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
1omalosa@uark.edu, 2lgarciar@uark.edu, 3jbalda@uark.edu
Sudip K. Mazumder,
NextWatt LLC
Hoffman Estates, IL, USA
sudipkumarmazumder@gmail.com

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 ANSYS™ 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 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 ANSYS™ 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 high-power 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 HFT 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 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:

\[
\frac{(h \cdot \rho \cdot B_{sat} \cdot \Delta T)^{1/3}}{\sqrt{4 \left( \frac{\rho \cdot k_{w} \cdot k_{a}}{\mu \cdot n} \right)^{1/3} \left( k_{w} \cdot f \right)^{1/2}}} = K_{a} \cdot f \cdot k_{u} \cdot k_{w} \cdot \sqrt{VA} \cdot \frac{1}{\sqrt{7}}.
\]

The minimum area product \( A_{p} \) is then calculated as follows [9]:

\[
A_{p} = \left( \frac{\sqrt{2} \cdot \sum VA}{K_{a} \cdot B_{opt} \cdot k_{u} \cdot k_{w} \cdot \sqrt{k_{s} \cdot \Delta T}} \right)^{1/7}.
\]

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

The temperature rise \( \Delta T \) as function of the area product \( A_{p} \) with assumed total power losses \( P_{losses} \) is estimated by [10] as:

\[
\Delta T = 450 \left( \frac{P_{losses}}{K_{c} \cdot \sqrt{A_{p}}} \right)^{0.826},
\]

where \( K_{c} = 39.2 \) 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 \( \Delta T \) 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 \(([P_{loss}/m^{2}] \cdot \text{volume})\). 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 \( \mu_{r} \) 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 cm². This core can be used to design a 64 kW (max) MFTs assuming \( \Delta T = 50 \degree C \). For a design with a higher rated power, designer should consider

Table I. Core Material Comparison [6]

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

Table II. Constant Values of Optimal Flux and Area Product

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>coefficient of heat transfer ( h ) (1000 W/m²°C)</td>
<td>10</td>
<td>Dimensionless</td>
<td>40</td>
</tr>
<tr>
<td>Initial wire resistivity ( \rho ) (Ω · m)</td>
<td>1.78 x 10⁴</td>
<td>Dimensionless</td>
<td>10</td>
</tr>
<tr>
<td>Window utilization factor ( k_{u} )</td>
<td>0.4</td>
<td>Dimensionless</td>
<td>5.6</td>
</tr>
<tr>
<td>Waveform factor square wave ( K_{s} )</td>
<td>4</td>
<td>Core stacking factor</td>
<td>0.95</td>
</tr>
<tr>
<td>Waveform factor ( k_{s} )</td>
<td>0.54 x 10⁻³</td>
<td>Expected temperature ( \Delta T ) (°C)</td>
<td>50</td>
</tr>
</tbody>
</table>

Table III. Material Coefficients [10]

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Nanocrystalline</th>
<th>Amorphous</th>
</tr>
</thead>
<tbody>
<tr>
<td>( K_{c} )</td>
<td>2.3 w/m²</td>
<td>1.3617 w/m²</td>
</tr>
<tr>
<td>( a )</td>
<td>1.32</td>
<td>1.51</td>
</tr>
<tr>
<td>( \beta )</td>
<td>2.12</td>
<td>1.74</td>
</tr>
<tr>
<td>( B_{sat} )</td>
<td>1.2 T</td>
<td>1.56 T</td>
</tr>
</tbody>
</table>
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 \( \frac{\mu}{\text{area}} \) and leakage inductance \( \frac{k}{\text{area}} \) when designing MFTs/HFTs is very challenging at high-power levels. The main goal is to keep \( \frac{\mu}{\text{area}} \) at the value constrained by:

\[
\frac{\mu}{\text{area}} < \frac{V_{\text{rms}}}{\pi f C},
\]

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., \( 0 < C_f \leq 1 \)), where its value depends on the type of isolated power converter. Lower values of \( C_f \) result in a significantly larger \( \frac{\mu}{\text{area}} \) 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 \( \frac{k}{\text{area}} \) is required for DAB-based MFTs/HFTs to maximize the transferred power [2]; i.e.:

\[
\frac{k}{\text{area}} = \frac{V_{\text{rms}}}{2 \pi f \Phi (\pi - \phi)},
\]

where \( \phi \) 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:

\[
L_m = \frac{N^2 S_e A \mu_e \mu_s}{l + l_g \mu_s} \left( 1 + \frac{l_g}{\sqrt{S_e A_e}} \ln \left( \frac{2C}{l_g} \right) \right),
\]

\[
L_k = \frac{\pi (M L T) N^2}{N_e} \left( \sum d_{\text{in}} + \frac{3}{2} d_e \right) 10^{-8} [H],
\]

where \( S_e \) is the number of cores needed to achieve the required area product \( A_e \); \( \mu_e \) and \( \mu_s \) are the air and material permeabilities; \( l \) is the mean length of the magnetic path; \( C \) is the window length of the core; \( M L T \) is the mean length of the turns; \( N \) is the number of turns [9]; \( l_g \) is the air-gap length; \( N_e \) is the portion of the dimension \( C \) covered by the windings; \( d_{\text{in}} \) is the distance between primary and secondary windings and \( d_e \) 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]:

\[
N = \frac{V_{\text{rms}}}{k_1 B_{\text{max}} A_{l} f},
\]

where \( k_1 \) 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:

\[
l_g = 2 N I_m \frac{\mu_e}{B_{\text{max}} - \mu_s}.
\]

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 \( \frac{\mu}{\text{area}} \) using (4) by selecting the proper value of \( C_f \).

Fig. 2. Main physical parameters of a MFT/HFT.
Step 2: The saturation flux density $B_{sat}$, and the Steinmetz coefficients $k$, $\alpha$ and $\beta$ 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 $l_g^*$ 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 $\Delta T$ 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]:

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

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 Vrms, $N_p/N_s = 1$). Based on the given

<table>
<thead>
<tr>
<th>Specified Parameters</th>
<th>Required</th>
<th>Calculated</th>
</tr>
</thead>
<tbody>
<tr>
<td>Magnetizing Inductance $L_m$</td>
<td>$\leq 68 \mu H$</td>
<td>$30 \mu H$</td>
</tr>
<tr>
<td>Leakage Inductance $L_k$ ($\mu H$)</td>
<td>Small as possible</td>
<td>1.11</td>
</tr>
<tr>
<td>Efficiency $\eta$ (%)</td>
<td>$\geq 99$</td>
<td>99.7</td>
</tr>
</tbody>
</table>
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 $L_m$ and $L_k$ 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, having a slightly larger magnetizing ($L_m > L_m^\prime$) is accepted as a tradeoff 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-gap of $l_g = 0.87$ mm was obtained from (6) as $L_m = 38.1 \mu$H which is very close to the measured prototype value $L_{exp} = 40.5 \mu$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 $L_d = 1.19 \mu$H where the experimental value is $L_{exp} = 1.04 \mu$H. The error between these two values was calculated as 14.4 % assuming that there is no space between turns which makes $N_g$ 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 $L_d$, 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 ($L_m$, $L_d$ and $C_{ps}$) have low changes (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

<table>
<thead>
<tr>
<th>Table V. Specified Parameters of the HFT Prototype</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameters</td>
</tr>
<tr>
<td>Rated Power</td>
</tr>
<tr>
<td>Primary DC Voltage</td>
</tr>
<tr>
<td>Primary RMS Current</td>
</tr>
<tr>
<td>Turns Ratio</td>
</tr>
<tr>
<td>Switching Frequency</td>
</tr>
</tbody>
</table>

Fig. 5. Prototype of 1020 kW, 120 V<sub>rms</sub> and 100 kHz high frequency transformer.
equations, a flyback converter capable of switching at 100 kHz with an input voltage of 120 VDC 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.

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.

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 voltage \( V_o \) 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 ANSYS™ 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