High-Frequency Circuit Modeling of Large Pin Count Packages

Stefaan Sercu, *Member, IEEE,* and Luc Martens

***Abstract*— In this paper, a technique is presented for the high-frequency circuit modeling of coupled conductor structures. The method is, in particular, very useful for the modeling of structures with a varying signal/ground configuration. Structures with a large number of conductors ($N+1$, $N>100$) are also easy to model, as the method reduces the modeling of the $2N$-port to the modeling of two- and four-port structures. Two- and four-port structures are much easier to model since their equivalent circuit model has fewer parameter values. Examples of multiconductor structures are high-density connectors and large pin count electronic packages. The model accurately simulates the electrical properties, such as reflection and transmission of all conductors, and the backward and forward crosstalk to all other conductors.**

***Index Terms*— Backplane connectors, circuit modeling, high frequency, interconnection structures, packages.**

I. Introduction

DEVELOPMENT of new technologies is the driving force for the increase of clock speed, and for the continuous increase of complexity in computer and telecommunication systems. This results in complex, dense, large pin count integrated-circuit (IC) packages and closely spaced interconnections between the different IC's. For high clock speeds, the mutual coupling between the different conductors is no longer negligible and can cause system failure. In order to incorporate all the effects of the passive devices, suitable circuit models are required. System designers prefer physical circuit models above black-box models (such as S-parameters) since they are more compact and can be incorporated in almost all circuit simulators such as SPICE. Recently, new research efforts are spent on the measurement of the resistance, inductance, conductance, and capacitance matrices of coupled interconnection structures [1]–[3]. However, in almost all approaches, a lumped-element model (the transmission-line basic G-cell [4]) is the result of the characterization. This means that the generated model is only valid for frequencies beneath the lumped-element model limit (interconnection length equal to 1/10th of the wavelength) [3]. At higher frequencies, more basic cells or sections are needed to model the electrical properties of the coupled structure and the described modeling methods are no longer valid.

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The authors are with the Department of Information Technology (INTEC), University of Gent, B-9000 Gent, Belgium.

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Fig. 1 shows a typical circuit model used for the high-frequency modeling of $N+1$ coupled conductors. All conductors are inductively and capacitively coupled. In Fig. 1, most coupling elements are omitted for clarity's sake. The number of sections K depends on the length of the conductors, the bandwidth of the model, and the desired accuracy of the model [4]. The longer the package leads or interconnection lines are, and the higher the frequency for which the model must be valid, the more sections are needed. The $\boldsymbol{R}$, $\boldsymbol{L}$, $\boldsymbol{C}$, and $\boldsymbol{G}$-matrices completely determine the electrical behavior of each section. For uniform coupled interconnections these matrices are equal for all sections. However, this is not the case for the nonuniform three-dimensional (3-D) structures considered here. All matrices can differ from section to section. This means that the equivalent circuit model of these structures has many different parameters. When the S-parameters of the structure (e.g., from measurements or full-wave simulations) are available, then the model parameters can be determined using an optimization process. Due to the large number of parameters to optimize (especially when many sections are involved) the optimization process may need large computer memory, may be time consuming, and may suffer from convergence problems.

In this paper a new theoretical method for the high-frequency circuit modeling of nonuniform coupled multiconductor structures is described. The method is, in particular, very useful for the high-frequency circuit modeling of structures with many conductors $(N+1,\ N>100)$. The problems associated with the determination of the large number of parameter values of the circuit model are avoided. A number of easier-to-model two- and four-port structures are modeled, not the $2N$-port structure. Further on, instead of determining the elements of the $\boldsymbol{L}$- and $\boldsymbol{C}$-matrices of the circuit model, the elements of the $\boldsymbol{Q}$- and $\boldsymbol{C}$-matrices are determined. The $\boldsymbol{Q}$-matrix is the inverse of the $\boldsymbol{L}$-matrix. Both $\boldsymbol{Q}$- and $\boldsymbol{C}$-matrices have the interesting property to be sparse. Only the coupling between neighboring conductors is not negligible and must be considered. By inverting the $\boldsymbol{Q}$-matrix, the dense $\boldsymbol{L}$-matrix is obtained. This means that the model also includes the inductive coupling between conductors at far distance. Section II describes this new modeling approach. To illustrate the method, a circuit model for a 48-lead tape automated-bonding (TAB) interconnection is derived in Section III. We verify the model by comparing circuit-model simulations with the original S-parameter data obtained from full-wave electromagnetic (EM) simulations.

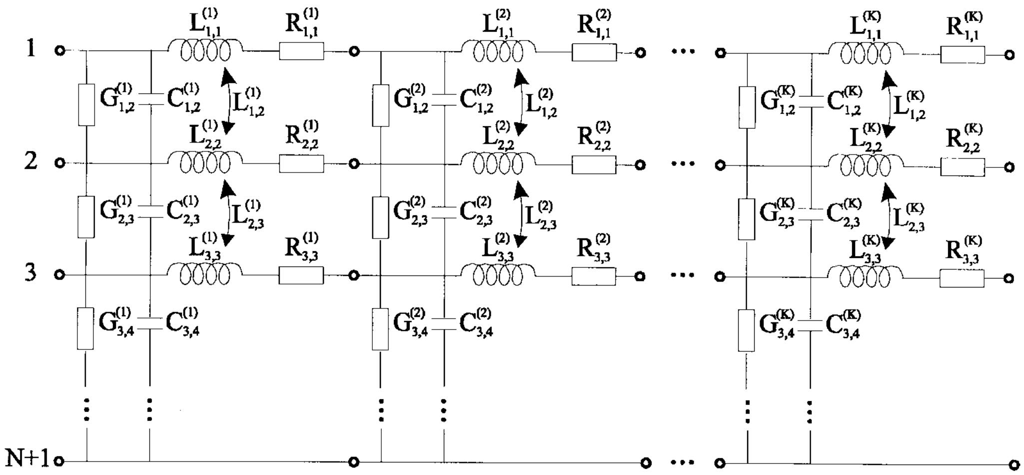

Fig. 1. Lumped-circuit model of a coupled interconnection structure.

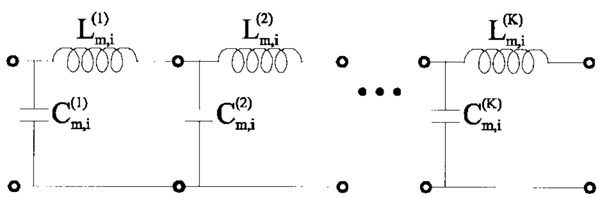

Fig. 2. Lumped-circuit model of a two-port interconnection consisting of K sections.

II. Modeling Procedure

We consider a multiconductor structure consisting of N nonuniform coupled signal conductors and one reference conductor. We assume that we can model the structure with the equivalent-circuit model of Fig. 1. The model consists of K sections. Each section is characterized by its $\boldsymbol{L}, \boldsymbol{C}, \boldsymbol{R}$ and $\boldsymbol{G}$-matrix as follows:

$$\boldsymbol{L}^{(k)} = \begin{pmatrix} L_{1,1}^{(k)} & \cdots & L_{1,N}^{(k)} \\ \vdots & & \vdots \\ L_{1,N}^{(k)} & \cdots & L_{N,N}^{(k)} \end{pmatrix}$$

$$\boldsymbol{C}^{(k)} = \begin{pmatrix} \sum_{i=1}^{N} C_{1,i}^{(k)} & \cdots & -C_{1,N}^{(k)} \\ \vdots & & \vdots \\ -C_{1,N}^{(k)} & \cdots & \sum_{i=1}^{N} C_{N,i}^{(k)} \end{pmatrix}$$

$$\boldsymbol{R}^{(k)} = \begin{pmatrix} R_{1,1}^{(k)} & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & R_{N,N}^{(k)} \end{pmatrix}$$

$$\boldsymbol{G}^{(k)} = \begin{pmatrix} G_{1,1}^{(k)} & \cdots & G_{1,N}^{(k)} \\ \vdots & & \vdots \\ G_{1,N}^{(k)} & \cdots & G_{N,N}^{(k)} \end{pmatrix}, \qquad (k = 1, \cdots, K).$$

The superscript (k) refers to the section of the circuit model. For simplicity, we assume that the coupled structures have no losses ($\boldsymbol{R} = \boldsymbol{G} = \boldsymbol{0}$), but the procedure is also applicable to lossy structures.

In the *first step* of the procedure we reduce the $2N$-port structure to a two-port structure by short-circuiting all conductors, except for one, at both sides with the reference conductor. We then characterize the two-port by its S-parameters, and propose an equivalent-circuit model for the two-port. The circuit model consists of K lumped sections. Each section has two parameter values: $L_{m,i}^{(k)}$ and $C_{m,i}^{(k)}$ ($k = 1, \cdots, K$, i = number of the conductor that is not short-circuited, the subscript m refers to the equivalent circuit model of the two-port). Fig. 2 shows the topology of the circuit model. An optimization process or another technique [5], [6] is used to determine the circuit-model parameters of each section. We notice that per section, we only have to determine two parameter values. The optimized parameter values do not belong to the $\boldsymbol{L}^{(k)}$- or $\boldsymbol{C}^{(k)}$-matrices of the complete model. In order to find the relationship between the parameters of the complete model and the parameters of the proposed two-port model, we short-circuit all ports of the complete model (except for the ports that belong to conductor i) with the reference conductor. Fig. 3 illustrates this short-circuiting (conductor $i = 2$). Using Kirchoff's laws and the requirement that both two-ports must have the same port voltages and currents, we obtain the relationship between both models. For the capacitances, we find

$$C_{m,i}^{(k)} = \sum_{j=1}^{N} C_{i,j}^{(k)} = (\boldsymbol{C}^{(k)})_{i,i}. \tag{1}$$

The optimized capacitance value of each section $(C_{m,i}^{(k)})$ is equal to the ith diagonal element of the $\boldsymbol{C}^{(k)}$-matrix. The relation between the modeled two-port inductance and the inductances of the complete model is more complex:

$$L_{m,i}^{(k)} = L_{i,i}^{(k)} - \boldsymbol{L}_{(i,-i)}^{(k)} \cdot [\boldsymbol{L}_{(-i,-i)}^{(k)}]^{-1} \cdot \boldsymbol{L}_{(-i,i)}^{(k)} \tag{2}$$

with the values of $\boldsymbol{L}_{(i,i)}^{(k)}$, $\boldsymbol{L}_{(i,-i)}^{(k)}$, $\boldsymbol{L}_{(-i,i)}^{(k)}$ and $\boldsymbol{L}_{(-i,-i)}^{(k)}$ shown at the bottom of the following page. However, if we define the $\boldsymbol{Q}^{(k)}$-matrix as the inverse of the $\boldsymbol{L}^{(k)}$-matrix

$$\boldsymbol{Q}^{(k)} = (\boldsymbol{L}^{(k)})^{-1} \tag{3}$$

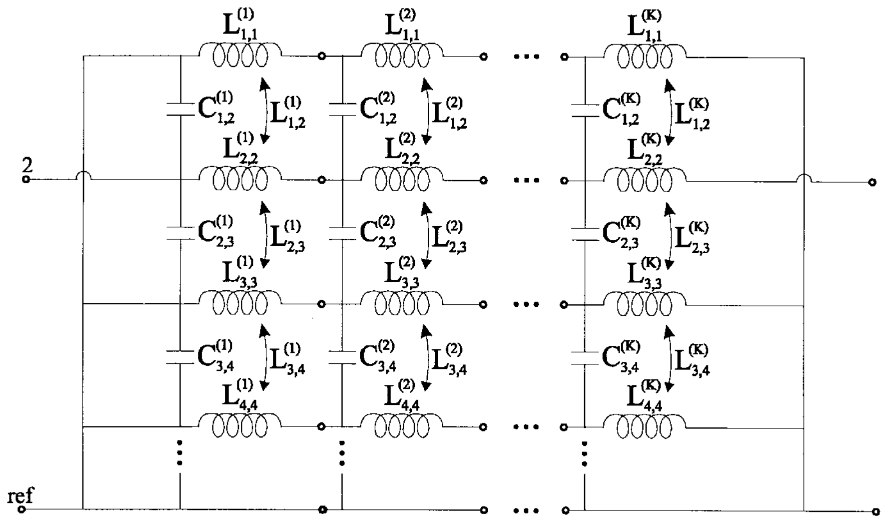

Fig. 3. Short-circuiting all conductors, except for conductor i of the complete model of Fig. 1 with the reference conductor (example for $i = 2$).

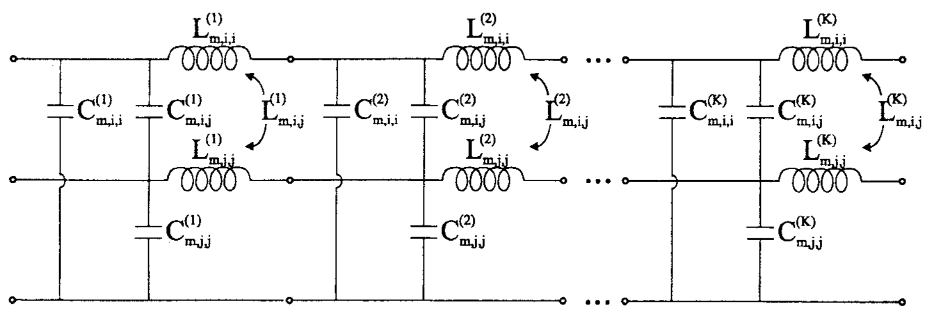

Fig. 4. Lumped-circuit model of a four-port interconnection consisting of K sections.

we find

$$\boldsymbol{Q}_{i,i}^{(k)} = (L_{m,i}^{(k)})^{-1}. \tag{4}$$

The inverse of the optimized inductance value $(L_{m,i}^{(k)})$ is equal to the ith diagonal element of the $\boldsymbol{Q}^{(k)}$-matrix. We repeat this first step for all conductors i $(i = 1, \cdots, N)$. We conclude that by short-circuiting all conductors except one, we find

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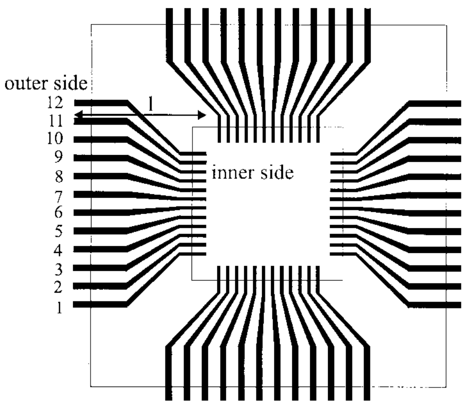
$$\boldsymbol{L}_{(i,i)}^{(k)} = (L_{i,i}^{(k)})$$

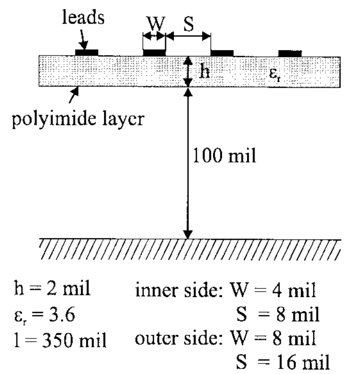
$$\boldsymbol{L}_{(i,-i)}^{(k)} = (L_{i,1}^{(k)} \quad \cdots \quad L_{i,i-1}^{(k)} \quad L_{i,i+1}^{(k)} \quad L_{i,N}^{(k)})$$

$$\boldsymbol{L}_{(-i,i)}^{(k)} = \begin{pmatrix} L_{1,i}^{(k)} \\ \vdots \\ L_{i-1,i}^{(k)} \\ L_{i+1,i}^{(k)} \\ \vdots \\ L_{N,i}^{(k)} \end{pmatrix}$$

$$\boldsymbol{L}_{(-i,-i)}^{(k)} = \begin{pmatrix} L_{1,1}^{(k)} & \cdots & L_{1,i-1}^{(k)} & L_{1,i+1}^{(k)} & \cdots & L_{1,N}^{(k)} \\ \vdots & & \vdots & \vdots & & \vdots \\ L_{i-1,1}^{(k)} & \cdots & L_{i-1,i-1}^{(k)} & L_{i-1,i+1}^{(k)} & \cdots & L_{i-1,N}^{(k)} \\ L_{1+1,1}^{(k)} & \cdots & L_{i+1,i-1}^{(k)} & L_{i+1,i+1}^{(k)} & \cdots & L_{i+1,N}^{(k)} \\ \vdots & & \vdots & \vdots & & \vdots \\ L_{N,1}^{(k)} & \cdots & L_{N,i-1}^{(k)} & L_{N,i+1}^{(k)} & \cdots & L_{N,N}^{(k)} \end{pmatrix}$$

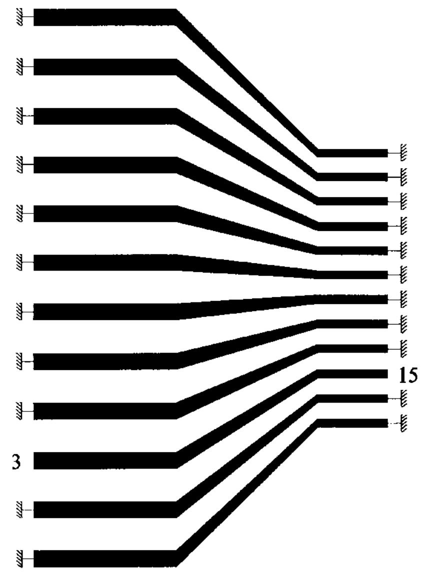


(a)

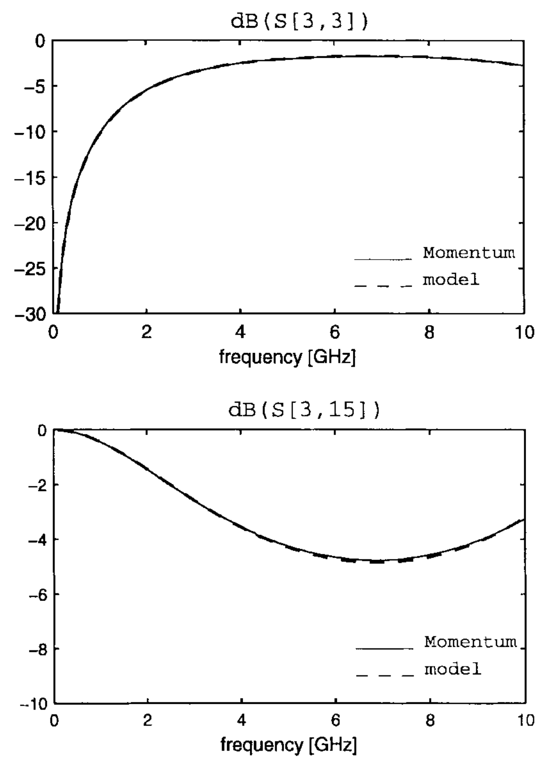


(b)

Fig. 5. (a) TAB structure with 48 leads, $l = 350$ mil, (b) layer description: $h = 2$ mil, $e_r = 3.6$, at inner side: $W = 4$ mil, $S = 8$ mil, at outer side: $W = 8$ mil, $S = 16$ mil.



(a)



(b)

Fig. 6. (a) Short-circuiting all leads except for lead 3 of the TAB at both sides with the reference plane. (b) Comparing simulated and modeled S-parameters (S[3, 3]) and S[3, 15]) of the structure of (a).

for each section the diagonal elements of the $\boldsymbol{C}^{(k)}$- and the $\boldsymbol{Q}^{(k)}$-matrices.

In the *second step* of the modeling procedure, we short-circuit all conductors except two (i and j) at both sides with the reference conductor. This results in a four-port structure. We characterize the four-port structure by its S-parameters and fit the circuit model of Fig. 4 to these S-parameters [7], [8]. Per section, the optimization process must only determine six parameters: $C^{(k)}_{m,i,i}, C^{(k)}_{m,i,j}, C^{(k)}_{m,j,j}, L^{(k)}_{m,i,i}, L^{(k)}_{m,i,j}$ and $L^{(k)}_{m,j,j}$ ($k =$ section, i and j refer to the conductors that are not short-circuited, the subscript m indicates that these parameters belong to the equivalent circuit model of the four-port). As in the first step, the parameter values of the four-port circuit model are not part of the $\boldsymbol{L}^{(k)}$- or $\boldsymbol{C}^{(k)}$-matrices of the complete model. The relation between the parameter values of both models is given by (5) and (6), shown at the bottom of the page. We can conclude that by short-circuiting all conductors except two, we find the diagonal elements (i,i) and (j,j) and the nondiagonal element (i,j) of the $\boldsymbol{C}^{(k)}$- and $\boldsymbol{Q}^{(k)}$-

$$\begin{pmatrix} (\boldsymbol{C}^{(k)})_{i,i} & (\boldsymbol{C}^{(k)})_{i,j} \\ (\boldsymbol{C}^{(k)})_{i,j} & (\boldsymbol{C}^{(k)})_{j,j} \end{pmatrix} = \begin{pmatrix} C^{(k)}_{m,i,i} + C^{(k)}_{m,i,j} & -C^{(k)}_{m,i,j} \\ -C^{(k)}_{m,i,j} & C^{(k)}_{m,j,j} + C^{(k)}_{m,i,j} \end{pmatrix} \tag{5}$$

$$\begin{pmatrix} (\boldsymbol{Q}^{(k)})_{i,j} & (\boldsymbol{Q}^{(k)})_{i,j} \\ (\boldsymbol{Q}^{(k)})_{i,j} & (\boldsymbol{Q}^{(k)})_{j,j} \end{pmatrix} = \begin{pmatrix} L^{(k)}_{m,i,i} & L^{(k)}_{m,i,j} \\ L^{(k)}_{m,i,j} & L^{(k)}_{m,j,j} \end{pmatrix}^{-1} \tag{6}$$

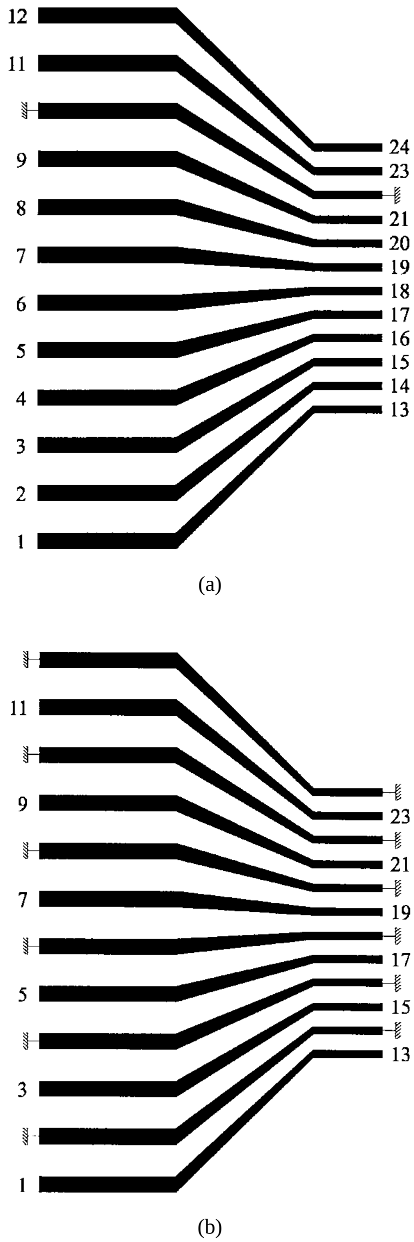


Fig. 7. Two considered signal/ground configurations and the associated port numbers. (a) Configuration 1: lead 10 is ground lead. (b) Configuration 2: leads 2, 4, 6, 8, 10, and 12 are ground leads.

matrix. If we now write the six parameters of the four-port as a function of the parameters of the complete model, we obtain

$$
\begin{aligned}
C_{m,i,i}^{(k)} &= (\boldsymbol{C}^{(k)})_{i,i} + (\boldsymbol{C}^{(k)})_{i,j} \\
C_{m,j,j}^{(k)} &= (\boldsymbol{C}^{(k)})_{j,j} + (\boldsymbol{C}^{(k)})_{i,j} \\
C_{m,i,j}^{(k)} &= -\,(\boldsymbol{C}^{(k)})_{i,j} \\
L_{m,i,i}^{(k)} &= \frac{(\boldsymbol{Q}^{(k)})_{j,j}}{(\boldsymbol{Q}^{(k)})_{i,i} \cdot (\boldsymbol{Q}^{(k)})_{j,j} - [(\boldsymbol{Q}^{(k)})_{i,j}]^2} \\
L_{m,j,j}^{(k)} &= \frac{(\boldsymbol{Q}^{(k)})_{i,i}}{(\boldsymbol{Q}^{(k)})_{i,i} \cdot (\boldsymbol{Q}^{(k)})_{j,j} - [(\boldsymbol{Q}^{(k)})_{i,j}]^2}
\end{aligned}
$$

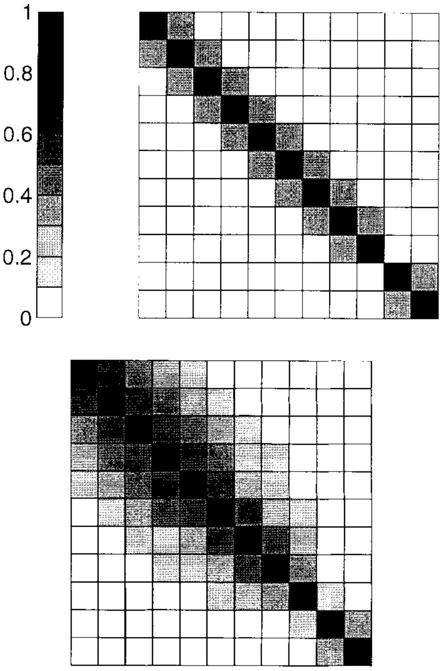


Fig. 8. Comparison of the normalized $Q^{(1)}$- and $L^{(1)}$-matrix of section one of configuration 1.

$$
L_{m,i,j}^{(k)} = \frac{-(\boldsymbol{Q}^{(k)})_{i,j}}{(\boldsymbol{Q}^{(k)})_{j,j} \cdot (\boldsymbol{Q}^{(k)})_{j,j} - [(\boldsymbol{Q}^{(k)})_{i,j}]^2}. \tag{7}
$$

Since $(\boldsymbol{C}^{(k)})_{i,i}, (\boldsymbol{C}^{(k)})_{j,j}, (\boldsymbol{Q}^{(k)})_{i,i}$, and $(\boldsymbol{Q}^{(k)})_{j,j}$ are known from the first step, (7) defines a relation of the six parameters to the unknown $(\boldsymbol{C}^{(k)})_{i,j}$ and $(\boldsymbol{Q}^{(k)})_{i,j}$. Instead of optimizing the six parameters, we introduce (7) in the optimizer and finally obtain as a result, the unknown $(\boldsymbol{C}^{(k)})_{i,j}$ and $(\boldsymbol{Q}^{(k)})_{i,j}$. We repeat this second step for all possible combinations of conductors i and j: (1, 2), (1, 3), $\cdots$, $(N-1, N)$.

In the *third step* of the procedure, we invert the $\boldsymbol{Q}^{(k)}$-matrix to find the inductance matrices $\boldsymbol{L}^{(k)}$ $(k = 1, \cdots, K)$ we are looking for.

In order to model an $(N+1)$-conductor interconnection or package, we need to model N two-port structures and $N \cdot (N-1)/2$ four-port structures. In practical applications, this number is much smaller after taking into account symmetry of the considered structures and after neglecting the small capacitive C- and Q-coupling between conductors at a far distance. We can neglect this coupling because the $\boldsymbol{C}^{(k)}$- as well the $\boldsymbol{Q}^{(k)}$-matrices have nondiagonal elements that decrease fast in amplitude as a function of the distance from the diagonal.

III. Results

In this section, we illustrate the procedure described above. As an example, we consider a 48-lead TAB interconnection structure (Fig. 5). Only the 12-leads at one side of the chip are considered. We assumed that the coupling between the leads on the different sides of the chip is negligible. The S-parameters of the complete structure are calculated with

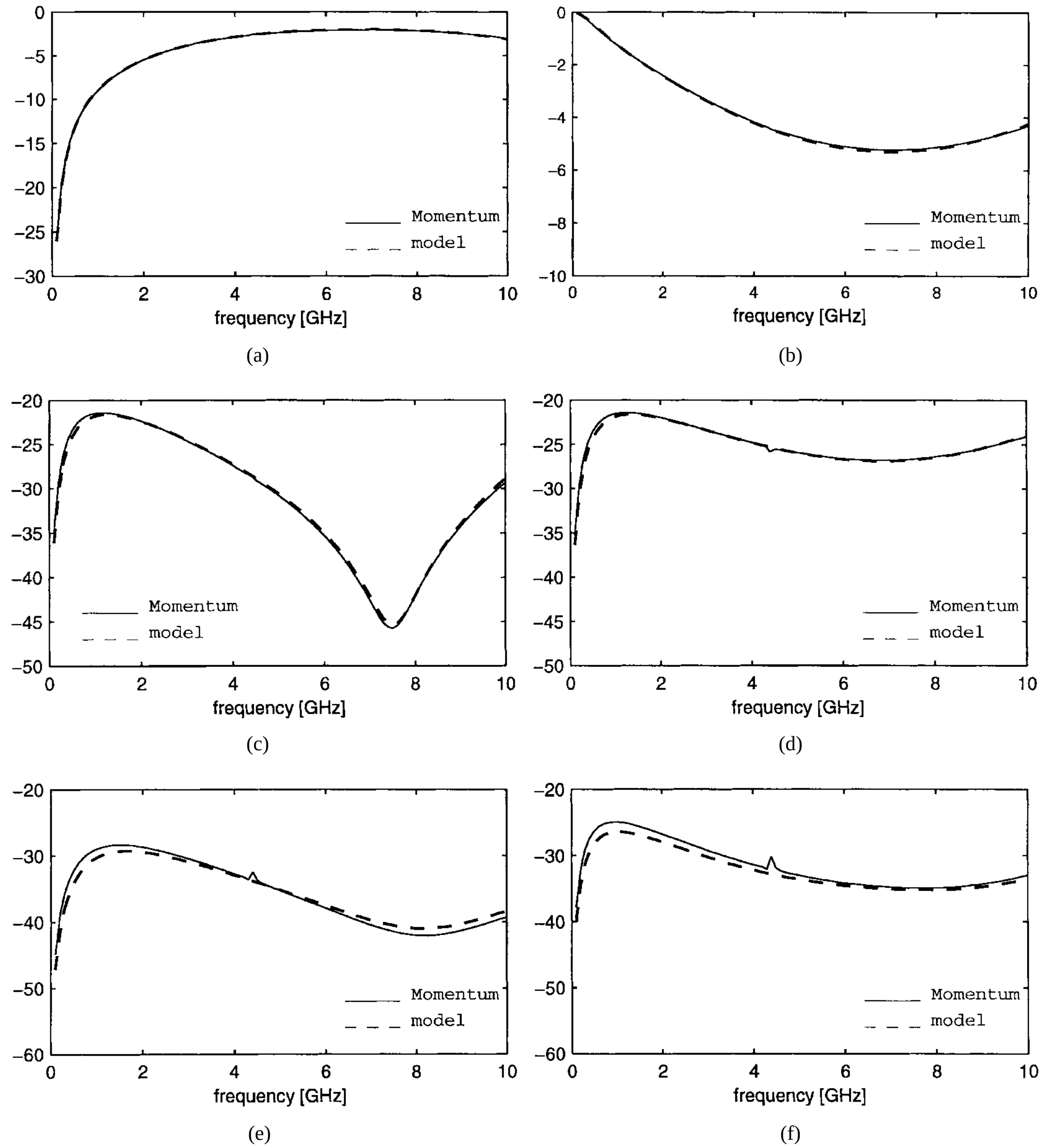


Fig. 9. Simulated and modeled S-parameters of the TAB structure (configuration 1). (a) Reflection at lead 3. (b) Transmission through lead 3. (c) Backward and (d) forward crosstalk between leads 3 and 5. (e) Backward and (f) forward crosstalk between leads 3 and 7.

the EM-field simulator HP-Momentum.[1] The reference plane necessary for the simulation is located under the leads at a distance of 100 mil.

The S-parameters of the two- and four-ports necessary for the modeling procedure are numerically derived from the S-parameters of the complete structure by terminating the appropriate ports with short-circuits and using the circuit simulator in the Microwave Design System (MDS) of Hewlett-Packard. We first derive a circuit model for the complete structure. We assume that all leads are signal leads and that no ground leads are present. The equivalent-circuit model of the structure with an arbitrary signal/ground configuration can easily be derived from the circuit model with no ground leads. The $\boldsymbol{C}^{(k)}$- and $\boldsymbol{Q}^{(k)}$-matrices of the complete model are determined following the procedure described in Section II. Three sections are required to model the structure up to 10 GHz. Fig. 6(a) shows one of the two-port structures used for the modeling. In this structure, all leads except for one (with port numbers 3 and 15) are short-circuited with the reference conductor. Fig. 6(b) shows the comparison between the S-parameters of the optimized two-port model of Fig. 2 and the S-parameters simulated with the MDS-software. The agreement between simulation and model is excellent.

We next assign the ground leads. To find the $\boldsymbol{C}^{(k)}$- and $\boldsymbol{Q}^{(k)}$-matrices of the specific configuration we only have to remove the row and column elements of the $\boldsymbol{C}^{(k)}$- and $\boldsymbol{Q}^{(k)}$-matrices corresponding with the ground leads. This means that when the signal/ground configuration changes, we immediately find the corresponding-circuit model. No extra calculations are needed. We illustrate this with two different signal/ground configurations (Fig. 7). The first configuration has only one

[1] Hewlett-Packard EEsof, Santa Rosa, CA.

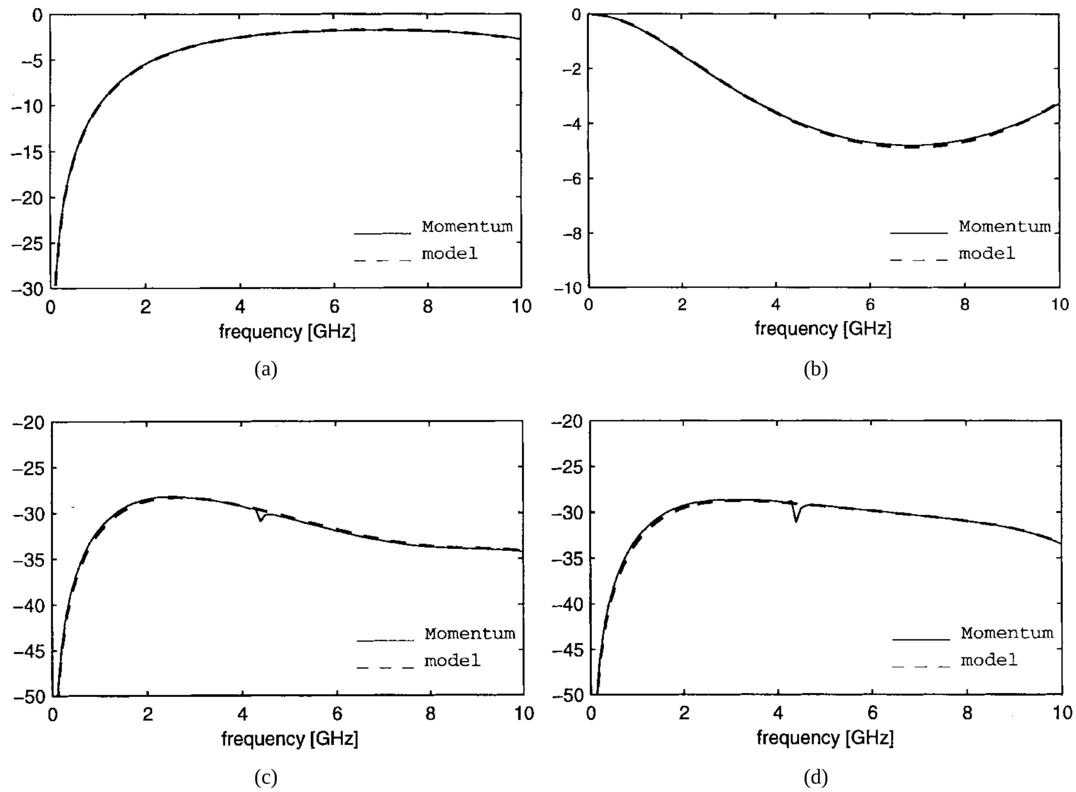


Fig. 10. Simulated and modeled S-parameters of the TAB structure (configuration 2). (a) Reflection at lead 3. (b) Transmission through lead 3. (c) Backward and (d) forward crosstalk between leads 3 and 5.

ground lead (number 10). The second configuration has six ground leads (numbers 2, 4, 6, 8, 10, and 12). Fig. 8 compares the normalized $\boldsymbol{Q}^{(1)}$- and $\boldsymbol{L}^{(1)}$-matrix of the first section of configuration 1. Both matrices are normalized to their maximum value. From Fig. 8, we can conclude that the $\boldsymbol{Q}^{(1)}$-matrix is sparse and that mainly the diagonal elements and the first nondiagonal elements are important. However, the corresponding $\boldsymbol{L}^{(1)}$-matrix is dense. The circuit model takes the inductive coupling to nonneighboring conductors into account. Figs. 9 and 10 compare simulated and modeled S-parameters for configuration 1 and configuration 2, respectively. The simulated S-parameters of the two considered specific configurations were obtained using the full-wave EM simulation and the MDS circuit simulator. On the other hand, the modeled S-parameters are derived from the complete model of Fig. 1 determined by the procedure of Section II. The agreement between model and simulation is excellent.

IV. Conclusions

In this paper, we described a new method for the circuit modeling of a large number of nonuniform coupled conductors which are, for example, present in packages. The structure is modeled by the circuit model of Fig. 1. A multiple number of sections must be used when a high-frequency model is required. The problems associated with determining the large number of parameter values of the model in an optimization process are avoided. Instead of modeling one $2N$-port with many parameters, we consider a number of easier-to-model two- and four-port structures. Per section, only two parameter values are unknown. The method is, in particular, very suitable for model structures with a varying signal/ground configuration. In order to illustrate the method, a TAB structure has been modeled. Two different signal/ground configurations for the TAB were considered. Comparing simulations of the original interconnection structure with the derived model verifies the model. The agreement is excellent.

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Stefaan Sercu (S'90–M'96) was born in Ieper, Belgium, on February 6, 1969. He received the degree in electrical engineering from the University of Gent, Gent, Belgium, in 1992, and is currently working toward the Ph.D. degree in electrical engineering in the Department of Information Technology (INTEC).

His research focuses on all aspects of time-domain reflectometry and circuit modeling of high-frequency and high-speed interconnections and connectors.

Luc Martens was born in Gent, Belgium, on May 14, 1963. He received the M.Sc. degree in electrical engineering and the Ph.D. degree from the University of Gent, Gent, Belgium, in 1986 and 1990, respectively.

Since 1991, he has been a permanent staff member of the Interuniversity MicroElectronics Centre (IMEC), University of Gent, Gent, Belgium, and is responsible for the research on characterization of packaging technologies with respect to high-frequency and EMC behavior, and for the research on experimental modeling of active devices in the Department of Information Technology (INTEC), and in 1993, he became a Professor in electrical applications of electromagnetism. His scientific work has focused on the physical aspects of hyperthermic cancer therapy and his research work has dealt with EM and thermal modeling, and the development of measurement systems for that application. He is also involved in the study of topics related to the health effects of wireless communication devices.