

Integrating magnetic component design and optimization into circuit simulator SIMPLORER

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Abstract - A new design and optimization methodology is presented for magnetic components used in switched mode power supplies for welding applications which links the system level simulator SIMPLORER and a magnetic component design and optimization tool, called CAEOMAG.¹

In this contribution the methodology underlying the new software tool is described, its benefits are summarized and selected example designs are given.

I. Introduction

Manufacturers of switched mode power supplies for welding applications are facing on the one hand a more competitive market and on the other hand a fast advance of technology. To maintain market share, the reduction of the time-to-market is inevitable. This was the motivation to set up a knowledge based system in the framework of the Brite-Euram-Craft project „Integrative CAE-tools for optimized development of welding power supplies with high power density - “CAE - WPS“ supporting topology preselection with respect to input / output specification of welding power supply to be developed, their DC and AC analysis for power circuit and controller design. The methodology of this knowledge based system is explained in [1].

Pre-parametrized simulation sheets are generated representing the starting point for the simulation aided iterative design. Design alterations of the passive components are accelerated and the choice of semiconductors are supported by a library so that iterations to attain a desired circuit behaviour are speeded up.

The only information available through a CAE design process is the electrical specification for magnetic devices comprising values for

1. inductances,

1. Funded by the European Community, Contract N° BRST-CT98-5310 (DG 12 - HIAS), Integrative CAE-tools for optimized development of welding power supplies with high power density“, For information search for „CAE-WPS“ in „http://dbs.cordis.lu/EN_GLOBALsearch.html“.

2. turn ratios,
3. allowable / desired leakage inductances,
4. allowable self and coupling capacitance,
5. the stress quantities, winding voltages and currents,

thermal and topological information like

1. number of windings,
2. unipolar or bipolar flux excursion.

But no information is available about the physical design of the magnetic components. State of the art CAE-tools for circuit simulation and design like SIMPLORER [6] or PSPICE [7] do not offer an embedded physical design and optimization of magnetic components. However, the information about the physical design is necessary to judge a circuit when compared. Additionally, magnetic components need to be designed individually with regard to their electrical, topological, thermal and mechanical specification. Without regarding the individual physical realization of a magnetic component at an early development stage unfavourable and unrealistic specifications occur. These disadvantages of existing tools are avoided by the new design and optimization tool **CAEOMAG**, which is integrated into the circuit simulator SIMPLORER.

Its structure and data flow, methodology and mathematical background and resulting benefits are described in this paper.

II. Structure and data flow of CAEOMAG

The data flow between CAEOMAG and SIMPLORER is illustrated in Fig. 1. CAEOMAG is called by a *graphical user interface* implemented as *SIMPLORER Wizard*. The Wizard belongs to a specific component topology. Currently, the following components are distinguished:

1. Linear, gapped inductors (Lopt).
2. linear two-winding transformers. (topt2W)
3. linear three-winding transformers (topt3W)

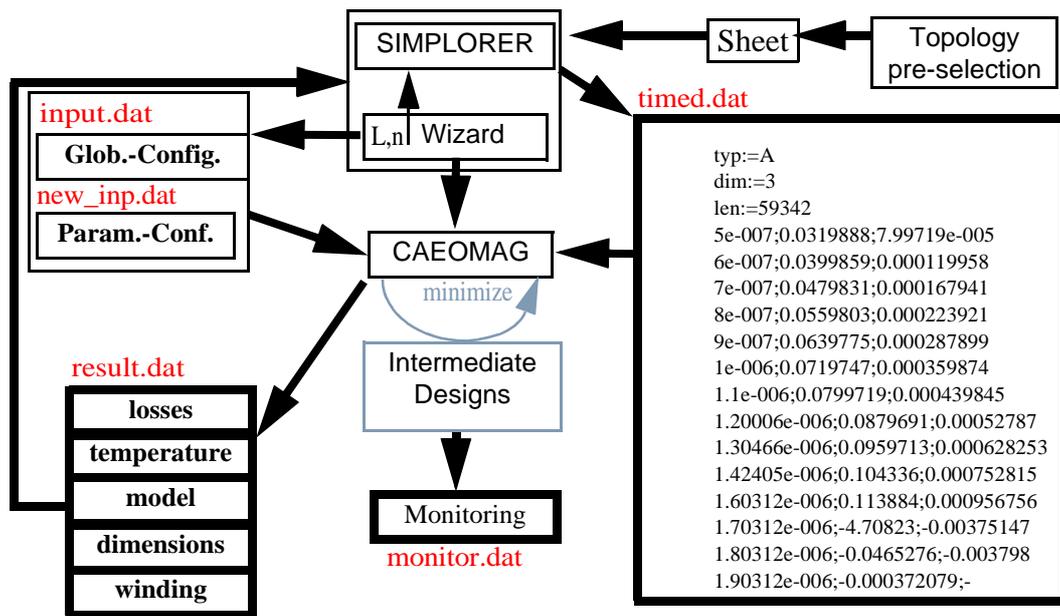
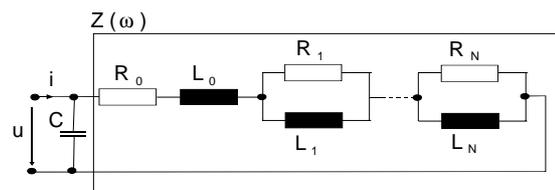


Fig.1:Dataflow between SIMPLORER and CAEOMAG.

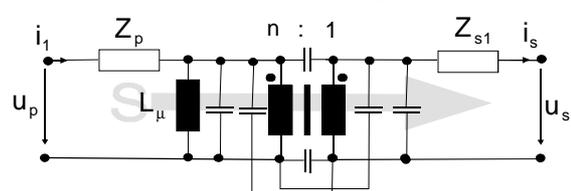
Each Wizard belongs to a circuit model, as indicated in Fig. 2. They consider eddy current effects in the winding and capacitive effects. Each impedance in the circuits is approximated by a partial fraction network of a linear inductor. The impedances are calculated from a quasi-stationary field analysis (only skin effect without displacement current) and are approximated by the network using a least square method for parameter fitting. The capacitive effects are accounted for by capacitances parameterised by an electrostatic analysis. For inductors, the self-capacitance is determined and for transformers the coupling capacitances are used for deriving the network elements between to windings depicted in Fig. 2 b).

The higher order elements ($n > 0$) and the parasitics can only be determined accurately once the first physical design is generated. Therefore, to predict the influence of the component on the circuit approximate values of the circuit properties are assumed for the first run. These are further refined iteratively by updating the model subsequently. Hence, the models generated by CAEOMAG regarding all parasitic elements are fed back to the simulator in order to fine tune the results. Before starting CAEOMAG a simulation has to be performed yielding the stress quantities at the terminals of the component. They are collected in a column oriented file (timed.dat) and represent thus time domain data. Due to the fact, that CAEOMAG uses frequency domain methods, a harmonic continuation is performed. This requires the selection of a time interval starting from the end of the simulation record used for

a) Circuit model of a linear inductor



b) Circuit model of a two winding transformer



c) Circuit model of a three winding transformer (Note: Capacitances are left out for brevity)

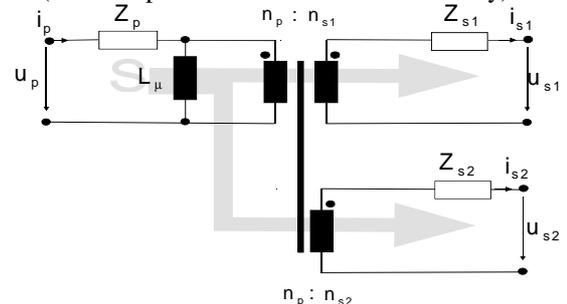


Fig.2:Circuit models of the Wizards.

fourier transform and harmonic analysis (time_from_end). It should reflect the steady state con-

dition of the circuit under investigation.

Usually, an unexperienced design engineer cannot conclude from stress quantities and circuit parameters to a physical design.

To ease the search for a good starting point the design and optimization tool allows a pre-optimization (PRE-OPT) of the device with respect to simulated stress quantities. This yields a core choice and starting values for the parameter model used by the succeeding parameter optimization [2],[3]. The parameter model allows the automated optimization of e.g. the layer thicknesses or core dimensions with respect to an objective function, which can be size, costs, etc. and constraints like temperature rise, maximum saturation induction etc.. Free design parameters can be selected from a menu. State-of-the-art genetic algorithms and a heuristic search method are used for optimization. They evaluate the core-winding set-ups representing intermediate designs automatically and vary parameters during an optimization in order to minimize an objective function to user demands.

III. Mathematical background and methodology

In order to apply an unconstrained optimization algorithm the objective function has to summarize the results of a calculation model into one value, which is minimized by the optimization.

The developed calculation models predominately use analytical methods in order to reduce the computation time for the evaluation of the objective function.

The following models are used:

1. A model for the winding: Resulting impedance matrix and winding losses.

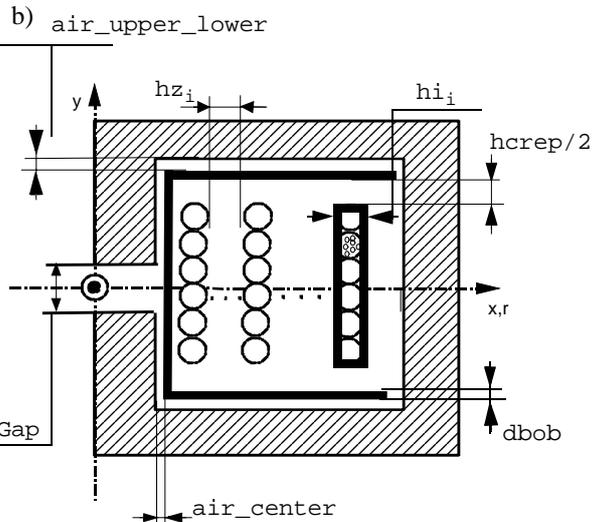
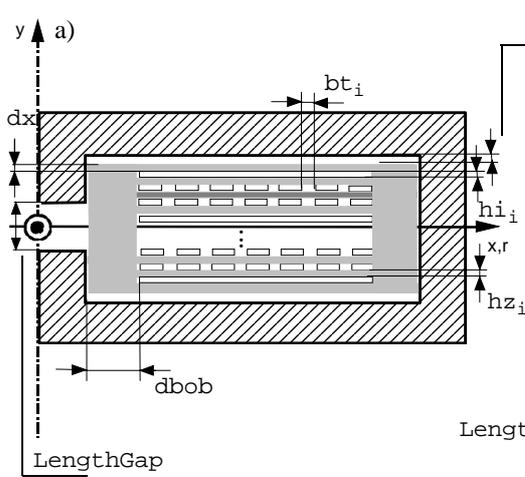


Fig.3: Cross-section of the supported winding structures: a) planar and b) solenoidal.

2. A model for the core: Resulting magnetic circuit and core losses.
3. A capacitance model: Resulting Self- and Coupling capacitances.
4. A thermal model: Resulting peak temperature rise at free and forced convection conditions.
5. An equivalent circuit model

The different models are described in the following subsections.

III.1 Model of the winding

Two different winding structures are supported: planar and solenoidal (see Fig. 3). A one dimensional field calculation in the frequency domain regarding eddy current losses is applied first ([2], [4]). This yields an impedance matrix of the component

$$\underline{Z}(j\omega) = \begin{bmatrix} R_{11} + j\omega L_{11} & \dots & R_{1L} + j\omega L_{1L} \\ \dots & \dots & \dots \\ R_{L1} + j\omega L_{L1} & \dots & R_{LL} + j\omega L_{LL} \end{bmatrix}. \quad (1)$$

The elements are calculated as follows:

1. Layer resistances:

$$R_{ii} = R_{DC,i} \cdot F_{sR_i}^{(M)} + N_i^2 \sum_{w=1}^{i-1} \frac{R_{DC,w}}{N_w^2} F_{pR_w}^{(M)} \quad (2)$$

2. „Proximity“-resistances:

$$R_{ij} = \frac{N_j}{2N_i} R_{DC,i} \cdot F_{pR_i}^{(M)} + N_i N_j \sum_{w=1}^{i-1} R_{DC,w} \cdot F_{pR_w}^{(M)} \quad (3)$$

3. Self inductances of the layers:

$$L_{ii} = \Lambda_K N_i^2 + N_i^2 \sum_{w=1}^{i-1} \Lambda_{Z,w} + L_{DC,i} \cdot F_{sL_i}^{(M)} + N_i^2 \sum_{w=1}^{i-1} \Lambda_{L,w} \cdot F_{pL_w}^{(M)} \quad (4)$$

4. Mutual inductances:

$$L_{ij} = \left[\begin{array}{c} \Lambda_K N_i N_j + N_i N_j \sum_{w=1}^{i-1} \Lambda_{Z,w} \\ \frac{N_i N_j}{2} \Lambda_{L,i} \cdot F_{pL_i}^{(M)} + N_i N_j \sum_{w=1}^{i-1} \Lambda_{L,w} \cdot F_{pL_w}^{(M)} \end{array} \right] \quad (5)$$

Here, R_{DC} is the DC-resistance, L_{DC} the DC-self-inductance, Λ_K the DC-reluctance of the core Λ_L and Λ_Z the DC-reluctance of layer and insulation.

The frequency dependent factors consider the following eddy current effects:

1. Power loss increase due to *inner* current density displacement: $F_{sR}^{(M)} (\geq 1)$
2. Power loss increase due to induced eddy currents caused by other layers: $F_{pR}^{(M)} (\geq 0)$
3. Decrease of stored magnetic energy due to inner current density displacement: $F_{sL}^{(M)} (\leq 1)$
4. Decrease of stored magnetic energy due to induced eddy currents caused by external fields: $F_{pL}^{(M)} (\leq 1)$.

The upper index M indicates the conductor type, implying different formulas for all wire types: Foil, Litz-wire and round wires.

Accounting for edge effects is performed by manipulating the skin-effect factor. FEA parameter studies similar to those described in [9] were performed for analysis of foil windings in solenoidal components and multi-turn windings in planar structures. Due to high currents in windings of magnetic components in the power train of welding applications usually single turn or in maximum 2 turn windings will be used in planar transform-

ers, so that the inaccuracies of the onedimensional field solution is not significant.

Gap effects are caused by the fringing of the magnetic field in the core when passing through air gaps. On the one hand, the reluctance of the airgap increases due to increase of effective cross section area, on the other hand the fringing field penetrates the layers in the winding window causing additional losses at higher harmonics of the inductor current. To account for these effects in the electrical model of inductors, a method derived in [2] and [3] was implemented for linear gapped, solenoidal inductors. It utilizes the current sheet technique and a separate field solution for the induced fields to compute the additional losses. They are superimposed to the losses determined from the method neglecting airgap effects. Again, windings of foils, round wires and litz wires are distinguished.

III.2 Model of the core

The magnetic circuit of the core is analyzed using the methods presented in [2]. The core is divided into straight and curved segments and the reluctances of these segments are calculated using mean cross section and path length formulations. The flux density and losses in all these segments are calculated and summarized to the total core losses. Rate dependencies of the losses are considered according to the method presented in [5]. The influence of airgap fringing on the gap reluctance is considered by applying conformal mapping techniques [2], yielding accurate formulas to predict the losses for legs with a circular or rectangular cross section.

III.3 Thermal model

After the heat sources are determined a thermal model is used to predict the temperature rise. From the Nusselt-numbers (free convection) and Reynold-numbers (at forced convection) of each surface facing the ambient thermal resistances core-to-ambient and winding-to-ambient are derived.

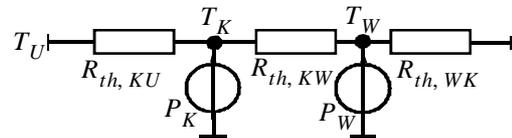


Fig.4:Two node thermal model.

Together with the heat transfer by conduction between core and winding an equivalent two node circuit is established (see Fig. 4). The core and winding temperature are understood as mean temperatures. The „temperature ripple“ in the core is neglected whereas the radial temperature gradient in the winding is considered by a layer model.

IV.1 Capacitance model

The capacitance model uses a plate capacitor approximation assuming a radial electric field for solenoidal and an axial for planar components (see Fig. 5).

IV.2 Objective Function and parameter optimization

The objective function collects the information generated from the model and summarizes the results in *one* scalar value.

$$F(\mathbf{x}) = \frac{1}{W} \left(w_1 \frac{P_{v,ges}(\mathbf{x})}{P_{v,ges}(\mathbf{x}_0)} + \dots + \sum P(g(\mathbf{x})) \right) + \sum \frac{Y_i(\mathbf{x})}{(w_i Y_i(\mathbf{x}_0) + w_K Y_i(\mathbf{x}_0))}$$

The model values $Y_i(\mathbf{x})$ are weighted with user defined weighting factors w_i and are superimposed to the penalty function $P(g(\mathbf{x}))$ punishing senseless parameter configurations or constraint violation resulting in the objective function $F(\mathbf{x})$ to be minimized.

The objective function is called by the parameter optimization algorithm, passing the free parameter vector \mathbf{x} and constant parameters to the objective function. Obviously, the objective function can be used for the evaluation of a given design if no parameter is varied by the optimizer. Typically, the parameter vector has

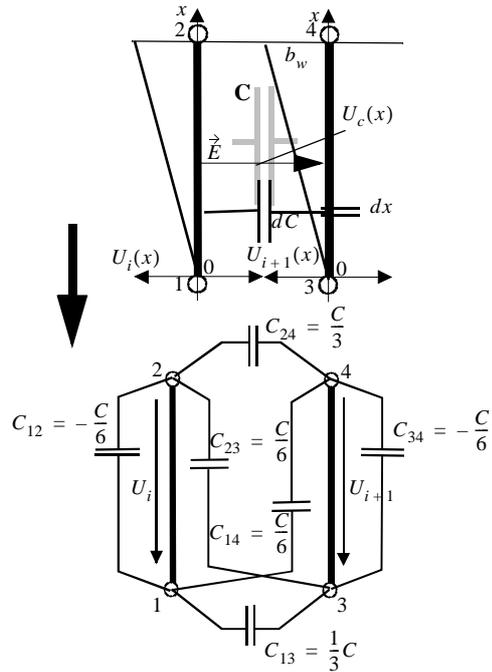


Fig.5: Plate capacitance model and circuit model. the structure $X = [d, l_L, \text{Core-geometry}]$ for inductors and $X = [d_1, d_2, \dots, \hat{B}, \text{Core-geometry}]$ for transformers. An evolutionary strategy using floating point number representation is used, representing a non deterministic global optimizer whereas the NELDER-

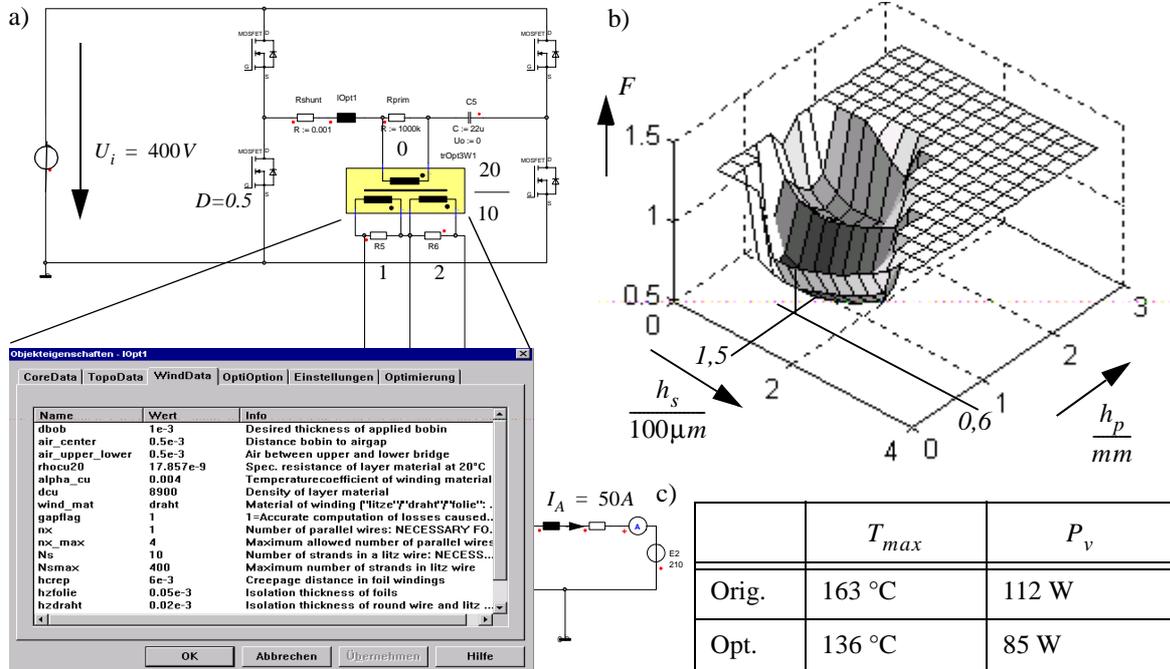


Fig.6: a) A part of a SIMPLORER-worksheet of the 5 kW-welder (10 kW peak) and the Wizards for a transformer b) Result of a parameter study and c) comparison of optimum and current design.

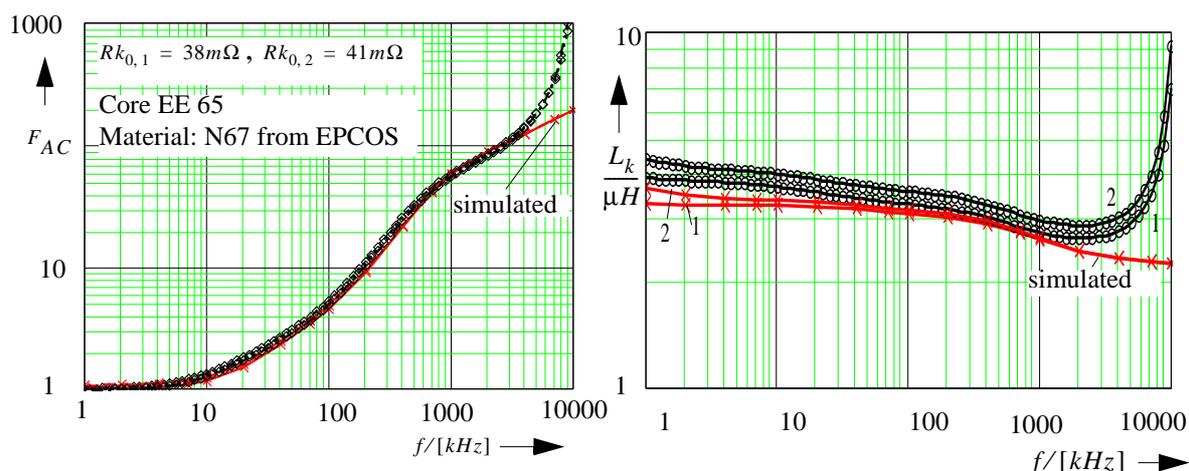


Fig.7: Measured and simulated resistance ratio of 5 kW - transformer design example and its winding data.

MEAD method is applied representing a deterministic local minimizer, which finds the next minimum efficiently and most exactly.

VI. Experimental results

Fig. 6 a) shows a schematic of a power stage of a 5 kW (10 kW peak) welding power supply and the user interface of the transformer design and optimization together the objective function (see Fig. 6 b)) with a constant core but variable primary and secondary conductor thickness. Note the two major subminima caused by the discontinuous change in parallel connected wires to fill the window as good as possible. Losses of the original design are reduced by at least 24 %, which leads to a temperature decrease of 27 °C. The model accuracy compared to measurements of the original transformer is depicted in Fig. 7 using the short circuit impedance as model quality indicator. The model reveals good agreement with the measured values as well for the resistance as for the inductance. Please note, that the winding consists of two different winding types, which complicates the modelling tremendously. The matrices Cs and Cp code the series and parallel connection scheme of the layers. Entries of 1 in the i-th and j-th column of a row in Cs indicate a series connection of these layers. Analogous accounts for the parallel connection.

VII. Conclusion

In this contribution the methodology and results of a simulator-coupled magnetic component design and optimization tool are outlined. Modelling results of a welding power supply transformer are presented and its optimization shows a power loss reduction for a selected transformer of at least 24 %.

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