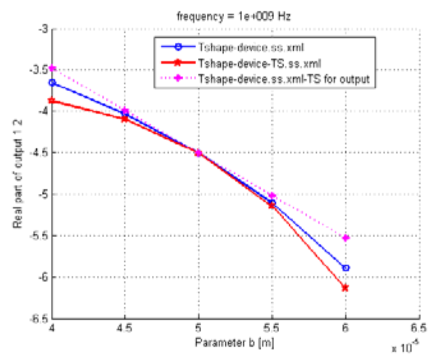


Fig.17 Reconstruction of the answer at 1GHz, from TS first order expansion

7. Conclusions

This paper proposed new approaches to model on-chip interconnects. These can be decomposed as transmission lines and as junction components. A method to compute the p.u.l. parameters from two field problems is presented. A new approach used to compute p.u.l. capacitance and conductance is described. A new method to obtain parametric models for transmission lines is proposed. It relies on field computations to extract line parameters and their sensitivities with respect to the parameters that vary. Next, a rational approximation in the frequency domain, obtained with Vector Fitting is combined with a first order Taylor Series approximation. The main advantage of this approach is that the final result is amenable to be synthesized with a small parameterized circuit.

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