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η - ρ Pareto Optimization of 3-Phase 3-Level T-type AC–DC–AC Converter Comprising Si and SiC Hybrid Power Stage

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Abstract—In this paper, the η - ρ (efficiency-power density) Pareto front of a 20 kVA Uninterruptible Power Supply (UPS). comprising a Si and SiC hybrid 3-level T-type rectifier and inverter stage, and input/output side filters, is determind. A multi-objective optimization procedure is detailed, which employs an electrical converter model and coupled electro-thermal component models. Each component model determines the corresponding losses, volumes, and temperature rises. A detailed description of the inductor model is given which considers high frequency winding losses as well as core losses, and calculates the temperature rises inside the inductor, e.g. the hot-spot winding temperature. Based on the determined η - ρ Pareto front, the most suitable realization of the proposed UPS system is determined, which is based on the resulting sensitivities of the design parameters (switching frequency, current ripples, magnetic materials, etc.) on the performance of power converter system (efficiency and power density). As a result, a switching frequency of 16 kHz, maximum relative input and output side current ripples of 20 %, amorphous core materials for DM inductors, and nanocrystalline core materials for CM inductors are selected in order to achieve a converter efficiency of 96.6% at a power density of $2.3 \, \mathrm{kVA/dm^3}$.

I. INTRODUCTION

The optimization of power converter systems is a multidomain procedure [1], which considers thermal effects and limitations besides electric and/or magnetic characteristics of active and passive power components. Early implementations of power converter optimizations are limited to the optimization with respect to a single performance index, e.g. power density ρ or efficiency η [2]–[4]. Simple-objective converter optimizations, however, often yield unsatisfying remaining converter characteristics, i.e. a converter optimized for high power density may generate high losses, due to increasing losses of high power density magnetic components. Thus, the multi-objective optimization of a PFC rectifier based on the η - ρ Pareto front is proposed in [5]. This Pareto front identifies the highest efficiency for a given power density (and vice versa) and can be used as basis for initial decisions concerning the converter design parameters. Details on the calculation of the η - ρ Pareto front and the extension to a third variable (e.g. cost) are presented in [6].

Examples of η - ρ Pareto optimizations of power converters include the realizations of an ultra-high efficiency DC–DC converter ($\eta = 99\%$ at $\rho = 2.3 \text{ kW/dm}^3$) [7] and an ultra-high efficiency PFC rectifier ($\eta = 99.2\%$ at $\rho = 1.1 \text{ kW/dm}^3$) [8]. Recent publications related to η - ρ Pareto optimizations further

TABLE I. CONVERTER SPECIFICATIONS		
Input phase voltage	$V_{\rm in}$	230 V RMS
Input frequency	$f_{\rm in}$	50 Hz
Max. input phase voltage ripple	$\gamma_{{f v},{ m in}}$	7.6%
EMI requirement		CISPR class A
Output phase voltage	$V_{\bullet ut}$	230 V RMS
Output frequency	$f_{\bullet ut}$	50 Hz
Output power	$S_{\bullet ut}$	20kVA
Max. output voltage ripple	$\gamma_{\mathrm{v,out}}$	1.0%
DC voltage	$V_{ m dc}$	720 V
Ambient temperature	T_{amb}	55 °C
Max. junction temperature	$\vartheta_{i,\max}$	150 °C
Max. inductor winding temperature	$\vartheta_{\rm whs,max}$	150 °C
Max. inductor core temperature	$\vartheta_{ m c,max}$	Depends on material.

confirm that this approach can be advantageously used for designing power converters, e.g. for the design of a $50 \,\mathrm{kW}$ bidirectional resonant converter in [9], and the selection of a suitable five-level inverter topology in [10].

This paper determines the η - ρ Pareto front for a complete 20 kVA UPS system consisting of a rectifier and an inverter stage and corresponding input and output filters in order to enable the converter design with respect to power density at a given efficiency. Section II presents the converter topology and specifications, Section III outlines the optimization procedure and the employed loss and volume models, and Section IV details the inductor models (loss and thermal models), which are verified based on experimental results. Section V discusses the obtained η - ρ Pareto front and the selection of optimum design parameters. Based on the results, a power density of 2.3 kVA/dm³ is calculated for a converter efficiency of 96.6%.

II. THREE-PHASE THREE-LEVEL T-TYPE CONVERTER

Fig. 1 depicts the rectifier and inverter stages of the considered UPS system. It consists of three-phase three-level T-type inverter and rectifier circuits, a mains-side EMI filter, and a two stage output filter. Both power stages employ hybrid Si and SiC power semiconductor configurations and reverse-blocking IGBTs in order to achieve minimum semiconductor losses, which is detailed in [11].

The input and output filters are both realized with two-stage differential mode (DM) filters and include damping networks in order to avoid instabilities due to resonances. The input filter, in addition, contains a fifth-order common mode (CM) filter with three CM inductors in order to fulfil the EMI requirements. **Tab.I** summarizes the converter requirements and specifications.



Fig. 1. Schematic of the 3-phase 3-level T-type AC-DC-AC converter comprising a Si and SiC hybrid rectifier and inverter stage.

TABLE II.	DESIGN	VARIABLES FOR	OPTIMIZATION
Switching frequen	су	$f_{\mathrm{sw,i},n}$	8 kHz 40 kHz

Max. input phase current ripple	$\gamma_{\mathrm{i,in},n}$	5% 40%	
Max. output current ripple	$\gamma_{\mathrm{i}, \bullet \mathrm{ut}, n}$	5% 40%	

III. MULTI-OBJECTIVE OPTIMIZATION PROCEDURE

The multi-objective optimization approach presented in this paper considers the converter and component models listed below:

- Electrical converter model: voltages and currents, predictions of CM and DM noise levels and filter component values to fulfil the design specifications; cf. [12], [13].
- Inductor model: winding and core losses, thermal model, boxed volume.
- Capacitor model: only boxed volume, capacitor losses are neglected; cf. [2].
- Power semiconductors: conduction and switching losses, thermal model; the volume of the semiconductors is not considered; cf. [11].
- Heat sink: thermal model for forced air cooling, power demand of the fan, boxed volume (heat sink plus fan); cf. [14].

Each component model facilitates the calculation of the respective losses, the temperature rises, and the component volumes. The losses and the temperature rises of the components are coupled, e.g. the inductors' winding and core losses depend on the respective winding and core temperatures (according to the material properties) and vice versa (according to the properties of the inductor's thermal network which includes the heat flux to the ambient). For this reason, not only accurate loss models, but also accurate thermal models are required. According to the list given above, a large number of publications related to component models is readily available. This is particularly true for the semiconductors [11] and the heat sink [14]. However, no detailed discussion of coupled electro-thermal models for inductors is available. Therefore, in this paper a coupled electro-thermal inductor model is developed, implemented, and verified.

Fig. 2 depicts the flow chart of the optimization procedure. It conducts fully automatic converter designs and calculates and stores the respective components' losses, temperatures, and volumes in a pool of design results. The models employed in this automatic design procedure are highly non-linear and, therefore, a high number of converter designs is carried out in order to find the global optimum. Thus, in a first step, the design space is created for the considered power converter, which, for the AC–DC–AC part, is based on Tab. II:¹

- switching frequencies $f_{sw,n}$: from 8 kHz to 40 kHz with a step size of 1 kHz;
- rel. input and output inductor current ripples γ_{i,in,n} and γ_{i,out,n}: from 5% to 40% with a step size of 5%; the ratio of the maximum peak-to-peak inductor current ripple, ΔI_{L,n}, to the peak fundamental current, Î, defines the relative current ripple:

$$\begin{aligned} \gamma_{\mathrm{i,in},n} &= \Delta I_{\mathrm{L}_{1,\mathrm{in}},n} / \hat{I}_{\mathrm{in}}, \\ \gamma_{\mathrm{i,out},n} &= \Delta I_{\mathrm{L}_{1,\mathrm{out}},n} / \hat{I}_{\mathrm{out}} \end{aligned}$$

If required, the optimization procedure could also consider further design variables, e.g. the numbers of semiconductors operated in parallel for each switch or diode (here: constant; the actual numbers are given in [11]), the base plate temperature of the heat sink (here, a base plate temperature of $\vartheta = 100 \,^{\circ}\text{C}$ is assumed), and the DC link voltage (here: 720 V).

In a second step the design procedure selects the *n*-th set of design variables from the design space and conducts a single converter design. It calculates the semiconductor losses, $P_{SC,n}$, and the respective junction temperatures for the assumed heat sink base plate temperature. With known semiconductor losses the heat sink can be optimized according to [14], which yields the total boxed volume of the cooling system, $V_{CS,n}$, and the power demand of the fans, $P_{CS,n}$. The values of the filter components are calculated according to [13] in order to fulfil the design specifications and design variables. The total losses and the boxed volumes of the DM and CM inductors ($P_{LDM,n}$, $P_{LCM,n}$, $V_{LDM,n}$, $V_{LCM,n}$) are calculated with the inductor optimization procedure detailed in Section IV, which employs an inductor model that considers electro-thermal and magneto-thermal couplings. Furthermore, the total volume of all DM

¹This paper is confined to the optimization of the AC–DC–AC part of the UPS, since the optimization of the DC–DC converter (not shown in Fig. 1) which is employed for interfacing the battery storage to the UPS DC link, can be implemented in an analogous manner.



Fig. 2. Block diagram of the optimization procedure used to determine the η - ρ Pareto front. $P_{SC,n}$, $P_{CS,n}$, $P_{LDM,n}$ and $P_{LCM,n}$ denote the total losses of the semiconductors with the *n*-th set of design variables, the fans of the cooling system, the DM inductors and the CM inductors, respectively. $V_{CS,n}$, $V_{LDM,n}$, $V_{LCM,n}$, and $V_{CDM,n}$ denote the total boxed volume of the cooling system (heat sink with fans), the DM inductors, the CM inductors, and the DM capacitors, respectively. The optimization of the DC-DC converter part which interfaces the UPS battery storage to the DC link can be perform in analogous manner.

capacitors, $V_{\text{CDM},n}$, is determined for each converter design (the total volume of all CM capacitors is found to be relatively small, i.e. negligible in a first step).

Based on all power densities and efficiencies calculated for all converter designs the η - ρ Pareto front can be generated which, finally, facilitates the identification of the optimum converter design, cf. Section V.

IV. INDUCTOR MODEL

A. Procedure overview and loss model

The calculation of the inductor losses is implemented according to the flow chart depicted in **Fig. 3**. The proposed procedure requires the specifications listed below:

- The inductance value L.
- The fundamental peak current \hat{I} and mains frequency f_{main} .
- The maximum peak-to-peak inductor current ripple $\Delta I_{\rm L}$ and the switching frequency $f_{\rm sw}$.
- The maximum allowed winding hot-spot and core temperatures $T_{\rm whs,max}$ and $T_{\rm c,max}$ and the ambient air temperature $T_{\rm amb}$.²

The implemented automatic inductor design procedure is based on a high number of different inductor designs and the automatic evaluation of the resulting ρ - η -Pareto front similar to the converter optimization procedure discussed in Section III. In a

 $^{^{2}}T_{\text{amb}}$ is the absolute ambient air temperature, e.g. $T_{\text{amb}} = \vartheta_{\text{amb}} + 273 \,\text{K}$. The same representation manner is applied for the other temperatures.



Fig. 3. Flow chart of the inductor design procedure.

first step, the inductor design space is created. The considered inductor design variables are:

- core material and geometry (selected from a predefined list of available magnetic cores),
- number of stacked cores n_{stack} (n = 1, 2, or 3),
- peak flux density $B_{\rm pk}$ (50% $B_{\rm sat}$ to 100% $B_{\rm sat}$ with a step size of 10%),
- filling factor k (50% to 100% with a step size of 10%).

The procedure starts with selecting the first set of design parameters from the inductor design space and calculates the inductor properties that can be directly calculated, such as the wire diameter, the number of turns, and the air gap length. In a next step, the procedure calculates the winding and core losses. The winding losses are separated into low (mains) frequency and high (switching) frequency losses. The low frequency losses are calculated with the DC resistance and the calculation of the high frequency losses is based on the mirroring method detailed in [12]. The inductor design considers two types of conductors, i.e. solid copper wire and high frequency litz wire. The core losses are calculated based on the improved Generalized Steinmetz Equations (iGSE) [15]. The winding and core losses, however, depend on the winding and core temperatures, which are determined using the thermal model discussed in the next Subsection IV-B. The winding and core losses and temperatures are calculated in an iterative loop according to Fig. 3. This loop is repeated until all differences between previously and currently calculated temperatures fall below 1 K. The losses and volumes resulting for both types of wires are evaluated with respect to the maximum allowable winding hot-spot and core temperatures ($T_{\rm whs}$ and $T_{\rm c}$, respectively). The design results of every inductor design, which features hot-spot temperatures less than the specified maximum values is stored in the pool of inductor design results. Thereafter, the design procedure selects the next set of design parameters and conducts the next inductor design. After processing the complete inductor design space, the results available in the pool of inductor design results is analysed in order to determine the optimal inductor according to the following procedure:

- 1. The inductor with minimum boxed volume, $V_{L,min} = \min(V_{L,j})$, is identified in the pool.
- 2. The inductor designs with $V_{L,j} > 1.2 \times V_{L,\min}$ are removed from the pool.
- 3. The inductor design with minimum losses in the remaining pool of inductor design results denotes the considered optimal choice.

The losses of the inductors contribute relatively little to the total losses, therefore, a major optimization criterion is maximum power density. Very compact inductor designs, however, yield comparably high losses. A considerable loss reduction is achieved if a somewhat higher boxed volume is allowable, which is taken into account by the procedure given above.

Name	$\mu_{ m r}$	$B_{\rm sat}$	$\vartheta_{ m c,max}$
EPCOS N87 (Ferrite)	2200	$0.25\mathrm{T}$	$120 ^{\circ}\mathrm{C}$
Magnetics Kool-Mu (Iron powder)	14125	$1.0\mathrm{T}$	$100 ^{\circ}\mathrm{C}$
Metglas 2605SA1 (Amorphous)	45000	$1.5\mathrm{T}$	$150 ^{\circ}\mathrm{C}$
Finemet FT-3M (Nanocrystalline)	70000	$1.23\mathrm{T}$	$150^{\circ}\mathrm{C}$



Fig. 4. Extended thermal network of the inductor. $P_{\rm w}$ and $P_{\rm c}$ denote the total winding losses and the core losses, respectively. $T_{\rm whs}$ and $T_{\rm c}$ denote the hot-spot winding temperature and the temperature inside the core. $T_{\rm ws}$, $T_{\rm cs}$ and $T_{\rm ps}$ are the surface temperatures of the winding, the core and the metal base plate, respectively. The base plate is considered only for the evaluation of the inductor model and it is neglected in the converter optimization in order to keep some safety margin for the inductor design.

For DM inductors the inductor design procedure considers E-core shapes and four different magnetic materials (Nanocrystalline, amorphous, ferrite, and iron powder, cf. Tab. III) with different magnetic properties, in particular different saturation flux densities ($B_{\rm sat}$), operating core temperatures ($\vartheta_{\rm c.max}$), and core loss characteristics. For CM inductors, toroidal cores made of *nanocrystalline material* are considered, due to the high permeability and the low core losses required. N.B.: Inductors using tape wound cores made of *amorphous* or nanocrystalline magnetic materials are subject to increased core losses in presence of orthogonal components of the magnetic flux density, i.e. if the direction of the magnetic flux vector is not aligned to the hard magnetic direction of the magnetic material, which particularly happens close to the inductor's air gap [16]-[18]. Currently, there is no loss model known to the authors, which accurately takes this effect into account. In order to still consider the expected increase of the core losses due to this effect, the core losses are multiplied with a correction factor $k_{\rm Pc} = 2$, which has been determined based on the results of [16].

B. Thermal model

The thermal model is implemented on the basis of [19] and estimates the hot-spot temperature inside the winding $(T_{\rm whs})$, the temperature inside the core $(T_{\rm c})$, the surface temperature of the winding $(T_{\rm ws})$, and the surface temperature of the core $(T_{\rm cs})$. It takes three different heat transfer mechanisms into account, i.e. *conduction, convection* and *radiation*. The thermal resistance network of [19] is extended by three additional thermal resistances in order to determine $T_{\rm c}$ and to consider the heat transfer from the inductor to the metal base plate the inductor is mounted on. **Fig.4** depicts the resulting thermal model.



Fig. 5. Simplified core dimensions (a) and base plate dimensions (b).

The thermal resistances $R_{\rm th,whs}$, $R_{\rm th,wc}$, $R_{\rm th,wa}$, and $R_{\rm th,ca}$ are calculated according to [19].

The calculation of the temperature inside the core, T_c , assumes a constant temperature along the core legs' center lines (cf. **Fig. 5(a)**). Thus, the additional resistance $R_{\rm th,c}$ represents the thermal resistance between the core legs' center lines and the core surface and is expressed as:

$$\frac{1}{R_{\rm th,c}} = \frac{1}{R_{\rm th,c,xy}} + \frac{1}{R_{\rm th,c,z}},\tag{1}$$

$$R_{\rm th,c,xy} = \frac{l_{\rm c}}{\lambda_{\rm c,xy}A_{\rm c,xy}},$$
(2)

$$R_{\rm th,c,z} = \frac{d_{\rm c}/2}{\lambda_{\rm c,z}A_{\rm c,z}},\tag{3}$$

where l_c is an effective thermal distance, $\lambda_{c,xy}$ and $\lambda_{c,z}$ are direction dependent effective thermal conductivities of the core (cf. Fig. 5(a) regarding x-, y-, and z-directions), $A_{c,xy}$ denotes the total surface area of the core in x- and y-directions (top, bottom, and both sides; $A_{c,xy}$ also includes the surface areas inside the core windows), and $A_{c,z}$ is the surface area of the core in z-direction (front and back). These parameters can be calculated using the core dimensions shown in **Fig. 5(a)**:

$$l_{\rm c} = \frac{w_{\rm c} - w_{\rm cw}}{4},\tag{4}$$

$$A_{\rm c,xy} = 2\{(h_{\rm c} + w_{\rm c}) + (h_{\rm cw} + w_{\rm cw} - w_{\rm cl})\}d_{\rm c}, \quad (5)$$

$$A_{\rm c,z} = 2 \{ h_{\rm c} w_{\rm c} - h_{\rm cw} (w_{\rm cw} - w_{\rm cl}) \}.$$
 (6)

The thermal conductivities of the core materials are given in **Tab. IV**.

The second additional resistance, $R_{\rm th,cp}$, represents the thermal resistance between the core and the base plate due to the combination of *conduction* and *radiation*. It is expressed as

$$\frac{1}{R_{\rm th,cp}} = \frac{1}{R_{\rm th,cond,cp}} + \frac{1}{R_{\rm th,rad,cp}},$$
 (7)

$$R_{\rm th,cond,cp} = \frac{l_{\rm eg}}{\lambda_{\rm air}A_{\rm cp}},$$
 (8)

$$R_{\rm th,rad,cp} = \frac{1}{h_{\rm rad,cp}A_{\rm cp}}$$
$$= \frac{T_{\rm cs} - T_{\rm ps}}{\epsilon_{\rm cs}\sigma \left(T_{\rm cs}^4 - T_{\rm ps}^4\right)A_{\rm cp}}.$$
(9)

TABLE V. TEST INDUCTOR DESIGN PARAMETE	RS
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Core material	Metglas 265SA1 Amorphous
Core size	AMCC06R3 (2 sets \times 2 stacked)
Air gap width	$2 \times 0.5 \mathrm{mm}$
Number of turns	$20 \text{ turns} (5 \text{ turns} \times 4 \text{ layers})$
$R_{\mathbf{d}c}$ of wire	$\approx 7.8 \mathrm{m\Omega}$ (at $25^{\circ}\mathrm{C}$)

An equivalent air gap length of $l_{\rm eg} = 0.2 \,\rm mm$ between the core surface and the base plate top surface is found to yield reasonably accurate and reproducible results for all inductors measured with the test set-up detailed in Subsection IV-C.³ Moreover, $\lambda_{\rm air} = 0.03 \,\rm W/(mK)$ is the thermal conductivity of air at 80 °C, $A_{\rm cp} = w_c d_c$ is the contact area of the core and the base plate, $\epsilon_c = 0.9$ is the assumed emissivity of the core surface, and σ is the Stefan-Boltzmann constant [= $5.67 \times 10^{-8} \,\rm W/(m^2K^4)$] [19].

The last additional resistance, $R_{\rm th,pa}$, represents the thermal resistance between the surface of the base plate and the ambient air and considers heat flux due to *convection* and *radiation*. The dimensions of the base plate are depicted in **Fig. 5(b)** and the expressions for calculating $R_{\rm th,pa}$ are given in [19]. Further parameters are: the assumed emissivity of the base plate surface, $\epsilon_{\rm p} = 0.04$ (aluminum, polished), the total open surface area of the base plate, $A_{\rm pa} =$ $2 \{w_{\rm p}d_{\rm p} + (w_{\rm p} + d_{\rm p})\} - A_{\rm cp}$, and the total distance passed by the air that cools the base plate, $L = d_{\rm p} + h_{\rm p}$.

C. Evaluation of Inductor Model

In order to evaluate the accuracy of the inductor model, i.e. the accuracies of the calculated losses and temperatures, the inductor's total losses and core and winding temperatures have been measured. The inductor test set-up and a test inductor are built as shown in Fig. 6. The inductor test set-up contains an aluminum base plate (dimensions: $220 \text{ mm} \times 300 \text{ mm} \times 5 \text{ mm}$), which supports a duct made of acrylic glass; the data logger used to process the measurement results is located below the base plate of the duct. The back side of the duct shown in Fig. 6 is terminated with a fan that can be used to control the air flow inside the duct. The test set-up is equipped with K-type thermocouples and air speed sensors to measure ϑ_{amb} and the air speeds at different locations inside the duct, respectively. The test inductor, shown in Tab.V, employs Amorphous core material and solid copper wire and is equipped with thermocouples, located inside the winding, at the surface of the winding, and at the surface of the core, in order to measure $\vartheta_{\rm whs}$, $\vartheta_{\rm ws}$, and $\vartheta_{\rm cs}$ (cf. Fig. 6(b).

The inductor losses are simultaneously measured with a calorimeter and a power analyzer for a low frequency (LF) sinusoidal current with a superimposed triangular high frequency (HF) AC current. The test conditions are specified in **Tab. VI**.

Fig.7 compares the obtained measured and calculated inductor losses and temperatures. A good agreement between

³Experimental results confirm the existence of a non-zero value of $R_{\rm th,cp}$, however, the exact value of the equivalent air gap length, $l_{\rm eg}$, is difficult to determine and may vary depending on the inductor core and the set-up.



Fig. 6. (a) Inductor test set-up with ambient air temperature sensors and air-speed sensors; (b) test inductor.

TABLE VI. TEST CURRENT CONDITIONS

HF component	$\Delta I = 9.5 \mathrm{A}$
(peak-to-peak)	$f_{\rm sw} = 16 \rm kHz$
LF component	$\hat{I} = 0, 20, 48 \mathrm{A}$
	$f_{\rm main} = 50 {\rm Hz}$

measured and calculated results for all considered operating points (losses and temperatures) is obtained if the core loss correction factor is reduced to $k_{\rm Pc} = 1.65$ (instead of $k_{\rm Pc} = 2$, cf. Subsection IV-A):⁴ The relative error for all considered total losses and winding hot-spot temperatures is less than 10 %.

V. OPTIMIZATION RESULTS AND DISCUSSION

Fig.8(a) shows the power densities and efficiencies calculated for all considered converter design points. Besides different switching frequencies and current ripples (cf. Section II) also different core materials are considered for the DM inductors. According to these results, highest efficiencies and highest power densities are achieved with *nanocrystalline materials*, due to the high saturation flux densities and the low core losses of *nanocrystalline materials* even including increased core losses with air gap. The high saturation flux densities of *amorphous materials* allows high power densities

⁴The value of k_{Pc} is expected to depend on the air gap length and is, therefore, not expected to be constant. For this reason, the optimization results presented in Section V are still based on $k_{Pc} = 2$ to avoid unrealistic inductor designs. Thus, further research related to a more accurate calculation of the core losses of tape-wound cores with air gaps is required.



Fig. 7. Comparison of measured and calculated inductor losses and temperatures. Measurement results are shown by black bars (meas.) and calculation results are shown by light gray bars (model). Ambient temperatures are shown by dark gray bars ($\vartheta_{\rm amb} \approx 33 \,^{\circ}{\rm C}$ inside the closed box of the calorimeter).



Fig. 8. η - ρ Pareto front of the proposed UPS system with four different magnetic materials (a) and Pareto front with *amorphous material* (b) used for everything the mapping of the design space into the system performance.

too, however, due to higher core losses, the achievable converter efficiencies are less than the efficiencies achievable for nanocrystalline materials. Also iron powder materials can be used to realize the inductors, however, the permeabilities of these materials are less than those of the other materials. In addition, the permeabilities drop with increasing flux densities, i.e. the inductances of inductors using iron powder cores show a non-linear dependency on the inductor current. As a consequence, the number of turns needs to be increased in order to maintain a given inductance value at an elevated current, which increases the winding losses. Due to the increased winding losses the inductor volume has to be increased in order to maintain the winding hot-spot temperature limit, which decreases the power density. Thus, for the given application, inductors using iron powder cores are found to be less suitable. For the given application also *ferrite cores* are found to give comparably low power density and efficiency values, which is due to the low saturation flux density of ferrite materials and the relatively low switching frequencies used by reason of the employed semiconductors. Due to the high cost of *nanocrystalline materials* $(23 \in /kg)$ [20] and the comparably small difference in efficiency and power density, the *amorphous material* ($16 \in /kg$) is selected for realizing the inductors of the converter.

Fig. 8(b) shows, how the design space is projected into the performance space:

• between points A and B on the led line in Fig.8(b), the *increase of the switching frequency* at constant current ripple of 20% causes the power density to increase and the efficiency to drop. If the switching frequency exceeds a certain



Fig. 9. Influence of the design parameter switching frequency f_{sw} on the efficiency (a), the loss distribution (b), the power density (c), and the volume distribution (d). In this figure, the current ripple is kept constant ($\gamma_i = 20\%$).

value, however, the power density drops (cf. **B–C in Fig.8(b)**), due to the additional converter volume needed to dissipate the heat (increased heat sink volume, additional volume of passive components due to thermal limitation).

• between points a and b on the blue line in Fig.8(b) the *increase of the current ripple* at constant switching frequency of 16 kHz as the selected design causes the power density and the efficiency to increase, due to the reduction of the inductances, which helps to reduce the inductor winding losses and volumes; between points b and c increasing current ripple start to deteriorate the overall efficiency and power density due to increase in the inductor losses and the capacitor volumes. With higher current ripple at same switching frequency, the inductor core losses is increased. In addition, in order to dissipate the increased heat, the inductor volumes can not be reduced (thermal limit). Moreover, the volume of the EMI filter (CM inductor and DM capacitor) increases, since the required filter attenuation increases with increasing current ripple.

Efficiencies, loss distributions, power densities, and volume distributions calculated for different switching frequencies, constant relative input and output side current ripples of 20%, and inductors made of *amorphous cores* are shown in **Fig.9**. According to Fig.9(a), the efficiency drops with an increase in switching frequency, which is mainly due to the switching losses of the power semiconductors. The power density improves with an increase in switching frequency due to the volume reduction in the passive components. However, it shows a maximum for a switching frequency of approximately



Fig. 10. Influence of the design parameter current ripple γ_i on the efficiency (a), the loss distribution (b), the power density (c), and the volume distribution (d). In this figure, the switching frequency is kept constant ($f_{sw} = 16 \text{ kHz}$).

29 kHz, and no further improvement is seen with increasing switching frequency, due to an increase in HF core losses of the DM inductors and the associated overall component surfaces needed to maintain the thermal limits. A switching frequency of 16 kHz is selected in order to achieve a high efficiency and to avoid audible noise. At $f_{\rm sw} = 16$ kHz, the contribution of the total semiconductor losses on the total converter losses is 65% and the contribution of the total passive components' losses is 27%. The contribution of the passive components (DM inductors, capacitors, and CM inductors) on the total converter volume is 49%.

Fig. 10 depicts the results calculated for different current ripples and constant switching frequency, $f_{sw} = 16 \text{ kHz}$. In this case the maxima of efficiency and power density occur for similar current ripples (15% for maximum efficiency and 20% for maximum power density) on the input and output sides. According to Fig. 10(b), the DM inductor losses decrease significantly between $5 \dots 15\%$, since a reduction of the inductance value significantly reduces the LF winding losses in the DM inductors. This helps to reduce the volume of the DM inductor and improves the overall power density (cf. Fig. 10(c) and (d)). However, the DM inductor losses slightly increase for current ripples between $15\% \dots 40\%$, due to increasing HF winding losses and core losses. Since the DM inductor losses are not decreasing, the DM inductor volume is also not decreasing in that range of the current ripple. However, the DM capacitor volume increases, since the voltage ripple is kept constant and the current ripple increases. Therefore, the power density drops if the relative current ripples exceed 20%. In order to achieve a high power density, a current ripple of 20% is selected.

VI. CONCLUSION

The optimization procedure presented in this paper employs a multi-domain approach which is based on component models that consider electric, magnetic, and thermal aspects. The paper especially details a coupled electro-thermal and magneto-thermal model of the inductors that features accurate calculation of losses, volume, and hot-spot temperature inside the winding. This presented results reveal that the global converter optimization based on the η - ρ Pareto front enables a high converter efficiency of 96.6% at a maximum power density of 2.3 kVA/dm³ for the investigated 20 kVA UPS system including the EMI input filter and the output filter. The relationship between the η - ρ performance space and the f_{sw} - γ_i design space is discussed and suitable design parameters of the converter system are determined by projecting selected performance points back into the design space. As a result, a switching frequency of 16 kHz, a current ripple of 20 %, and dedicated inductor realizations (DM inductors: amorphous cores and solid copper wires; CM inductors: nanocrystalline cores and solid copper wires) are selected.

Future research may focus on further improvements related to the inductor model, e.g. modelling of the increased core losses in the tape wound core in presence of an air gap, and including forced air cooling in the thermal model of the inductor. In summary, the multi-objective optimization approach gives a clear picture of the achievable overall converter performances (efficiency, power density, etc.) as coordinates in the performance space including all power components. It, therefore, provides a very good basis for decision making and helps to shorten the development time for an optimized power electronics system.

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