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A High-Frequency Transformer(HFT) design for a Dual Active Bridge(DAB) converter

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Abstract

This paper provides a detailed description of the design approach for a 1 kW high-frequency transformer for use in dual-active-bridge converter applications. The unique features of the dual-active-bridge converter, such as bidirectional power transfer, soft-switching, and high efficiency, largely depend on the design of the isolation transformer. In dual-active-bridge power transfer, leakage inductance plays an important role. To increase efficiency, the selection of leakage inductance is essential, especially under low-load conditions. For efficient operation, the leakage inductance must be properly selected, especially in low-load conditions. It includes selecting the core, designing the winding, calculating the turns ratio, magnetising inductance and controlling the leakage inductance. ANSYS Maxwell was used to create a model of the same and to determine the leakage inductance and losses. Based on the comparison results, a suitable winding/core structure is determined. The inductances and turns ratio, as described above, are employed in a 400/48-volt converter system operating at a switching frequency of 100 kHz under Single-Phase Shift modulation. The work presented in this paper will be very beneficial for engineers and researchers working in the field of HFT.

Keywords: Dual Active Bridge converter, High Frequency Transformer, inductance and losses.

I INTRODUCTION

Growing requirements for high performance, high power density, and high compactness in power conversion systems, driven by their use in batteries, statistical capabilities, renewable energy sources, and electric cars, have created an attractive opportunity to advance digital power conversion techniques. Among all available topologies, the Dual Active Bridge (DAB) converter is of immense significance due to its attractive advantages, including flexibility, galvanic isolation, bidirectional energy-flow capability, and soft-switching operation across various modes [1].

High-frequency galvanic isolation transformers are essential to the functionality and physical integration of DAB converters. Transformer design significantly impacts the converter's efficiency, thermal profile, electromagnetic compatibility (EMC), and total power density; it is not merely a passive component for voltage scaling and isolation [2]. High-frequency transformer design difficulties arise as switching frequencies increase into the hundreds of kilohertz and even megahertz ranges, due to the use of Wide Bandgap (WBG) semiconductors such as silicon carbide (SiC) and gallium nitride (GaN) [3]. Nevertheless, the core loss in the HFT will likewise rise with the switching frequency. Loss optimisation [4] is thus another crucial design criterion for the HFT utilised in a DAB.

A number of aspects need to be carefully considered when designing HFTs for DAB converters, including the choice of core material, winding configuration, leakage inductance optimisation, and parasitic capacitance minimisation [5]. The DAB converter requires the transformer's leakage inductance for operation, which plays an important role in power transfer capability, soft-switching, and efficiency [6]. Leakage inductance is usually minimised in conventional transformer designs, but DAB applications frequently require precise control of this value to achieve optimal performance across a range of load conditions [7]. For DAB converters, HFT design must address several technical challenges, including Core Loss Management, Winding Design Optimisation, Thermal Management, and Electromagnetic Compatibility.[8-10].

A number of studies on the design of HFTs for DAB converters have recently been carried out, identifying a variety of optimisation techniques. One of these studies, conducted by Smith et al. [11], proposed a general approach to HFT design, emphasising the importance of considering the efficiency-power density trade-off in DAB systems. As regards core materials, Johnson and Lee [12] assessed the efficiency of ferrite, nanocrystalline, and amorphous cores and concluded that nanocrystalline cores exhibit better properties at frequencies above 50 kHz. Likewise, Brown et al. [13] have investigated how the core shape influences the loss distribution and thermal characteristics of high-frequency transformers. The optimisation of winding design has been extensively researched recently.

The optimisation of windings has likewise been a priority; specifically, the model developed by Wang and Chen [14] predicts high-frequency winding losses, accounting for the effects of skin and proximity. Additionally, the research direction is moving towards a combination of gadgets and education; in other words, Rodriguez et al. [15] have shown that optimising transformer characteristics (especially leakage inductance) improves converter performance by 3%. Besides, Kim and Park [16] addressed the problem of compactness and investigated the thermal coupling process between the transformer and the electrical semiconductor.

The main contributions of this research work are attributed to the HFT design methodology, which is outlined and tested in Ansys Maxwell. The proposed design methodology includes selecting appropriate core materials and winding configurations, followed by their optimisation based on the specified power rating and operating frequency range. In addition, the transformer is analysed with respect to key performance parameters, including core losses, copper losses, and leakage inductance under different winding topologies.

The remainder of this paper is organised as follows: Section II presents the DAB converter needs, while Section III provides information on the selection procedure for the high-frequency transformer centre and associated winding arrangements. Section IV describes the implementation and simulation of these designs in Ansys Maxwell, and Section V provides the study's conclusions.

II DESIGN SPECIFICATIONS

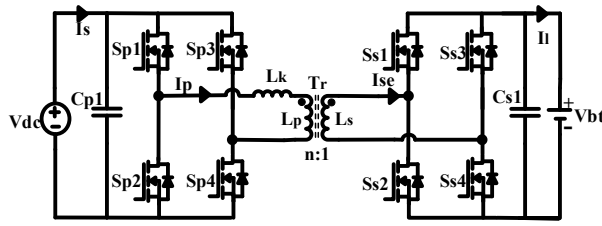


Figure. 1 Circuit diagram for Dual Active Bridge Converter

The Dual Active Bridge (DAB) converter architecture comprises two full-bridge (H-bridge) circuits connected through an over-frequency transformer (HFT). Regarding the mechanism of active power transfer in the HFT, the governing relationship is described as follows [17]

$$P = n \frac{V_i V_o}{2\pi f_s L} \phi \left(1 - \frac{\phi}{\pi}\right) \quad (1)$$

Based on equation (1), the variables are as follows: n denotes the transformer turns ratio, and V_i and V_o represent the input and output voltages, respectively. The phase shift between the primary and secondary bridges is denoted by ϕ . In addition, f_s represents the switching frequency, and L denotes the leakage inductance, which can be realised either as a discrete external inductor or as the inherent leakage of the high-frequency transformer.

The configuration parameters for the Dual Active Bridge (DAB) converter are summarised in Table 1. The maximum allowable inductance is calculated to be 200 μH , assuming a maximum phase shift of 90 degrees for the DAB converter's 1kW power rating. Therefore, 200 μH represents the upper limit of the configuration for L . It is found that an increase in inductance leads to higher circulating currents, resulting in elevated losses and reduced standard performance. Finally, the magnetisation current is limited to 2%-5% of the nominal current [18], as this current no longer contributes to the energy-efficient electrical switching mechanism.

Table I : Design parameters of DAB

Parameter	Value
V_i	400V
I_i	2.5A
P_r	1kW
V_o	50V
I_o	20A
f_s	100kHz

The literature [19] presents numerous methods for designing high-frequency transformers. The design procedure followed in this work follows a systematic progression: first, the magnetising inductance is determined; Subsequently, the core material, geometry, and specific winding configurations are selected. Based on the converter specifications, the current requirements for both the primary and secondary windings are calculated, and the results are summarised in Table I.

III DESIGN PROCESS:

(A) Selection of winding :

To facilitate natural air cooling, a current density of 4 A/mm² is set as the design criterion for calculating the total cross-section area of the transformer windings. Given the converter's high-frequency operation, skin and proximity effects become important design considerations that can significantly affect overall performance. To minimise these parasitic effects, the use of multi-stranded Litz wire is recommended. For optimal performance, the radius of each individual strand within the Litz wire should be kept under the calculated skin depth (δ), as expressed by the following relationship:

$$\delta = \frac{1}{\sqrt{\pi f_s \mu \sigma}} \quad (2)$$

Where f_s denotes the switching frequency, while σ and μ represent the conductivity and permeability of the copper, respectively. Substituting a switching frequency of 100 kHz into equation (2) gives skin depths of approximately 0.2 mm. Thus, AWG 33 wire is chosen for the individual strands of the Litz wire [20]. To meet current requirements, the number of strands is set to 40 for the primary winding and 200 for the secondary winding, yielding total cross-sectional areas of 1 mm² and 5 mm², respectively.

(B) Selection of core :

(i) Material selection

Table II: Magnetic materials and their properties

Material	Model	Kc	α	β	Core loss mW/cm ³	Curie temperature °C	Permeability	Peak flux density
Ferrites	Epcos N87	16.9	1.25	2.35	684	210	2200	0.49
Nano crystalline	Viproperm 500F	2.3	1.32	2.1	311	600	1500	1.2
Amorphous	Metglas 2605	0.053	1.81	1.74	3614	399	10000-150000	1.56
Si iron	Unisil 23M3	3.388	1.70	1.90	50338	745	5000-10000	2.0
Permalloy	Magnetics Permalloy 80	0.448	1.56	1.89	1349	460	20000-50000	0.82
Powdered iron	Micro-metals 75 μ	1798	1.02	1.89	10807	665	75	06-1.3

According to the study presented in [18], Table II provides a comparative overview of many transformer core materials and their respective electromagnetic properties. In view of the converter design requirements, the switching frequency is fixed to 100 kHz, and the peak

magnetic flux density ($B_{\max}$) is set to 0.2 T in line with general company practices. To determine the core losses, the Steinmetz equation is used, which is defined in equation (3)

$$P_{fe} = k f_s^\alpha B_{\max}^\beta \quad (3)$$

In this expression, P_{fe} denotes the time-averaged core loss per unit volume, while k , α , and β are empirical constants derived from the manufacturer's technical datasheets.

Magnetic core materials exhibit different performance characteristics that make them suitable for different operating environments. Ferrites (e.g., Epcos N87) provide balanced performance with remarkably low core losses (684 mW/cm³) and permeability of 2200; however, they are limited by a lower Curie temperature (210°C) and a peak flux density of (0.49 T). Nanocrystalline materials (e.g., Vioperm 500F) offer superior thermal stability at excessively high Curie temperatures (600°C) and excellent low-loss characteristics (311 mW/cm³), which makes them moderately permeable (1500).

The Amorphous materials have a high flux density of 1.56 T and a wide permeability range. However, it exhibits high core losses (3614 mW/cm³), resulting in low efficiency at high frequencies. Silicon iron (Unisil 23M3) exhibits the highest Curie temperature (745°C) and peak flux density (2.0 T), making it suitable for high-power applications, but at the cost of extremely high core losses (50,338 mW/cm³). Permalloy (Magnetics Permalloy 80) offers the highest permeability range (20,000-50,000) with reasonable core losses (1349 mW/cm³), making it excellent for low-field applications, though its peak flux density is limited (0.82 T). The powdered iron materials have very low permeability and high core loss when compared to amorphous materials. Moreover, they are suitable for distributed air-gap applications, which provide better efficiency.

The selection of magnetic core material depends on the following factors: operating frequency, thermal effects, magnetic field strength, and efficiency. Based on the above parameter constraints, a comparison of different magnetic materials is conducted, with ferrite being the better choice for this high-frequency converter topology. The various ferrite cores and specifications are represented in Table III.

(ii) Core shape

Magnetising current in a high-frequency transformer (HFT) does not contribute to the transmission of power energy within a Dual Active Bridge (DAB) converter. Thus, optimising transformer inductance involves a vital trade-off: while a higher magnetising current may degrade light-load performance, choosing a very low magnetising inductance increases both the required wire gauge and the transformer's physical footprint. To balance these issues, the optimal inductance is determined by adhering to industry practice, which maintains magnetising current at or below 2% of nominal rated current [21], as defined by the following expression:

$$L_m = \frac{V_{in}}{4f_{sw} I_m} \quad (4)$$

After obtaining the magnetising inductance from equation (4), the number of turns of the primary winding is determined by the core shape. The mathematical expression of the above

is $N_1 = \sqrt{\frac{L_m}{A_1}}$. The secondary winding turns are determined through the turns ratio. Figure 2 shows the high-frequency transformer design methodology as a flowchart.

Different magnetic cores can be used for high-frequency transformers. To identify which shape works best for this system, a comparative analysis was conducted using ETD, UI, and toroidal cores. The primary (N_1) and secondary (N_2) winding turns are determined by the core dimensions and converter specifications. These values are ultimately used to determine the window space factor (W_x) and the maximum magnetic flux density (B_m), calculated from equations (5) and (6), respectively. A core is considered suitable for the installation if these parameters fall within the specific limits - specifically, $W_x \leq 0.3$ and $B_m \leq 0.2$ T - according to established industrial design standards [22]. The governing mathematical expressions for W_x and B_m are described as follows:

$$W_x = \frac{N_1 A_i + N_2 A_i}{A_w} \quad (5)$$

$$B_m = \frac{V_l D}{4 f_s N_1 A_e} \quad (6)$$

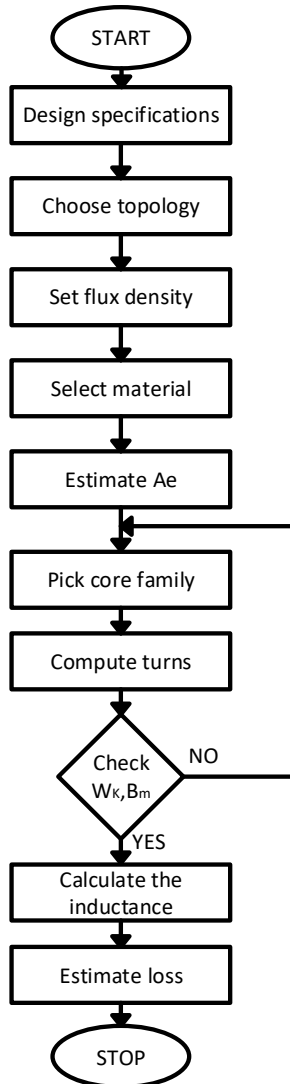


Figure 2. Flowchart for the HFT core selection.

Here, A_w is the window area of the core, D is the duty cycle of the input voltage waveform, and V_1 and N_1 are the input voltage and the number of primary winding turns, respectively. The winding parameters are represented in Table IV.

Table III: Ferrite core Specifications

Part number	Material	Inductance Factor $A_i(\text{nH/turns})$	Effective Area (A_e) mm^2	Effective Length (L_e) mm	Effective Volume (V_e) mm^3
ETD 39/20/13	3C95	3650	125	92.2	11500
UI 93/76/30	3C90	6400	840	258	217000
T 80/20/50	PC40	4320	295	197	58000

Table IV: Winding parameters

Part number	N_1	N_2	Window space factor (W_x)	Maximum flux density (B_m)
ETD 39/20/13	80	10	0.012	0.05 T
UI 93/76/30	56	8	0.075	0.01 T
T 80/20/50	68	8	0.328	0.02

(C) Winding arrangement

The leakage inductance (L_k) is limited to a maximum of 200 μH at the maximum phase shift angle. In accordance with this criterion, the winding arrangements were systematically evaluated across the selected core geometries. To achieve the specified transformer leakage inductance, individual winding configurations are designed. The technique for calculating the leakage inductance referred to in [23] is shown below. The leakage inductance of the primary winding (L_{kp}) and the secondary winding (L_{ks}) can be expressed in equations (7) and (8):

$$l_{kp} = L_{pp} - kM \quad (7)$$

$$l_{ks} = L_{ss} - \frac{M}{k} \quad (8)$$

Where L_{pp} , L_{ss} and M are the self inductances of primary, secondary and mutual inductances. The equivalent leakage inductance referred to the primary expressed in equation (9)

$$l_{12} = l_{kp} + k^2 M \quad (9)$$

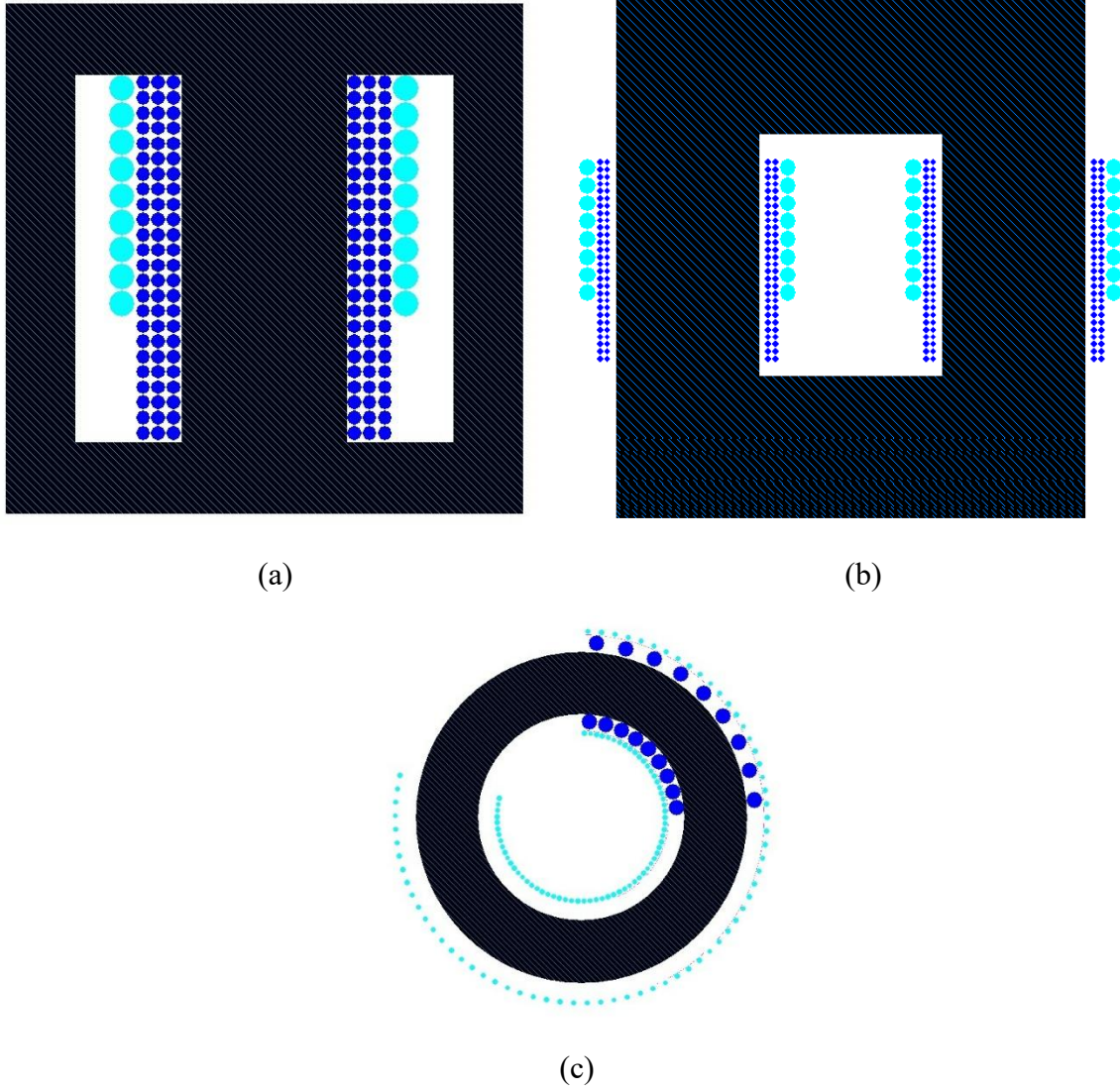


Figure. 3 Winding arrangements of (a) ETD core, (b) UI core and (c) Toroidal core

To determine the leakage inductance of the various core applicants, 2D finite element analysis (FEA) was performed using ANSYS PEmag, as shown in Figure 3. Given that particular winding configurations produce different leakage inductance values, the analysis focused on designing within the configured limits. Once a configuration was found that met those requirements and achieved the desired leakage inductance, the winding arrangement was finalised.

IV RESULTS AND DISCUSSION

A. Electromagnetic Performance and Leakage Inductance

The FEA-based analysis confirms that the proposed ETD core geometry effectively controls the leakage inductance required for the 1kW DAB converter. Based on the 2D FEA calculations, the equivalent leakage inductance referred to the LV side is $2.06\mu\text{H}$, while the HV-referred value is $131.92\mu\text{H}$. The above values enable stable bidirectional power transfer in the DAB converter. The winding configuration of the ETD core yields primary and secondary self-inductance values of 22.42 mH and $350.36\text{ }\mu\text{H}$, respectively.

B. Loss Analysis and Efficiency Estimation

To evaluate the transformer's thermal robustness, a comprehensive loss analysis was performed.

Copper loss: Using the calculated equivalent resistance $R_{eq,LV} \approx 5.43 \text{ m}\Omega$ which is shown in Figure 4 and the rated output current ($I_o=20\text{A}$), the total copper loss is 2.17W. The uneven current density distribution observed in the FEA simulations shows that using multi-stranded litz wire is vital for reducing the effects of proximity to excessive frequency.

Core loss: The transient analysis is performed for three core shapes in Ansys Maxwell at 100kHz, yielding an average core loss of 1.8W for the ETD core, as shown in Figure 5. The waveforms show that the pattern remains consistent over time and follows the Steinmetz equation. Moreover, the N87 ferrite material provides better thermal performance for the high-frequency application.

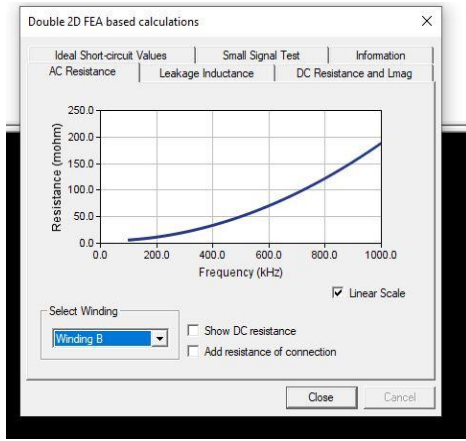


Figure.4. AC resistance

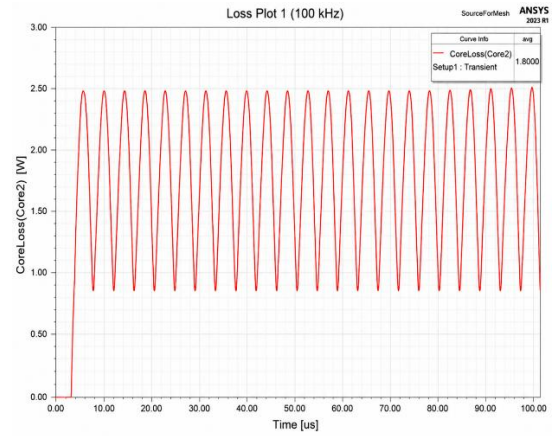


Figure.5. Core loss plot

C. Magnetic Field and Flux Distribution

The magnetic field distribution was analysed to ensure efficient current coupling and to detect possible saturation points inside the ETD centre. As shown in Figure 6, the maximum flux density reached 33,008 μT . This value remains well within the linear working range of the N87 ferrite material, confirming that the core is operating safely within saturation limits

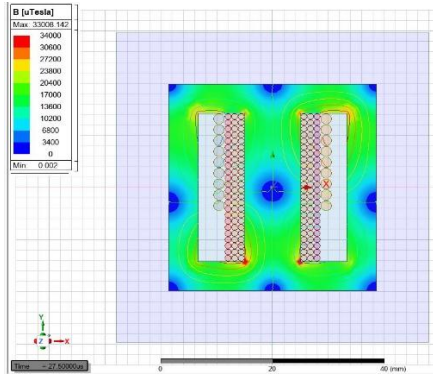


Figure.6. Magnetic field density

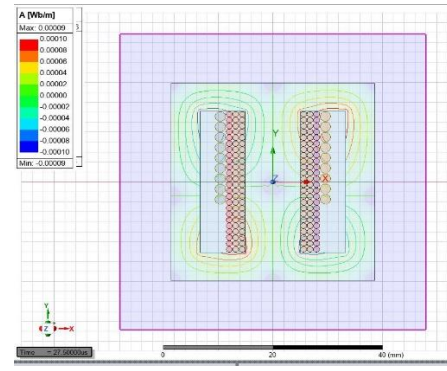


Figure.7. Flux distribution

at nominal 1kW. The magnetic flux line distribution shown in Figure 7 confirms the high coupling efficiency between the primary and secondary windings. This configuration is essential to minimise leakage and ensure the powerful power transfer required for the DAB topology. Figure 8 shows that the maximum flux density across the core at 25us is approximately 0.18-0.19 T. The value is well within the design constraint value of 0.2T to avoid saturation. Moreover, Figure 9 shows that the current density of the core is $8.746 \times 10^6 \text{ A/m}^2$, also within safe limits. The above is achieved by selecting multistrand litz wire. This transformer design is well-suited for operation at 100 kHz due to reduced skin and proximity effects. The comprehensive design undertaken in consideration of company standard values results in a highly reliable design for the 1 kW Dual Active Bridge converter, thereby providing a solid foundation for maximum performance.

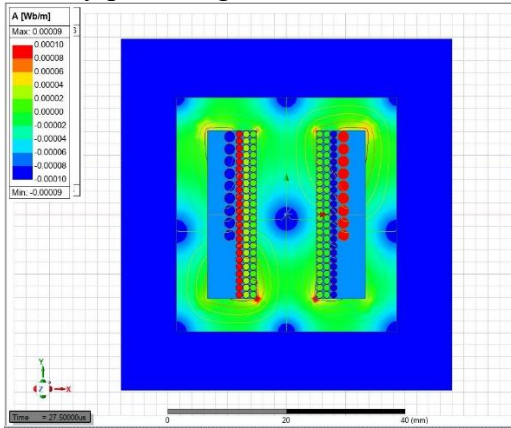


Figure. 8. Peak flux density

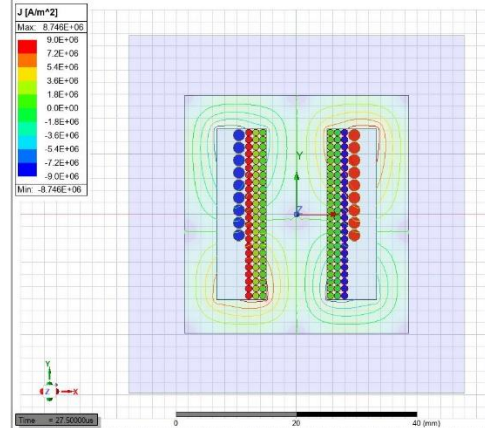


Figure. 9. Current density

D) Implementation of transformer parameters in DAB topology

The simulation results in LTspice and the corresponding typical waveforms confirm the overall theoretical performance of the designed 1 kW Dual Active Bridge converter. As illustrated in the circuit schematic in Figure 10, the system utilises a high-frequency transformer interface with primary and secondary H-bridge configurations, incorporating leakage inductance (Lk) to facilitate energy exchange via single-phase-shift control.

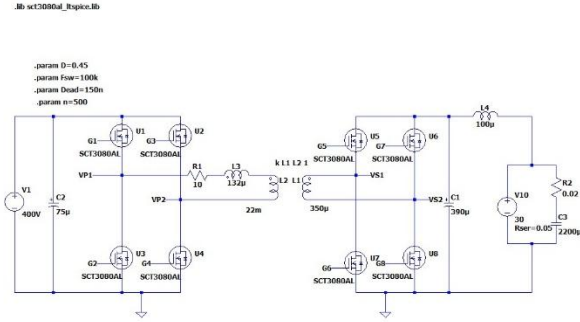


Figure. 10. DAB Schematic

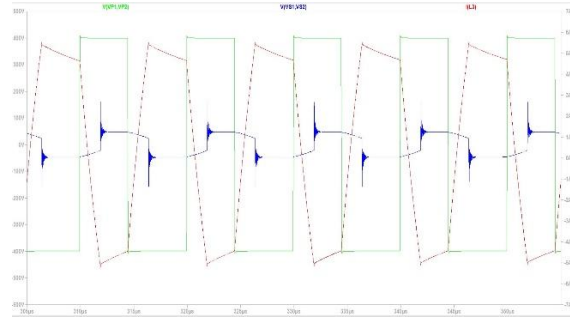


Figure. 11. Waveforms of DAB

The Steady-state waveforms in Figure 11 illustrate the phase-shifting operation between the primary bridge voltage $V((VP1,VP2))$ (green) and the secondary bridge voltage $V((VS1,VS2))$ (blue). The transformer current $I(L3)$ is shown in red, which justifies the choice of leakage inductance; consequently, the power flow is transferred effectively. The inductor current waveform is the resultant of the difference between the primary and secondary voltages of the transformer using Single Phase Shift Modulation. The output voltage waveform exhibits brief spikes due to high-frequency switching. Such behaviour is maintained within safe operating limits, allowing the converter to operate in a stable condition. This correlation between the Ansys-simulated parameters, such as leakage inductance, and the circuit-simulation waveform justifies the winding arrangement and core selection for the high-frequency transformer of the DAB converter. Moreover, the design method is suitable to meet the overall performance objectives.

E) Comparison with latest studies

Compared with Choi et al [24], Guo et al. [25] represent a completely different design philosophy regarding the methodology used in this paper for high-frequency transformers in Dual Active Bridge (DAB) converters. In particular, while Choi et al. use a fully integrated transformer design to increase the power density, Guo et al. use a wide-input-current-fed configuration. Choi et al. use insertion-depth adjustment to dynamically adjust leakage inductance over a wide range of loads to achieve zero-voltage switching, but both methods introduce additional complexities or strict criteria for magnetic coupling. On the other hand, the current manuscript presents a novel method that is reliable, easily predictable and applicable in an application using an ordinary ETD 39/20/13 core. This design works around the problems of high-voltage insulation mentioned by Guo et al. as well as the fatigue of movement as mentioned by Choi et al., resulting in a transformer loss of 3.97 W. Use of multi-stranded AWG 33 Litz wire successfully eliminates both the problems of skin effect and proximity effect as well as keeps the current density at a safe value of $8.746 \times 10^6 \text{ A/m}^2$. Moreover, the structure maintains a maximum magnetic flux density of 0.18-0.19 T to avoid saturation. All these factors increase its strength, thermal stability, and efficiency while ensuring no movement in the structure.

Conclusion

This paper explains the design method for a transformer for high-frequency applications in a 1 kW dual-active-bridge converter and compares it with the simulated results. Through meticulous Finite Element Analysis (FEA) utilising ANSYS PEmag and steady-state analysis of LTspice, this finding clearly indicates that transformer optimisation is achieved through appropriate calibration of leakage inductance and coil configuration. The choice of the geometry of the core ETD, along with the usage of the multi-stranded Litz wire, was very important for the maintenance of proper operating parameters within protection limits, particularly the maintenance of magnetic flux density around 0.003 T. Both the core transformation model and waveforms from the converters prove the dependability of the design methodology approach. The thermal equilibrium of the core is an entry point for its utilisation in high-power-density and novel converter architectures.

Author contributions

Rajesh Sethuraman conceived and designed the study, developed the high-frequency transformer design methodology, and performed the core, winding, and leakage inductance calculations. Rajesh Sethuraman carried out the finite element analysis simulations in ANSYS Maxwell/PEmag and performed the LTspice circuit-level validation of the DAB converter. Rajesh Sethuraman and Mageshvaran Rudhramoorthy analysed and interpreted the simulation results. Rajesh Sethuraman wrote the manuscript. Mageshvaran Rudhramoorthy supervised the research and reviewed the manuscript. All authors read and approved the final version of the manuscript.

Data availability statement

The datasets generated and analysed during the current study are available in the Zenodo repository at <https://doi.org/10.5281/zenodo.22293225>, but access is currently restricted as the work has not yet been published. Data are available from the corresponding author upon reasonable request and will be made openly accessible after publication.

Competing interests

The author(s) declare no competing interests.

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