Design Trade-Offs for Medium- and High-Frequency Transformers for Isolated Power Converters in Distribution System Applications

Obaid Aldosari¹, Luciano A. Garcia Rodriguez², Juan Carlos Balda³, Department of Electrical Engineering University of Arkansas Fayetteville, AR, 72701, USA ¹omaldosa@uark.edu, ²1garciar@uark.edu, ³jbalda@uark.edu

Abstract — Medium- and high-frequency transformers (MFTs/HFTs) are a fundamental component in many isolated power-converter topologies proposed for electric distribution applications (e.g., solid-state power substations). Previous work presented detailed transformer design methodologies and addressed core loss limitations for different core materials and operating frequencies. However, MFT/HFT designs become significantly challenging for high power levels that are typical of distribution systems (e.g., greater than 100 kVA). Furthermore, few references include specific requirements in the design methodology like desired leakage and/or magnetizing inductances (which are normally specified for high-power applications). A design methodology for MFTs/HFTs is presented in this paper that accounts for tradeoffs like having a given leakage inductance for maximum power transfer (e.g., in the case of dual-active bridges (DABs)) or a given magnetizing inductance (to either attain a certain power transfer or to limit the power semiconductor currents). The design methodology is verified via finite-element analysis (FEA) using ANSYSTM and an experimental prototype.

Keywords — Medium- and high-frequency transformer design, Solid-state transformer, Finite element analysis

I. INTRODUCTION

Nowadays, several applications of power electronics in electric power distribution systems are envisioned due to advances in high-voltage power semiconductor technologies, in particular, those enabled by wide bandgap power devices [1]. High-voltage silicon-carbide (SiC) **MOSFETs** are commercialized up to 1.7 kV with 3.3 kV and 6.5 kV devices to be commercialized soon. These devices have lower switching losses than silicon IGBTs so they can be operated at much higher switching frequencies leading to reductions of passive component sizes. These applications of power electronics often require stepping down or up a particular voltage level using a transformer. Examples are (a) solid-state transformers based on DABs requiring a certain value of the transformer leakage inductance for maximum power transfer [2], and (b) flyback converter topologies, or (c) input-output continuous converter topologies [3] where the transformer that store energy requires a

Sudip K. Mazumder, NextWatt LLC Hoffman Estates, IL, USA sudipkumarmazumder@gmail.com

given magnetizing inductance and ideally no leakage inductance to avoid adverse effects [4].

New cores based on amorphous and nanocrystalline materials enable size reductions of inductors and transformers in the medium-frequency range due to their higher flux densities when compared to ferrite cores. However, high-power applications require stacking several cores of these materials (due to the size limitations of commercial cores) or, if possible, use large expensive custom cores to satisfy area-product requirements. As a result, the design becomes more complex due to several tradeoffs among the transformer specifications. With the goal of simplifying the design, a methodology for high-power MFTs/HFTs considering system specifications (e.g., desired leakage and/or magnetizing inductances) and constraints (e.g., temperature rise, operating frequency, or volume) is presented and verified via ANSYSTM finite-element analysis (FEA) and experimental results.

The paper is organized as follows: Section II reviews different core materials currently available for high-power applications and illustrate a simple technique to estimate temperature rise as a function of rated power, Section III describes the proposed design methodology, Section IV considers a case study and evaluates FEA results, Section V presents a scaled-down prototype as a verification of the proposed tradeoff strategies and Section VI provides the research conclusions.

II. MATERIALS SUITABLE FOR HIGH-POWER MFTS/HFTS

Selecting the proper core material is a critical decision leading to a successful MFTs/HFTs design. In the last decades, intensive research has been done on a variety of different magnetic materials (e.g., nanocrystalline, amorphous and ferrite) in terms of cost, power loss, and size [5]. However, it is very challenging and time consuming to choose the right material to meet specific application's requirements, especially at highpower levels. For this reason, the following subsections will provide an overview and comparison between these magnetics materials (A), and describe a simple technique for selecting the core material over a wide range of power ratings subject to a specified temperature rise (B).

A. Core Material Review

Well-known materials for designing MFTs/HFTs are nanocrystalline, amorphous and ferrite [5]. Table I shows a general comparison between these materials [6]. At high-power levels, nanocrystalline and amorphous are the two main materials for designing MFTs (e.g., f = 20 kHz) due to their low core losses (low eddy current losses), high saturation flux and high permeability [6],[7]. The low prices and flux densities (0.3-0.5 T) of the soft ferrite material make them only a suitable choice for low-power HFTs applications [8].

B. Temperature Rise Considerations

It is initially desired for simplicity to design MFTs/HFTs for high-power levels without a detailed design of the thermal management system (e.g., forced convection, liquid cooling). However, the temperature rises with different slopes for different magnetic materials as the rated power increases. To have good estimations about temperature changes as function of the rated power, the following assumptions are made:

- Turns ratio N = 1 (minimum area product, worst-case scenario).
- Desired temperature rise $\Delta T = 50 \,^{\circ}\text{C}$.
- Operation frequency f = 20 kHz.
- Power losses (in cores and windings) are a percentage of the maximum output power which is based on evaluating multiple design results for different power levels and the considered core material. The assumed values were 0.28 % for nanocrystalline and 0.54 % for amorphous.

The optimal flux density B_{opt} is given by [9] as:

$$B_{opt} = \frac{\left(h_c k_a \Delta T\right)^{\frac{2}{3}}}{\sqrt[3]{4} \left(\rho_w k_w k_u\right)^{\frac{1}{12}} \left(k_c K_c f^{\alpha}\right)^{\frac{7}{12}}} \left(\frac{K_v k_{sf} k_u f}{\sum V A}\right)^{\frac{1}{6}}.$$
 (1)

The minimum area product A_p is then calculated as follows [9]:

$$A_{p} = \left(\frac{\sqrt{2}\sum VA}{K_{v}B_{opt}k_{f}K_{i}f\sqrt{k_{u}\Delta T}}\right)^{\frac{5}{7}},$$
(2)

TABLE I. CORE MATERIAL COMPARISON [6]

Material Pros		Cons	
Nanocrystalline	High B_{sat} (1.2 T) Low losses @ high power High permeability	High cost	
Amorphous	High B_{sat} (1.55 T) High permeability Reasonable cost	Medium losses and large sizes @ high power levels	
Ferrite	Low losses and Low cost @ low-power levels	Low <i>B_{sat}</i> (0.5 T) Production difficulty and large sizes <i>@</i> high power levels	

TABLE II. CONSTANT VALUES OF OPTIMAL FLUX AND AREA PRODUCT

Variable	Value	Variable	Value
coefficient of heat transfer $h_c(10w/m^2)$	10	Dimensionless quantity k_a	40
Initial wire resistivity $\rho_w (\Omega * m)$	1.78*10-8	Dimensionless quantity k_w	10
Window utilization factor k_u	0.4	Dimensionless quantity k_c	5.6
Waveform factor square wave K_v	4	Core stacking factor k_{sf}	0.95
$K_t = \sqrt{h_c k_a / \rho_w k_w}$	48*10-3	Expected temperature ΔT (°C)	50

where the values of the constants are shown in Table II [9].

The temperature rise ΔT (°C) as function of the area product A_p with assumed total power losses P_{losses} is estimated by [10] as:

$$\Delta T = 450 \left(\frac{P_{losses}}{K_s \sqrt{A_p}} \right)^{0.826},$$
(3)

where $K_s = 39.2$ is a constant used to calculate the surface area for C cores [10]. The material specifications and properties obtained from manufacturer datasheets are shown in Table III.

A MATLAB[®] code was generated to evaluate the above equations and approximate the temperature rise at different power levels. Fig. 1 shows that the nanocrystalline material has a lower ΔT compared to the amorphous material, making it a better choice for designing a 120 kVA and 20 kHz MFT [11].

In general, the core and winding losses increase due to the increase of the area product as the rated power increases. Furthermore, the amorphous material results in a low B_{opt} so the area product is large requiring a large number of turns due to its large cross sectional area. In addition, the core losses will be large due to the resulting product of $[(P_{loss}/m^3)*volume]$. As a result, the numerator in (3) is relatively larger than the denominator, which makes the temperature rise higher than that for the nanocrystalline material. The main drawback of the latter is its relatively high price. Efficiency and cost are important tradeoffs between these two materials.

King Magnetics[®] provides C cores which are made from nanocrystalline ribbon materials that have a high saturation flux density B_{sat} , low magnetostriction, low noise and relative magnetic permeability μ_r higher than 30,000 H/m [12]. The largest commercial core is 85x106x171 mm, weights 6,600 grams and has an area products of 4193.3 cm⁴. This core can be used to design a 64 kW (max) MFTs assuming $\Delta T = 50$ °C. For a design with a higher rated power, designer should consider

TABLE III. MATERIAL COEFFICIENTS [10]

Parameters	Nanocrystalline	Amorphous
K_C	2.3 w/m ³	1.3617 w/m ³
α	1.32	1.51
β	2.12	1.74
B_{sat}	1.2 T	1.56 T



Fig.1. Estimated temperature rise ΔT as a function of rated power using nanocrystalline (blue) and amorphous (red) core materials.

stacking cores in parallel to increase the area product if the window area is enough to fit the need number of turns.

III. DESIGN METHODOLOGY FOR HIGH-POWER MFTS/HFTS

A. Magnetizing and Leakage Inductance Requirements

Obtaining the specified magnetizing inductance L_m^* and leakage inductance L_k^* when designing MFTs/HFTs is very challenging at high-power levels. The main goal is to keep L_m^* at the value constrained by:

$$L_m^* = \frac{V_{rms}}{2\pi f\left(I_{rms}C_f\right)},\tag{4}$$

while maintaining the flux density close to its optimal value. The new variable C_f introduced in (4) has a range between 0 and 1 (i.e., $0 < C_f \le 1$), where its value depends on the type of isolated power converter. Lower values of C_f result in a significantly larger L_m^* value, which is necessary to avoid a high magnetizing current as in the case of DAB converters where C_f is at least 0.25 (which means that the magnetizing current I_m should be less than 25 % of the primary current I_p [2]). However, a certain leakage inductance L_k^* is required for DAB-based MFTs/HFTs to maximize the transferred power [2]; i.e.:

$$L_k^* = \frac{V_p}{2d\pi^2 P_o f} \phi(\pi - \phi), \qquad (5)$$

where ϕ is the phase-shift between primary and secondary transformer voltage waveforms, V_p is the primary voltage, d is the duty cycle of the voltage waveforms, and P_o is the output power. The L_m and L_k inductances depend on the core physical dimensions and winding arrangements around the core [9], [10]; specifically:

$$L_m = \frac{N^2 S_F A_c \mu_o \mu_r}{l_c + l_g \mu_r} \left(1 + \frac{l_g}{\sqrt{S_F A_c}} \ln\left(\frac{2C}{l_g}\right) \right) [H], \quad (6)$$

$$L_{k} = \frac{\pi (MLT) N^{2}}{N_{L}} \left(\sum d_{iso} + \frac{\sum d_{a}}{3} \right) 10^{-9} [H].$$
(7)

Where S_F is the number of cores needed to achieve the required area product A_p ; μ_o and μ_r are the air and material permeabilities; l_c is the mean length of the magnetic path; C is the window length of the core; *MLT* is the mean length of the turns; N is the number of turns [9]; l_g is the air-gap length; N_L is the portion of the dimension C covered by the windings; d_{iso} is the distance between primary and secondary windings and d_a is the diameter of the Litz wire as shown in Fig. 2.

However, the core and winding geometries in the case of DAB-based converters should be a tradeoff in favor of achieving the required magnetizing inductance since the leakage inductance can be increased by adding an external inductor. The number of turns is determined by [10]:

$$N = \frac{V_{rms}}{k_{\nu}B_{max}A_{c}f},$$
(8)

where k_v is the waveform factor.

In the case of topologies like flyback converters, C_f is 1.0, which implies that the magnetizing current I_m is equal to the primary current I_p . As a result, the magnetizing inductance L_m in this case is much smaller than for the DAB case. From (6), L_m can be reduced to the desired value by introducing an air-gap of length l_g according to:

$$l_g = 2NI_{rms} \frac{\mu_o}{B_o} - \frac{l_c}{\mu_r}$$
(9)

Equation (9) considers the case of a shell-type transformer implemented by using two C-cores [13]. However, the value of the leakage inductance should be minimal to avoid undesired voltage spikes across the converter's power devices.

B. MFTs/HFT Design Steps

The general objective of the transformer design is to maximize efficiency while minimizing transformer weight or volume subject to the specified temperature rise and inductances. Thus, the MFTs/HFTs design largely follows the procedure illustrated in Fig. 3.

The suggested main steps for designing MFTs/HFTs are the following:

• Step 1: Design parameters are specified based on the application's requirements that take into account the desired magnetizing inductance L_m^* using (4) by selecting the proper value of C_f .



Fig. 2. Main physical parameters of a MFT/HFT.



Fig. 3. MFTs/HFTs design flow chart.

- *Step 2*: The saturation flux density B_{sat} , and the Steinmetz coefficients k, α and β are obtained from the manufacturer datasheet of the selected core material.
- *Step 3*: The optimal flux density *B*_{opt} is calculated using (1) that considers that minimum loss point (i.e., copper and core losses are equal).
- Step 4: The minimum area product A^{*}_p is calculated using
 (2) and selecting a core whose area product A_p > A^{*}_p.
- *Step 5*: Winding arrangements and dimensions are made based on the application specifications as explained in the above subsection (A).
- Step 6: Air-gap length l_g is calculated using (9) to meet the desired L^{*}_m that was obtained in (4).
- Step 7: Magnetizing inductance evaluation based on the structure of the cores and winding coil dimensions around the central core. The actual L_m can be calculated using (6), and then compared to the desired L_m^* (4).
- *Step 8*: Leakage inductance evaluation based on the structure of the cores and the distance between primary and secondary coils, the actual *L_k* can be calculated using

(7), and then compared to the desired L_k^* . The windings should be as close as possible to each other to minimize the leakage inductance in case of flyback converters. For that reason, the secondary windings are wound around the primary windings with the minimum distance constrained by the required isolation distance.

- Step 9: Finite-Element Analysis (FEA) is performed for design verification of the previous steps. The flux density distribution B through the cores is visualized using ANSYS[™] simulations as it can be seen in the following section. If the cores experience high magnetic fields B, the core dimensions and the air-gap length should be modified (increased) based on (9) to reduce the flux fields towards its optimal level.
- *Step 10*: Volume calculation is very important, particularly, for applications characterized by space limitations. If the volume specification is not fulfilled, a new optimization design iteration can be considered by selecting a different operating flux density in terms of the optimal volume in *Step 3*. However, there is a tradeoff between efficiency and power density.
- *Step 11*: Core and winding losses are the two main transformer losses. There are many approaches to calculate power core losses (i.e., separation of losses, Preisach model and Jiles-Atherton model based on the hysteresis model). The core loss calculation here is based on the Improved Generalised Steinmetz Equation (IGSE) due to the square waveform excitation [14].
- Step 12: The resulting temperature rise ΔT is calculated using (3) to ensure that the specified limit (in Step 1) is not exceeded. In addition, the required isolation level is evaluated by calculating the distance between conductors d_{iso} per the procedure given in [15]:

$$d_{iso} = \frac{V_{iso}}{k_{iso}E_{ins}},$$
(10)

where V_{iso} is the isolation voltage level and it should comply with ANSI/IEEE C57.12.01, k_{iso} is the safety margin specified by the designer based on the application criteria and E_{ins} is the dielectric strength for the material isolating the primary voltage from the secondary voltage. If the requirements are not fulfilled, a new iteration is initiated considering two options: changing the core dimensions in *Step 4*, or winding arrangement in *Step 5*.

IV. HIGH-POWER CASE STUDY DESIGN RESULTS

Table IV presents the specifications and requirements of the case study for designing a MFT where energy storage is required (120 kVA, 20 kHz, 1020 V_{rms}, $N_p/N_s = 1$). Based on the given

TABLE IV. SPECIFICATIONS AND RESULTS FOR MFT/HFT

Specified Parameters	Required	Calculated
Magnetizing Inductance L _m	\leq 68 μ H	30 <i>µ</i> H
Leakage Inductance L_k (μH)	Small as possible	1.11
Efficiency η (%)	≥ 99	99.7



Fig. 4. ANSYSTM flux density values inside the cores.

specifications and requirements, the optimal flux density B_{opt} was calculated as 0.147 T, being a good compromise between efficiency and volume. Fig. 2 illustrated the main physical dimensions of the selected shell-type transformer to achieve the required magnetizing inductance and a low leakage inductance. In case of a topology like the flyback converter, the isolation distance d_{iso} determines the minimum distance between windings in order to meet the leakage inductance requirement. The flux density distribution inside the cores is shown in Fig. 4 where the flux densities B are within the material allowable limits. The sharp edges of the core are the regions where the higher magnetic fields are located, while most of the core structure experience density fields between (0.161 T green) and (0.096 T blue) which is a perfect range for the calculated optimal flux. The temperature rise of the designed MFT was calculated as low as 48 °C due to nanocrystalline material which has low losses at the operation frequency of 20 kHz. As the power level increases, challenges arise in terms of obtaining the required L^*_{μ} and L^*_{μ} inductances. Substantial power losses in the cores and windings may require complex cooling systems [16] which compromise the transformer size and weight. Also, the required voltage isolation level increases with increased power levels.

V. HFT SCALE-DOWN PROTOTYPE AND RESULTS

A scale-down prototype of a HFT based on a flyback converter topology was built for verification of the proposed design methodology. The main specification parameters of the 1020-W high-frequency transformer are given in Table V. Leakage inductance has to be as small as possible to avoid unwanted voltage spikes across the terminal of the transformer and converter's switches. Meanwhile, the magnetizing inductance values satisfies (4) to guarantee power transfer from primary to secondary side at rated voltage and current. However,

TABLE V. SPECIFIED PARAMETERS OF THE HFT PROTOTYPE

Parameters	Values
Rated Power	1020 W
Primary DC Voltage	$120 V_{DC}$
Primary RMS Current	8.5 A _{RMS}
Turns Ratio	1:1
Switching Frequency	100 kHz

TABLE VI. PHYSICAL PARAMETERS FOR THE HFT PROTOTYPE

A	10 mm	F	53 mm	Mean path length	12.8 cm
В	11 mm	d_f	1.25 mm	Eff. cross area	1.56 cm ²
С	33 mm	d_a	2.07 mm	Weight	146 g
D	20 mm	diso	0.0762 mm	Number of turns	8
Ε	31 mm	d _{air}	11.251mm	Area product	7.26 cm ⁴

having a slightly larger magnetizing $(L_m > L_m^*)$ is accepted as a tradeoffs to keep leakage inductance as small as possible. The main physical parameters of the cores been used for this prototype (Fig. 2) are shown in Table VI. The calculated value for the magnetizing inductance with an air-gab of $l_g = 0.87$ mm was obtained from (6) as $L_m = 38.1 \,\mu\text{H}$ which is very close to the measured prototype value $L_{exp} = 40.5 \,\mu\text{H}$ measured by the 4192A LF impedance analyzer which means an error of less than 6.5%. The calculated value for the leakage inductance was calculated from (7) as $L_k = 1.19 \,\mu\text{H}$ where the experimental value is $L_{exp} = 1.04 \,\mu\text{H}$. The error between these two values was calculated as 14.4 % assuming that there is no space between turns which makes N_L at its minimum value. The built high frequency transformer prototype is shown in Fig. 5.

The primary-to-secondary stray capacitance between the windings must be very small to avoid undesired interactions between primary and secondary windings. For that reason, a copper shield was inserted between the two windings to reduce the total capacitance. Measurements were performed over a wide range of frequencies to make sure the transformer parameters are constant. Fig. 6 shows (a) the magnetizing inductance L_m , (b) the leakage inductance L_k , and (c) the stray primary-to-secondary windings capacitance C_{ps} as a function of frequency. From 50 kHz to 150 kHz, the passive components (L_m , L_k and C_{ps}) have low changes as (10 %, 0.9 % and 0.4 %) respectfully.

After verifying that the values of the magnetic components are close to the designed values calculated from the design



Fig. 5. Prototype of 1020 kW, 120 $V_{\rm rms}$ and 100 kHz high frequincy transformer.



Fig. 6. (a) Magnetizing inductance L_m , (b) leakage inductance L_k and (c) primary-to-secondary stray capacitance C_{ps} as function of the frequency.

equations, a flyback converter capable of switching at 100 kHz with an input voltage of 120 V_{DC} and a peak current of 15 A was constructed as shown in Fig. 7. Table V provides the main specifications of the built HFT and Table VI its main dimensions. Fig. 8 shows the transformer primary and secondary currents at rated conditions of delivering 450 W to a 35 Ω resistive load.



Flyback Converter

Fig. 7. Flyback converter experimental setup .



Fig. 8. Primary I_p and secondary I_s flyback transformer currents when the input voltage V_{in} is 120 V.



Fig. 9. SiC MOSFET drain-to-source voltage V_{ds} and flyback converter output votlage V_o when the input voltage V_{in} is 120 V.

The selected operating condition for the flyback converter was the boundary conduction mode (BCM) due to its improved performance in comparison with the continuous and discontinuous modes of operation [17]. Because of the BCM operation with a duty cycle of 50 %, the voltage conversion ratio is 1, so the output voltage is the same as the input voltage as shown in Fig. 9. The drain to source voltage of the SiC MOSFET is also shown in Fig. 9 where it is seen that the voltage spikes across the transistor are due to the presence of the leakage inductance L_k . Those voltage spikes can be reduced to a desire lower level at the expense of reducing the efficiency of the converter [18]. For this particular case, the voltage spikes were limited to 450 V using a snubber based on passive components since the rated voltage of the implemented devices was 1200 V.

VI. CONCLUSIONS

Designing MFTs/HFTs for high power levels is challenging due to system specifications and constraints (e.g., magnetizing and leakage inductances, voltage insulation level, temperature rise, efficiency) which drive the design in opposite directions so the designer must make several tradeoffs. A new simple technique to compare different core materials in terms of temperature rise was presented followed by a designed methodology that considers inductance specifications by specifying the value of the new variable C_f . The feasibility of the proposed ideas were verified through ANSYSTM simulations and a scaled-down prototype. The experimental results agreed fairly well with the theoretical calculations and simulations.

ACKNOWLEDGMENTS

This material is based upon work supported by the Department of Energy under Award Number DE-OE0000853, awarded to NextWatt LLC in 2017.

DISCLAIMER

This paper was prepared as an account of work sponsored by an agency of the United States Government. Neither the United States Government nor any agency thereof, nor any of their employees, makes any warranty, express or implied, or assumes any legal liability or responsibility for the accuracy, completeness, or usefulness of any information, apparatus, product, or process disclosed, or represents that its use would not infringe privately owned rights. Reference herein to any specific commercial product, process, or service by trade name, trademark, manufacturer, or otherwise does not necessarily constitute or imply its endorsement, recommendation, or favoring by the United States Government or any agency thereof. The views and opinions of authors expressed herein do not necessarily state or reflect those of the United States Government or any agency thereof."

REFERENCES

- Haitham Abu-Rub; Mariusz Malinowski; Kamal Al-Haddad, "Recent Advances in Power Semiconductor Technology," in *Power Electronics* for Renewable Energy Systems, Transportation and Industrial Applications, 1, Wiley-IEEE Press, 2014, pp.832.
- [2] Y. Du, S. Baek, S. Bhattacharya and A. Q. Huang, "High-voltage high-frequency transformer design for a 7.2kV to 120V/240V 20kVA solid state transformer," *IECON 2010 36th Annual Conference on IEEE Industrial Electronics Society*, Glendale, AZ, 2010, pp. 493-498.
- [3] S.K. Mazumder, "A modular and flexible high-frequency-link transformer with a reduced device count and zero high-side devices", *Final Project Report, submitted to the U.S. Department of Energy*, April 2018.
- [4] G. N. Wooding and A. S. De Beer, "The effect of leakage inductance and snubbing on electromagnetic interference generated by a flyback converter," *AFRICON*, 2011, Livingstone, 2011, pp. 1-5.
- [5] R. Burdt et al., "Evaluation of Nanocrystalline Materials, Amorphous Alloys and Ferrites for Repetitive-Magnetic Pulse Compression Applications," 2005 IEEE Pulsed Power Conference, Monterey, CA, 2005, pp. 843-847.
- [6] S. Vaisambhayana, C. Dincan, C. Shuyu, A. Tripathi, T. Haonan and B. R. Karthikeya, "State of art survey for design of medium frequency high power transformer," 2016 Asian Conference on Energy, Power and Transportation Electrification (ACEPT), Singapore, 2016, pp. 1-9.
- [7] T. Kauder and K. Hameyer, "Performance Factor Comparison of Nanocrystalline, Amorphous, and Crystalline Soft Magnetic Materials for Medium-Frequency Applications," in *IEEE Transactions on Magnetics*, vol. 53, no. 11, pp. 1-4, Nov. 2017.
- [8] M. Jafari, Z. Malekjamshidi, G. Lei, T. Wang, G. Platt and J. Zhu, "Design and Implementation of an Amorphous High-Frequency Transformer Coupling Multiple Converters in a Smart Microgrid," *IEEE Transactions* on *Industrial Electronics*, vol. 64, no. 2, pp. 1028-1037, Feb. 2017.
- [9] W. G. Hurley, W.H. Wölfle, Transformers and Inductors for Power Electronics: Theory, Design and Applications, 1st ed., Wiley, 2013.
- [10] C. William, T. McLyman, Transformer and Inductor Design Handbook, 4th ed., Taylor & Francis Group, 2011.
- [11] K. Warnakulasuriya, F. Nabhani and V. Askari, "Development of a 100kW, 20 kHz Nanocrystalline Core Transformer for DC/DC Converter Applications," *PCIM Europe 2016; International Exhibition and Conference for Power Electronics, Intelligent Motion, Renewable Energy and Energy Management*, Nuremberg, Germany, 2016, pp. 1-8
- [12] [Online]. Available: http://www.kingmagnetics.com/nanocrystalline-ccore.html [Accessed Oct. 04, 2017]
- [13] N. Mohan, T. Underland and W. Robbins, Power Electronics: Converters, Applications, and Design, 3rd Edition.
- [14] J. Li, T. Abdallah, C.R. Sullivan, "Improved calculation of core loss with nonsinusoidal waveforms," in *Proceedings of the IEEE Industry Applications Society Annual Meeting*, pp. 2203–2210, October, 2001.
- [15] G. Ortiz, J. Biela, J. W. Kolar, "Optimized design of medium frequency transformers with high isolation requirements," *36th IEEE Industrial Electronics Society Conference, IECON 2010*, pp. 631-638, Nov. 2010
- [16] A. Atalla et al., "Advancements in high power high frequency transformer design for resonant converter circuits," 2016 IEEE Energy Conversion Congress and Exposition (ECCE), Milwaukee, WI, 2016, pp. 1-8.
- [17] L. A. Garcia Rodriguez and J. C. Balda, "A comparison of isolated DC-DC converters for microinverter applications," 2013 Twenty-Eighth Annual IEEE Applied Power Electronics Conference and Exposition (APEC), Long Beach, CA, USA, 2013, pp. 2084-2091.
- [18] E. O. Lindstrom, L. A. Garcia-Rodriguez, A. R. Oliva and J. C. Balda, "Designing an optimum non-dissipative LC snubber for step-up flyback converters in DCM," 2017 IEEE 8th Latin American Symposium on Circuits & Systems (LASCAS), Bariloche, 2017, pp. 1.