

Article

Design and Analysis of a Three-Phase High-Frequency Transformer for Three-Phase Bidirectional Isolated DC-DC Converter Using Superposition Theorem

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Abstract: Battery energy storage systems based on bidirectional isolated DC-DC converters (BIDCs) have been employed to level the output power of intermittent renewable energy generators and to supply power to electric vehicles. Moreover, BIDCs use high-frequency transformers (HFTs) to achieve voltage matching and galvanic isolation. Various studies have recently been conducted using soft magnetic materials, such as nanocrystalline, amorphous solids, and ferrite, to develop more compact and effective transformers with superior power densities. The HFTs in three-phase BIDCs are composed of three magnetic cores. However, this leads to low power density and high cost. Besides, the three-phase (3P) ferrite core has not been investigated for high-power converters such as 3P-BIDCs. This paper presents the design and development of a 3P-EE ferrite magnetic core for 3P-BIDCs. The area product design method was used to determine the core and winding design. The paper also proposes the use of the superposition theorem in conducting a magnetic circuit analysis to predict the flux density and magnetising inductance of the transformer core. Moreover, the use of the superposition theorem allowed the required air-gap length for balancing the distribution of flux density and magnetizing inductance in the transformer core to be determined. The balanced flux distribution and magnetizing inductance resulted in a uniform core loss and temperature in the transformer. This paper also presents the experimental results of the designed HFT operated in a 300-V, 3-kW 3P-BIDC. The experimental results showed that the proposed HFT achieved a balanced flux density and magnetizing inductance with a high power density and low cost. Moreover, the transformer performed at a maximum efficiency of 98.67%, with a decrease of 3.33 °C in the overall temperature of the transformer as compared to the transformer without air gaps.

Keywords: bidirectional isolated DC-DC converter; high-frequency transformer; battery energy storage system; three-phase electromagnetic devices; magnetic circuit analysis; superposition theorem; magnetic equivalent circuit



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1. Introduction

In light of the global energy crisis and increasing environmental concerns about fossil fuels, there is a shift towards generating electricity from renewable energy sources such as solar and wind. According to the reports by [1,2], 71% of all new PV installations (102.4 GW) globally in 2018 were utility-scale PV systems. As solar PV and wind energy are intermittent in nature, utility-scale battery energy storage systems are required to ensure that the power systems can function reliably and meet load demand. Moreover, electrification of transportation is seen as one of the solutions to reducing the use of fossil fuel and slowing down the effects of global warming. Battery energy storage systems based on a DC-DC converter should have bidirectional power transfer capability, galvanic

isolation, high-power density and high efficiency. The power density of a DC-DC converter can be increased by applying a higher switching frequency (f) to downsize the magnetic components. Nevertheless, the frequency level is constrained by a higher switching loss [2].

DC-DC converters are essential for promoting energy sustainability due to their ability to enhance the efficiency of renewable energy systems and electric vehicles; they significantly minimize energy losses and enable power management among various sources. Furthermore, they support decentralized energy solutions, including microgrids, which are increasingly vital in modern energy infrastructure [3]. In electric vehicles, bidirectional DC-DC converters enable regenerative braking and enhance energy yield, thereby reducing emissions [4]. By making renewable energy systems more scalable and economically viable, these converters are instrumental in decreasing dependence on fossil fuels and fostering a sustainable, low-carbon energy future [5].

Although non-isolated DC-DC converters are frequently employed because they are simple, these converters do not provide galvanic isolation. Isolated DC-DC converters that come with transformers are utilised for voltage matching and to provide galvanic isolation. They are also extensively utilised in contemporary power systems, including electric cars and uninterruptible power supplies. The lower voltage levels produced by sustainable energy sources, such as photovoltaic (PV) modules and fuel cells, necessitate step-up converters with high gain to export power to the grid or to serve the load.

Bidirectional isolated DC-DC converters (BIDCs) or dual active bridges (DABs) are the most promising isolated DC-DC topologies. These converters ensure stable and sustainable power transmission across various energy sources while enabling high power conversion efficiency [5]. Their application as solid-state transformers in all-electric ships and aircraft, as well as in energy storage for electric cars and renewable energy systems, has been the subject of several recent studies [2,6–10]. The two most popular types of BIDC are the single-phase (1P-BIDC) and three-phase (3P-BIDC) models. The advantages of these converters include isolation features, soft switching, and bidirectional power flow. A 3P-BIDC can have smaller input and output filters, a lower root mean square (RMS) current in its transformer, and a higher power density than a 1P-BIDC. Moreover, the 3P-BIDC with its greater efficiency and power density, lower backflow power, and lower switching stresses, is becoming increasingly more popular than the 1P-BIDC [11–14].

A high-frequency transformer (HFT), which is the key component of a BIDC, serves three purposes: (i) ensuring galvanic isolation between high and low-voltage windings for safe operation; (ii) matching the voltages on the DC sides by using the transformer winding ratio; and (iii) maximising power transfer by exploiting the leakage flux of the transformer and using it as an inductor for soft switching [15].

The design of isolated DC-DC converters that have both high efficiency and high power density has recently received significant attention in the industry. One of the key elements contributing to the power density of isolated converters is transformer design. For the transformer design process, the weight and volume must be minimised, and efficiency must be optimised. Compact designs are created by downsizing the self-inductance and core size at higher switching frequencies [16]. Along with temperature increase, winding proximity, and skin effect, the core loss also increases with higher frequencies. Therefore, the design algorithm must be able to minimise losses and temperature, while achieving high power density and low weight.

Numerous attempts to design HFTs have been documented in the literature, but the majority of these studies focused only on the design of 1P-HFTs for 1P-BIDCs as in [16–21] and were extended even further for use in 3P-BIDCs. As in [11], three discrete ferrite cores were considered in a 3P-BIDC for use in charging stations. On the other hand, three toroidal nanocrystalline cores were examined in [22] as an isolation and filtering network for use in a multiport 3P-BIDC. Similar to this, ref. [23] used three single-core transformers with two windings to produce the inductance to operate a 3P-BIDC. However, each transformer had losses and increased the overall temperature. Moreover, the number of components in the structure was increased, which may have reduced the overall power density [23].

There have been attempts to design a compact 3P-transformer for use as a 3P-HFT. As reported in [24], a five-legged core was utilised to prevent core saturation and eliminate harmonic flux and stray flux from unbalanced voltages. A silicon steel core was created by [25] and optimised for a 3P-BIDC converter for off-shore wind turbines at $f = 1$ kHz. To maximise the efficiency of the transformer while minimising losses, magnetising current (I_m), and weight, an objective function was used in the optimisation process. A multi-objective method for designing a grain-oriented electrical steel (GOES)-core transformer operating at 1 kHz for a 3P-BIDC was proposed by [26]. Nevertheless, the HFT design was constrained by the use of silicon steel since it generated a high flux density (B) at low frequencies, which increased the transformer size and core loss.

Conversely, ferrite is most frequently used as an HFT core due to its saturation flux density (B_{sat}) of 0.5 T and relatively low core losses [27]. However, 3P-ferrite cores are not commercially available. As a consequence, 3P-HFTs with ferrite are designed with combinations of different shapes of ferrite. The authors of [28] suggested the use of a 1P-EI core with a middle limb (Limb B) that was wider than the outer limbs (Limbs A and C). To achieve equal cross section area (A_c), Limb B was evenly cut out on both sides [28]. However, the three limbs still suffered from a high flux density that rendered it unsuitable for a core material such as ferrite. Similar to this, a three-phase LLC resonant converter in [29] used a 1P-EC core shape intended for a 3P transformer. Here, the outer core legs had a higher flux density because of the small cross-sectional area. A 3-kW, 20-kHz prototype made of a five-legged ferrite core based on two sets of 1P-EE cores was also tested in [30], and a 3P-HFT with an EE core shape made of stacked “I” cores was built in [31]. On the other hand, a symmetrical triangular core was built by [32] using a cut ETD ferrite core. Nevertheless, these methods produced a 3P-core with several air gaps between the stacked cores, which complicated the control of the magnetising inductance and flux density.

Figure 1 summarises the limitations of the state-of-the-art 3P-HFT structure for the 3P-BIDC. In order to reduce the number of magnetic components for high power density, a compact 3P-EE core was developed in this paper in accordance with the requirements of a 300-V, 20 kHz, 3-kW 3P-BIDC, keeping in mind that all the core limbs had the same A_c . Since ferrite material is less expensive and has a lower core loss, it was chosen for the core material and Litz wire was used to reduce the proximity and skin effects at high frequency.

The main concerns in designing the 3P-HFT were efficiency and power density. The shape of the 3P-EE core suffered from unequal magnetic mean path length (MPL) among the core limbs, which affected the flux density distribution (B) and magnetizing inductance (L_m) [33]. The middle limb exhibited the highest flux density, resulting in an elevated core loss and temperature. Likewise, the magnetizing inductance on the middle limb was higher compared to the outer core limbs, which had a negative impact on the performance of the 3P-BIDC.

The influence of magnetizing inductance on the power transfer capability, reactive power, RMS current, and zero voltage switching of a 1P-BIDC must be considered if the L_m is ≤ 5 times the leakage inductance value, as in [32,34]. However, for a 3P-BIDC, an unbalanced L_m in the 3P-transformer results in unbalanced AC voltages across the transformer windings. Therefore, the power transfer among the 3P-BIDC's legs would be unequal.

Introducing an air gap in the magnetic MPL affects the B and L_m , and a magnetic circuit analysis is typically employed for this purpose. As in [35], a magnetic circuit analysis was proposed for the 1P-transformer, and the air gap length (l_g) could be easily calculated to control the B and L_m . In [36], a magnetic circuit analysis of a 3P-transformer was presented, and the calculated l_g was used to reduce the B in the core limbs and yoke. However, the distribution of B among the core limbs was asymmetrical and an unbalanced inductance was achieved. Although the magnetic circuit analysis proposed for the EC core achieved a balanced L_m , a higher B was exhibited at Limbs A and C of the 3P-transformer [29].

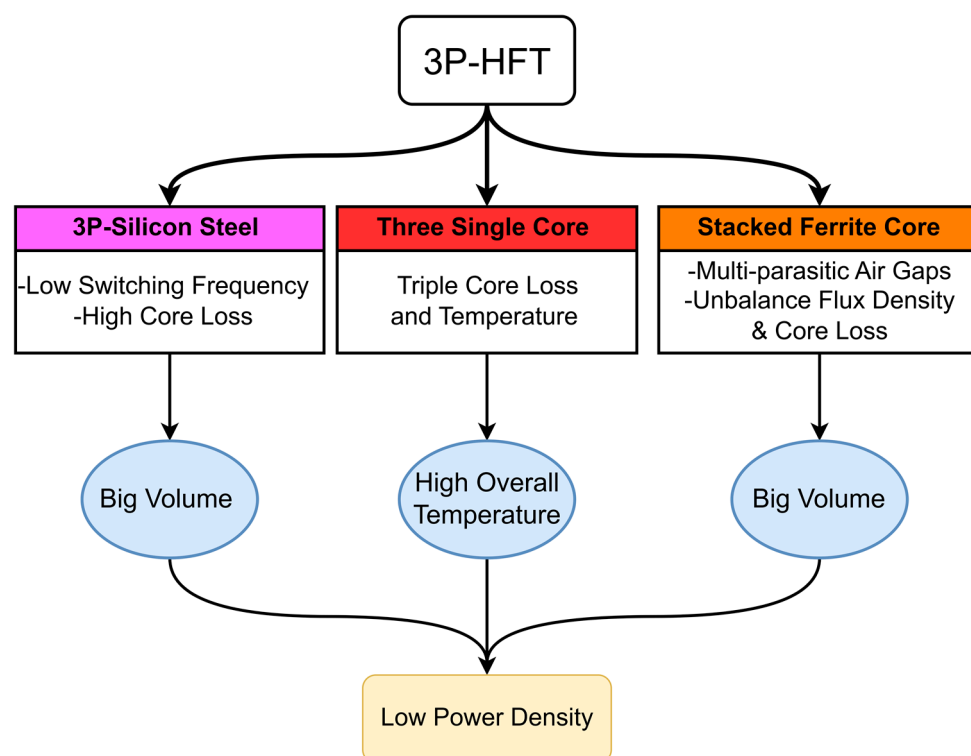


Figure 1. A summary of the review of the literature on 3P-HFTs [8,19,22,23,27,28].

This paper presents a new magnetic circuit analysis method based on the superposition theorem that serves two purposes: (1) accurate prediction of core loss and temperature during the design phase by predicting the B as well as the L_m ; (2) calculation of the necessary l_g for each core limb in order to achieve a uniform distribution of B in the HFT, thereby lowering the core loss and overall temperature. Furthermore, a balanced L_m must also be attained. This paper presents the theoretical analysis of the proposed EE core for a 3P-HFT based on the superposition theorem to gain a better understanding of the magnetic circuit behaviour. Moreover, a finite element analysis (FEA) simulation was performed to achieve a design with high power density and minimal core loss and temperature. Additionally, a 3P-BIDC with a power of 3 kW, input (V_{in}) and output voltages (V_{out}) of 300 V, and switching frequency of 20 kHz was used for an experimental evaluation to assess the designed 3P-HFT. A uniform distribution of B , core loss and temperature were observed in the proposed 3P-transformer, which outperformed previous 3P-HFTs in terms of volume, weight, power density, and cost.

The rest of the paper is organised as follows: Section 2 deals with the formulation of the 3P-EE core problem associated with the magnetic MPL effect. The general design process and magnetic circuit analysis of the 3P-HFT are discussed in Section 3. The theoretical, simulation and experimental results of the designed 3P-HFT combined with a 3P-BIDC are presented in Section 4. Section 5 concludes with a summary of the contributions of this paper.

2. Problem Formulation

Figure 2 shows the circuit diagram of a 3P-BIDC operating in six-stage switching mode with a duty cycle of 50% for each 3P branch. The phase branches have a phase shift of 120° from each other at the switching frequency. The 3P-BIDC has two active voltage-source bridges connected via an HFT. The primary-side converter inverts the input DC voltage into a high-frequency AC voltage via three half bridges, which is then applied to the transformer. The secondary-side converter rectifies the voltage at the secondary

terminals of the transformer [6]. The converter specifications used in this article are shown in Table 1.

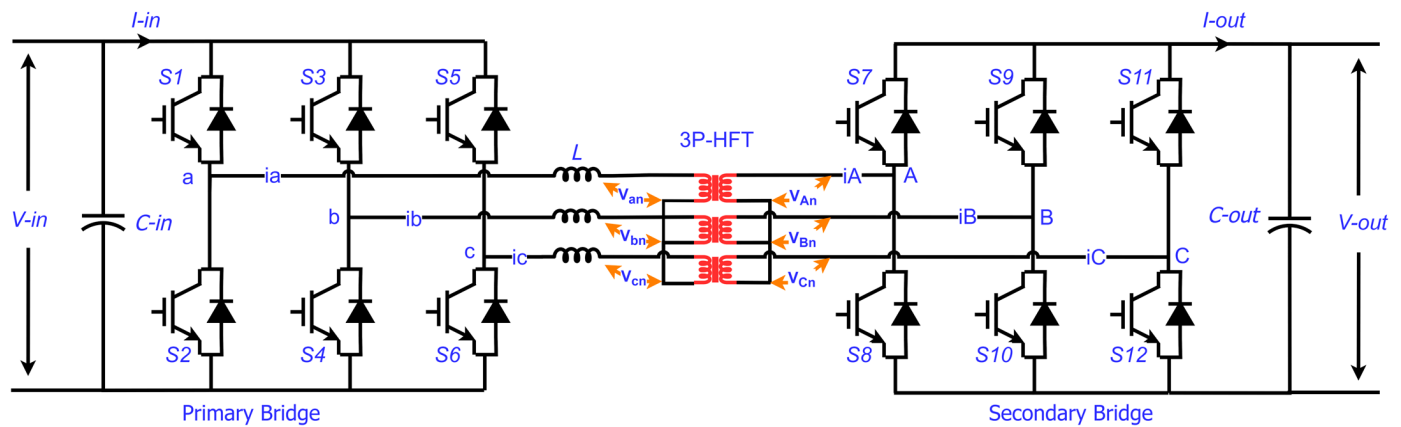


Figure 2. The 3P-BIDC equivalent circuit.

Table 1. The specifications of the 3P-BIDC.

Parameter	Value
Input Voltage (V_{in})	300 V
Output Voltage (V_{out})	300 V
Rated Power (P)	3 kW
Frequency (f)	20 kHz
Phase Shift Angle (ϕ)	4°
Transformer connection	Star-Star
Leakage Inductance (L)	$10.65 \mu\text{H}$
Input Capacitor (C_{in})	$0.5 \mu\text{F}$
Output Capacitor (C_{out})	$0.5 \mu\text{F}$

The transformer B is commonly computed using the applied RMS phase voltage (V_{pn}), f , number of primary turns (N_p), and A_c [35], as follows:

$$B = \frac{V_{pn}}{k_f N_p f A_c} \quad (1)$$

where the flux density is represented by B , and k_f is the waveform coefficient (4.44 for sinewave, and 4 for square wave). Figure 3 shows a 3P-EE core with three limbs and an equal A_c , where each core limb generates its own B according to Equation (1) and shares it with the other core limbs based on the core magnetic reluctance of the magnetic circuit of the transformer.

In considering the equivalent magnetic reluctance of each core limb, Limb B had a shorter magnetic MPL than Limbs A and C, and according to the 3P-cores in [37], the difference in MPL among the core limbs was about 28.3%. The B and L_m values were inversely proportional to the MPL, as shown in Equations (2) and (3), respectively [35]:

$$B = \frac{\mu_0 \mu_r N_p I_m}{MPL} \quad (2)$$

$$L_m = \frac{N_p^2 \mu_0 \mu_r A_c}{MPL} \quad (3)$$

where μ_0 is the permeability of air, μ_r is the relative permeability of the core material, N_p is the primary number of turns, I_m is the magnetising current, L_m is the magnetising inductance, and the A_c is the core cross-section area.

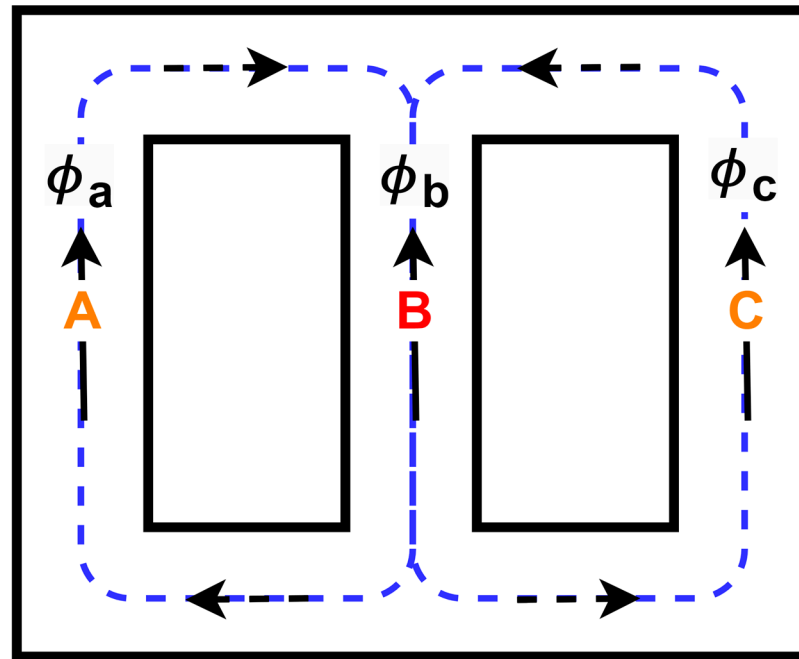


Figure 3. The 3P-EE core.

The MPL of Limbs A and C were equal and the MPL of Limb B was the shortest. Therefore, Limbs A and C were considered to have the same B and L_m . On the other hand, the value of B in Limb B was about 24.4% higher, as the core loss was directly proportional to f and B . As will be mentioned later in Steinmetz's equation, the core loss and temperature in Limb B would be higher, and so on. Likewise, the L_m at Limb B was about 28% higher.

This imbalance in L_m can cause an imbalance in the I_m drawn by each phase. The voltage across the magnetising inductor V was as follows:

$$V = I_m X_m \quad (4)$$

where X_m is the magnetising reactance, given as follows:

$$X_m = 2\pi f L_m \quad (5)$$

As seen in Equations (4) and (5), an imbalance in current and L_m leads to an imbalance in voltage.

Figure 4 shows the simulated AC phase voltages of the 3P-BIDC based on the converter specifications in Table 1. The maximum AC phase voltage in the 3P-BIDC was equal to $\frac{2V_{in}}{3}$, and according to the converter specifications, the peak phase voltage should be 200 V. However, the maximum AC voltage could not reach 200 V in each phase due to the different values of L_m among the core limbs. The voltage in phase B was greater than 200 V due to the higher L_m in Limb B of the 3P-HFT.

This issue led to an unbalanced power transfer among the converter legs, which in turn affected the performance and power quality of the 3P-BIDC. Therefore, a magnetic circuit analysis for the simultaneous balancing of the B distribution and L_m was necessary to improve the core loss distribution and temperature of the HFT. Moreover, the performance and power quality of the 3P-BIDC was improved by achieving a balanced L_m , and the rated power of the converter was transferred evenly among the converter branches.

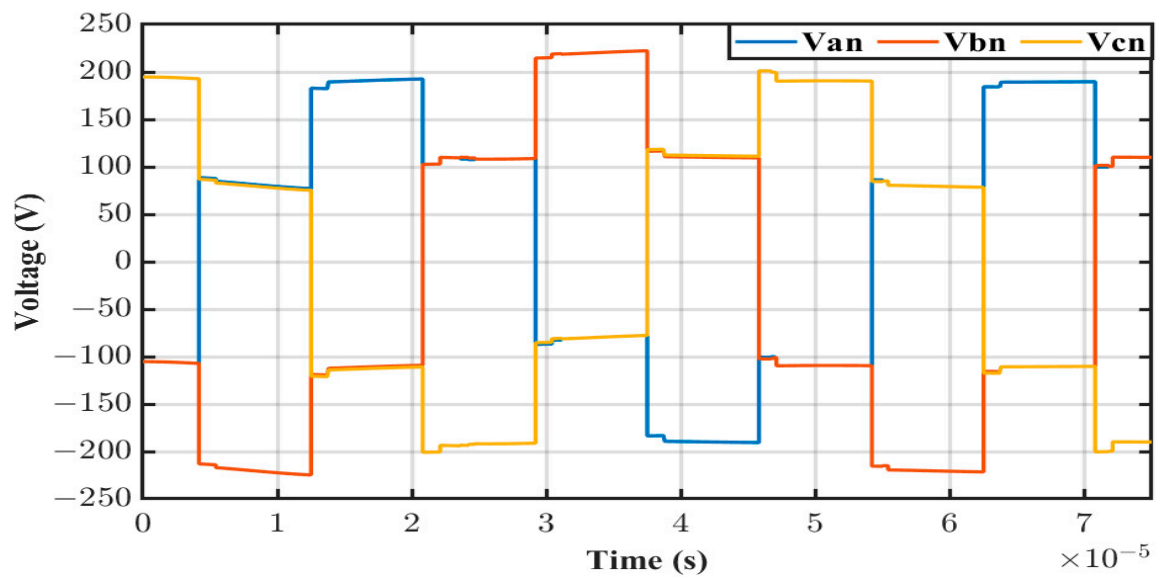


Figure 4. The simulated AC phase voltage with unbalanced L_m .

3. Design Methodology

Figure 5 shows the flowchart of the design methodology. The area product design method was used for the core and winding design as it is considered an effective HFT design method for application in power electronic converters. The initial stage of the HFT design involved choosing an appropriate magnetic material.

Soft magnetic materials like ferrite, nanocrystalline, and amorphous materials are commonly used as HFT cores, which are responsible for concentrating and shaping the magnetic flux (Φ). Each material has its own optimal cost, size, switching frequency, and performance characteristics. Ferrite cores are often used at high frequencies, particularly in power electronic applications such as transformers and inductors. In this paper, ferrite PC40 was used as the core material [27]. According to the B-H curve shown in the datasheet [27], the B-H relationship was linear when the magnetic flux density was <0.3 T at temperatures of 25°C and 60°C . Table 2 lists the properties of the PC40 ferrite, including its B , material density, and Steinmetz coefficients as determined by its curve fitting.

Table 2. The properties of the PC40 ferrite material.

Parameter	Value
Saturation flux density (B_{sat})	500 mT
Density	4800 kg/m ³
Steinmetz coefficients	
K_c	3.48
α	1.36
β	2.2

The optimal flux density (B_{opt}) was calculated by considering the design specifications and the selected material. Then, the core area product was calculated to select the core and define the core size and dimensions, such as A_c , window area (W_a), and core volume (V_c). The primary number of turns (N_p) was calculated based on the applied voltage and the A_c while the number of secondary turns (N_s) was calculated based on the transformer's turn ratio. In order to determine the B and L_m in each core limb, the superposition theorem was used for the magnetic circuit analysis when the maximum operating B of the transformer was <0.3 T. Then, the air gaps were calculated based on balancing the B and L_m , while considering that the final B should be $<B_{\text{sat}}$. Finally, the designed transformer was evaluated based on the calculated core loss and copper loss and overall rise in temperature.

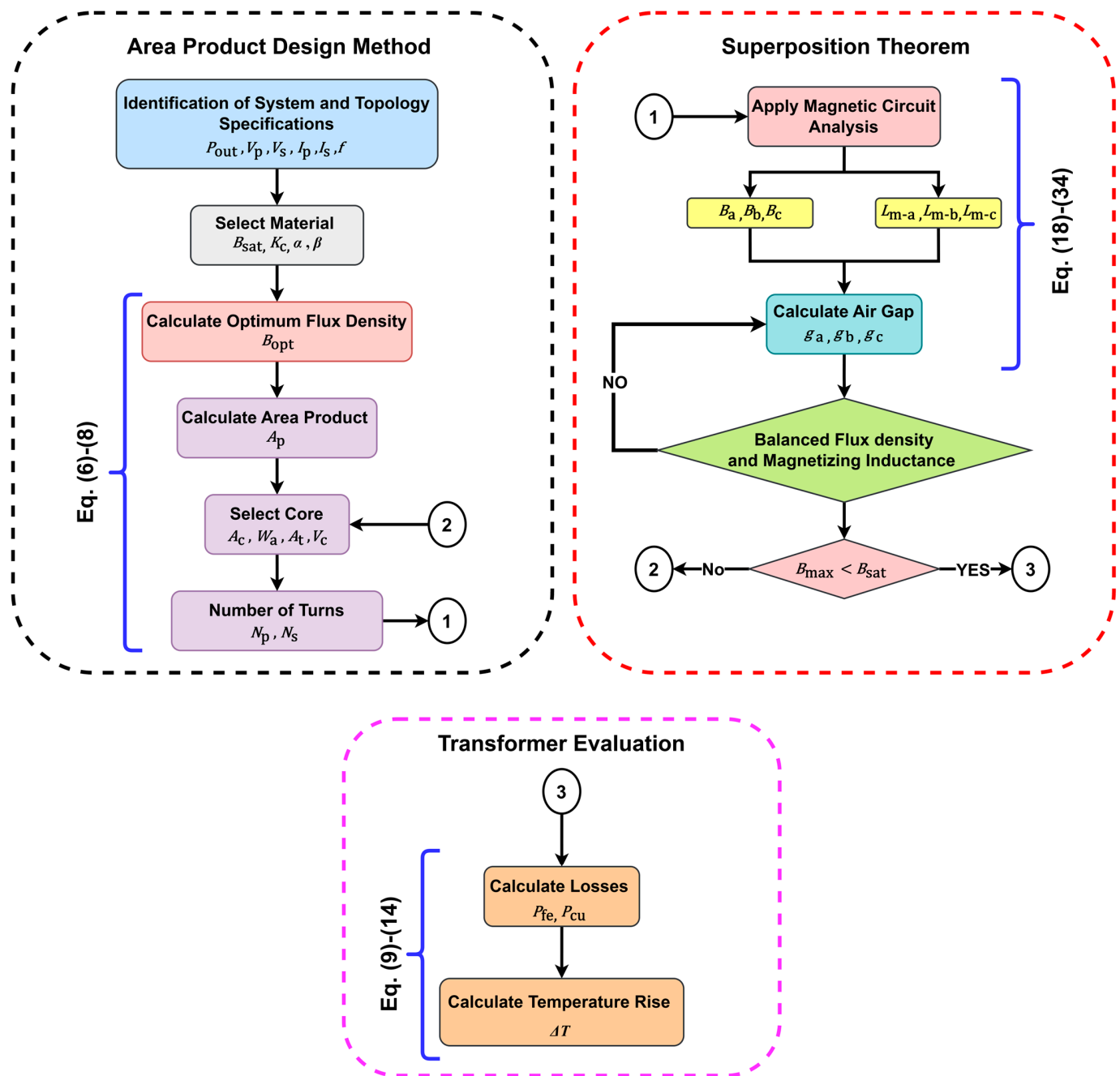


Figure 5. A flowchart of the design methodology.

3.1. Transformer Design

High efficiency is a valuable property for HFTs in power electronic converters because it means lower losses and a higher overall performance. The design method presented by [35] was adopted in this study. The optimal design criteria were based on the selection of B_{opt} at which the core and copper losses were at a minimum:

$$B_{opt} = \frac{[h_c k_a \Delta T]^{2/3}}{2^{2/3} \times [\rho_w k_w k_u]^{1/12} [k_c K_c f \alpha]^{7/12}} \times \left[\frac{K_v f k_t k_u}{\sum VA} \right]^{1/6} \quad (6)$$

where ρ_w is the wire resistivity; k_a, k_c, k_w are dimensionless coefficients acquired from dimensional research of various kinds of cores [35]; h_c is the coefficient of heat transfer;

k_u is the window utilisation factor; $\sum VA$ is the total transformer power rating; ΔT is the temperature rise; and k_f is the stacking factor pertaining to the effective A_c .

According to the area product, the core size was determined by considering the apparent power, f , and B_{opt} , as in Equation (7):

$$A_p = \left[\frac{\sqrt{2} \sum VA}{K_v f B_{opt} k_f \sqrt{\frac{h_c k_a}{\rho_w k_w}} \sqrt{k_u \Delta T}} \right]^{8/7} \quad (7)$$

To maintain a lower copper loss, a larger core with a larger A_c was chosen, resulting in a lower number of turns (N_p), as calculated in Equation (8). In addition, sufficient window area (W_a) was required for the 3P-transformer windings:

$$N_p = \frac{V_{pn}}{4f B_{opt} A_c} \quad (8)$$

The copper loss of the 3P-HFT was calculated as follows:

$$P_{cu} = 3 \left(I_{p,rms}^2 R_{ac} + I_{s,rms}^2 R_{ac} \right) \quad (9)$$

where $I_{p,rms}$, and $I_{s,rms}$ are the RMS current of the primary and secondary sides, respectively, and R_{ac} is the AC winding resistance, which is calculated by Equation (10) as reported in [38]. A Litz wire winding was used to reduce the resistance under high-frequency operation.

$$R_{ac} = R_{dc} A_{str} \left(\frac{\sinh 2A_{str} + \sin 2A_{str}}{\cosh 2A_{str} - \cos 2A_{str}} + \frac{2}{3} \left(m^2 n_{str} - 1 \right) \frac{\sinh A_{str} - \sin A_{str}}{\cosh A_{str} + \cos A_{str}} \right) \quad (10)$$

where $A_{str} = \frac{1}{\delta} \left(\frac{\pi}{4} \right)^{0.75} \sqrt{\frac{d_b^3}{d_w}}$; δ is the skin depth, $\delta = \frac{6.62 \times 10^{-2}}{\sqrt{f}}$; d_b and d_w are the bare and outer diameter of a single strand of Litz wire, respectively; m is the number of layers, n_{str} is the number of strands; and R_{dc} is the DC resistance of the winding, calculated as follows:

$$R_{dc} = (MLT) N \frac{r_{str}}{n_{str}} \quad (11)$$

where N is the number of turns; r_{str} is the resistance density a single strand of Litz wire; and MLT is the mean length turn of the winding.

For core loss, the improved generalised Steinmetz equation (IGSE) was used to estimate the core loss density, with non-sinusoidal excitation applications [39]:

$$P_{fe} = \frac{1}{T} \int_0^T k_i \left| \frac{dB(t)}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} dt \quad (12)$$

where T is the time period, ΔB is the peak-to-peak B , and k_i is the following:

$$k_i = \frac{K_c}{2^{\beta+1} \pi^{\alpha-1} \left(0.2761 + \frac{1.7061}{\alpha+1.354} \right)} \quad (13)$$

and K_c , α , and β are the Steinmetz coefficients as in Table 2. The selection of the core was based on the value of $B_{opt} = 0.174 T$, which was the B generated by each core leg, whereas the actual B was higher than that in the 3P-HFT.

Figure 6 shows the design points of the proposed 3P-HFT. The dotted red curve represents the variation of flux density with the number of windings turns at a fixed core cross-sectional area ($A_c = 3.385 \text{ cm}^2$), an operating frequency ($f = 20 \text{ kHz}$), and applied RMS voltage ($V_{pn} = 141.42 \text{ V}$). The dotted blue curve represents the change of the

total transformer loss with the number of turns. The copper loss becomes the dominant transformer loss when the number of winding turns is more than 42.

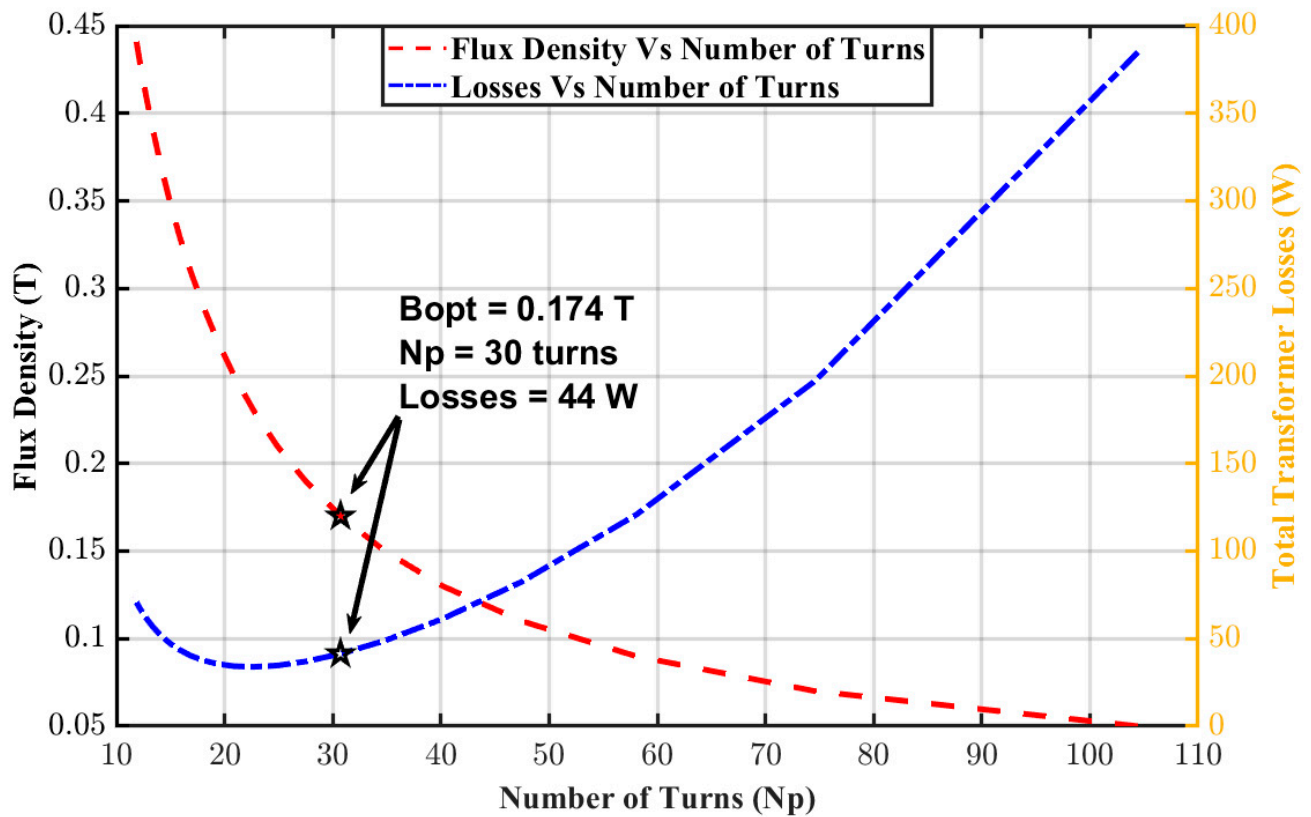


Figure 6. The design points of the 3P-HFT.

The total core loss was calculated based on Equation (12) using the peak-to-peak value of the B_{opt} , while the copper loss was estimated based on Equation (9). The estimated change in temperature caused by the total transformer loss was obtained by Equation (14), considering the surface area (A_t) of the core [40]. The estimated temperature had to be lower than the thermal dissipation capability of the component at the desired temperature, which could be estimated during the design phase.

$$\Delta T = 450 \left(\frac{P_{fe} + P_{cu}}{A_t} \right)^{0.826} \quad (14)$$

3.2. Magnetic Circuit Analysis

A magnetic circuit analysis simplifies the complex design of electrical machines and transformers by modelling the magnetic field behaviour like an electrical circuit [35]. In this manner, the magnetomotive force (MMF), similar to voltage in an electrical circuit, drives the magnetic flux Φ , which is analogous to the electrical current. The MMF, measured in ampere-turns, is influenced by the current and number of turns in the magnetic core winding, generating the flux accordingly [35].

$$\mathcal{F} = Ni = \Phi \mathfrak{R}_{eq} \quad (15)$$

and

$$\Phi = B_{opt} A_c \quad (16)$$

where $\mathfrak{R}_{eq}$ is the equivalent reluctance of each core limb, which obstructs the flow of flux in the magnetic circuit, similar to the resistance in an electrical circuit. The reluctance in a

magnetic field is directly proportional to the magnetic MPL and inversely proportional to the A_c :

$$\Re = \frac{MPL}{\mu_o \mu_r A_c} \quad (17)$$

Figure 7 shows the magnetic circuit of a 3P-transformer with three MMFs, which arises from the current flow in the primary winding of each phase. Since the 3P-core has multiple MMF sources, the superposition theorem was proposed as a magnetic circuit analysis technique, which could be used to determine the flux, flux density, and magnetizing inductance of each magnetic core limb during the design process.

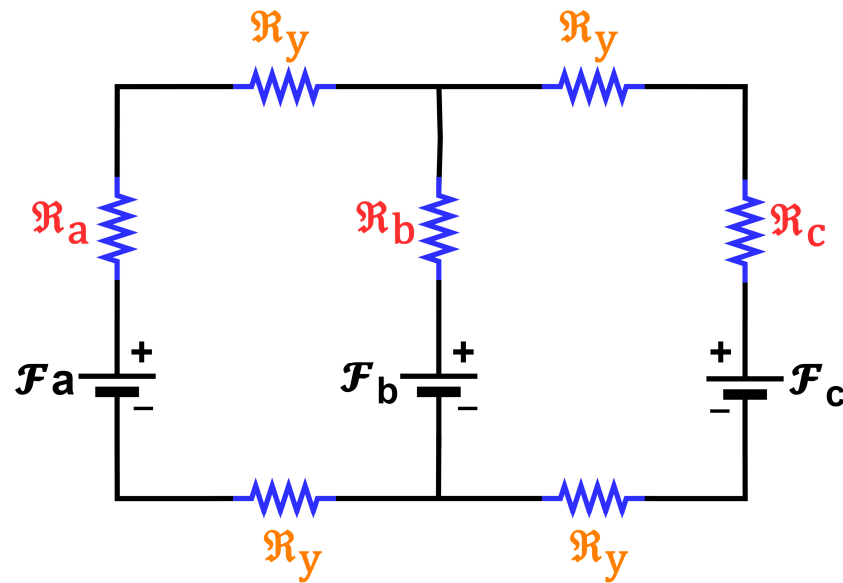


Figure 7. The magnetic circuit of the 3P-transformer.

The equivalent reluctance was calculated with one MMF being activated in the core limb, while the other two MMFs were considered to be zero. This was performed for Phases A, B, and C, as follows:

$$\Re_{eq-a} = \Re_a + \Re_y + \Re_y + \frac{(\Re_y + \Re_c + \Re_y) \times \Re_b}{(\Re_y + \Re_c + \Re_y) + \Re_b} \quad (18)$$

$$\Re_{eq-b} = \Re_b + \frac{(\Re_y + \Re_a + \Re_y) \times (\Re_y + \Re_c + \Re_y)}{(\Re_y + \Re_a + \Re_y) + (\Re_y + \Re_c + \Re_y)} \quad (19)$$

$$\Re_{eq-c} = \Re_c + \Re_y + \Re_y + \frac{(\Re_y + \Re_a + \Re_y) \times \Re_b}{(\Re_y + \Re_a + \Re_y) + \Re_b} \quad (20)$$

As shown in Figure 7, $\Re_a$, $\Re_b$, $\Re_c$ are the reluctances of limbs A, B, and C, respectively, and $\Re_y$ is the reluctance of the core yoke. Each reluctance value was calculated using Equation (17), while Equations (18)–(20) were used to calculate the phase equivalent reluctances. The L_m of each core limb was then calculated as follows:

$$L_{m-x} = \frac{N_p^2}{\Re_{eq-x}} \quad (21)$$

where x stands for A, B, or C. To determine the total flux due to the current flowing in the 3P-windings, the three cases shown in Figure 8 were considered according to the superposition theorem, as follows:

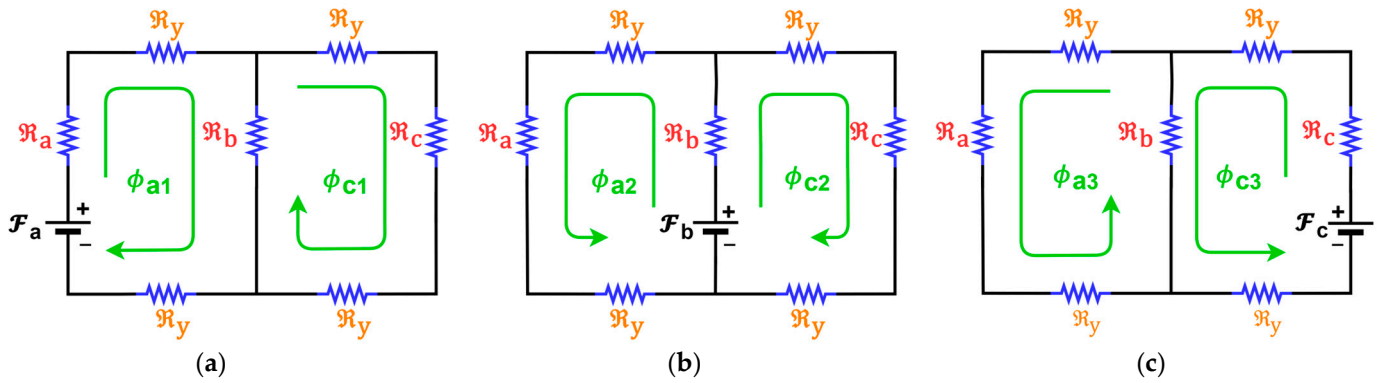


Figure 8. Three cases of superposition theorem: (a) Flux flow due to MMF in limb A, (b) Flux flow due to MMF in limb B, (c) Flux flow due to MMF in limb C.

Case 1: Considering only $\mathcal{F}_a$ (Figure 8a):

$$\phi_{a1}(\mathcal{R}_y + \mathcal{R}_b + \mathcal{R}_y + \mathcal{R}_a) - \phi_{c1}(\mathcal{R}_b) = \mathcal{F}_a \quad (22)$$

$$-\phi_{a1}(\mathcal{R}_b) + \phi_{c1}(\mathcal{R}_y + \mathcal{R}_b + \mathcal{R}_y + \mathcal{R}_c) = 0 \quad (23)$$

$$\phi_{b1} = \phi_{a1} - \phi_{c1} \quad (24)$$

Case 2: Considering only $\mathcal{F}_b$ (Figure 8b):

$$\phi_{a2}(\mathcal{R}_y + \mathcal{R}_b + \mathcal{R}_y + \mathcal{R}_a) + \phi_{c2}(\mathcal{R}_b) = \mathcal{F}_b \quad (25)$$

$$\phi_{a2}(\mathcal{R}_b) + \phi_{c2}(\mathcal{R}_y + \mathcal{R}_c + \mathcal{R}_y + \mathcal{R}_b) = \mathcal{F}_b \quad (26)$$

$$\phi_{b2} = \phi_{a2} + \phi_{c2} \quad (27)$$

Case 3: Considering only $\mathcal{F}_c$ (Figure 8c):

$$-\phi_{a3}(\mathcal{R}_b) + \phi_{c3}(\mathcal{R}_y + \mathcal{R}_c + \mathcal{R}_y + \mathcal{R}_b) = \mathcal{F}_c \quad (28)$$

$$\phi_{a3}(\mathcal{R}_y + \mathcal{R}_a + \mathcal{R}_y + \mathcal{R}_b) - \phi_{c3}(\mathcal{R}_b) = 0 \quad (29)$$

$$\phi_{b3} = \phi_{c3} - \phi_{a3} \quad (30)$$

where ϕ_a , ϕ_b , ϕ_c , are the fluxes of limbs A, B, and C, respectively, and $\mathcal{F}_a$, $\mathcal{F}_b$, $\mathcal{F}_c$ are the MMFs of each primary winding in Phases A, B, and C, respectively. The total flux value of each core limb was determined by simultaneously solving Equations (22)–(30) while B was determined by dividing the obtained flux on the A_c according to Equation (16).

3.3. Air Gap Calculation

The introduction of an air gap in the magnetic path led to a reduction in the slope of the B - H curve and a delay in the onset of the effect of magnetic saturation on the L_m [40]. The presence of an air gap led to the formation of a fringing flux at the air gap boundary. The magnitude of the fringing flux was influenced by the size and distance of the air gap relative to the winding as well as the size, shape, and position of the winding [40]. Increasing the size of the air gap caused the fringing flux to increase. Conversely, winding a coil tightly around the core and air gap resulted in a reduction in the fringing flux as it pushed the surrounding flux back into the core. The reluctance of the air gap was calculated using the following formula:

$$\mathcal{R}_g = \frac{l_g}{\mu_0 A_g} \quad (31)$$

where l_g is the air-gap length, and A_g is the air-gap cross-sectional area, which is assumed to be equal to the A_c . Firstly, the reluctance of each air gap was predefined. Then, using the

core limb reluctance along with air gap reluctance in the series, the total reluctance ($\mathfrak{R}_T$) of each core limb was determined, as shown in Figure 9.

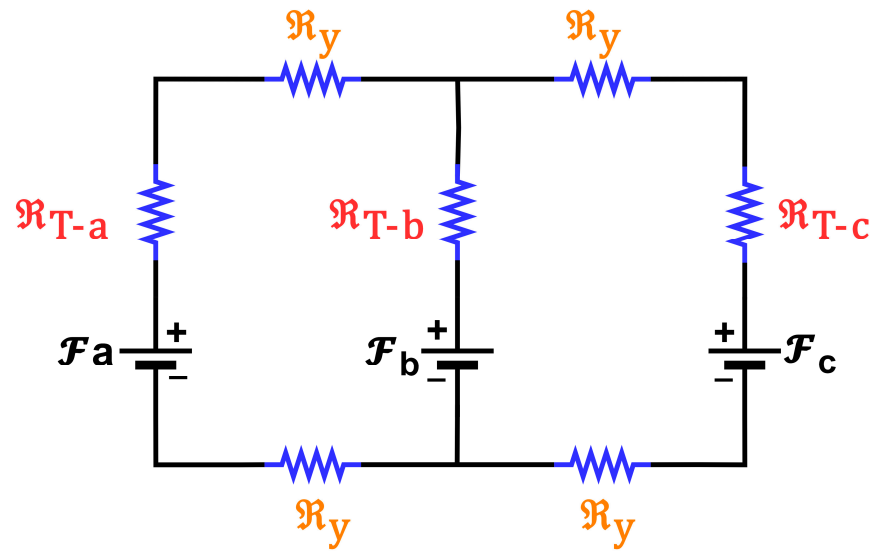


Figure 9. The magnetic circuit with air gaps.

The $\mathfrak{R}_T$ of each core limb was calculated as follows:

$$\mathfrak{R}_{T-a} = \mathfrak{R}_a + \mathfrak{R}_{ga} \quad (32)$$

$$\mathfrak{R}_{T-b} = \mathfrak{R}_b + \mathfrak{R}_{gb} \quad (33)$$

$$\mathfrak{R}_{T-c} = \mathfrak{R}_c + \mathfrak{R}_{gc} \quad (34)$$

In order to consider the effect of the air gap on the magnetic circuit of the 3P-HFT, the equivalent reluctance of each core limb was recalculated using Equations (18)–(20). Then, Equation (21) was used to determine the new L_m of each phase based on the air gap. The new MMF value was determined using Equation (15) with an air gap. Then, the flux of each core limb was calculated by following three cases of the superposition theorem, considering $\mathfrak{R}_T$ of each core limb instead of $\mathfrak{R}_a$, $\mathfrak{R}_b$ and $\mathfrak{R}_c$. The new B was then determined using Equation (16). Once the flux density is below B_{sat} and L_m were balanced, the air gap length (l_g) was fixed.

4. Results and Discussion

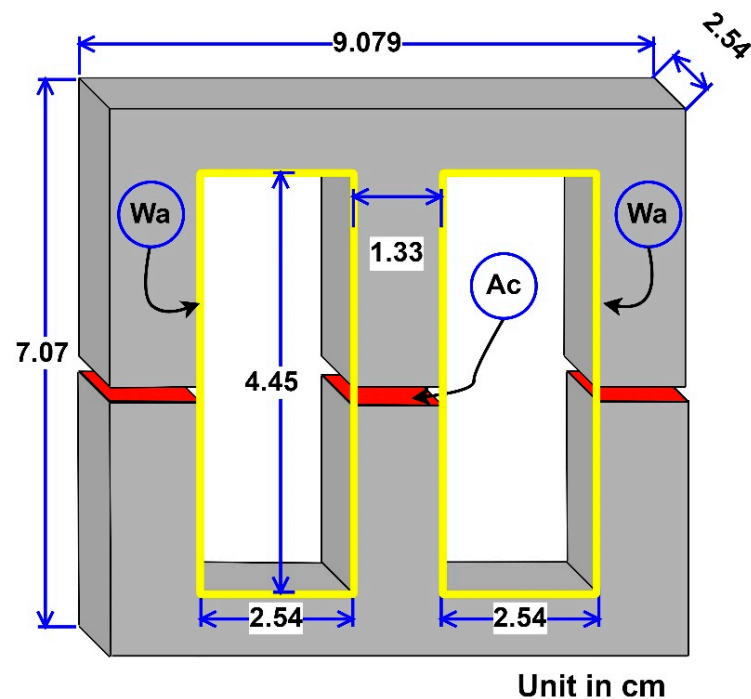
4.1. Theoretical Results

Table 3 shows the details of the proposed 3P-HFT design. The core size and dimensions were selected from [37] as in Figure 10. The selection of the core was based on the B_{opt} and the area product method, which indicated how the selected core would be able to handle the rated power with an acceptable rise in temperature.

Using Equation (8), the number of primary turns (N_p) was calculated to be 30 based on the obtained B_{opt} and the A_c , while the secondary number of turns (N_s) resulted from the transformer turns ratio. A larger core W_a was required to accommodate 30 turns with the selected Litz wire. Therefore, a two-layer winding structure was implemented to maintain a high power density while maintaining a compact volume (V_c). This configuration also increased the leakage inductance of the transformer, thereby eliminating the need for external inductors to regulate the power flow of the 3P-BIDC.

Table 3. The design parameters of the proposed HFT.

Parameter	Value
Optimum flux density (B_{opt})	0.174 T
Core Part Number	ME1570-05 [37]
Core cross section area (A_c)	3.385 cm ²
Core window area (W_a)	11.303 cm ²
Core surface area (A_t)	444 cm ²
Core Volume (V_c)	106.6 cm ³
Winding Parameters	
Current Density (J_o)	6 A/mm ²
Primary number of turns (N_p)	30
Secondary number of turns (N_s)	30
Wire Diameter	2 mm ² (insulated)
Strand Diameter	0.1 mm
Number of Strands	170 strands

**Figure 10.** The dimensions of the transformer's core.

Within the 3P-EE core, one window was shared by two phases. The W_a of the chosen core was 11.303 cm², and the areas occupied by the primary and secondary windings in each window were $0.0314 \times 30 = 0.942$ cm² and $0.0314 \times 30 = 0.942$ cm², respectively. Therefore, the $K_u = (2 \times 0.942 + 2 \times 0.942)/11.303 = 0.34$. As demonstrated in the section on the comparison of the results, a lower K_u improved the thermal performance of the 3P-transformer.

By applying the superposition theorem on the selected core and windings, the B and L_m in each core limb were determined, and the air gap was selected based on the balance between B and L_m . Table 4 compares the B and L_m of the transformer with and without air gaps. Given the shorter MPL at Limb B, a higher B of 0.417 T was recorded, resulting in a 2 mH disparity in L_m between Limb B and Limbs A and C, as listed in Table 4.

Table 4. The theoretical results of the proposed 3P-HFT.

Parameter	Without Air Gap			With Air Gap		
	Limb A	Limb B	Limb C	Limb A	Limb B	Limb C
Air Gap (μm)	-	-	-	6	41	6
MPL (cm)	17.58	12.55	17.58	22.44	22.67	22.44
Flux density (T)	0.313	0.417	0.313	0.348	0.348	0.348
Magnetizing inductance (mH)	5	7	5	3.9	3.9	3.9
Core Loss (W)		8.091			5.43	
AC Resistance ($\text{m}\Omega$)		108			108	
Copper Loss (W)		37.125			37.125	
Leakage Inductance (μH)		10.65			10.65	
Temperature ($^{\circ}\text{C}$)		68.19			64.86	

This discrepancy led to notable differences in the peak phase voltages of the three phases of the 3P-BIDC, as shown in Figure 4. However, the selected air gap values effectively balanced the B and L_m , reducing the overall B by 0.069 T in Limb B. The core loss without the air gap was estimated based on the highest B found at Limb B at 0.417 T, while the core loss with the air gap was reduced by approximately 2.66 W, which was associated with a reduction in B .

The AC winding resistance was calculated using the Litz wire details as in Table 3. The leakage inductance was also calculated and shown in Table 4. Finally, the temperature of the 3P-HFT was taken into account, where the temperature was calculated in relation to the copper and core losses. Due to the chosen air gaps, which helped to reduce the B and core loss, the overall HFT temperature was expected to be reduced by approximately 3.33 $^{\circ}\text{C}$ and the overall temperature would be the same at each core limb.

4.2. Simulation Results

The proposed 3P-HFT design was validated by FEA using ANSYS Maxwell 2022 R1 before the experiments were conducted. The dimensions in Figure 10 and the winding details in Table 3 were used to model the 3P-HFT. Two 3P-HFTs, one with and one without air gaps, were included in the simulation to perform the results of the proposed analysis method and demonstrate the impact of the designed air gaps on the L_m , B , core loss and temperature.

Table 5 contains a list of the simulation parameters. The measured L_m of both the 3P-HFT models was higher than the expected value due to the simulation accuracy. However, the air gaps provided reduced both the inductance value and the variation of the L_m in each core limb.

Table 5. The simulation results of the proposed 3P-HFT.

Parameter	Without Air Gap			With Air Gap		
	Limb A	Limb B	Limb C	Limb A	Limb B	Limb C
Magnetizing Inductance (mH)	5.7	8.1	5.7	4.37	4.27	4.37
Core Loss (W)		8.4			5.12	
AC Resistance ($\text{m}\Omega$)		109			109	
Copper Loss (W)		37.125			37.125	
Leakage Inductance (μH)		10.86			10.86	

Figure 11 displays the distribution of B inside the 3P-HFT core. The strategically designed air gaps resulted in a balanced distribution of magnetic flux density B among the core limbs (Figure 11b). The B was slightly lower than the expected value. In contrast to the model without air gaps (Figure 11a), the highest B was observed in Limb B due to its shorter MPL.

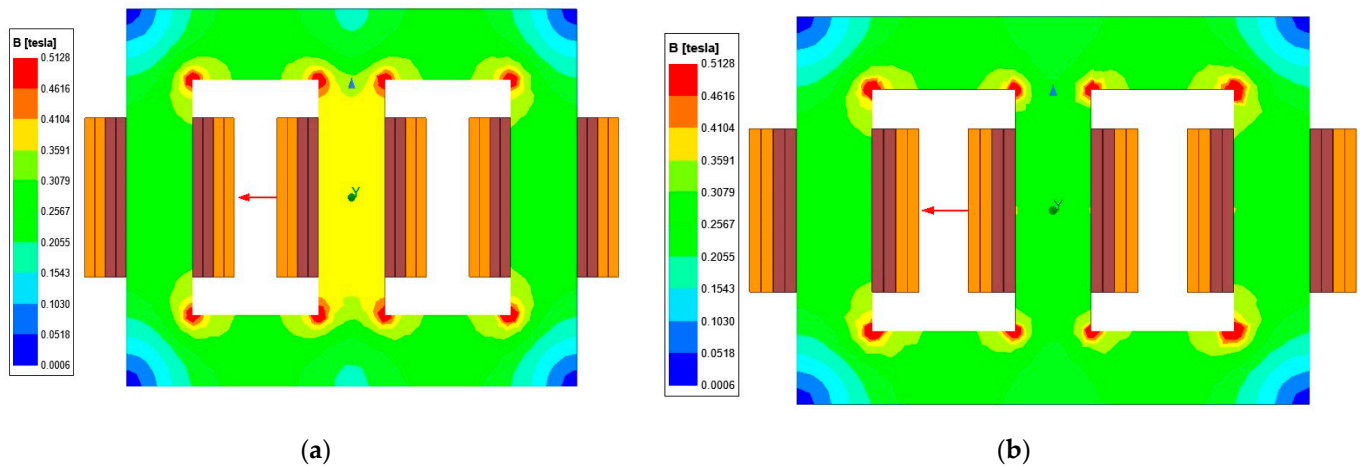


Figure 11. The flux density distribution in the simulation: (a) without air gap, (b) with air gap.

The distribution of the volumetric core loss, as determined by the ANSYS Maxwell, is shown in Figure 12. As depicted in Figure 12a, Limb B had a higher core loss density than Limbs A and C. On the other hand, in Figure 12b, each core limb had the same core loss, which was due to the uniform distribution of B achieved according to the designed air gaps.

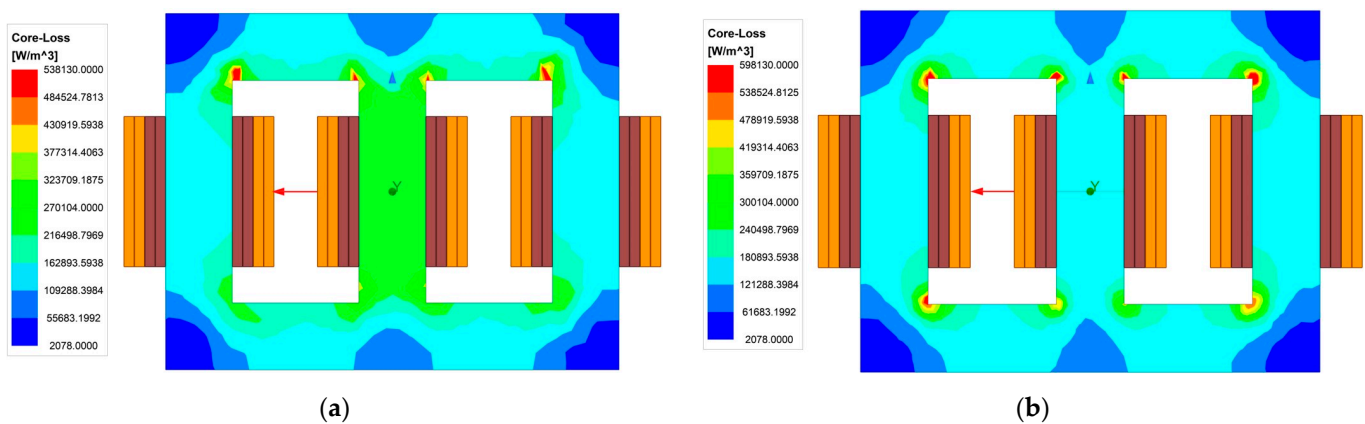


Figure 12. The distribution of the volumetric core loss density: (a) without air gap, (b) with air gap.

The transformer temperature is shown in Figure 13 through the use of an ANSYS Maxwell and Icepack coupling simulation. The temperature rise obtained during full load in the FEA was caused by the copper and core losses. As seen in Figure 13a, the highest temperature was 67.99 °C in Limb B, which was due to the higher core loss of the 3P-HFT without air gaps. However, as can be seen in the model with air gaps in Figure 13b, the overall temperature was reduced due to the lower core loss achieved. Furthermore, the temperature difference between the core limbs was about 2 °C in contrast to 1 °C achieved in the 3P-HFT without air gaps.

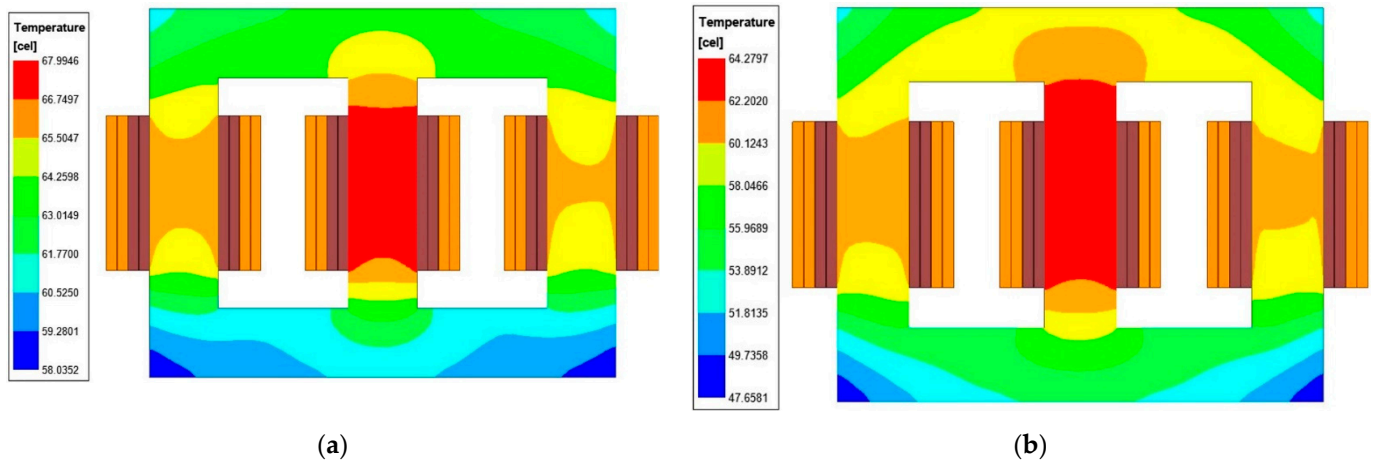


Figure 13. The temperature of the transformer in the simulation: (a) without air gap, (b) with air gap.

4.3. Experimental Results

As mentioned in Section 1, ferrite EE with a 3P structure is not currently available on the market. Thus, the proposed 3P-EE core was fabricated as a laboratory prototype according to the core dimensions and characteristics shown in Figure 10. All three limbs of the 3P-core had an equal A_c , considering the air gaps calculated by the proposed analysis method, as listed in Table 4. Figure 14 shows the 3P-HFT prototype featuring the same Litz wire parameters as in Table 3.

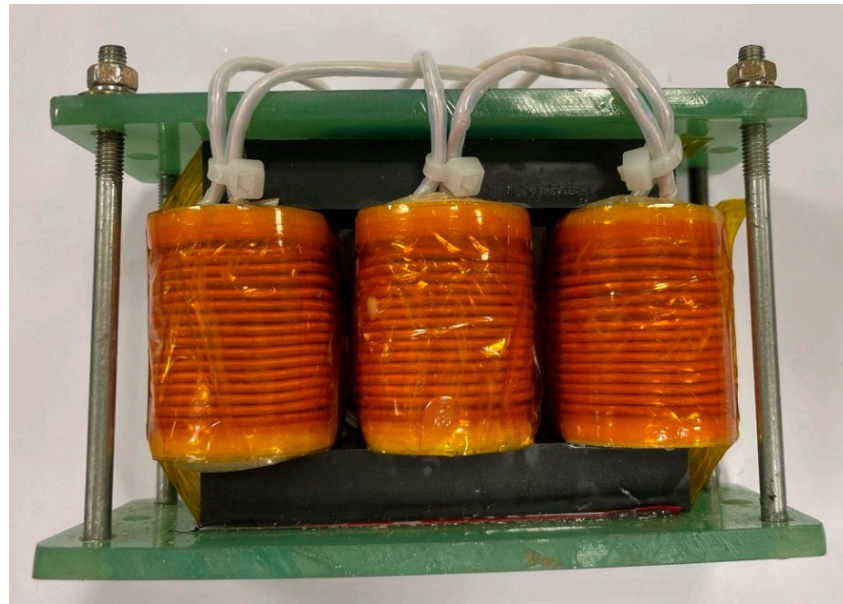


Figure 14. The 3P-HFT prototype.

An open circuit measurement setup was carried out by connecting the 3P-HFT to the primary bridge of a 3P-BIDC with a left secondary-side open circuit supplied by a DC input voltage of 300 V and an operating f of 20 kHz at the primary bridge. This setup was used to measure the magnetising B of each core limb individually. A 200-MHz WaveSurfer 3024z oscilloscope was employed to measure the induced phase voltage of the secondary side. The built-in integration function of the oscilloscope was used to obtain the flux in each limb, after which B was obtained by applying Equation (35).

$$B(t) = \frac{1}{N_s A_c} \int v_s(t) dt \quad (35)$$

Figure 15 shows the B of each core limb plotted by using MATLAB R2023b, according to the flux data collected from the oscilloscope and Equation (35).

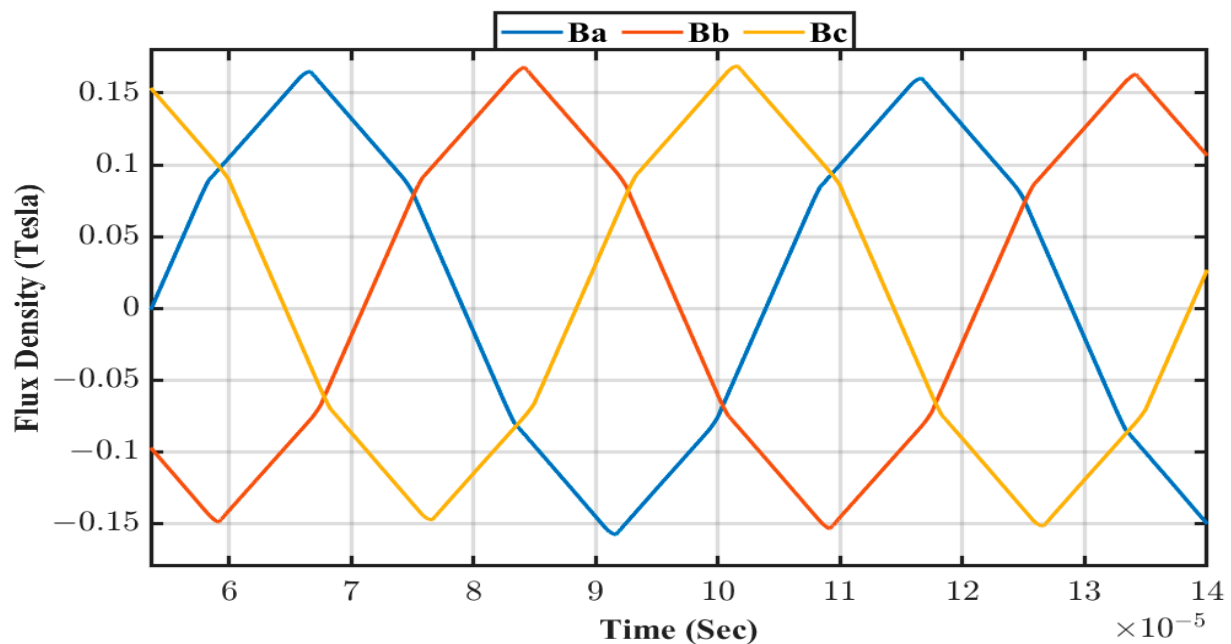


Figure 15. The flux density of each core limb.

This B was considered as the B generated by each core limb. In the 3P-core, the generated B was shared among the core limbs based on the MPL of each core limb. Therefore, the total B in each limb was calculated by using Equations (36)–(38), in accordance with the superposition theorem, as presented in Table 6.

$$B_{a-(\text{final})} = B_a + 0.7B_b + 0.3B_c \quad (36)$$

$$B_{b-(\text{final})} = B_b + 0.5B_a + 0.5B_c \quad (37)$$

$$B_{c-(\text{final})} = B_c + 0.7B_b + 0.3B_a \quad (38)$$

Table 6. The total flux density of each core limb.

Core Limb	Flux Density	
	Generated Value (T)	Total Value (T)
Limb A	0.164	0.332
Limb B	0.168	0.333
Limb C	0.168	0.334

The obtained flux density in each core limb was lower than the expected value calculated by the proposed analysis method. This was due to the variation between the designed and practical air gap. The practical air gap could not be exactly the same as the designed value as the air gap was in micrometres. As shown in Table 6, the final B was reduced and a uniform distribution was achieved.

In terms of the core loss measurement, the secondary phase voltage and primary no-load current along with the power factor angle between them were measured for each phase individually, and the RMS values of the voltage and current were obtained using an oscilloscope, as presented in Table 7.

Table 7. The results of the open circuit measurements.

Phases	RMS Voltage (V)	RMS Current (A)	Power Factor Angle (°)	Loss (W)
Phase A	140.3	0.0998	83.27°	1.64
Phase B	140.1	0.0965	82.36°	1.797
Phase C	140.1	0.0931	82.96°	1.598
Total Core Loss		5.035 W		

The no-load losses that occurred at each core limb were close to each other with a difference of around 6%. On the other hand, the total losses obtained were lower than the theoretical value because the actual B was lower than the theoretical value.

A GW Instek-LCR 8101G meter with a switching frequency measurement range of 20 Hz to 1 MHz was used to measure the winding resistance, magnetising, and leakage inductance of the proposed 3P-HFT, as presented in Table 8. To measure the L_m of the primary side, the secondary winding was an open circuit, whereas the leakage inductance and windings resistance were measured by a short-circuit test on the secondary side. The L_m was almost balanced, achieving a balanced AC phase voltage on the primary and secondary windings of the 3P-HFT to ensure that the power was transferred equally via the 3P-BIDC legs. The leakage inductance was based on the winding geometry, and it could be used as the power transfer element of the 3P-BIDC, instead of adding an auxiliary inductor, for high power density. The measured AC resistance matched well with the calculated value. The measured RMS current was lower than the theoretical value at $\varphi = 3.9^\circ$. Therefore, based on the same copper resistance, the copper loss was lower than the theoretical value.

Table 8. The results of the experiment.

Parameters	Limb A	Limb B	Limb C
Magnetizing Inductance (mH)	4.17	4.23	4.17
Leakage Inductance (μ H)	11.37	11.36	11.23
AC Winding Resistance (m Ω)	108	108	108
Copper Loss (W)		29.08	

A 3P-BIDC was constructed and tested to evaluate its proper operation with the proposed 3P-HFT. The 3P-BIDC prototype used six Infineon IGBT dual modules with a voltage rating of 1.2 kV and current rating of 50 A, 12 snubber capacitors, and six film capacitors. Figure 16 shows the complete prototype of the converter with ratings of 20 kHz, 3 kW, and 300 V. To operate the 3P-BIDC under full-load conditions, the power circuit was arranged so that the output is connected in parallel to the input to test the converter at full power with minimum energy consumption [41,42]. The input to the DC-DC converter was supplied from a DC power supply. The DC-DC converter was operated in power regulation mode using the DC voltage and current transducers, as shown in Figure 17. This configuration was feasible only when the transformer turns ratio was 1:1. Consequently, a DC power supply was required to supply the 3P-BIDC losses.

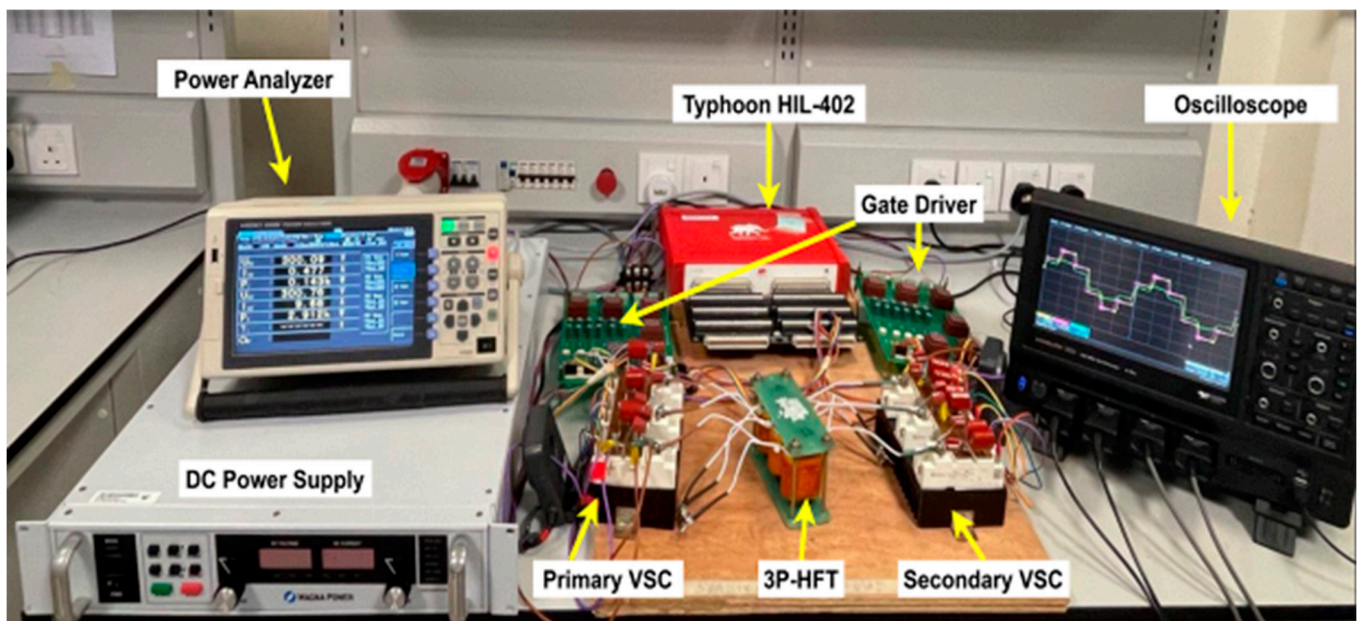


Figure 16. The setup of the 3P-BIDC prototype.

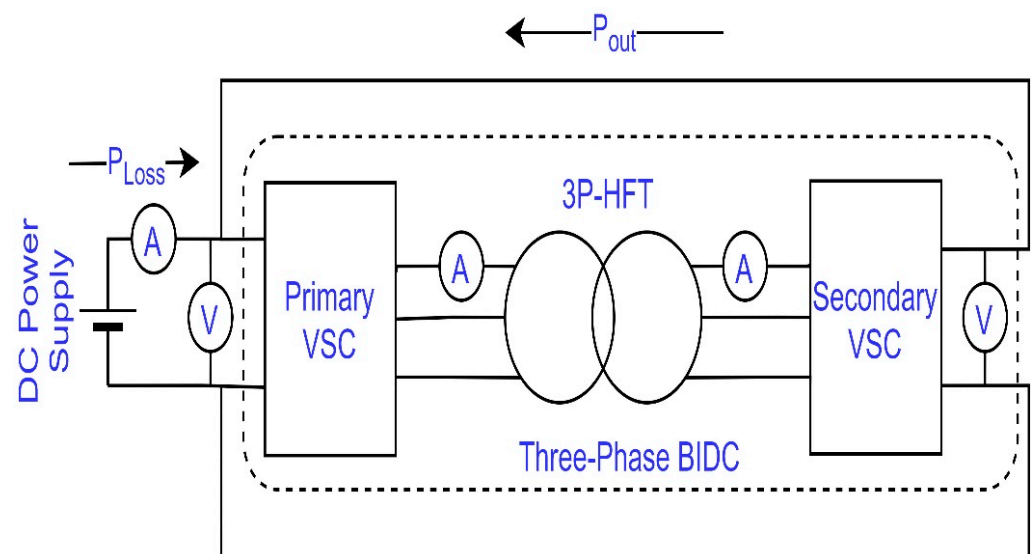


Figure 17. The experimental setup.

Figure 18 depicts the experimental waveform of the AC terminal phase voltages across the primary and secondary bridges, where the rated power was achieved at $\varphi = 3.9^\circ$. The waveform revealed that the primary voltage led the secondary voltage by $\varphi = 3.9^\circ$, indicating that power flowed from the primary bridge to the secondary bridge.

Furthermore, the measured 3P voltages of the primary windings are shown in Figure 19. The peak phase voltage was 200 V, which was equal to $\frac{2V_m}{3}$. This also indicated the balance of L_m inside the 3P-HFT because it had a negative impact on the phase voltage of the 3P-BIDC, as shown in Figure 4.

Figure 20 shows the symmetrical 3P current of the primary side operating at full load. The peak value was 10 A, where this value was achieved according to the leakage inductance inside the 3P-HFT without the use of an external inductor. Moreover, each leg of the 3P-BIDC transferred power equally, according to the phase voltage in Figure 19 and phase current in Figure 20.

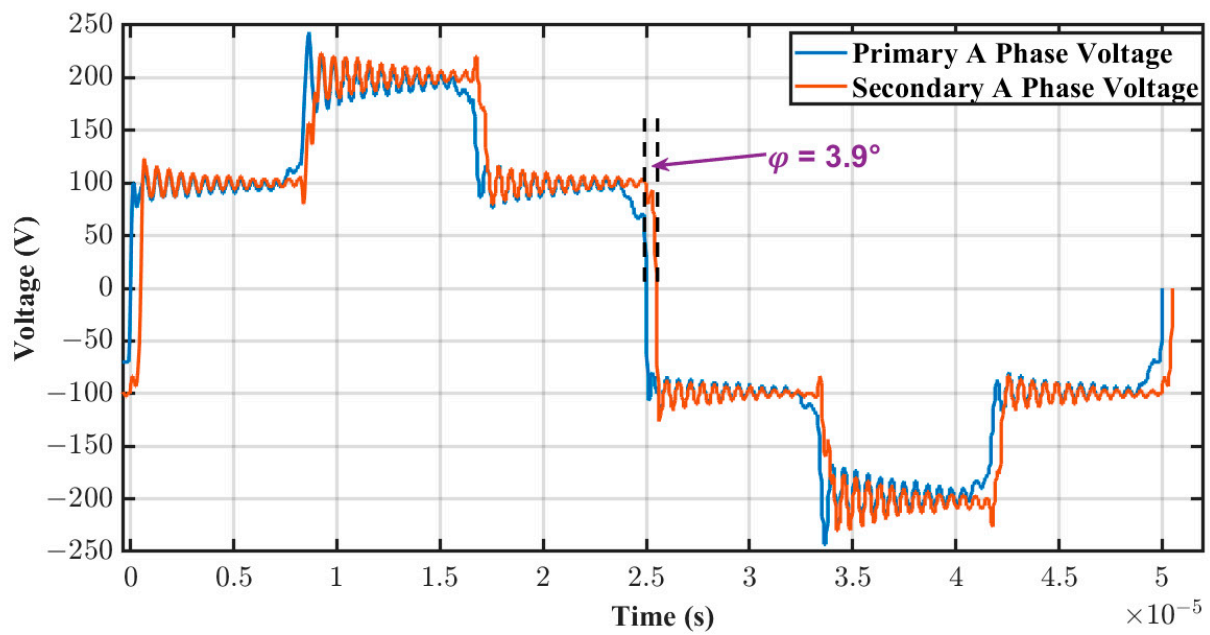


Figure 18. The measured AC voltage of the primary and secondary Phase A.

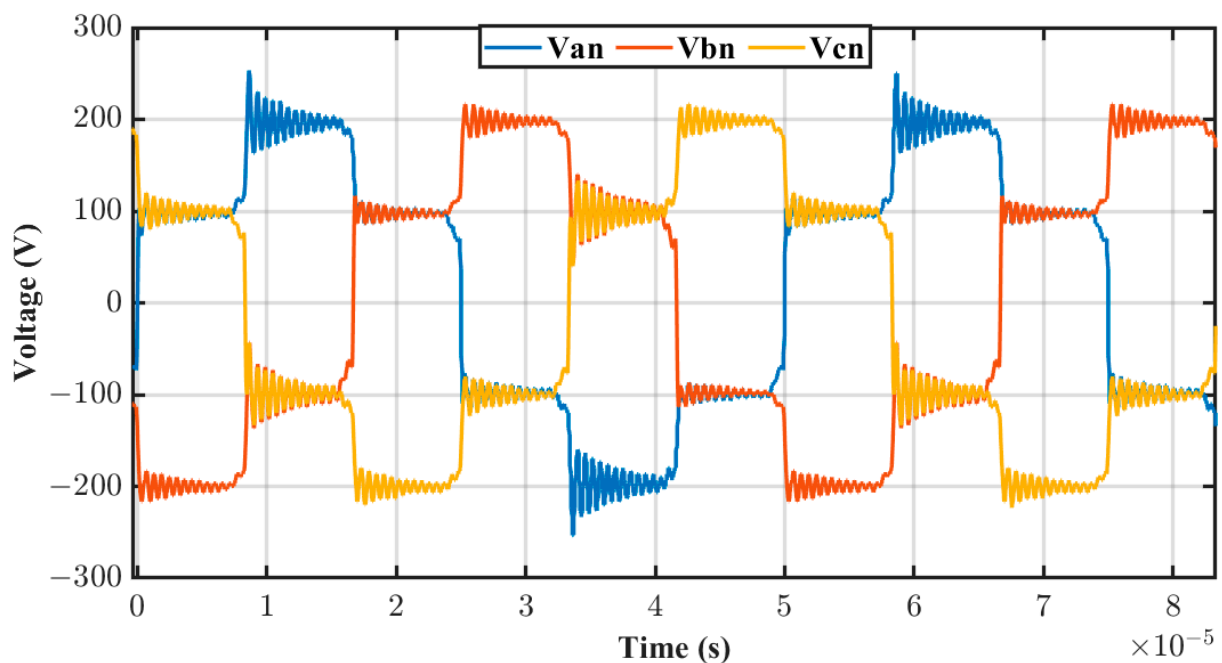


Figure 19. The 3P voltages of the primary winding.

According to the full-load setup in Figure 17, the V_{in} and V_{out} along with the input and output currents were measured, as shown in Figure 21. The V_{in} and V_{out} were both 300 V, while the output current was achieved about 8.79 A, and input current, which was used to supply the converter losses, including the 3P-HFT and power semiconductor switches losses, was 0.72 A. Therefore, the total power from the DC power supply was about 216 W, which was considered to be the total converter losses, and according to the measured core and copper losses of the 3P-HFT, the total 3P-HFT loss was about 34.115 W, which indicated that the transformer efficiency was 98.67%, and the converter efficiency was measured to be about 92.43%.

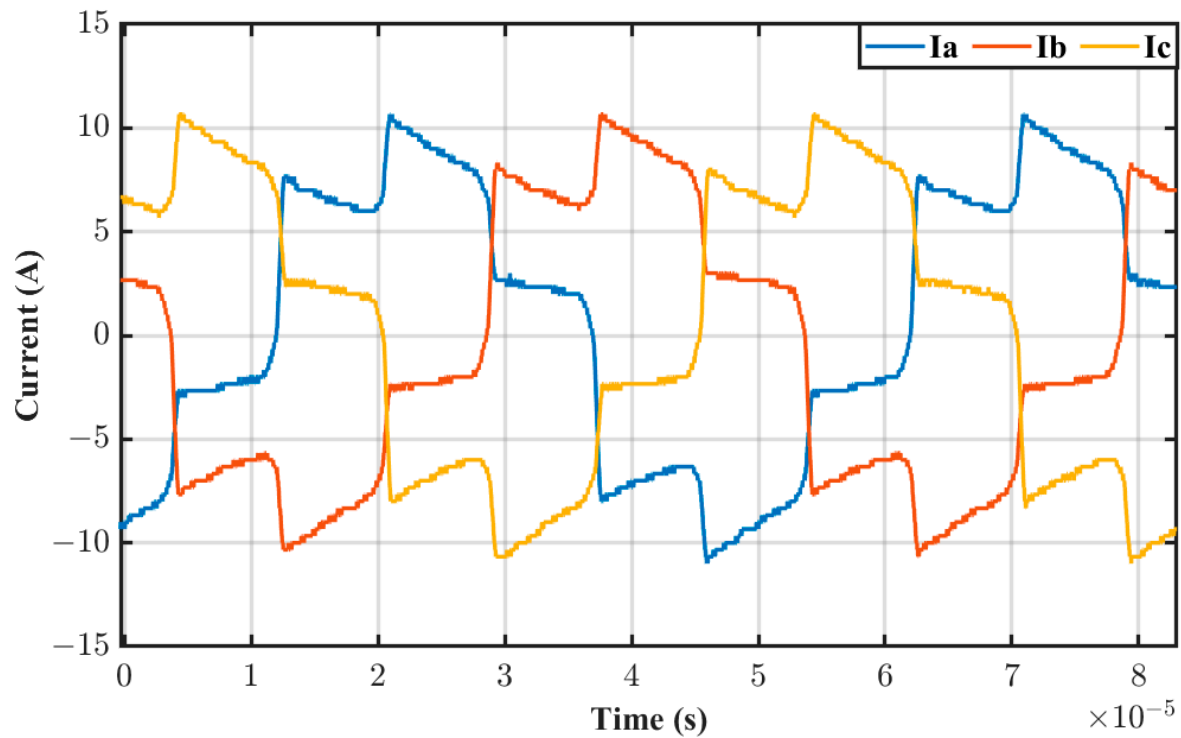


Figure 20. The 3P current of the primary winding.

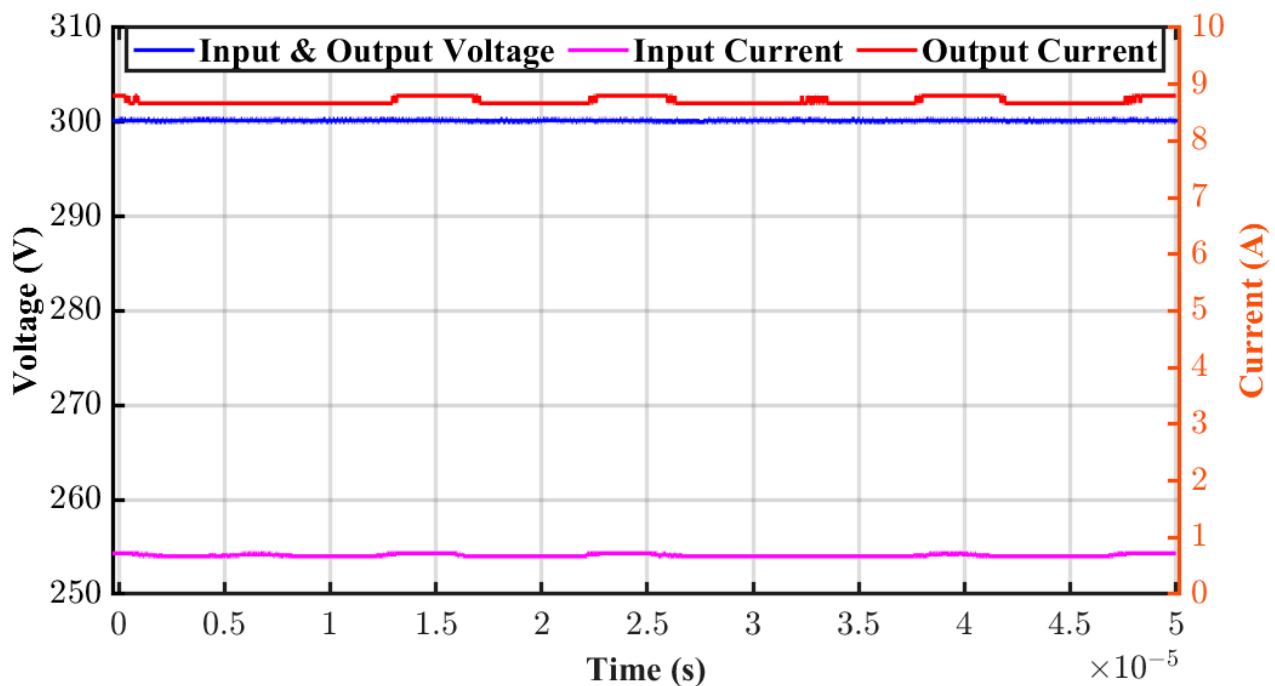


Figure 21. The measured DC voltage with input and output current.

The converter was operated under a full load for 20 min and a FLUKE Ti401 PRO Thermal Imager was used to measure the temperature of the proposed 3P-HFT. Figure 22 shows that the highest recorded temperature of the 3P-HFT was approximately 64.6 °C at a steady state point. The temperature distribution among the core limbs of the transformer was seen to be even because of the distribution of B inside the transformer limb, which resulted in similar core losses within the core limbs. Compared with the calculated temper-

ature without air gaps, the total 3P-HFT temperature was reduced by 3.33 °C due to the reduction in the overall B and core loss brought about by the designed air gaps.



Figure 22. The temperature results of the prototype 3P-HFT.

4.4. Results Comparison and Validation

This section reports on the merits of employing a 3P-EE ferrite core for a 3P-HFT from a practical point of view. The merits such as core cross sectional area, volume, mass, power density, heat dissipation and cost are included in the comparison, where the proposed 3P-EE core has been compared with a 1P-EI core used for a 3P-transformer, as reported in [28], a 1P-EC core as in [29], and a three-toroidal core, which is considered the most popular structure for 3P-BIDCs [23].

Table 9 shows the cores that were selected for a 300-V, 20-kHz, 3-kW 3P-HFT from the datasheet found in [43,44]. A high A_c of the EI and EC cores was selected to prevent a higher B at Limbs A and C because they had a smaller A_c than Limb B, whereas for the proposed 3P-EE core all the core limbs had the same A_c . A lower V_c was achieved with the EI core, and this value was lower than that of the proposed core by 5.26%. On the other hand, the proposed core achieved a 19.9% reduction in volume compared to the three discrete cores, which was enough to achieve a higher power density.

The rise in temperature of the transformer depends on the external thermal resistance (R_{ET}) (°C/W) of the core, which can be estimated as a function of W_a using a simple formula based on the rule of thumb [45] during an average situation of natural convection cooling, as follows:

$$R_{ET} = \frac{36}{W_a[\text{cm}^2]} \text{ } ^\circ\text{C/W} \quad (39)$$

It was deduced that the 3P toroidal core gave a better thermal performance than the other cores; however, each core had its own temperature, which may have led to an increase in the overall temperature, and this core achieved a lower power density as it was the heaviest and most expensive. The proposed EE core was considered as the best solution for the 3P-core because it achieved a high power density, low weight, and low cost compared to other cores.

Table 9. A comparison of 3P-HFT cores.

Parameter	Proposed Core	EI70/64/31N [43]	EC90/90/30 [43]	T72/20/48 [44]
Core cross section area A_c (cm ²)	3.385	6.95	6.26	3×2.37
Core window area W_a (cm ²)	11.3	10.71	14.2	3×66.35
Volume (cm ³)	106.6	101.2	138.27	3×43.42
Mass (g)	511.9	519	635	3×217
Power Density (W/cm ³)	28.126	29.96	21.67	23.03
R_{ET} (°C/W)	3.184	3.36	2.53	3×0.5425
Cost (USD)	50	130 [46]	98 [46]	3×200 [46]

As stated previously, the core loss of the proposed core was reduced by the air gaps calculated by the proposed analysis method, which lowered the overall temperature. On the other hand, a large air gap was required for the EI and the EC core in order to achieve a balanced B inside the core, which led to a decrease in the L_m and affected the power transfer of the 3P-BIDC.

Table 10 presents the validation of the theoretical, simulation and experimental results of the proposed 3P-HFT with air gaps. The B predicted by the proposed method was close to the experimental measurements with some variations due to the micrometre-sized air gaps and the difficulty in accurately achieving the designed air gaps. As a result, the difference in L_m of Limb B was at its maximum at approximately 8.4%. The results of the proposed method, which predicted the B and L_m , were sufficiently accurate for the 3P-HFT design, and it dealt with the core loss and temperature reduction. Therefore, the proposed 3P-HFT outperformed the previous research by designing a new 3P-EE core based on the requirement of a 3P-BIDC with high power density, lower losses and temperature, and low cost, while the previous research structure produced HFTs with a lower power density, requiring a high volume and high material cost.

Table 10. Results validation.

Parameter	Theoretical	Simulation	Experimental		
			Limb A	Limb B	Limb C
Flux density (T)	0.348	0.31	0.332	0.333	0.334
Magnetizing Inductance (mH)	3.9	4.37	4.17	4.23	4.17
Leakage Inductance (μH)	10.85	10.86	11.37	11.36	11.23
Core Loss (W)	5.43	5.12	1.64	1.797	1.598
Copper Loss (W)	37.125	37.125	9.7	9.7	9.7
Temperature Rise (°C)	64.86	64.27	64.2	64.5	64.6

5. Conclusions

In this study, a 300-V, 3-kW, 20-kHz 3P-HFT was designed, constructed, and examined for a 3P-BIDC based on a new EE ferrite core structure. The area product method was used to design the 3P-HFT. In addition, the authors proposed a superposition theorem for the magnetic circuit analysis to predict the B and L_m and to determine the air gap that balanced the B and L_m so that the core loss and temperature in the core limbs could be distributed equally. The simulation and experimental results verified the theoretical design. The experimental results showed that the B of the transformer was uniformly distributed and was below the B_{sat} of the ferrite material. The three L_m values were also equal. As a result, the core loss was reduced and equally distributed inside the transformer core, which also helped to decrease the overall maximum 3P-HFT temperature. The proposed 3P-HFT achieved a high power density at a low cost and exhibited an efficiency of 98.67% at the rated power. A high power density 3P-BIDC was achieved using one 3P-HFT instead of three 1P-cores. On the other hand, the window area of the 3P-core was smaller than that of the 1P-core, thereby making the 1P-core better in terms of heat dissipation.

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