THE ELECTROMAGNETIC CIRCUIT ELEMENT – THE KEY OF MODELLING ELECTROMAGNETICALY COUPLED INTEGRATED COMPONENTS

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Key words: Full wave electromagnetic analysis, Electromagnetic circuit element, (EMCE), Electric and magnetic terminals, Domain partitioning.

This paper shows how the electromagnetic circuit element can be used to model parasitic inductive couplings in integrated circuits. The advantages of this approach are the reduction of computational complexity for the model extraction process, the inherent parallelism and the possibility of using different, independent models in several sub-domains, adapted to the analyzed structure.

1. **INTRODUCTION**

In the transition to the nanoscale era, the RF designers need improved IC automation tools to model and simulate full blocks, taking into account the electromagnetic (EM) coupling among the down-scaled individual devices integrated on one chip. At the high frequencies that are now envisaged, the couplings and loss mechanisms, including EM field coupling and substrate noise are becoming too strong and too relevant to be neglected, whereas more traditional coupling and loss mechanisms are more difficult to describe given the wide frequency range involved and the greater variety of structures to be modeled [1]. All this will cause extra design iterations, over-dimensioning or complete failures, unless appropriate solutions are found to resolve these design issues. These problems were addressed in the FP6/Chameleon-RF project (www.chameleon-rf.org) whose general objective is that of developing methodology and prototype tools that take a layout description of typical RF functional blocks that will operate at RF frequencies up to 60 GHz and transform them into sufficiently accurate, reliable electrical simulation models taking EM coupling and variability into account. The RF block is partitioned in basic devices (active and passive) and their compact models are augmented with connectors (called also "hooks") that allow EM interaction with and representation of their environment.

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The “hooks” that were searched for, turned out to be the boundary conditions proposed in 1971 by Al. Timotin [2] who introduced the concept of passive electromagnetic circuit element, a generalization of the multipolar circuit element proposed in 1966 by the Academician Remus Răduleț, Alexandru Timotin and Andrei Țugulea [3]. The “missing link” was internationally promoted again in [4] and the first very promising results in the frame of the Chameleon-RF project were obtained by using this concept for the simulation of EM coupling [5].

This paper shows how the electromagnetic circuit element can be used to model parasitic inductive couplings in integrated circuits. Results obtained for a simple test problem as well as for one benchmark tests from the Chameleon project are shown. For the latter one, the final check consists of the comparison between simulation results and measurements.

2. **THE ELECTROMAGNETIC CIRCUIT ELEMENT**

One of the main theoretical problems encountered in the modeling of RF components is the difficulty to define a unique terminal voltage, independent of the integration path. This independence is compulsory, since it is the only one that allows the connection of the component to an electric circuit, where the voltage does not depend of the path shape. Apparently, Kirchhoff voltage law is not valid in the case of RF structures. The solution found in our approach is to adopt appropriate boundary conditions for the field problem associated to the analysed components, in order to allow the consistency with the electrical circuit that contains this component. The only restriction is to consider only simple connected domains as components, with voltage integration paths included in the domain boundary (not going inside component). Moreover, Kirchhoff current law is valid in case of RF structures if it is written for the total current (conduction + displacement).

The correct definition of the component terminals (used for intentional interconnections) and connectors ("hooks" which represent parasitic couplings), which was an important challenge of the research, is based on the correct formulation of the EM field fundamental problem, particularly on the appropriate boundary conditions which ensure the uniqueness of the problem solution. Fortunately, the theoretical basis was set almost forty years ago by Al. Timotin [2] who rigorously introduced the concept of passive electromagnetic circuit element (EMCE), a generalization of the multipolar circuit element proposed by the Remus Răduleț, Alexandru Timotin and Andrei Țugulea [3]. The concept of Electro-Magnetic Circuit Element is briefly recalled here.

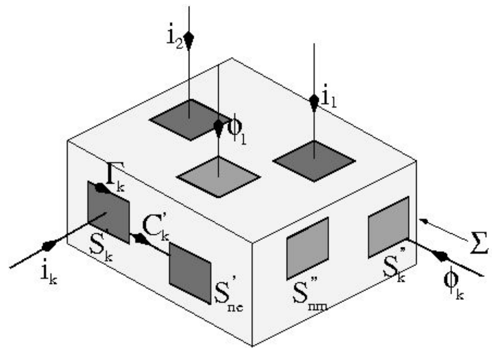

Fig. 1 – The Electro Magnetic Circuit Element – a simply connected domain with special boundary conditions on electric and magnetic terminals.

By definition, an EMCE (Fig. 1) is a simply connected domain *D* bounded by a fixed closed surface **Σ** on which there are *ne* disjoint parts $S_1', S_2', \ldots S_{ne}'$, called electric terminals and *nm* disjoint parts $S_1'', S_2'', \ldots S_{nm}''$ called magnetic terminals (hooks) on which the following conditions hold:

a) $\boldsymbol{n}\cdot$curl $\boldsymbol{E}(P,t) = 0$, for any *P* in $\Sigma - \bigcup S_k''$;

b) $\boldsymbol{n}\cdot$curl $\boldsymbol{H}(P,t) = 0$, for any *P* in $\Sigma - \bigcup S_k'$;

c) $\boldsymbol{n} \times \boldsymbol{E}(P,t) = \boldsymbol{0}$, for any *P* placed on the electric terminals $\bigcup S_k'$;

d) $\boldsymbol{n} \times \boldsymbol{H}(P,t) = \boldsymbol{0}$, for any *P* placed on the magnetic terminals $\bigcup S_k''$,

where $\boldsymbol{n}$ is the unitary vector, orthogonal to the boundary Σ, in the point *P*.

These conditions are less restrictive than usual approximations of the electric and magnetic circuit theory, because they are related only to the boundary and not to the internal structure or field in the defined circuit element. Condition a) interdicts the inductive couplings through the domain boundary, excepting for the magnetic terminals. This condition can be complied by enlarging the boundary, so that the magnetic field has a negligible normal component or it may be considered perpendicular to the magnetic terminals. Condition b) interdicts the conductive and capacitive couplings through the boundary, excepting for the electric terminals. Condition c) interdicts the variation of the electric potential over every electric terminal, allowing its connection to a node of an external electric circuit. Consequently, the current lines are orthogonal to the electric terminal surfaces. This condition is automatically satisfied, if these terminals are perfect conductors. Finally, condition d) interdicts the variation of the magnetic potential over every magnetic terminal, allowing its connection to a node of an external magnetic circuit. Consequently, the magnetic field lines are orthogonal to the magnetic terminal surfaces. This condition is automatically satisfied, if these terminals are made of perfect magnetic materials. Consequently, the magnetic field lines are orthogonal on the magnetic terminal surfaces. It is important to note that according to the EMCE definition, the electric and magnetic terminals cannot be overlapped.

With these boundary conditions, the interaction between the EMCE and its environment is completely described by two scalar variables for each terminal (for an electric terminal, its current and voltage, and for a magnetic terminal, its magnetic flux and magnetic voltage). From the point of view of boundary conditions, we will use the terminology of *terminals* either electric or magnetic. The term of *hooks* appeared from the necessity of showing that these boundary conditions will allow the parasitic coupling of this element with electric or magnetic circuits, or with other compatible compact models. The term hooks will not be used for intentional electric terminals.

For each electric terminal k, its **current** is defined as the magnetic field loop-integral $i_k(t) = \int_{\Gamma_k} \boldsymbol{H} \cdot \mathrm{d}\boldsymbol{r}$, where $\Gamma_k = \partial S_k^{'}$ is a closed curve, the boundary of the surface $S_k^{'}$ (representing the total current) and its **voltage** is defined as the line integral $v_k(t) = \int_{C_k} \boldsymbol{E} \cdot \mathrm{d}\boldsymbol{r}$ along an arbitrary open curve C_k, included in $\Sigma - \bigcup S_k^{"} - \bigcup T_k^{"}$ which is a link between a point on $S_k^{'}$ and a point on $S_n^{'}$. Here $T_k^{"}$ is a path belonging to Σ which links a point on $S_k^{"}$ with a point on $S_{k+1}^{"}$.

For each magnetic terminal k, its **flux** is defined so that its time derivative is $\dot{\varphi}_k(t) = \int_{\Gamma_k} \boldsymbol{E} \cdot \mathrm{d}\boldsymbol{r}$, where $\Gamma_k = \partial S_k^{"}$ is the contour - boundary of the magnetic terminal surface $S_k^{"}$, and its **magnetic voltage** is defined as the line integral $u_k(t) = \int_{C_k} \boldsymbol{H} \cdot \mathrm{d}\boldsymbol{r}$ along an arbitrary open path C_k, included in $\Sigma - \bigcup S_k^{'} - \bigcup T_k^{'}$ which is a link between a point on $S_k^{"}$ and a point on $S_n^{"}$. Here $T_k^{'}$ is a path in Σ, which links a point on $S_k^{'}$ with a point on $S_{k+1}^{'}$. A uniqueness theorem has also been formulated [5]. This theorem is a direct consequence of the expression of electromagnetic power transferred by means of its boundary from outside to inside of any EMCE: $P = \sum_{k=1}^{ne-1} v_k i_k + \sum_{k=1}^{nm-1} u_k \frac{\mathrm{d}\varphi_k}{\mathrm{d}t}$.

Details on the implementation of this concept in the Finite Integration Technique (FIT) [6] are given in [7].

3. DOMAIN PARTITIONING IN INTEGRATED STRUCTURES

Integrated components and systems with complex structures generate complex EM field problems that are difficult to solve. An efficient approach to manage this complexity is to decompose (partition) the computational domain in sub-domains, generate simpler field problems for each sub-domain and couple the resulting models to obtain a simpler model of the initial complex structure. We will refer to this methodology as domain partitioning (DP).

The numerical approach we propose is based on the domain decomposition of the RF block in its active and passive components as well in the „environmental" components, for instance the substrate and the upper air (Fig. 2). Each of these simple connected sub-domains is assumed to satisfy EMCE boundary conditions. Finally, the components will be interconnected by means of several hooks. In our approach, each of these components are analyzed independently. For instance, a compact or a reduced order model can be extracted. The compact models are described as equivalent circuits of these models which are re-connected together to generate the model of the entire RF block. The electric environment can be represented by an electric circuit, and the magnetic environment by a magnetic circuit. These two circuits are coupled together by means of controlled sources, representing e.g. induced voltages. The electric terminals allow the modeling of the electric interaction whereas the magnetic terminals allow the modeling of the inductive interaction. Thus, the EMCE boundary conditions allow the coupling of the component model with its EM environment. The component model can be a field model generated by any numerical method (finite or boundary elements, finite differences etc). From the coupling point of view, the models can be interchanged if they were derived from the EMCE boundary conditions.

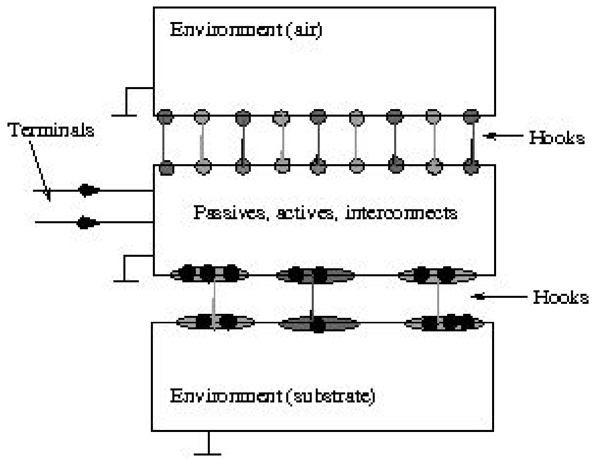

Fig. 2 – Domain partitioning in integrated structures.

The coupling between sub-models can be carried out in various ways, depending on the model representation. For instance, assuming that the system was partitioned in two parts, and the two EMCEs are described by semi-state-space representations,

$$\boldsymbol{C}_k \frac{\mathrm{d}\boldsymbol{x}_k}{\mathrm{d}t} + \boldsymbol{G}_k \boldsymbol{x}_k = \boldsymbol{B}_k \boldsymbol{u}_k , \quad \boldsymbol{y}_k = \boldsymbol{L}_k \boldsymbol{x}_k , \tag{1}$$

where k is 1 or 2 then, the global representation can be easily obtained by combining sub-blocks of the state space matrices. For instance, each matrix $\boldsymbol{C}_k$ is partitioned as $\boldsymbol{C}_k = [\boldsymbol{C}_{k1} \quad \boldsymbol{C}_{k2} \quad \boldsymbol{C}_{k3}]$ according to the partition of the state space vector $\boldsymbol{x}_k$ in three, the first part representing the inner degrees of freedom, the second part representing the output quantities associated to terminals that will not be connected and the third part representing the output quantities associated to terminals that will be connected. Note that the output quantities $\boldsymbol{y}_k$ are also placed on the last positions of the state space vector $\boldsymbol{x}_k$, and therefore the matrix $\boldsymbol{L}_k$ is merely a selection matrix. The global matrix $\boldsymbol{C}$ is thus given by

$$\boldsymbol{C} = \begin{bmatrix} \boldsymbol{C}_{11} & 0 & 0 & \boldsymbol{C}_{13} & \boldsymbol{C}_{12} & 0 \\ 0 & \boldsymbol{C}_{21} & 0 & -\boldsymbol{C}_{23} & 0 & \boldsymbol{C}_{22} \end{bmatrix}. \tag{2}$$

Similar reasoning can be applied for the rest of matrices [7].

If the systems are described by means of transfer matrices $\boldsymbol{H}_k$, where

$$\boldsymbol{H}_k = \boldsymbol{L}_k \left(\boldsymbol{G}_k + \mathrm{j}\omega \boldsymbol{C}_k \right)^{-1} \boldsymbol{B}_k \tag{3}$$

links the complex representations of outputs to the complex representation of inputs $\underline{\boldsymbol{y}_k} = \boldsymbol{H}_k \underline{\boldsymbol{u}_k}$, and these transfer matrices are split according to the number of terminals that are connected together $\boldsymbol{H}_k = \begin{bmatrix} \boldsymbol{H}_{11}^{(k)} & \boldsymbol{H}_{12}^{(k)} \\ \boldsymbol{H}_{21}^{(k)} & \boldsymbol{H}_{21}^{(k)} \end{bmatrix}$ then, the global transfer matrix $\boldsymbol{H}$ defined by $\underline{\boldsymbol{y}} = \boldsymbol{H} \underline{\boldsymbol{u}}$ can be easily computed, as

$$\boldsymbol{H} = \begin{bmatrix} \boldsymbol{H}_{11}^{(1)} & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{H}_{11}^{(2)} \end{bmatrix} - \begin{bmatrix} \boldsymbol{H}_{12}^{(1)} \\ \boldsymbol{H}_{12}^{(2)} \end{bmatrix} \left(\boldsymbol{H}_{22}^{(1)} + \boldsymbol{H}_{22}^{(2)} \right) \left[\boldsymbol{H}_{21}^{(1)} \quad \boldsymbol{H}_{22}^{(2)} \right]. \tag{4}$$

In brief, terminals act as hooks between sub-domains. They allow independent meshing, and even using independent PDE or different numerical method in each sub-domain. If adjoin sub-domains have conformal aligned meshes, the number of hooks can be increased up to the limit when each node on the interface is an independent terminal. In this degenerate case, the interface does not perturb the field solution, thus being numerically transparent. According to the convergence theorem, in correct discretizations, the numerical solution tends to the exact one, when the norm of the mesh goes to zero. In these conditions, the number of hooks tend to infinity and the interface becomes perfectly transparent [8].

4. NUMERICAL EXAMPLES

The first validations were carried out on simple benchmark tests, such two U-shaped coupled coils, placed in a Si layer, above a SiO_2 substrate. The reference result was obtained by simulating this passive device with a computational domain including the substrate below and the air above. The problem was then partitioned in a top part which includes the coils and the air above, and a bottom part consisting only of the substrate. On the cutting interface EMCE boundary conditions have been imposed. Table 1 gives the results obtained for various settings of the hooks. It can be noticed that if node-hooks are used, i.e. if each node on the interface is defined as a distinct terminal, the error due to DP is zero when the union of the grids used for the submodels is exactly the grid used for the full simulation. In this case each electric and magnetic node on the common surface represents a distinct terminal – the surface is in fact transparent for the EM field. Since the full simulation perfectly overlaps the simulation with DP and with node-by-node interconnection between the models this validates the theory of hooks and its implementation in the framework of the FIT numerical method. However, in such a case, the number of inputs/outputs increases drastically (more than 200 for this simple example which used a coarse grid). This affects the CPU time needed for the evaluation of each sub-model, and this choice would not be effective for real cases. That is why it is very important to decrease the number of hooks.

Table 1

Results obtained by using various settings of the hooks

Test	ne	nm	No.of Degrees of Freedom		No. of I/O		Rel. err [%]
			Top	Bottom	Top	Bottom	
All elmag node hooks	120	99	3399	1349	220	218	0
Only el node hooks	120	0	3301	1251	122	120	39
Only mag node hooks	0	99	3279	1299	100	98	5.36
Only 15 mag surface-hooks	0	15	3091	1041	16	14	6.13
Only 9 mag surface-hooks	0	9	3087	1037	10	8	6.12

The second test kept only the electric node-hooks whereas the third test kept only the magnetic hooks. It can be noticed that the use of magnetic hooks is very important, which was expected for this configuration in which the coils are close and inductive coupling is important. The final two tests used less number of hooks obtained by clustering the magnetic nodes, according to the expected magnetic field pattern. For instance, in the last test, the magnetic hooks are placed as shown in Fig. 3. The result has an acceptable global accuracy, of about 6 % (Fig. 4).

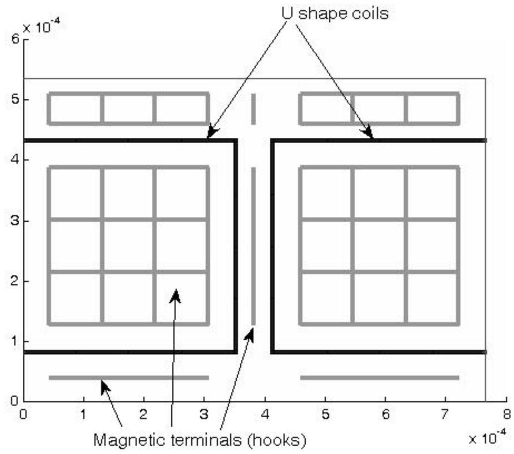

Fig. 3 – Interface between sub-models, with 9 magnetic hooks (8 surface hooks and 1 node hook).

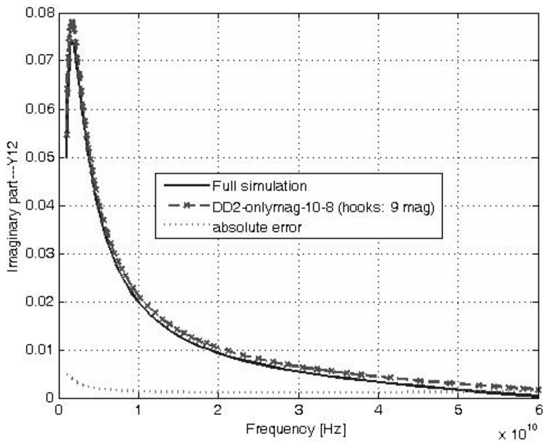

Fig. 4 – Imaginary part of the admittance component. Full simulation vs. simulation with hooks.

This simulation methodology was also applied for real benchmarks, realized and characterized at our industrial partners. Fig. 5 shows the layout of a coupled pair of coils embedded into a layered structure made and characterized at austriamicrosystems (www.austriamicrosystems.com). In this case three separate EMCE models were computed, two of them corresponding to the environmental components (top part – the air; bottom part – the substrate), whereas the third one (middle part) included the coils and their neighborhood. On each interface, 14 hooks were used, the models obtained having respectively the following number of degrees of freedom: 17138 (top) 81453 (middle) and 15427 (bottom). Fig. 6 shows the comparison between the measurements (scattering parameters) and the results obtained from this simulation.

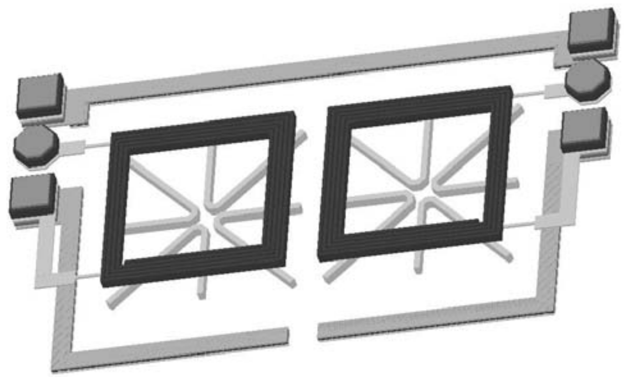

Fig. 5 – Real benchmark.

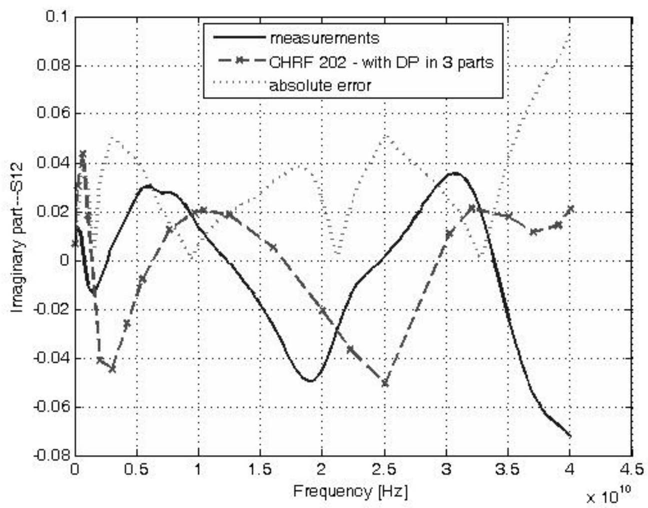

Fig.6 – Measurements vs. simulation.

5. **CONCLUSIONS**

The technique we propose for the modeling of ICs is based on domain partitioning and use of the EMCE formulation. Its main advantages are the reduction of computational complexity for the model extraction process and the possibility of using different, independent grids in several sub-domains, locally refined and adapted to the local modeled structure. In this manner the main drawback of numerical methods based on the rectangular, uniform grids, such as FIT and FDTD is eliminated. The difficulties come from the fact that the use of hooks introduces a new numerical error, the interface being no longer transparent. The hooks technique has practical importance only when their number is reduced to 1…10. With such values, the sub-domains having different shapes can be modeled independently and in parallel. Next, the reduced size models (represented as matrices – frequency dependent circuit functions, state equations or reduced order Spice circuits) are interconnected, aiming to obtain a model for the global system. The global modeling effort is then reduced, replaced by the independent model extraction for each sub-domain. In order to identify the hooks, nodes on interface have to be merged in a minimal number of clusters, so that approximation error be kept below an acceptable level. Hence, the pseudo-optimal hooks identification, problem related to so called "terminal reduction" has to be formulated as a discrete optimization problem. In addition to the clustering algorithms, heuristic rules may be also applied for this reduction. For instance, as suggested by the simple example shown, the placement of magnetic hooks in the holes of spiral inductors. Other tests showed that placement of electric hooks near conductors, allow a proper modeling of capacitive couplings, especially at high frequencies. As a general rule, the interface should be as close as possible to a constant potential surface, orthogonal to the field lines. Investigation related to this optimization task will be carried out in our future research.

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