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Far field extrapolation from near field interactions and shielding influence investigations based on a FE-PEEC coupling method

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Abstract -- Considering standards, the main source of far field emitted by power electronic structures is the common mode current generated by the floating potentials tracks. Moreover, due to the frequency and power increases, the compliance to EMC becomes more and more difficult. In consequence, it is necessary to investigate practical solutions to reduce the emitted field. A useful solution is the shielding of wounded components which are near field influent sources. But this solution may be not efficient for far field reduction and investigations concerning component placement can be a better and cheaper solution. A FE-PEEC hybrid method was developed and can be used to investigate near field interactions and shielding influence. The method is applied to a structure combining a common mode filter with the floating potentials of an inverter. Using the hybrid method, the near field interactions and the influence of the common mode inductance shielding are analyzed on the floating potentials which are the sensitive part of the structure.

Index Terms -- Electromagnetic compatibility, Electromagnetic coupling, Modeling, Power conversion

I. INTRODUCTION

Many studies are carried on the EMC modeling of power electronic devices. Indeed, this field of research is closely linked to industrial constraints and needs, inherited from standards compliance. That is why, considering EMC cost reducing by less time consuming during the conception stage, several Computer Aided Design (CAD) tools have been developed [1-4]. Unfortunately, far field models of industrial products with complex geometries are still very difficult to obtain. In the radiated EMC domain, the usual modeling methods are necessarily limited by the considered frequency range and/or by the geometry complexity, and/or by the medium heterogeneity. That is why many works are focused on modeling methods able to take into account all these constraining elements [5]-[6]-[12]. By this way, new modeling solutions have to be proposed to engineers in order to improve their modeling approach during the conception stage. In other words, modeling environments must be robust, and implemented in CAD tools with easy-use interfaces.

Moreover, engineers must have practical solutions reducing the emitted field levels. Layout techniques with investigations upon placement of components have the advantage of do not decreasing rendering factor by increasing losses and being effective in high frequency range [7-11]. The developed models must be enough accurate and flexible

to be used to improve power electronic structures reducing the number of prototypes.

From this point of view, an analysis is developed in this paper in order to extrapolate the far field from the near field interactions by using an original modeling method consisting in the coupling between the Finite Elements and PEEC methods [12]. These methods are implemented in commercialized software's [13]. The study of the shielding is analyzed too. Indeed, it is interesting to show that this very usual solution is not always the good one to adopt and that cheaper solutions have better performances. The method is applied to a case of study which is the common mode filter and the output tracks of the inverter of a commercialized variable speed drive. The aim is to evaluate the near field interactions between the tracks, which have complex geometries, and the magnetic component, which is the common mode inductance of the common mode filter.

In a first part, the mechanisms linked to the generation of EMC perturbations in power converters are presented. The aim is to explain how the near field interactions can impact the far field levels. Our case of study is detailed here. In a second part, the usual modeling methods are presented. It is interesting to show their limitations and why it is necessary to improve the classical approaches by proposing new solutions. The modeling method FEM-PEEC used in our study is presented in the next section. Finally, the results showing the influence of the component placement and analyzing the shielding influence are detailed in a fifth section.

II. EMC STUDY OF STATIC CONVERTERS

A. EMC mechanisms

The increase of power and frequency amplifies the disturbance level. Also, for marketing considerations, an increase of performances must not be threatened by the level of pollution of new structures. The means of disturbance reduction depend on the frequency range which extends from some Hertz to a few Gigahertz. Electromagnetic disturbances can be separated into two bands. Below 30MHz, the conducted disturbances are preponderant, beyond is the domain of radiated disturbances. In our case, the power electronic structures which are studied are static power converters. Due to thermal constraints, multilayer printed

circuit boards are often used. In consequence, modeling of the geometries can be complex. Moreover, depending on the applications and the power range, the switching frequency is usually comprised between 10 kHz and 100 kHz. Conducted band studies show that the spectrum shapes are closely linked to the driver configurations. Indeed, EMC perturbations of a static converter are due to the differential and common mode currents [9]-[14]. These last ones are generated by the floating potential tracks with their stray capacitances referenced to the ground planes during the switching stages.

Moreover, the far field is correlated to the common mode. This is illustrated by measuring the far field of a boost converter at 3 meters and its common mode voltage (Fig. 1). Only the field emitted by the converter is measured. The influence of cables, power supply, load and driver is neglected since they are placed behind copper planes. The two obtained curves are closed each other. In conclusion, to reduce or control the far field emissions, it is necessary to control the common mode generation [9]. In consequence, it is essential to preserve the floating potential tracks from electromagnetic disturbances.

Considering now the near field environment of static converters, eddy currents which are generated by near field couplings, can increase the common mode generation (Fig. 2). For example, wounded components are important near field sources of power electronic structures and interactions with the layout are potentially responsible of eddy currents generation. A useful solution is to shield the main sources, here the common mode inductance. But this drastically increases the weight and cost of the final product. In order to quantify the impact of the couplings and shielding on the floating potentials, models need to be developed.

III. STATE OF THE ART ON THE MODELING METHODS

Electrical equivalent models are necessary in order to take into account EMC constraints during the design process.

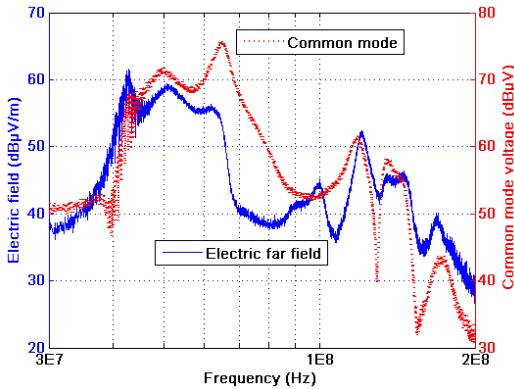


Fig. 1. Correlation between far field and common mode

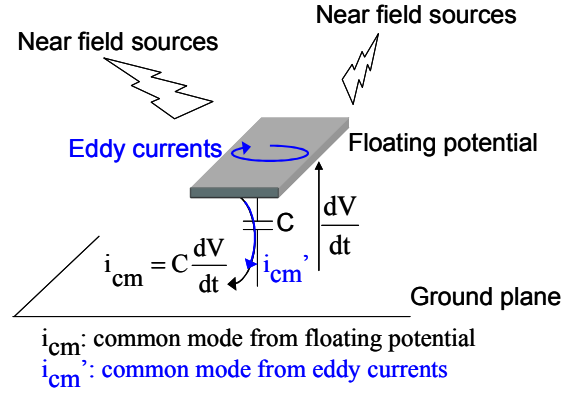


Fig. 2. Mechanism of common mode generation

According to the studied component, the modeling method which will be used to evaluate EMC performances will be different, depending on its nature, the required accuracy, the level of the model.

Models have to consider heterogeneity of media (magnetic components, copper tracks, substrates such as FR4 for example and air) and multilayer topologies. On one hand, Method of Moments (MoM) and method of Partial Element Equivalent Circuits (PEEC) are useful because they are not expensive in computational time. Indeed, they do not require meshing the whole considered space [15-16]. On the other hand, the Finite Element Method (FEM) takes into account inhomogeneous environments and magnetic materials [17]. Unfortunately, a fine meshing is needed to obtain precise results. Moreover, all space is considered and computational time can become important for complex geometries. An alternative could be the development of coupling methods. In consequence, works are carried on development of modeling methods making it possible to take into account these elements [5]-[6]-[12]. Moreover, due to the considered geometries, dedicated tools with adapted interfaces are really needed.

IV. PROPOSED MODELING METHOD

To model a problem constituted by many thin conductors and ferromagnetic materials like motor drivers, using FEM requires a lot of meshes and so can lead to too big problems so that no solution can be found. For power electronic applications, this step has to answer two very strict conditions:

- close to the conductors, variations of magnetic field is very important [18], so to take into account this aspect, the meshing must be dense;
- geometry of conductors is very difficult to mesh with a good quality since it can be constituted by very thin planes like DBC (Direct Copper Board).

Indeed, to take into account the skin effect, a good quality

meshing requires two meshes into the skin depth.

On the other hand, because of three dimensional ferromagnetic materials, it is very difficult to solve this problem by PEEC method. Hence a hybrid method dealing with this type of problem was developed [12]-[18]. The idea of the hybrid method is to benefit of the main advantage both of FEM and PEEC method:

- PEEC method takes easily into account interactions between complex 3D massive conductors;
- FEM takes into account the interaction between conductors and ferromagnetic materials.

Conductors are meshed by PEEC mesh type to take into account skin and induced effects. Each element of conductor's mesh is considered in the coupling as an inductor whose current density is uniform. In other words a conductor is represented by a set of inductors. Air around and magnetic regions are meshed using finite elements.

Consider a general problem comprising inductors and ferromagnetic materials like in Fig. 3. To adapt FEM to the coupling, the magnetic scalar potential formulation is used.

In this formulation the magnetic field is calculated by a total scalar potential ϕ and a reduced field $\mathbf{T}_0$ due to all inductors as (1).

$$\begin{aligned}\mathbf{H} &= \mathbf{T}_0 - \text{grad } \phi \text{ in } \Omega_0 \\ \mathbf{H} &= -\text{grad } \phi \text{ in } \Omega_1\end{aligned}\quad (1)$$

$\mathbf{T}_0$ can be expressed in terms of $\mathbf{t}_{0k}$ - the source field created by 1 A in the inductor k and its current I_k : and because $\text{curl } \mathbf{H} = \mathbf{j}$ in Ω_0 where $\mathbf{j}$ is local current density, $\text{curl } \mathbf{T}_0 = \mathbf{j}$. If $\mathbf{j}_{0k}$ is current density of inductor k fed by a current of 1A, $\mathbf{t}_{0k}$ has to satisfy (2) and (3).

$$\begin{aligned}\text{curl } \mathbf{t}_{0k} &= \mathbf{j}_{0k} \text{ in } \Omega_0 \\ \mathbf{t}_{0k} \times \mathbf{n} &= 0 \text{ on } \Gamma_{01}\end{aligned}\quad (2)$$

$$(3)$$

Equation (3) is imposed because $\mathbf{H}_t$ is continuity on the interface Γ_{01} . We express $\mathbf{t}_{0k}$ by (4).

$$\mathbf{t}_{0k} = \mathbf{h}_{0k} - \text{grad } \delta\phi_k \quad (4)$$

In equation (4), $\mathbf{h}_{0k}$ is the Biot and Savart field and $\delta\phi_k$ is the jump between scalar and reduced potential [11] which allows satisfying (3). Note that this formulation doesn't require any meshing of the inductors but needs to estimate the source field $\mathbf{t}_{0k}$ before the finite element resolution. The weak form of $\mathbf{T}_0$ - ϕ formulation leads to the finite element matrix system (5).

$$[A]\{\phi\} = -[C]\{I\} \quad (5)$$

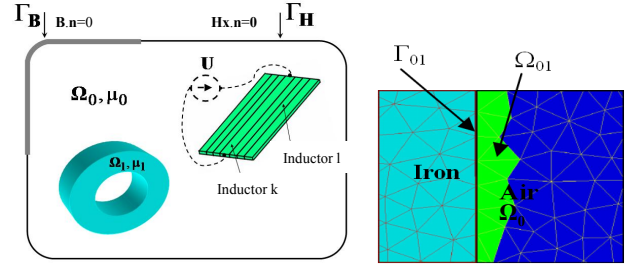


Fig. 3. General description of FEM-PEEC coupled method

In equation (5), terms of matrices are defined by (6).

$$\begin{aligned}A_{ij} &= \int_{\Omega} \mu \text{grad } \alpha_i \cdot \text{grad } \alpha_j \, d\Omega \\ C_{ik} &= - \int_{\Omega_0} \mu_0 \text{grad } \alpha_i \cdot \mathbf{t}_{0k} \, d\Omega\end{aligned}\quad (6)$$

If the inductor k is fed by a voltage source U_k , the current I_k is unknown and we must add a circuit relation [20] (7).

$$U_k = R_k I_k + j\omega \int_{\Omega_0} \mathbf{t}_{0k} \cdot \mathbf{B} \, d\Omega \quad (7)$$

The magnetic induction in Ω_0 (air) can be written using (8).

$$\mathbf{B} = \mu_0 \mathbf{H} = \mu_0 \left[\sum_{k=1}^m \mathbf{t}_{0k} I_k - \text{grad } \phi \right] \quad (8)$$

Combining (5) and (7) gives (9).

$$\begin{bmatrix} A & C \\ C^t & D + R \end{bmatrix} \begin{Bmatrix} \phi \\ I \end{Bmatrix} = \begin{Bmatrix} 0 \\ U \\ \frac{U}{j\omega} \end{Bmatrix} \quad (9)$$

In equation (9), terms of matrices are defined by (10).

$$D_{kl} = \int_{\Omega_0} \mu_0 \mathbf{t}_{0k} \cdot \mathbf{t}_{0l} \, d\Omega \quad R_{kk} = \left[\frac{R_k}{j\omega} \right] \quad (10)$$

Computations of D_{kl} require a fine meshing around inductors because the variation of $\mathbf{t}_{0k}$ and $\mathbf{t}_{0l}$ is very strong around them. To avoid this problem, a coupling of this formulation with PEEC method was proposed. This consists in calculating D_{kl} by the mutual inductance M_{kl} that can be exactly determined by PEEC method. Substituting (4) into (10) gives (11).

$$D_{kl} = \int_{\Omega_0} \mu_0 \mathbf{h}_{0k} \cdot \mathbf{t}_{0l} \, d\Omega - \int_{\Omega_0} \mu_0 \text{grad } \delta\phi_k \cdot \mathbf{t}_{0l} \, d\Omega \quad (11)$$

The first term of right hand of (11) represents the mutual inductance in vacuum between inductor k and l [20]. The second term can be transformed into surface integral by applying divergence theorem (12).

$$D_{kl} = m_{kl} - \int_{\Gamma_{01}} \mu_0 \delta\phi_k \mathbf{h}_{0l} \cdot \mathbf{n} d\Gamma + \int_{\Omega_{01}} \mu_0 \text{grad} \delta\phi_k \cdot \text{grad} \delta\phi_l d\Omega \quad (12)$$

Note that the terms C_{ik} in (6) can be transformed as follow (13).

$$C_{ik} = - \int_{\Gamma_{01}} \mu_0 \alpha_i \mathbf{h}_{0k} \cdot \mathbf{n} d\Gamma + \int_{\Omega_{01}} \mu_0 \text{grad} \alpha_i \cdot \text{grad} \delta\phi_k d\Omega \quad (13)$$

This avoids estimating $\mathbf{t}_0$ in entire domain. One needs to estimate $\mathbf{t}_0$ only on the interface Γ_{01} . In conclusion, the hybrid method proposed permits to avoid pre-calculating $\mathbf{t}_0$ in entire domain and to relax the mesh around conductors. This was validated and gives a stable result [12]-[18]. A case of study is defined in order to analyze this phenomenon using this original modeling method.

V. APPLICATION AND RESULTS

The chosen case of study is a simplified structure extracted from the geometry of an industrial variable speed drive. It is composed of a three phase common mode filter with the floating potential tracks of an inverter (Fig. 4).

The toroidal core inductance material is the FT-1KM nanocrystalline, manufactured by Hitachi.

Due to the fact that this material has a higher permeability than ferrite, it is first-rated for common mode filtering applications. The initial permeability μ_r is equal to 16000 at 20°C and for a frequency equal to 100 kHz. There are five turns for each winding.

The theoretical value of the inductance is 1.2mH. Due to the winding turns, the inductance is the main near field source.

The PCB is composed of two layers. Phase 1 and 2 are crossed. Phase 3 and the floating potentials U, V and W are routed on the top layer.

The copper width is equal to 70 μ m. The resolution frequency is equal to 10.6 kHz which is the same than the switching one.

The aim of the study is to evaluate the interactions between the filter and the floating potentials which are the most influent part considering common mode generation and by this way, far field emissions (Fig. 5).

The floating potentials tracks are supplied with a 1V voltage source and loaded with 1 $\mu\Omega$ in order to easily observe the eddy currents induced by the interaction with the filter which is supplied with current sources equal to 2A.

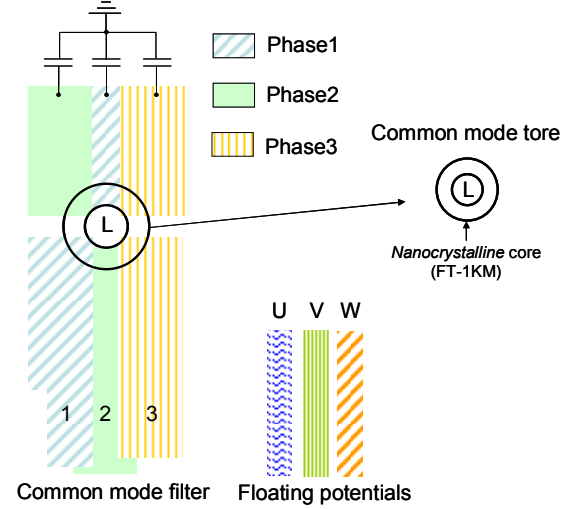


Fig. 4. Description of the studied layout

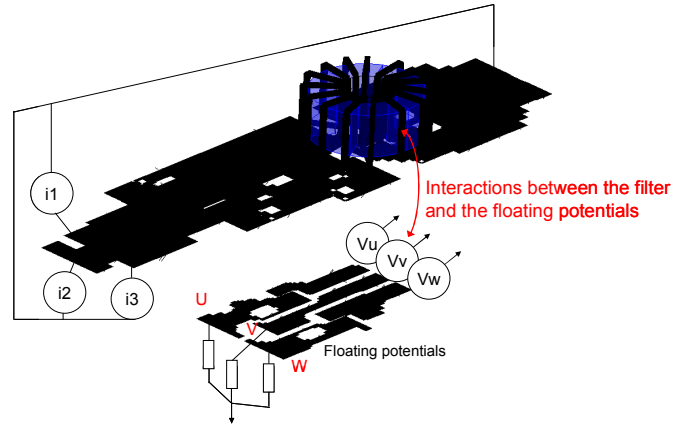


Fig. 5. Electrical topology to model interactions

The simulation frequency is equal to 10 kHz and is done using the classical method FEM and using the FEM-PEEC one [12]. The performances are compared (Table 1). It is clear that the coupling method offers better performances for such problems and gives same accuracy of results.

TABLE 1
COUPLING AND SHIELDING INFLUENCES ON THE CURRENTS OF FLOATING POTENTIALS TRACKS

	FEM	FEM-PEEC
Meshes	39726	
Time (s)	37777	6139
Memory (Mo)	944.7	331.3

The currents of floating potentials tracks are computed without common mode filter. Then, the influence of the couplings between the floating potential tracks and the common mode filter is analyzed. The influence of the shielding is also studied (Table 2 and Fig. 6).

The couplings are important between the common mode filter and the floating potentials tracks with unbalanced currents. The geometry is influent and routing process with component placement need to take into account these elements in order to reduce the possible increase of the common mode current.

Moreover, the shielding does not reduce the eddy currents. Its influence is localized around the inductance (Fig. 7) and its efficiency decrease is too fast to reach the floating potentials (Fig. 8). In this case, the common mode inductance shielding is not an efficient solution to reduce the common mode generation.

TABLE 2
COUPLING AND SHIELDING INFLUENCES ON THE CURRENTS OF FLOATING POTENTIALS TRACKS

Without filter	U	V	W
	328.628	373.487	403.316
With filter	U	V	W
Without shielding (A)	465.60	527.09	570.53
With shielding (A)	465.59	527.04	570.48

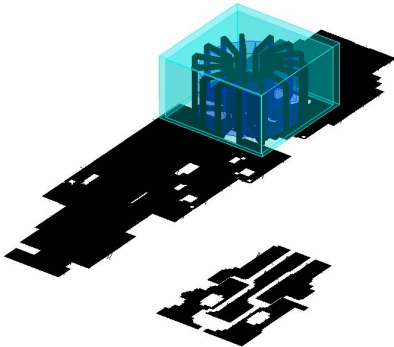


Fig. 6. Common mode inductance shielding

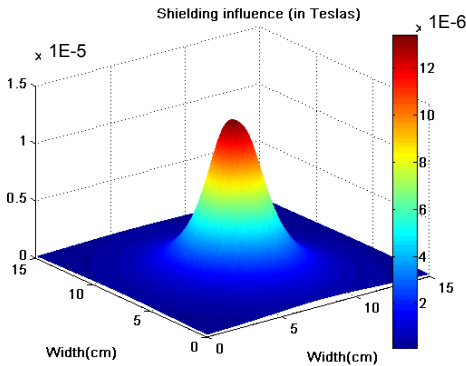


Fig. 7. Shielding influence repartition

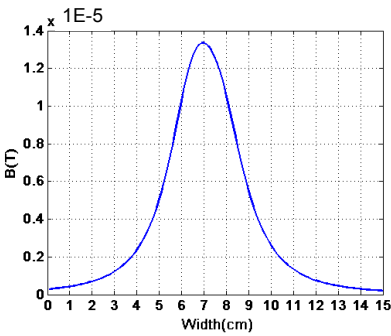


Fig. 8. Shielding influence decreasing

The coupling influence is verified by measuring the far field of variable speed drives with the studied configuration. The emitted field by the studied configuration is compared with a topology where the inductance is placed closer to the floating potential tracks (Fig. 9). As expected, the emitted field is more important for this last configuration (Fig. 10). The interactions are favored by the proximity between the inductance and the floating potential tracks. The common mode currents are more influent due to the adding of induced currents in floating potential tracks U, V and W.

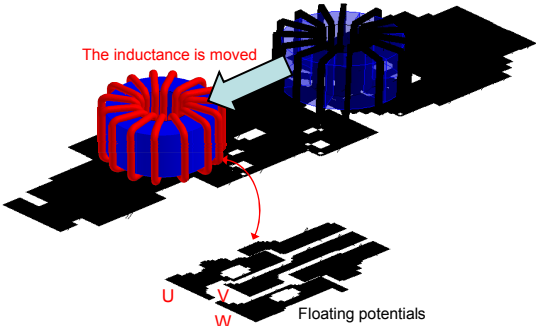


Fig. 9. Two positions of the common mode inductance

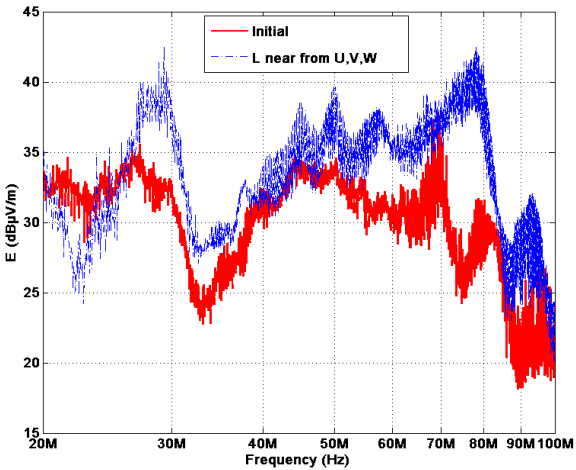


Fig. 10. Far field measurement showing the influence of the interactions

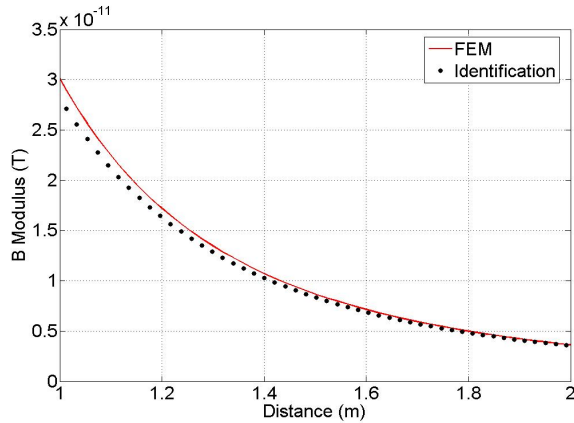


Fig. 11. Magnetic field decrease (modulus) vs the distance from the source

It appears that the couplings can not be neglected in a near field computation. The question which can be asked is to determine if it is the case in far field. In consequence, it is now needed to compare modeling methods in order to transpose the study in the far field domain. Works are in progress to compare FEM, PEEC and multipole expansion models in far field. And for example, a static study of the inductor using FEM and multipole approach is presented on Fig. 11 [21]. The magnetic field evaluation using a multipole approach instead of FEM is computed. This result shows the pertinence to use FEM for near field and multipole approach, which requires less memory and time solving, for far field studies.

VI. CONCLUSION

In this paper, an analysis was presented. Focused on the far field extrapolation from near field interactions, several conclusions can be enounced. First, these interactions can not be neglected during the modeling of power electronic devices. This is demonstrated thanks to an original modeling method based on the coupling between FEM and PEEC methods. A second conclusion is that the shielding of most polluting sources is not always the best solution. An adequate placement of the components and more particularly, of the floating potential tracks, is a priority. Finally, this kind of modeling approach is very useful to test some improvements on the structure in order to satisfy the standards and to improve the EMC and EMI performances of power electronic devices. New investigations are carried on in order to improve the approach by using multipolar expansion for the far field studies.

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