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EMC Modeling of an Industrial Variable Speed Drive with an Adapted PEEC Method

Vincent ARDON^{1,2}, Jérémie AIME^{1,3}, Olivier CHADEBEC¹, Edith CLAVEL¹,
Jean-Michel GUICHON¹, Enrico VIALARDI²,

¹Grenoble Electrical Engineering Lab., Université J. Fourier / Grenoble-INP / CNRS, 38402 Grenoble, France

²Cedrat, Technological Department-Software, 38246 Meylan, France

³Schneider and Toshiba Inverter Europe, 27120 Pacy sur Eure, France
vincent.ardon@g2elab.grenoble-inp.fr

Abstract—This paper presents an adapted PEEC-based (Partial Element Equivalent Circuit) methodology applied to the modeling of interconnections of power electronics devices. Although this method is known, the originality of this work is its use to model a device presenting an industrial complexity. To make possible this modeling, two adapted integral methods, based on two different meshings are presented. They compute respectively parasitic inductances and capacitances and lead to an equivalent circuit of the system. From a time-domain simulation of this circuit, current and voltage sources can be extracted and used to compute the radiated near magnetic field. This approach has been applied to model a real industrial static converter, a variable speed drive.

I. INTRODUCTION AND STATE OF THE ART

Today power electronics devices are more and more compact and small and the working frequency range is rising. Consequently the compliance to EMC standards is not guaranteed. That is why power interconnections need accurate modeling that takes into account some magnetic constraints (i.e., the skin and proximity effects) and the most important electric couplings between the conductors of the device. According to the sizes of the typical industrial application, the propagation phenomena can still be neglected because the frequency range is from 150kHz to 200MHz and the longest conductor size is around 10cm ($\lambda/10=15\text{cm}$).

To model such industrial devices, constituted of complex conductors (multi-layer tracks close to each others), the finite element method would be few adapted regarding the difficulty of volume air region meshing and conductors discretization to properly take into account skin/proximity effects. On the contrary, the PEEC method (i.e. extraction of equivalent circuit components thanks to integral approach) is known to be well adapted to model complex geometries with an important surrounded air region [1]. However, the full-PEEC approach, based on dual inductive and capacitive meshings (and a 3D meshing of dielectrics), would be also very memory consuming and not adapted.

Thus, in the next part, we propose an adapted PEEC formulation where two integral methods lying on two separated meshes are combined to build a parasitic equivalent circuit of the power interconnections. In the last section, this approach is tested by modeling an industrial variable speed drive and comparisons between simulations and measurements are carried out: the behavior of the harmonic response and the emitted near magnetic field.

II. EXTRACTION OF (R-L-M-C) EQUIVALENT CIRCUIT

By considering the integral form of the Maxwell's equations and by assuming quasi-static conditions and surface location of the free-charges ρ in a medium of permittivity ϵ and permeability μ , the external applied electric field E in a point r of space can be rewritten as:

$$\begin{aligned} \mathbf{E}(\mathbf{r}) &= \frac{\mathbf{J}(\mathbf{r})}{\sigma} + j\omega\mathbf{A}(\mathbf{r}) + \nabla\phi(\mathbf{r}) \\ &= \frac{\mathbf{J}(\mathbf{r})}{\sigma} + j\omega\mu \iiint_V G(\mathbf{r}, \mathbf{r}') \mathbf{J}(\mathbf{r}') dV + \frac{1}{\epsilon} \nabla \iint_S G(\mathbf{r}, \mathbf{r}') \rho(\mathbf{r}') dS \end{aligned} \quad (1)$$

where \mathbf{J} is current density, σ the conductivity of conductors, \mathbf{A} the magnetic vector potential, ϕ the electric potential and $G=1/(4\pi\|\mathbf{r}-\mathbf{r}'\|)$ the Green function. The three parts of (1) are respectively due to resistive, inductive and capacitive effects.

In our adapted PEEC approach, these effects are modeled by using two different meshings of conductors and dielectrics (Fig. 1). Parasitic resistances, inductances and mutual-inductances (R-L-M) are computed from a volume meshing of conductors, and capacitances (C), from a non-necessarily conformal surface meshing of conductors and dielectrics.

A. Extraction of resistive and inductive elements (R-L-M)

Each mesh element presents a resistance and a self inductance (analytical formulas). Mutual inductances between different elements are numerically computed thanks to a Galerkin method [1-2], which consists in integrating the magnetic vector potential created by one volume element on the other.

Depending on the directions of the flowing current, conductors can be either unidirectional (thin or long tracks) or bidirectional (large tracks and ground planes). The conductors belonging to the first type are meshed in the skin depth but not in their length (gain in number of elements), whereas the bidirectional ones are discretized in two directions. These meshes are adapted to model skin and proximity effects from which an inductive dense and square matrix and a resistive

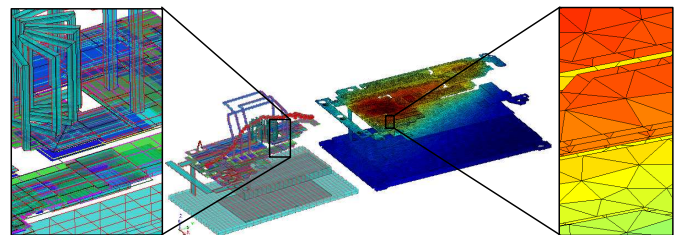


Fig. 1. Resistive-inductive (left) and capacitive (right) meshings

diagonal matrix are extracted. The number of mesh elements is equal to the size of the (R-L-M) matrix.

B. Extraction of capacitive elements (C)

For the modeling of capacitive effects, an electrostatic formulation is used and consequently the meshing is adapted to the surface free-charge distribution (side effects). Thinner mesh elements than the inductive ones are required to take into account accurate spatial variation of surface charges which is due to the uniform surface distribution of electric potentials.

Unlike the use of a Galerkin method, a 0-order point matching method is sufficient to accurately compute the potential coefficients $P_{i,j}$ and the normal field coefficients $E_{i,j}$. These coefficients are respectively calculated between each conductor or dielectric elements and all (conductor and dielectric) elements of the meshing. Consequently, they compose the square and dense interaction matrix P/E [3].

The N*N capacitance matrix (N << number of elements) which models the electric couplings between the N conductors of the device (or part of conductors) is computed by properly adding free-charge sets. These ones (Q) are obtained by performing N resolutions of the linear system $V=[P/E]*Q$ with N different potential distributions V (0 or 1V on each mesh). The worst inconvenience of this method is the storage of the matrix and the integration time which increases in $O(n^2)$.

To overcome this issue of modeling large devices, the algorithm AMLFMM [4] is used: in fact, it accelerates the computation of interaction coefficients ($O(n \cdot \log(n))$) and is low-memory consuming thanks to the use of a truncated multipole decomposition of interactions. This method subdivides the geometry in cubes of different levels with an adaptive octree algorithm which controls the interaction computations and leads to a compact matrix-vector product. Finally, the problem is solved with an iterative solver, a left preconditioned GMRESR(m) algorithm [5].

III. APPLICATION TO AN INDUSTRIAL VARIABLE SPEED DRIVE

A. Extraction of the equivalent circuit

This methodology has been applied to obtain an equivalent circuit of an industrial static converter: a marketed variable speed drive. This complex device is composed of four copper-track layers. Figure 1 shows the two different meshings used to extract the lumped elements. The square resistive-inductive matrix (6,800*6,800 elements) is computed with InCa3D [6]. The capacitive matrix (27*27 elements) is computed with AMLFMM and 27 GMRESR(m) solvings of a linear system of 48,500 unknowns (the conductor charges).

The complete PEEC circuit of the interconnections of this device is built in a system-level software (Saber) and is composed of an (R-L-M) macroblock and all the capacitances which are manually linked themselves.

B. Harmonic response and near magnetic field studies

The supply chain, the load (a three-phase motor) and the command circuit are then added to the previous PEEC circuit. Thanks to time-domain simulations and Fourier transforms of the results, it is possible to analyse the parasitic equivalent circuit by means of a harmonic response study in the

conducted frequency range (150kHz-30MHz). This study has validated our method because the frequency resonances obtained by simulation and measurement are the same (detailed results of this analysis will be reported in the final version of the paper).

Then, a second analysis of the emitted near magnetic field, just over the static converter, has been led. The Fourier transform of currents and voltages of the circuit, which have been obtained by a time-domain simulation, have been introduced as sources in our model and the emitted magnetic field has been computed at a given frequency by means of the Biot and Savart law. Figure 2 shows the comparison between the vertical component of the magnetic field Hz at 32kHz simulated and measured and a good adequation between the two color shades can be appreciated. It means that the current loops are correctly modeled.

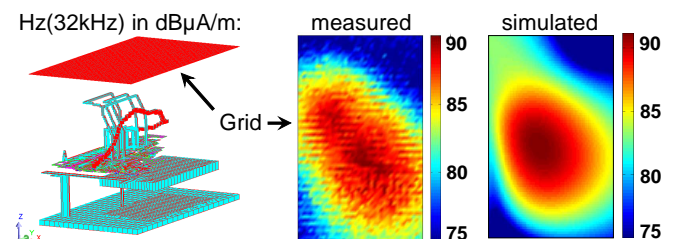


Fig. 2. Radiated near magnetic field H_z measured and simulated at 32kHz

IV. CONCLUSIONS

In this paper, it has been presented the coupling between two adapted integral methods from two adapted meshings allowing, respectively, the extraction of resistive-inductive and capacitive equivalent elements of power interconnections of industrial complexity.

It has also been highlighted that the complete (R-L-M-C) equivalent circuit can be exported into a SPICE-like tool where time-domain analysis can be performed. Then, the obtained Fourier transforms of the currents and voltages have been introduced as sources in the 3D geometric representation of the interconnections for analyzing the EMC performances of the system on a wide frequency range. The harmonic response and the emitted near magnetic field comparisons with measurements have validated this approach on a real industrial variable speed drive.

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