

Simulating Electromagnetic Interactions in High Power Density Converters

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Abstract— In high power density inverters with integrated functions like filters and control electronics electromagnetic interaction between components becomes an important issue. Especially filters for electromagnetic interference are very sensitive to electromagnetic fields and therefore critical to design. Nevertheless only few guidelines for optimizing placement of filter components were published up to now. This paper introduces a simulation method for predicting stray fields of components and conducting structures, which is verified experimentally for different passive components and setups. The examples demonstrate a possibility of handling interaction problems.

I. INTRODUCTION

A major advantage of the progress in power semiconductor technology is the possibility, to increase power density for energy conversion devices by reducing space for energy storage and cooling. Especially in the power range up to 100 kW converter size is an important issue. Manufacturers minimize the size by integrating power components, heat sink and control electronics. Thereby severe problems arise due to electromagnetic interaction between close components, which often lead to malfunction and EMC problems. Especially EMC filters being very sensitive to electric or magnetic fields, cause extra cost for screening or for additional space [1].

Reducing component interaction in the stage of development requires knowledge about the magnitude and direction of stray fields of every component and conducting structure. Using this information a designer is able to choose an optimum placement in order to avoid coupling effects. Therefore methods are required to simulate and measure stray properties of single components. The result gained enables a complete prediction of circuit behaviour including the effect of electromagnetic field interaction in real 3D setups.

In this paper different methods of simulating field effects in power electronic devices are discussed and a suggestion for an appropriate method is made. This method is tested for predicting external fields and coupling behaviour mainly of passive components, e.g. capacitors, rod core inductors and common mode inductors. The predictions are verified experimentally and simplifications are suggested to reduce simulation complexity. The results show the way to predict the influence of electromagnetic fields in power electronic devices.

II. FIELD SIMULATION ENVIRONMENT

State of the art for simulating fields is the finite element method (FEM). It allows the calculation of all electromagnetic effects, including time dependent field propagation, inhomogeneous material properties and material losses. The FEM works by defining a discrete mesh for the investigated setup and solving Maxwells equations for every point of the mesh. To obtain correct results, a fine discretization is required, resulting in a high number of elements to be calculated. Although the FEM can be applied in 2D and 3D, the number of elements rises disproportional in 3D setups, easily exceeding the calculation capability of nowadays computers. The first possibility to reduce simulation complexity therefore is the selection of a simulation method which does not require the discretization of the whole investigated space. Integral equation methods like the Method of Moments (MoM) or the Partial Element Equivalent Circuit method (PEEC, [2]) fulfil this requirement by only considering the surface of conducting structures (MoM) or the whole conducting structure (PEEC). As the PEEC method is customized for lower frequencies where far field radiation is no major issue, it is applied in this paper. For simulating electromagnetic properties with this method, conducting structures are partitioned into strait filaments and the equivalent resistance, the inductance and the coupling capacities and inductances to every other filament are calculated for each of them. These results can directly be used in a circuit simulator to calculate circuit behaviour under influence of field coupling.

A drawback of MoM and PEEC is the missing capability to model the effect of inhomogeneous permeability. As chokes are mostly built with ferrite cores, a possibility to bypass this restriction is essential for the application of PEEC. In this paper a solution for this problem is suggested.

III. MODELLING LEAKAGE FIELDS OF PASSIVE COMPONENTS

As modelling of conductors is widely examined, focus of this paper are passive components.

A. Capacitors

Capacitors influence their environment by electric and magnetic leakage fields. The evaluation of electric fields is quite straight forward, as the outer electrode of the capacitor acts as electrode of a stray capacitor. Fig. 1 shows the coupling capacitance of a film capacitor to an adjacent conductor.

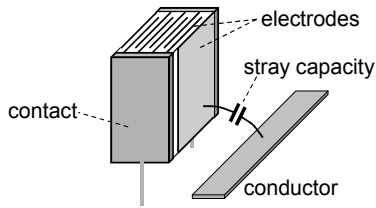


Fig. 1: Stray capacity of the outer electrode of a film capacitor to an adjacent conductor

The magnetic fields caused by the current flow through the capacitor need a more detailed analysis. Starting at one contact, the current flows from one electrode via the dielectric to the other electrode and its contact. As the number of electrodes is very high, an exact modelling of this path is far too complex. Therefore we suggest a simplification that only models the total magnetic field of these currents. This is done by treating the structure of electrodes and dielectric as homogeneous material with properties being the mixture of the different materials. The resulting magnetic field has the same direction and strength as a solid conductor with the outer dimensions of the foil stack.

In Fig. 2 an example for the model of a foil capacitor on a printed circuit board (PCB) is shown. The capacitor in this test setup is modelled by one block, representing the foil stack with electrodes and dielectric. The capacitors pins are placed at both sides of the block, connecting the foil stack to the PCB. The first circuit on the PCB provides current for the capacitor, the second one acts as antenna loop. Connecting a signal source to the capacitor and a measurement instrument to the antenna loop enables the measurement of induced voltage.

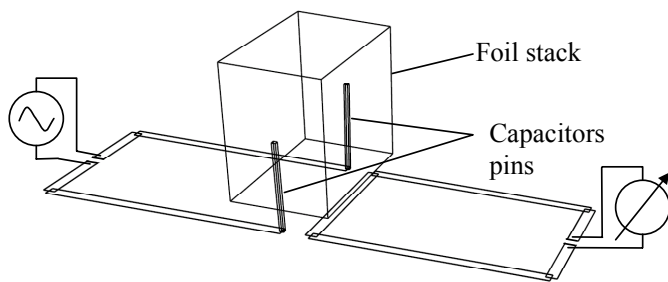


Fig. 2: Simulation model of test structures and test objects

A comparison between simulation and measurement was carried out, the results are shown in Fig. 3. A more detailed description of experiments on capacitors can be found in [3].

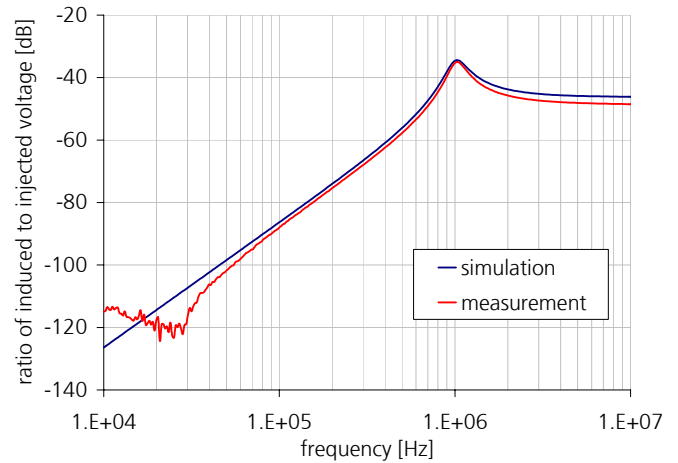


Fig. 3: Comparison of measured and simulated magnetic coupling between capacitor and conductor

The good coincidence proves the suitability of the proposed method. The simplification of the foil stack allows a fast and efficient calculation of the equivalent series inductance and its coupling to other components.

B. Solenoid with ferrite rod core

To model inductors as part of a complex system some simplifications are necessary to reduce complexity of the structures. One common simplification for wound components is the reduction of the number of turns to a solid copper tube. The copper tube has the outer dimensions of the winding and the current around the tube is the current of the inductor times the number of turns in order to generate the same field strength (Fig. 4).

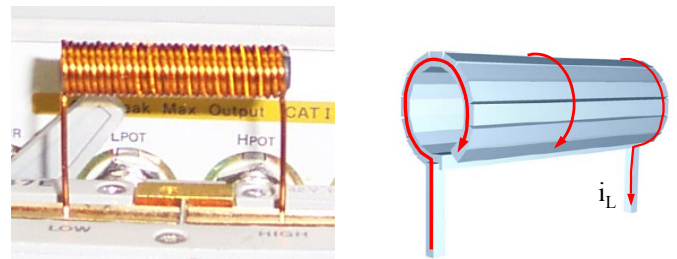


Fig. 4: Solenoid as test object and as PEEC model with current directions added

Thereby the calculation effort can be reduced significantly. This simplification is validated by field measurements, using a test setup for 3D field characterization (Fig. 5). The test fixture is made of non conducting material and allows measurements in all directions. Test signals are produced by a HP4194A gain/phase analyser. The magnetic field is measured by a coil with diameter of 5 mm and 1mm length, which is connected to the analyzer via a preamplifier. The measurement frequency is 5 MHz.

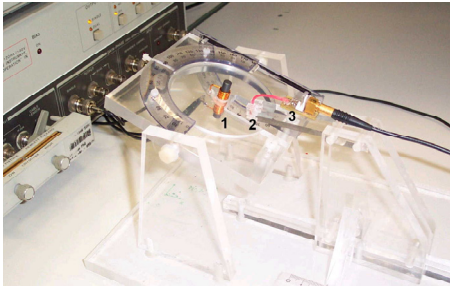


Fig. 5: Test setup for characterization of magnetic leakage fields, 1: device under test, 2: magnetic field detector, 3: preamplifier

The major question when calculating magnetic fields with PEEC solver is how to consider inhomogeneous permeability. The PEEC method does not take this into account, nevertheless one constraint resulting from practical setups allows to overcome this restriction. The constraint results from the fact, that the designer normally tries to place components without interaction between each other to prevent interaction. Therefore obvious coupling placement, e.g. two rod core coils next to each other, normally is not implemented. Consequently magnetic stray fields will propagate at least half of their way through air or non ferromagnetic material. From this we derive the assumption, that the direction of stray fields is not distinctly influenced by the ferromagnetic material of the components themselves, as the magnetic resistance of stray field path is dominated by high impedance of non ferromagnetic material. Only inside components a substantial part of the magnetic path propagates through material with high permeability.

For the prediction of coupling effects this assumption allows significant simplifications, which make integral equations methods suitable for setups including inhomogeneous permeability. As long as the assumption is valid, magnetic coupling between components is calculated as if no ferromagnetic material is present. If for example the coupling between a rod core inductor and a line is calculated, two inductances and a mutual inductance between both are the result when using the PEEC method. The influence of the rod core is considered in the next step by adapting the simulation results. The core increases the inductance of the solenoid inductor by a factor called effective permeability (μ_{eff}). Snelling [4] calculates this factor from geometric data. In order to not alter the field simulation results the factor is included in a circuit represented by two sources connected to the inductance of the rod core inductor and the rest of the circuit (Fig. 6).

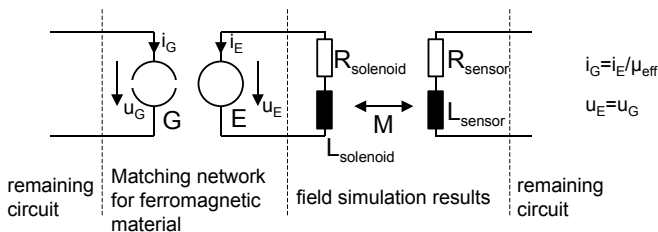


Fig. 6: Equivalent circuit considering ferromagnetic material of a component

The two sources are a current and a voltage source. Their task is to consider the effect of the ferromagnetic core, which increases the inductance of the solenoid. As long as the direction of magnetic field lines is not changed by the core, a constant voltage connected to the solenoid will lead to a constant voltage transferred to the field sensor. Only the current flowing into the solenoid will be reduced. The formulas describing the sources reflect this behaviour: The current of the current source (i_G) equals the current of the voltage source (i_E) divided by the effective permeability of the inductor. The matching network thereby models the reduction of current necessary to generate the same field strength. The voltage of the voltage source (u_E) equals the voltage of the current source, as the voltage needed to build up a defined flux is not altered by the magnetic resistance of the flux path.

Fig. 7 shows the stray field in radial direction measured at different distances for a solenoid without core (6 mm diameter and 20 mm long), a solenoid with a ferrite core and the corresponding simulation result. In this setup a voltage with constant amplitude is injected into the solenoid. Under this circumstance a constant flux is generated, resulting in a constant stray field (Fig. 7).

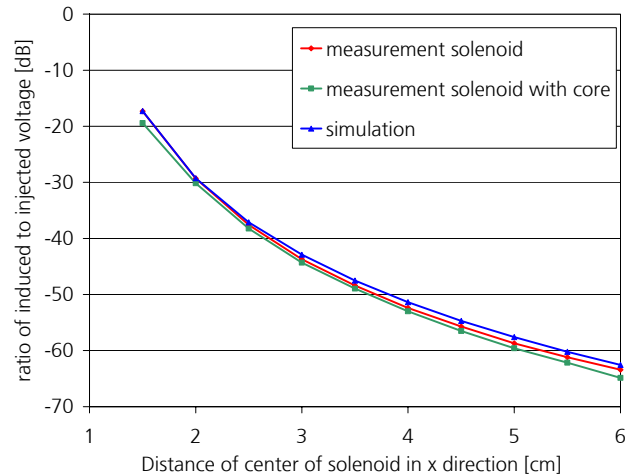


Fig. 7: Amplitude of the induced voltage in the field sensor, simulated and measured with and without ferrite core

The good correspondence between the induced voltage with and without core verifies the assumption, that the direction of field lines is not distinctly altered by the ferromagnetic material and therefore the magnetic coupling between coil and field sensor keeps constant. The simulation results match with the measurements.

C. Common mode inductor

The common mode inductor consists of two or three windings on a toroidal core. Magnetic stray fields of this component are mainly caused by differential mode signals [5] and depend particularly on the angular coverage of windings and the style and size of winding. An impression of the stray field with a differential mode signal gives a two dimensional FEM simulation in Fig. 8.

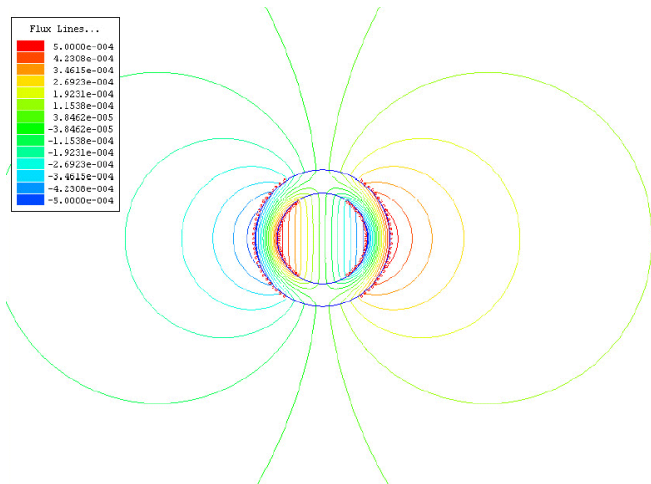


Fig. 8: Magnetic stray field of a two phase common mode inductor

The idea of modelling this component is similar to the solenoid: magnetic properties of the windings are modelled without core and the influence of the ferrite material is added in the second evaluation step, the circuit simulation. Fig. 9 shows the test object, a two phase common mode inductor on a core with 36 mm diameter and the corresponding CAD model.

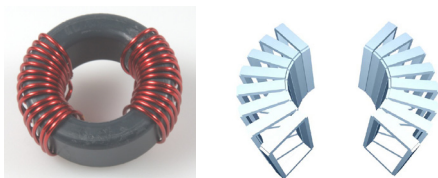


Fig. 9: Common mode inductor, test object and CAD model

Instead of every turn a number of turns is modelled together in order to reduce calculation effort. The coupling inductances calculated from this model are transferred to a circuit simulator by the equivalent circuit in Fig. 10, taking into account the strong coupling between both windings and the effective permeability of stray fields.

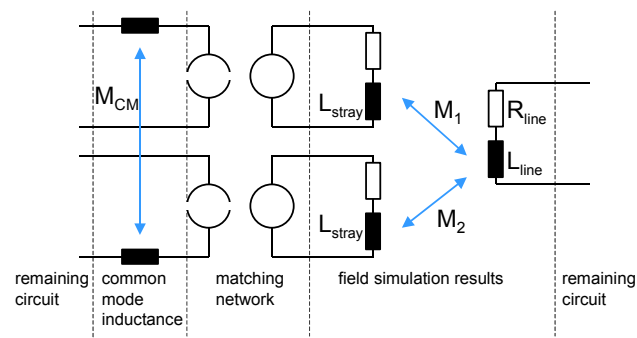


Fig. 10: Equivalent circuit for common mode inductor

The common mode inductance is modeled by two fully coupled inductances. The influence of the effective permeability is considered by the matching network as described for the solenoid. Stray inductances are calculated as results of PEEC simulation. The effective permeability of a

common mode choke can not be derived as easily as of a solenoid. Additional factors have to be taken into account: the geometry of the toroid, the angular coverage of the winding, the number of windings and the opposite orientation of their fields for differential mode signals. We therefore suggest to calculate the effective permeability with formulas commonly used for the solenoid, modified by an experimentally derived correction factor. For a solenoid, diameter, length and permeability of the rod core are the relevant quantities. Since toroidal cores have a nearly square cross section, a diameter is not helpful. As permeability is proportional to cross section, we may equate the expressions for round area and square area to derive the relation

$$D = 2\sqrt{\frac{A_e}{\pi}} \quad (1)$$

where,

D is an equivalent diameter of a square cross section toroid
 A_e is the cross sectional area of the toroidal core

For the core length, one may substitute the effective path length of the toroid divided by the number of windings. The effective permeability is either derived from a chart [4] or for technical toroids from the expression [5]

$$\mu_{eff} = 2.5 * \left(\frac{l_t}{2n} \sqrt{\frac{\pi}{A_e}} \right)^{1.45} \quad (2)$$

where,

l_t is the magnetic length of the toroid and
 n the number of windings (2 for two-phase, 3 for three-phase).

To consider the influence of angular coverage of each winding and the reduction of effective permeability by the opposite windings the correction factor AC was derived from experimental results, carried out on a toroid with two windings [Fig. 11].

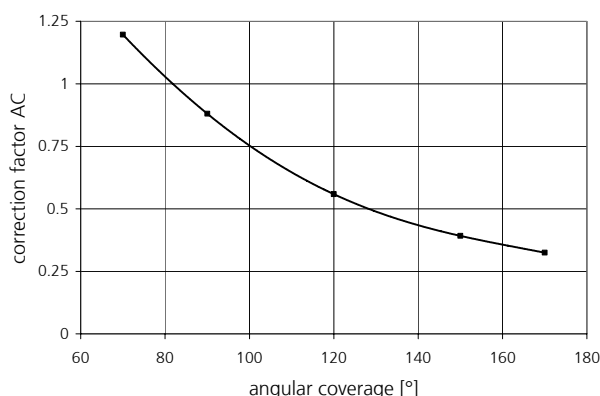


Fig. 11: Correction factor considering the angular coverage of winding for common mode inductors

This factor will be topic of additional publications.

To compare the simulated results with measurements the test

setup shown in Fig. 5 is utilized. Fig. 12 shows the results in two directions in space.

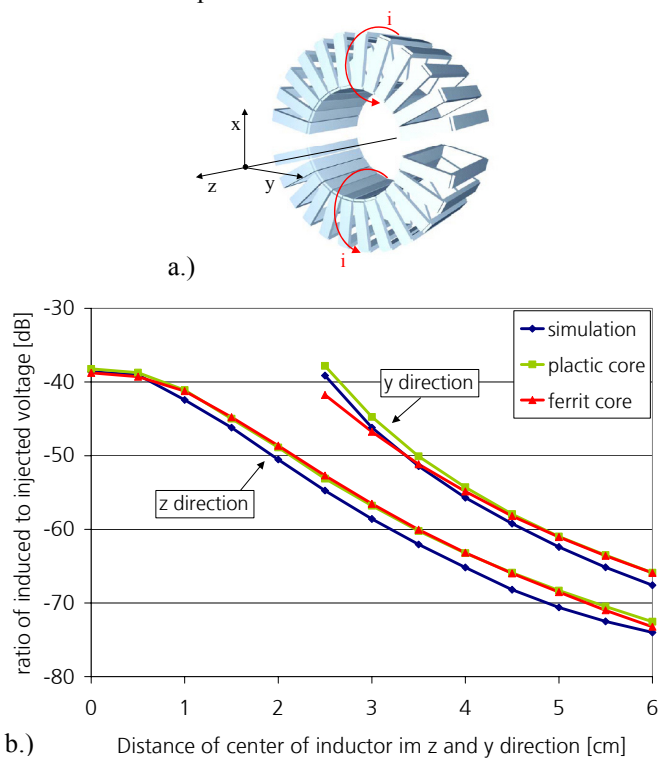


Fig. 12 a.) space directions, b.) Simulated and measured stray fields, measured with plastic and ferrite core

The comparison of a plastic and a ferrite core inductor proves the initial assumption, that ferrite material does not alter the direction of stray fields of technical components significantly. In z direction there are no differences between both, while in y direction in proximity of the inductor differences are visible. The small deviation between plastic core measurement and simulation indicates the need to refine the model slightly.

IV. EXAMPLE: INTERACTIONS IN EMI FILTERS

From our experience EMI filters are the components affected primary by electromagnetic interactions in high power density inverters. As example for the prediction of interaction we chose an EMI filter comparable to the one published in [1], consisting of a common mode choke and two capacitors, built up on a printed circuit board.

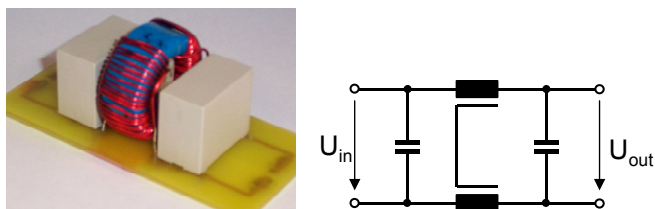


Fig. 13: EMI filter as test object with two capacitors (1.5 μ F) and a common mode choke (5mH)

First level of filter performance prediction is the simulation with ideal components, leading to the transfer function shown in Fig. 14.

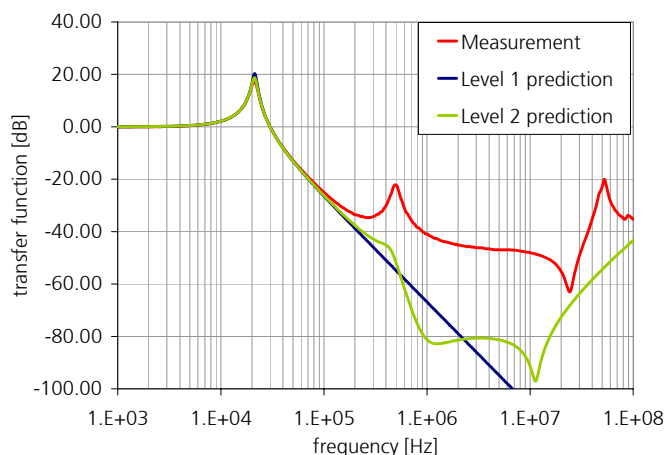


Fig. 14: Prediction of filter performance with ideal components (level 1) and with added parasitics (level 2)

The function is calculated and measured with a ideal voltage source at the input and a 50 ohm resistor at the output. As measurement instrument the HP4194A is used.

Second level prediction includes the parasitic properties like equivalent series resistance and inductance (ESR and ESL) for the capacitor, the frequency dependent behavior of ferrites and winding capacitance for the inductor and inductance and resistance of lines. This stage of prediction is state of the art, nevertheless there are severe differences between predicted and measured performance (Fig. 14).

The methods presented in this paper are used for a third level prediction, regarding the magnetic coupling of components. The PEEC simulation is carried out with the model shown in Fig. 15, generating 12 inductances and resistances and 66 mutual inductances. The calculation time is about 4 seconds on a CPU with 1.6 GHz, which is far less than for any three dimensional finite element method. The equivalent circuit elements are transferred to a circuit simulator to simulate the filter performance.

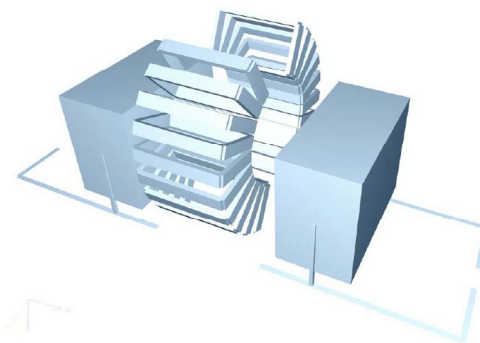


Fig. 15: Three dimensional model of filter for PEEC simulation

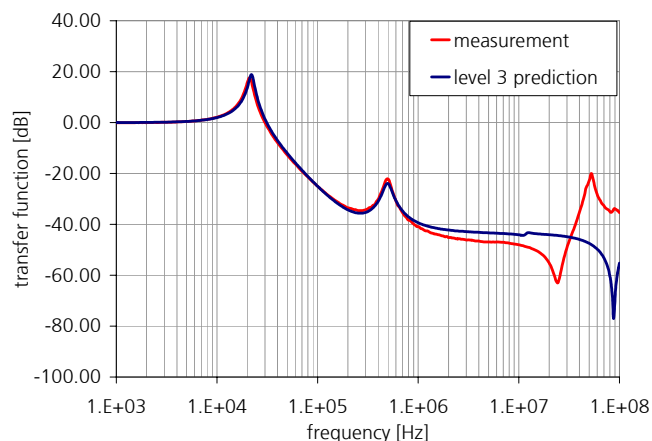


Fig. 16: Level 3 filter prediction including magnetic coupling, capacitive coupling neglected

The simulated performance meets the measured result, except for the frequency range above 20 MHz. The reason for the difference are the neglected capacitive couplings, in this example the capacities of each turn of the choke to the ferrite core. The good correlation shows the way how to predict electromagnetic interaction in high power density devices with a reasonable effort.

V. CONCLUSION

In this paper the Partial Element Equivalent Circuit (PEEC) method is suggested for the prediction of electromagnetic interaction in high power density devices. Using this method complex three dimensional setups can be handled and it presents the results in form of equivalent circuits. For the main drawback, the missing ability to consider inhomogeneous permeability, a solution is developed. The methods are tested on passive components, comparing simulated and measured results. An EMI filter as a device very sensitive to electromagnetic interaction is examined to prove the capability for prediction of interaction on system level. The promising results show the suitability of the approach to accelerate the development of high performance power electronic devices.

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