

SIMULATION ASSISTED DESIGN OF INDUCTORS IN POWER ELECTRONICS SYSTEMS

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ABSTRACT

Inductors are main components in power electronics systems, e.g. switching mode power supplies used in grid-connected inverters for photovoltaic systems. Inductive components mainly contribute to losses, therefore there is a big potential to increase the overall efficiency. In addition to the electric requirements due to the system architecture, energy efficiency and thermal behavior are the main aspects for an optimal inductor design. To achieve this it requires an accurate prediction of the occurring electromagnetical and thermal effects and their mutual interactions. The design procedure is a rather complex task and the result is highly dependent on the experience of the design engineer. To reduce the time to market and in order to standardize the design process, a computer aided design tool was developed. The aim is an accurately and fast prediction of the inductors physical behavior and the optimization of all significant design parameters.

Keywords: power electronics systems, inductors, energy efficiency, model-based design, optimization

1. INTRODUCTION

A main aspect in the design of switching mode power supplies is to increase the operating frequency and to reduce the size of the inductive components, e.g. transformer and variable chokes, and therefore decrease the energy density and the overall size of the product. As a consequence high frequency effects like eddy currents inside the conductor and increased core losses due to non-sinusoidal current waveforms become the major challenge to achieve an effective design of the required inductive components.

2. MODELING PROCESS OF MECHATRONIC MULTIDOMAIN SYSTEMS

For an optimal design of power electronics systems the coupling of several physical domains has to be taken into consideration. Energy efficient designs mostly depend on multiple physical effects and their mutual coupling, as shown in figure 1. Therefore, the design process of mechatronic systems is a rather complex and time-consuming challenge which also needs a lot of experience.

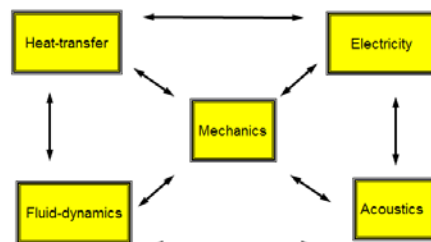


Figure 1: Mutual coupling of physical domains

Finding an optimum system configuration in a short time is a particular challenge to the engineer. There are a lot of parameters that need to be optimized in an early stage of the design process. In this situation a computer aided (automated) design tool could help to speed-up the design process (see figure 2, green line).

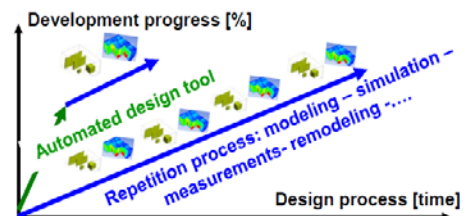


Figure 2: Automated design tool

3. INDUCTIVE POWER ELECTRONIC COMPONENTS

In figure 3 a typical topology of a switching mode power supply being part of power electronics system is shown.

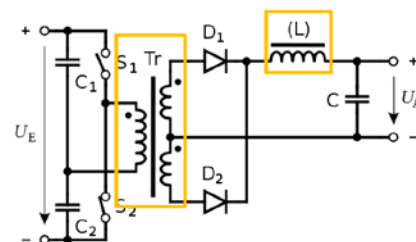


Figure 3: Topology of a switching mode power supply with inductive components

In the Push-Pull converter topology depicted in figure 3, there is a lot of potential for optimization, e.g. capacitors, power electronics semiconductors, noise-filters and inductive components.

As a matter of fact, inductive components mainly contribute to losses and therefore there is a great potential to increase the overall energy efficiency. The requirements to optimize inductive components are:

- Taking into account electromagnetic and thermal effects
- Understand the power loss mechanisms in the inductor winding and the magnetic core
- Consider the mutual interactions of physical effects

4. POWER LOSS MECHANISMS

In the design of high frequency inductive components, e.g. output choke, performing multi-field simulations of electro-magnetically, thermal and possibly structural field interaction is inevitable. For an optimal design of such devices one has to understand the mechanisms of losses within the device and the interaction with electrically connected components.

A main element in the design of chokes is the calculation of the heat losses, due to transformation of part of the electric energy into thermal energy and to determine the resulting temperature-rise of the choke. In practice energy is dissipated due to the resistance of the windings (known as winding or copper losses) and due to magnetic effects mainly attributable to the core (known as core or iron losses). Figure 4 shows the total losses occurring inside an inductor and how they are calculated. It is split up into losses occurring inside the winding and the magnetic core:

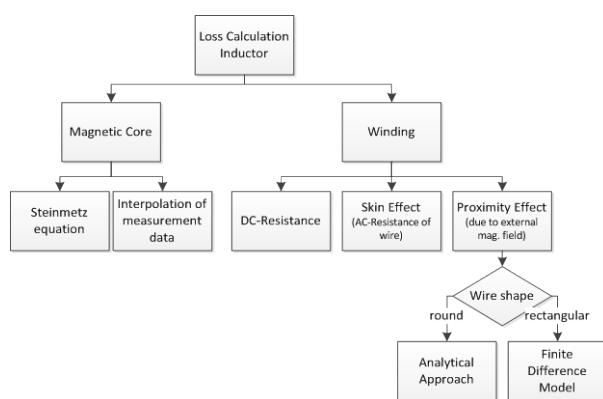


Figure 4: Inductor loss calculation

4.1. Core Losses

Ferromagnetic materials used in chokes typically are magnetically soft and preferable have a low coercive field strength and low power losses. Commonly used materials are based on iron, cobalt and nickel and its alloys and can be divided into polycrystalline and amorphous materials. The proper selection of the

magnetic material depends on both the magnetic characteristics like saturation flux density and power loss and more important price. Most of the data is provided by the manufacturers or can be calculated using formulas describing measurement data.

Core or iron losses can be split into hysteresis- and eddy current losses. Hysteresis losses are related to the energy loss in each alternating-cycle of the magnetic field in the core. Whereas induced eddy currents cause resistive heating in the magnetic core. The main parameters of core losses are the magnetic flux density B , the operating frequency f and the temperature T . A well-known and frequently used analytical method calculating core losses is the equation of Steinmetz (Eq. 1) (Reinert 2000).

$$P_v = c_m f^\alpha B^\beta g(T) \quad (1)$$

The parameters c_m , α , β and temperature polynomial $g(T)$ are provided in datasheets or are identified through fitting characteristic curves. The Steinmetz equation is valid for sinusoidal current waveforms without dc-bias. A method to calculate core losses for arbitrary current waveforms is presented in (Albach 1996). To consider the influence of dc-bias measurement data of 3F3 Ferrite material presented in (Brockmeyer 1996) is normalized on the saturation flux density and used for all other ferrite materials in the calculation routine.

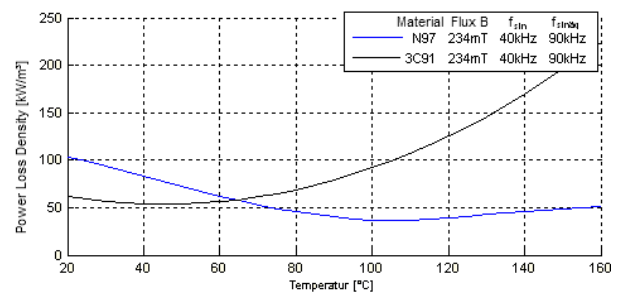


Figure 5: core losses of ferrite material N97 & 3C91

Figure 5 shows the power loss density of the two different ferrite materials N97 (Epcos) and 3C91 (Ferroxcube), excited with the same flux density. The material 3C91 has its power loss minimum at around 50°C whereas the material N97 is best used at 105°C.

4.2. Winding Losses

Winding losses occur because of the electric current flowing through the windings of the inductor. Winding losses can be split up into regular dc-losses and ac-losses caused by induced eddy-currents (see figure 4 and 7). The determination of dc-losses is rather easy. They result from the dc-resistance of the wire and the rms-value of the current flowing through it.

The alternating current produces a time-varying magnetic field, which again induces eddy currents that

try to oppose the magnetic field. If those eddy currents occur inside the wire itself, they oppose the original current and therefore increase its ac-resistance. This is called skin-effect, and can be calculated for round wires using Bessel-functions (Albach 2000).

The increase of ac-resistance because of eddy currents induced by currents flowing in nearby wires and by the fringing field of air gaps is called proximity-effect.

Skin- and proximity-effect depend on the frequency of the current flowing and the conductivity of the wires. Proximity-effect usually is considerable larger than skin-effect and in most cases can't be neglected in the design process. To calculate the losses for alternating currents two different approaches are used depending on the type of cross-section of the wire.

Round wires structures:

The winding losses for round winding structures are determined using a two-dimensional analytical calculation procedure as described in (Albach 2000). First the magnetic field inside the winding window due to the air gaps is calculated. To reduce the calculation area only to the winding window we replace the air gaps by linear current loads, which produce the same field inside the winding window as the fringing flux of the air gaps. Figure 6 shows a cross-section of a typical EE-core configuration.

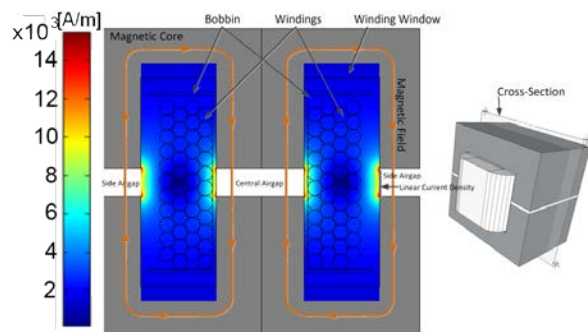


Figure 6: magnetic field strength of 1st harmonic inside winding window of typical EE-core configuration

The following steps lead to the proximity losses in each wire:

1. Express the external field strength on the wire boundary as function of radius r and angle φ (transformation of coordinates)
2. Transform field strength into Fourier series
3. Solve boundary value problem of Helmholtz equation for a current carrying conductor with arbitrary distribution on its boundary to find field strength and current density distribution inside the conductor
4. Calculate power loss by means of the Poynting vector

The detailed calculation procedure is described in (Albach 2000). Figure 7 shows the ratio between ac-losses and dc-losses. The influence of the air gap on the

wires next to it leads to a significant increase of total power loss inside the winding.

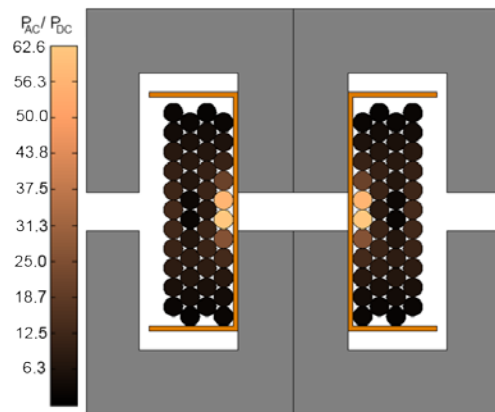


Figure 7: Ratio between AC-losses and DC-losses

Rectangular wire structures:

The field problem arising from rectangular wires is solved using a finite difference model. The finite difference method is convenient, because calculation domain requires a regular quad-mesh discretization, which is good applicable for rectangular wires but not for round wires. If the general formulation of the finite difference method with variable step size is used, a grid as shown in figure 8 is generated.

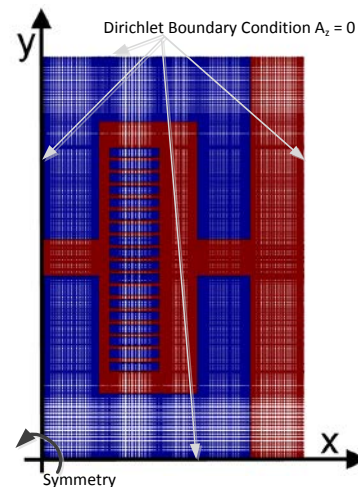


Figure 8: mesh of a typical EE-core configuration for FDM-model

The partial differential equation being solved is the Maxwell's amperes law for the time-harmonic case:

$$j\omega\sigma A_z + \nabla \times \mu_0^{-1} \mu_r^{-1} \nabla \times A_z = J_z^e \quad (2)$$

with ω as the angular frequency, σ as the electrical conductivity, μ_0 as the permeability of vacuum, μ_r as the relative permeability and j as the imaginary unit. The external current density J_z^e is the source for the magnetic field and has to be chosen so that the total current (sum of induced currents and external current)

in every wire is the net current flowing through all wires. The resulting linear equation system in form of

$$A x = b \tag{3}$$

where x is the unknown vector potential A_z , A is the coefficient matrix and b is the source vector. Since the coefficient matrix is sparse and band, the equation system is solved with an iterative solver using the quasi-minimal residual method and an incomplete LU-preconditioner. The magnetic field strength results from the magnetic potential

$$\nabla \times A_z = B = \mu_0 \mu_r H \tag{4}$$

The induced current density is calculated by

$$j \omega \sigma A_z = J_{ind} \tag{5}$$

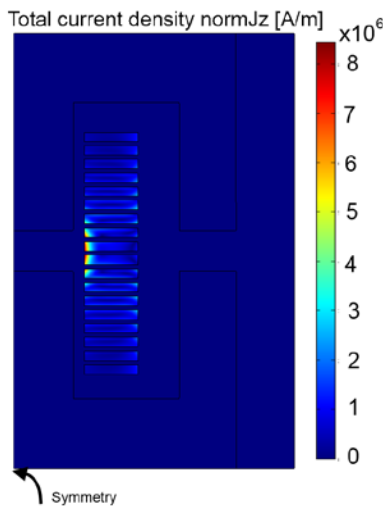


Figure 9: Total current density of EE-core configuration

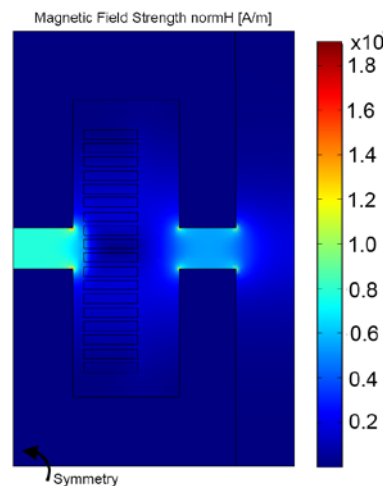


Figure 10: Magnetic field strength of EE-core configuration

Figure 9 shows the resulting total current density distribution inside the wires. The calculated field strength is shown in figure 10.

5. THERMAL MODELING

Optimum heat transfer in electronic enclosures, done by conduction, radiation and convection is of particular interest in the design process of power electronics in general. Especially in the design of inductive components e.g. chokes, a main element is the calculation of the heat losses, due to transformation of part of the electric energy into thermal energy and to determine the resulting temperature- rise of the choke. The losses depend on the magnetic flux density B , the frequency f and the Temperature T as shown in Eq. 1. These effects do interact with each other and it is necessary to solve the coupled equations iteratively as explained in chapter 6. Even the thermal calculation itself is to be solved iteratively, because of the temperature dependent heat transfer coefficients and material properties.

In addition to selecting one of the three possible mounting positions displayed in Figure 12, the following calculation-options can be chosen in the 'Choke calculator' design tool:

- Natural Convection
- Natural Convection and Radiation
- Forced Convection
- Forced Convection and Radiation
- Manual Input of Heat Transfer Coefficients

To balance the heat fluxes, a network of thermal resistances was generated as shown in Figure 11 for 'E-core shapes' and in Figure 13 for 'U-cores shapes' for example.

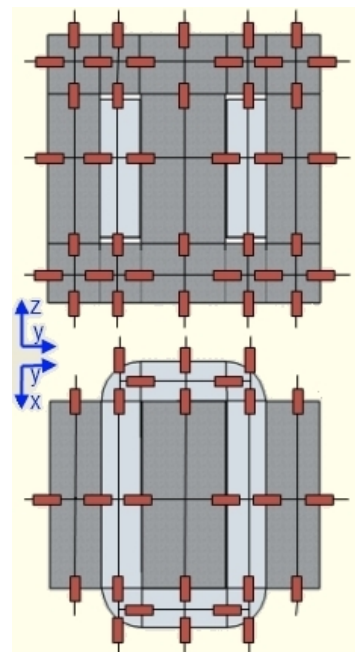


Figure 11: Network of thermal resistances ('E-core')

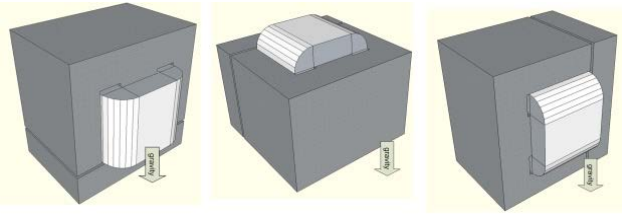


Figure 12: Mounting positions

The thermal resistances are calculated, using empirical equations given in (VDI 2006). So the thermal model of the ‘E-core shapes’ has an accuracy of 21 temperature-nodes and the model of the ‘U-cores’ consists of 26 nodes.

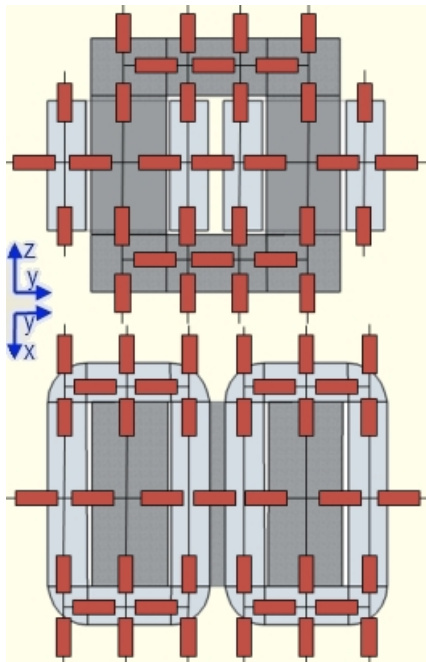


Figure 13: Network of thermal resistances (‘U-core’)

In Figure 14, the circuit diagram of these network models with n - Temperature nodes is shown. This model leads to a system of equations, generated by the node potential method. Then the set of equations is solved iteratively because of the temperature dependency of the thermal resistors.

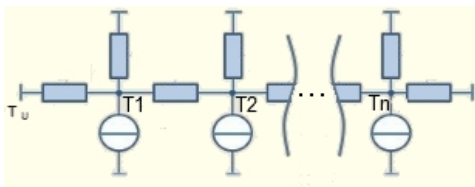


Figure 14: Circuit diagram of the thermal network model

Even when the radiation model is selected, the solution is converged in less than half a minute and show good agreement as verified e.g. in (Jungwirth 2011).

6. AUTOMATED DESIGN TOOL

To reduce the time to market and in order to standardize the design process, a computer aided design tool was developed. The tool named ‘Choke calculator’ is programmed in MATLAB with a user friendly graphical user interface (GUI). The workflow of the modeling process and the schematic design of the tool are shown in Figure 15.

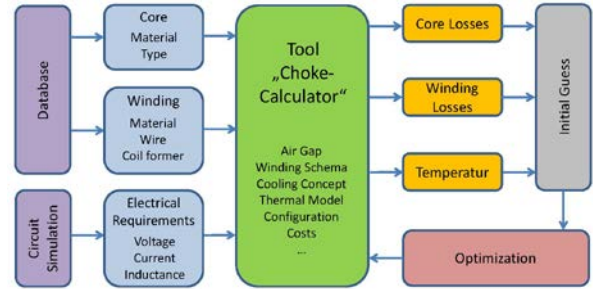


Figure 15: Workflow of the modeling process

The tool is characterized by low computing time and minimal effort in geometry modeling because of its connection to a component database. A ‘Microsoft Access’ database contains the required data of core materials, coil formers and winding materials of relevant suppliers and can be extended in any order. The construction of the database, taking into account various data formats of individual suppliers, can already be seen as a first step towards standardization. The communication to MATLAB is programmed by SQL-statements.

Figure 16 shows a flowchart of the calculation tool. The green frame marks the Electromagnetic calculation, coupled to the Thermal calculation, shown in the yellow section.

Traditionally, the design of inductors has been based on sinusoidal current waveforms operating at low frequencies. In modern switching power supplies a movement towards higher power density continues. Therefore non-sinusoidal excitation and high frequency skin and proximity effects must be considered as described in chapter 4.

The design of an inductor usually begins with the specification of the desired inductance value and the maximal required current flowing through it. Depending on the application variable current waveforms can be chosen.

The next step is a preselecting of an appropriate core. A selection of a manufacturer, the core-material, the core-shape and the core-type is possible. The dependency of the core-losses on the temperature can be plotted and a comparison of the different types can easily be done.

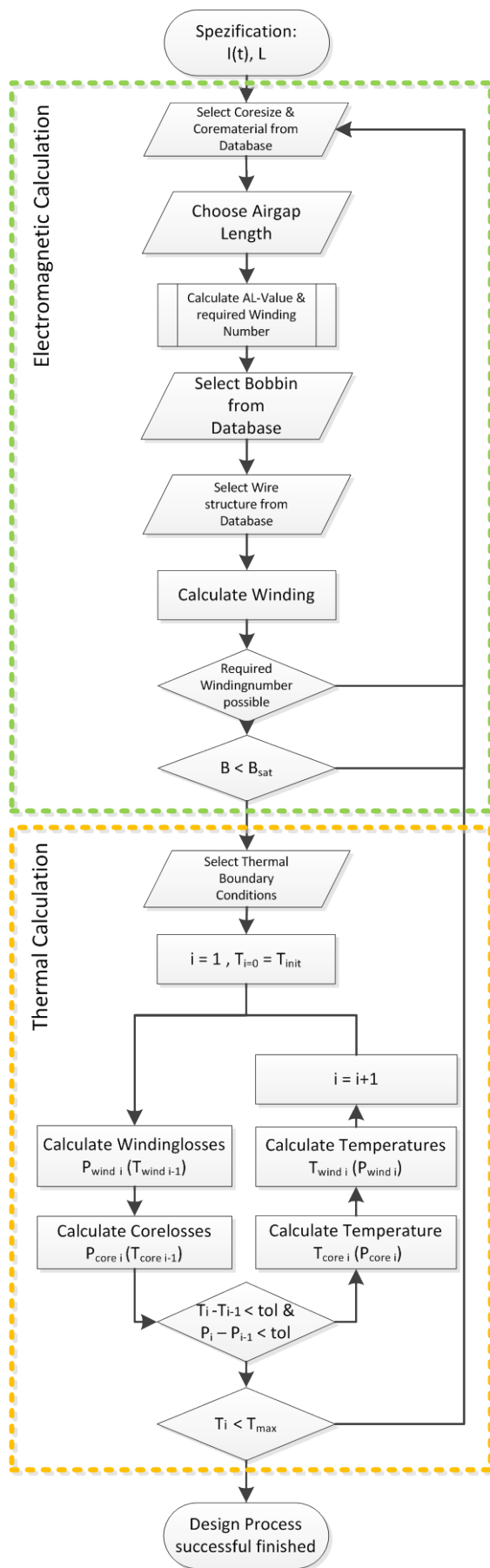


Figure 16: Flowchart of the design tool

To obtain the desired inductance value, chokes usually require air gaps to define the resistance of the magnetic circuit so it stores the correct amount of magnetic energy. To do so the size of the air gaps is varied and corresponding required winding numbers are computed.

A small air gap has the advantage that only a few turns are required to obtain the desired inductance, however, the low magnetic resistance results in a large magnetic flux. As a consequence the core might saturate and the iron loss increase. On the other hand a large air gap requires more turns, which is spatially limited by size of the winding window. So the right choice of air gap size and winding number is a tradeoff between magnitude of the magnetic flux and core losses on the one side and copper losses on the other side.

For the thermal calculation, the boundary conditions have to be defined, as described in chapter 5. After initializing by a starting temperature, the core losses and winding losses can be calculated. Then a new temperature distribution can be determined for the calculated losses of the actual iteration step. The next iteration loop can be started, where the temperature dependent losses will be calculated using the values from the previous step and so on. The iteration process is continued until the changes in temperature and power dissipation are smaller than a specified tolerance.

The GUI of the design tool is intuitive to use and subdivided into the following six major steps

- Specification
- Selection of the core
- Selection of the winding
- Electromagnetically calculation
- Thermal calculation
- Post processing

Figure 17, shows for example the visualization of the winding calculation.

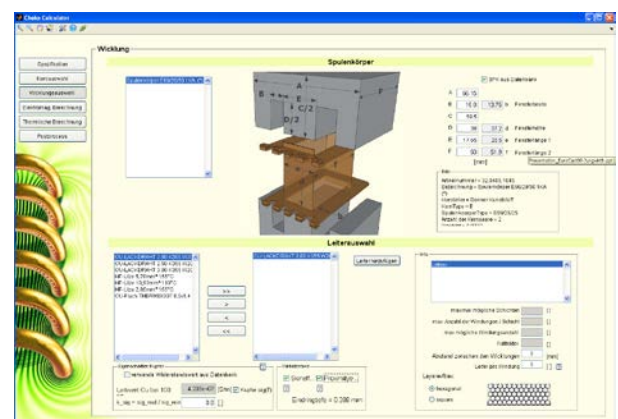


Figure 17: GUI of the design tool

7. VERIFICATION STUDY

In order to describe the benefit of the easy-to-use design tool, a typical workflow from specifications to optimal results is presented.

A choke as shown in figure 18 should be optimized.

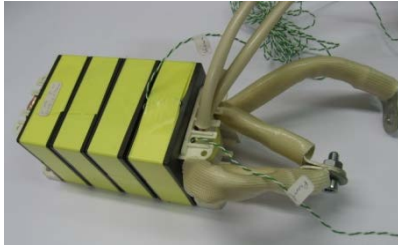


Fig 18: Original choke to be optimized (courtesy of Fronius International GmbH)

Firstly, the following specifications are given:

- required inductance $L=180\mu\text{H}$
- DC-current 10A / ripple-current 12A (see figure 19 for current wave-forms)
- Switching frequency $f=40\text{kHz}$

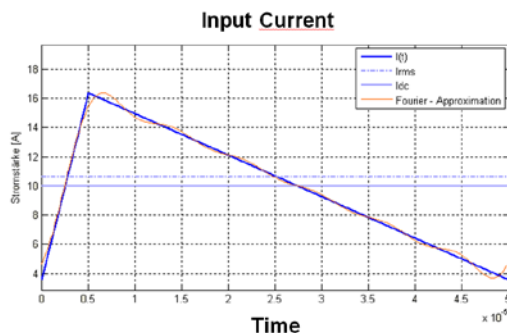


Fig 19: Input current (10A DC/12A ripple)

After setting the specifications one has to define the core-type, the type of wire and some parameter variations, e.g. variation of the air-gap (1mm to 5mm).

The calculation results are shown in figure 20. Therein, for each variation of the airgap calculations on the losses (core and winding losses), magnetic flux density and suitable winding turns are shown. The optimal result is indicated by the red line:

- Inductance $L=195\mu\text{H}$
@ 2mm airgap and $N=23$ winding turns
- Losses:
 - $P_{\text{rms}}=1,2\text{W}$
 - $P_{\text{prox}}=3,2\text{W}$

- $P_{\text{skin}}=0,1\text{W}$
- $P_{\text{core}}=0,8\text{W}$

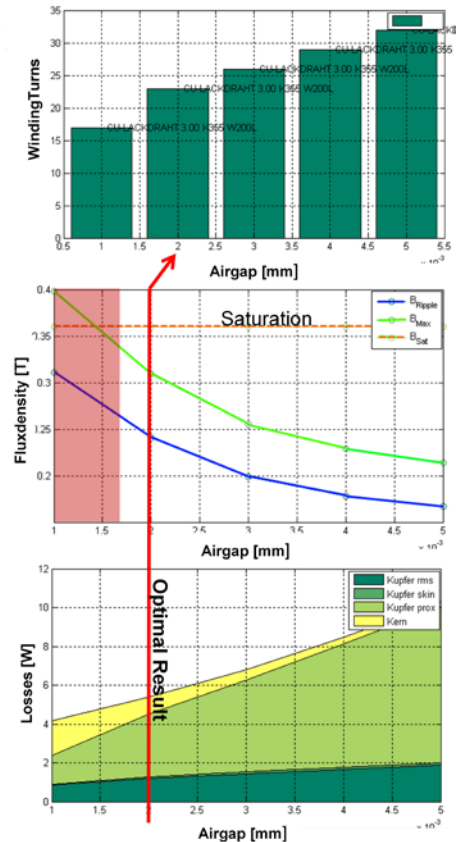


Fig 20: Optimal result of verification study (red line)

8. CONCLUSIONS AND FURTHER PROSPECTS

An automated design tool based on physical effects and material parameters for inductive components in power electronics systems was presented and verified. This tool could be very useful giving the engineer an initial guess in the early design process.

As a next step the design tool could be extended to optimize a complete switching mode power converter with capacitors, semiconductors and noise-models in the different mode of operations.

ACKNOWLEDGMENTS

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REFERENCES

- Albach, M. and Durbaum, Th., 1996. Calculating Core Losses in Transformers for Arbitrary Magnetizing Currents, *Power Electronics Specialists Conference, PESC '96 Record., 27th Annual IEEE.*
- Albach, M., 2000. Two-dimensional calculation of winding losses in transformers, *Proceedings of the*

- IEEE Power Electronics Specialists Conference*, pp. 1639- 1644, Galway, Ireland.
- Brockmeyer, A. and Reinert, J., 1996. Experimental Evaluation of the Influence of DC-Premagnetization on the Properties of Power Electronic Ferrites, *Proceedings of Applied Power Electronics Conference and Exposition*, pp. 454 - 460 vol.1, San Jose, CA
- Jungwirth, M. and Weinzierl, H., 2011. Efficient Computer Based Optimization of Thermal Management in Electronic Systems, *Proceedings of Thirteen International Conference on Computer Aided Systems Theory*, Las Palmas, Spain.
- Reinert, J., Brockmeyer, A., De Doncker, R.W., 2001. Calculation of losses in ferro- and ferromagnetic materials based on the modified Steinmetz equation, *IEEE Transactions on Industry Applications*, vol. 37, pp. 1055 – 106.
- Verein deutscher Ingenieure (VDI), 2006. VDI Wärmeatlas, *Springer*, Heidelberg.

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