

Ultra-High-Power Density of LLC Transformer Based on Unbalanced-flux Magnetics Approach

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Abstract— The MHz design approach for high-frequency magnetics has been widely adopted to increase power density in power converters. However, this approach presents several limitations, including reduced operating magnetic flux density B , increased electromagnetic interference (EMI) due to high switching frequency, and a strong dependence on soft-switching operation, which limits its applicability in hard-switching converters. The recently introduced unbalanced-flux magnetics (UBFM) concept offers a promising alternative to overcome these limitations. A first experimental implementation demonstrated a power density three times higher than that achieved with conventional MHz-based magnetic designs. Building on this concept, this paper presents a second experimental demonstration of UBFM applied to a 1-kW, 400/24-V LLC converter. The realized power density of the LLC transformer reaches 230 W/cm³, while the converter achieves a maximum efficiency of 96.07% at 500 W, and efficiencies of 95.3% and 94.94% at rated and peak power (1 kW and 1.1 kW), respectively. These results demonstrate the potential of UBFM to significantly increase magnetic power density while maintaining high efficiency.

Index Terms—unbalanced-flux magnetics, LLC transformer, power density.

I. INTRODUCTION

In recent years, significant efforts have been devoted to increasing the power density and efficiency of DC–DC power converters. The emergence of GaN semiconductor devices together with the soft-switching capability of resonant converter topologies has been a key enabler for these improvements. These technologies allow converters to operate at switching frequency extending from hundreds of kHz to the MHz range, which has long been considered a necessary condition for reducing the size of magnetic components, the main obstacle to achieving high power density in switching power converters [1]. The MHz design approach relies on two complementary aspects. The first is related to the converter topology, particularly resonant converters, which enable soft-switching operation and therefore reduce switching losses that would otherwise severely degrade converter efficiency and thermal performance. The second aspect is technological and associated with the development of wide-bandgap semiconductor devices, which support higher switching frequencies and faster switching transitions. Despite these advances, the improvements in magnetic power density have remained limited compared to the expected potential of MHz

operation. Even with practical techniques such as flux cancellation and fractional turns, the reported power density of high-frequency magnetics typically remains around 55–70 W/cm³ in single-phase 1-kW 400-V LLC converters [2]–[8]. A detailed review of these works is presented in [10].

The main limitation of the MHz approach arises from the strong dependence of magnetic losses on both frequency and flux density. As switching frequency increases to the MHz range, the magnetic flux density B must be significantly reduced to limit core losses and thermal stress. However, reducing B directly decreases the power capability of magnetic components, since magnetic power is proportional to B^2 . Consequently, the expected gains in power density from higher switching frequencies are fully offset by the required reduction in flux density.

Our previous studies [9–10] have shown that improving power density is fundamentally linked to minimizing magnetic losses, not only to improve efficiency but primarily to allow operation at higher magnetic flux density. In most existing high-frequency designs, magnetic materials are significantly underutilized, with operating flux densities typically limited to 15–30% of the allowable maximum. Increasing B would therefore provide a direct path to improving magnetic power capability. However, this objective conflicts with the fact that magnetic losses also increase approximately with the square of B , making magnetic power and magnetic loss inherently conflicting design objectives. Traditionally, this tradeoff has been addressed using mathematical optimization techniques applied to winding structures and magnetic geometries.

Unlike these approaches, which mainly rely on increasing switching frequency and optimizing winding arrangements, the concept introduced in [9–10] proposes a fundamentally different design philosophy for high-frequency magnetics. This concept divides the magnetic core into two functional regions: an active region, responsible for energy transfer, and a passive region, responsible for dissipating magnetic losses. Each region operates at a different magnetic flux density. The flux in the active region, referred to as the active flux, is intentionally maximized to increase the power capability of the magnetic component. In contrast, the passive flux in the passive region is optimized to minimize energy loss.

This concept, referred to as unbalanced-flux magnetics (UBFM), demonstrated a breakthrough result in its first experimental implementation, achieving a magnetic power density of 224 W/cm³, corresponding to an approximately 300% increase compared with conventional MHz-based magnetic designs [10]. The concept was validated in a 1-kW LLC converter with diode rectification, achieving 94.5% efficiency, while the transformer delivered 750 W with forced-air cooling. The transformer could not deliver the full

rated power mainly due to suboptimal geometric design. In particular, the ratio between the outer and central cross-sections of the magnetic core, defined as $A_r = A_o/A_c$, was equal to 0.8, which lies outside the optimal range of 1–1.25 as shown in [9].

The Unbalanced-Flux Magnetic (UBFM) concept introduces a novel design paradigm for high-frequency (HF) magnetic components, offering an alternative to the prevailing MHz-based design approach. The core principle of UBFM is to prioritize an increase in flux density B , rather than switching frequency, as the primary means to enhance power density. This approach has enabled a significant improvement in the achievable power density of HF magnetics.

To further validate the proposed concept, it is essential to evaluate its performance across different applications and design conditions. In this context, this paper presents a second experimental demonstration of the UBFM concept. In this implementation, the area ratio A_r is adjusted to operate within its identified optimal range, and the diode rectifier used in the previous prototype is replaced with synchronous rectification to improve efficiency. Additionally, the output voltage is reduced from 48 V to 24 V to better align with typical low-voltage, high-current applications.

The main contributions of this work, which distinguish it from the state-of-the-art and from our previous work in [10], are summarized as follows:

- An experimental demonstration of a UBFM-based LLC transformer achieving a power density three times higher than the average power density obtained with MHz-based designs
- Experimental validation of the optimal A_r ratio in the range of 1–1.25, demonstrating improved thermal performance and increased power capability of the transformer compared to [10], which uses $A_r=0.8$
- Detailed optimization design procedure of the transformer and the secondary MOSFETs and experimental characterization of the thermal performance of the converter, together with comparisons with existing high-frequency magnetic designs.

This paper is structured as follows: section II describes the design procedure of the UBFM transformer and the thermal dissipation approach for the synchronous rectifier. Section III presents the experimental setup and discusses the obtained results. Finally, the conclusions summarize the main findings of this work.

II. DESIGN OF UBFM LLC TRANSFORMER

The designed half-bridge LLC circuit is shown in Fig.1. The resonant tank consists of L_r and the series resonant capacitors C_{r1} and C_{r2} . The transformer is implemented using two integrated center-tapped transformers. The output rectification stage employs synchronous MOSFETs to reduce conduction losses and improve efficiency. The main specifications of the converter are summarized in Table I.

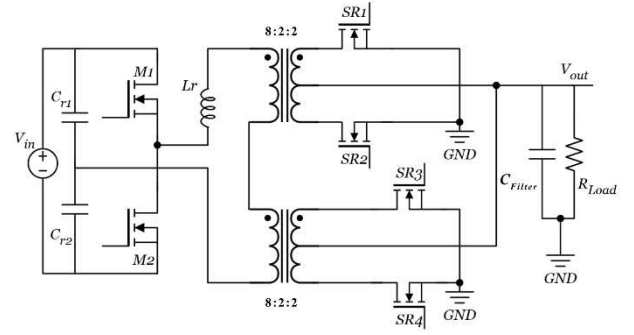


Fig. 1. Designed LLC circuit.

Tab. I. Converter Specifications

	Unit	Value
Input voltage	V	400
Output voltage	V	24
Output power	kW	1
Resonant frequency	kHz	350

A. Transformer optimization

The primary objective of the transformer optimization is to minimize the total transformer losses while maintaining a minimum volume and ensuring that the temperature rise remains within the maximum allowable limit. Transformer losses consist of winding losses and core losses, which constitute the objective functions to be minimized. These losses depend on several parameters such as switching frequency, number of turns, magnetic flux density, winding cross-sectional area, winding length, and core cross-sectional area.

In the considered design, the switching frequency and the number of turns are fixed at 350 kHz and 16 turns, respectively. Therefore, the central optimization variable is the central leg diameter, which determines all other geometric parameters of the transformer. Since the primary objective of this work is to maximize power density rather than efficiency, the winding width is dimensioned according to the rated current and the selected copper layer thickness of 5 oz (175 μm). This copper thickness is suitable for the operating switching frequency and helps minimize skin-effect losses. Although oversizing the winding reduces winding losses, it negatively affects power density and may increase core losses. For the UBFM core structure, the outer cross-section A_o is selected to be 1.23 times the central cross-section A_c . This value lies within the optimal outer-to-center cross-section ratio range $A_r \in [1, 1.25]$, which was shown in [9] to provide improved thermal performance and reduced losses.

The magnetic losses are calculated using Steinmetz equation adopted for UBFM core concept as follows [9]:

$$P_{c_unb} = k f^\alpha B_m^\beta V_{cb} \left[\left(\left(\frac{1}{2 A_r} \right)^{\beta-1} - 1 \right) \delta + 1 \right] \quad (1)$$

Where δ is the outer-to-central core volume ratio and A_r is the outer-to-center cross section ratio defined as

$$\begin{cases} A_r = \frac{A_o}{A_c} \\ \delta = \frac{V_{co}}{V_{cb}} \end{cases} \quad (2)$$

Where V_{cb} is the core volume of the balanced flux core given by

$$V_{cb} = A_c L_c \quad (3)$$

The winding loss are the sum of the primary and secondary winding losses and are expressed as

$$P_w = R_p I_p^2 + R_s I_s^2 \quad (4)$$

Where R_p and R_s are the primary and secondary windings resistance.

Equations (1) and (4) are the objectives functions to be minimized. The following subsections present the intermediate design relations required to determine the variables involved in the loss calculations.

1) Intermediate design function for core loss

The magnetic flux density (B) is calculated using Faraday's law as

$$B = \frac{V_{in}}{4 N_p A_c f} \quad (5)$$

The expression of B requires determining the cross-sectional area A_c , which is given as follows while the remaining variables are given data:

$$A_c = \pi R_c^2 \quad (6)$$

Where R_c is the radius of the central leg. The core base area (A) is given as follows:

$$A = W_c L_A \quad (7)$$

Where W_c and L_A are the width and the length of the base. The core window length l_w is given as follows:

$$l_w = 2R_c + 2w_w \quad (8)$$

Where w_w is the winding width selected based on the rated current. The base length is calculated by equation (9):

$$L_A = l_w + 2\frac{A_o}{W_c} \quad (9)$$

The base width is calculated by equation (10) to enclose the winding within the core:

$$W_c = l_w \quad (10)$$

The core height (H) is given by:

$$H = h + 2\frac{A_o}{w_c} + m \quad (11)$$

Where h is the PCBs height and m is a safety margin. The core length of the UBFM L_{cu} is equal to:

$$L_{cu} = 2\left(l_w + \frac{A_o}{W_c}\right) + h + \frac{A_o}{w_w} \quad (12)$$

The core volume is given as follows:

$$V_c = 2A \frac{A_o}{w_w} + h A_c \quad (13)$$

The box volume is given as follows:

$$V_{box} = A H \quad (14)$$

2) Intermediate design function for winding loss

The AC resistance is defined by:

$$R_{DC} = K \frac{\rho l}{S} \quad (15)$$

Where K is a Dowell factor to account for the skin and proximity effects. The winding length (l) is expressed as function of the R_c , w_w and A_c as follows:

$$l = 2\pi N \left(R_c + \frac{w_w}{2}\right) = 2\sqrt{\pi} N \left(\sqrt{A_c} + \frac{w_w}{2}\right) \quad (16)$$

The layer cross section is the product of the layer thickness (l_{th}) and turn width (w_w):

$$S = l_{th} w_w \quad (17)$$

The secondary current depends on the output current, the number of transformers connected N_t in parallel on the secondary side and the winding configuration k_w (Centre Tapped or conventional). For center tapped- transformer and $N_t=2$, it is written as follows:

$$I_s = K_w \frac{1}{N_t} \frac{\pi}{2} I_o = K_w \frac{1}{N_t} \frac{\pi P_o}{2 V_o} \quad (18)$$

The primary current is approximated as follows:

$$I_p = \frac{P_o}{\eta V_p} \quad (19)$$

Where V_p (200 V) is the primary voltage and η is the efficiency (0.95).

3) Design constraints

$$B_{max} = 0.3 T \quad (20)$$

4) Optimization results

Fig. 2 presents the optimization results showing the effect of the central leg diameter D_c on the magnetic flux density B , the total transformer loss P_t , and the core volume for different values of N_p .

As shown in Fig. 2(a), for a given flux density B , increasing the number of primary turns N_p leads to a reduction in the required central diameter D_c . Consequently, increasing N_p is beneficial for minimizing the magnetic core size while satisfying the flux density constraint. For instance, when $N_p = 8$, a diameter of $D_c = 6\text{mm}$ satisfies the design constraint ($B_{max} < 0.3 T$) while achieving the minimum total loss. In

contrast, when $N_p = 4$, the diameter required to satisfy the same flux density constraint is approximately 9mm. This value is about 50% larger than that obtained with $N_p = 8$, which results in an increase in core volume of approximately two times, as illustrated in Fig. 2(c).

It has been reported in several studies that fractional-turn transformer designs can improve magnetic power density. However, under our design conditions considered in this work (350 kHz switching frequency and high flux density $B \approx 0.25$ T), this assumption does not hold. In particular, a half-turn secondary winding configuration obtained with $N_p = 4$ results in a lower power density than a full-turn configuration with $N_p = 8$.

Another important observation from Fig. 2(b) is that the minimum total loss is weakly dependent on the number of turns. This indicates that the transformer power density can be increased by selecting a smaller central diameter D_c while maintaining nearly the same efficiency.

The optimal central leg diameter of 6 mm, obtained using the proposed design approach, is consistent with the value reported in [10] obtained by optimization using Genetic Algorithms. Based on the findings presented in [9], the ratio A_r is increased from 0.8, as used in [9], to 1.23. The main geometric differences between the first core version and the proposed design are illustrated in Fig. 3. As can be observed, increasing A_r results in a larger core volume (3.42 cm³ compared with 2.17 cm³). However, the higher A_r improves the thermal dissipation capability of the magnetic core, enabling the transformer to handle higher power levels and ultimately achieve a higher power density. This design modification allows the transformer to operate safely at higher flux density while maintaining acceptable thermal performance, which is essential for achieving the high power density demonstrated in this work.

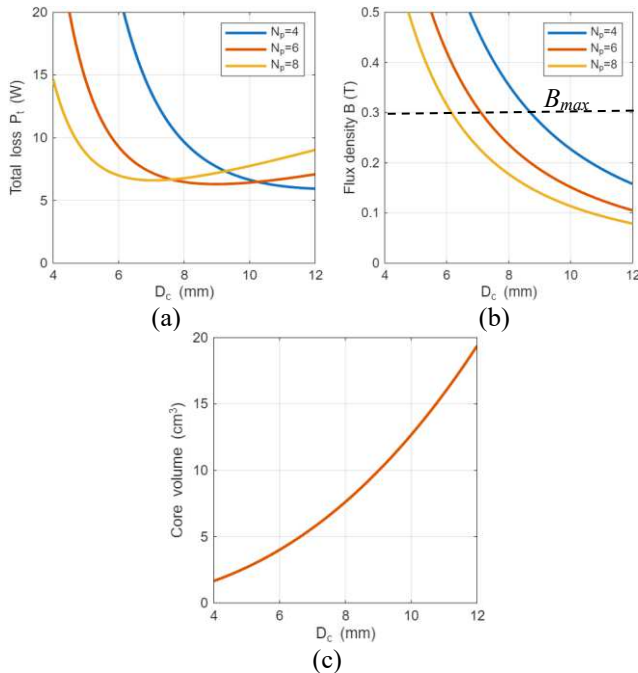


Fig. 2. Effect of D_c for different N_p (per transformer) on: (a) magnetic flux density, (b) total transformer loss and (c) core volume.

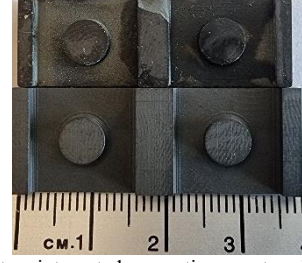


Fig. 3. Designed two integrated magnetic cores: top is 1st version in [10], bottom is new version [this work].

B. Thermal Dissipation of Secondary MOSFET Losses Using Thermal Vias

Copper conduction losses increase with the square of the current. Therefore, in high-current applications, the dissipation of conduction losses in switching devices can significantly affect both efficiency and thermal performance. Consequently, careful thermal design of the cooling structure is required.

In the proposed converter, the thermal dissipation of the secondary-side MOSFETs is achieved using thermal vias placed beneath the drain pad. The via structure must be optimized to minimize the thermal resistance while satisfying the geometric constraints imposed by the available drain pad area.

The optimal solution involves determining two interdependent design variables: the number of thermal vias and their diameter. The structure of the drain-pad via array is illustrated in Fig. 4, where d_v denotes the via diameter and e represents the spacing between two adjacent vias, which is determined by PCB manufacturing constraints. Let N_l and N_2 denote the number of vias along the horizontal and vertical directions, respectively, and N_{tv} the total number of vias.

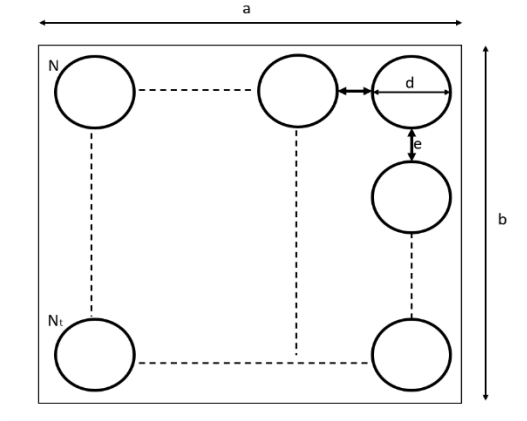


Fig. 4. Vias structure for thermal dissipation of synchronous Mosfet.

The MOSFET power loss is dominated by conduction loss, while switching loss is negligible due to soft-switching operation. The conduction loss is given by:

$$P_{cond} = R_{ds} I_{MOSFET}^2 \quad (21)$$

where R_{ds} is the drain-to-source resistance and I_{MOSFET} is the RMS current through the MOSFET. The drain dimensions (a, b) of the MOSFET is (5, 6) mm.

The total drain area A is given by the product of length and width:

$$A = a b \quad (22)$$

The spacing between two successive vias e is given by:

$$e = \frac{a - N_1 d_v}{N_1 - 1} = \frac{b - N_2 d_v}{N_2 - 1} \quad (23)$$

where: N_1 is the number of vias in the horizontal direction, N_2 is the number of vias in the vertical direction and $N_{tv} = N_1 N_2$ is the total number of vias.

The number of vias can be expressed from (23) as:

$$N_1 = \frac{1 + \frac{a}{e}}{1 + \frac{d_v}{e}}, N_2 = \frac{1 + \frac{b}{e}}{1 + \frac{d_v}{e}} \quad (24)$$

Thus, the total number of vias is:

$$N_{tv} = \left[\frac{1 + \frac{a}{e}}{1 + \frac{d_v}{e}} \right] \left[\frac{1 + \frac{b}{e}}{1 + \frac{d_v}{e}} \right] \quad (25)$$

The thermal resistance of a single via is given by:

$$R_{thv} = \frac{h}{\sigma \pi t (d_v + t)} \quad (26)$$

where h is the PCB thickness, d_v is the via diameter, t is the copper thickness, and σ is the thermal conductivity of copper. Increasing the via diameter reduces the thermal resistance of an individual via, however, it reduces the total number of vias that can be placed within the available drain area. Therefore, the overall thermal resistance must be evaluated considering both effects.

Since the vias form parallel thermal paths, the equivalent thermal resistance is given as follows:

$$\frac{1}{R_{th}} = \sum_{i=1}^{N_{tv}} \frac{1}{R_{thv_i}} = \frac{N_{tv}}{R_{thv_i}} \quad (27)$$

Substituting (25) into (27), we get:

$$R_{th} = \frac{h}{\sigma \pi t (d + t)} \left[\frac{1 + \frac{a}{e}}{1 + \frac{d_v}{e}} \right]^{-1} \left[\frac{1 + \frac{b}{e}}{1 + \frac{d_v}{e}} \right]^{-1} \quad (28)$$

Equation (28) is nonlinear function of via's diameter. By fixing the via spacing according to PCB manufacturing constraints, the optimal via diameter can be determined numerically.

The optimization results indicate that a via diameter $d_v = 0.6$ mm and $N_v = 16$ are the optimal solutions for thermal dissipation for our case.

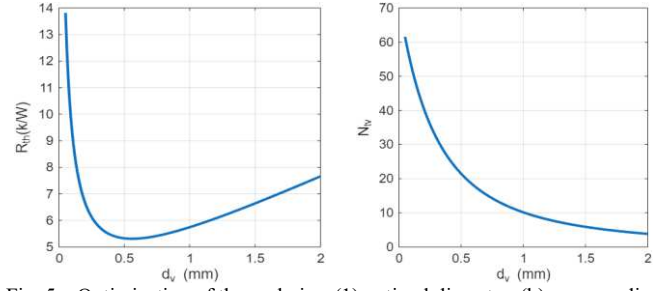


Fig. 5. Optimization of thermal vias: (a) optimal diameter, (b) corresponding optimal N_v .

C. Winding design

The layouts of the primary and secondary windings are shown in Fig. 6 and Fig. 7, respectively. In the proposed design, a partial interleaving technique is employed between the primary and secondary windings in order to reduce proximity-effect losses.

For the primary winding, each transformer uses two turns per layer. All the turns are connected in series, resulting in 8 turns per transformer and 16 turns for the full-integrated transformer.

For the secondary winding, each transformer employs one turn per layer. The top turn in layer 1 is connected in series with the top turn in layer 4. A symmetric configuration is implemented on the bottom side, and both together form the first center-tapped secondary winding. As illustrated in Fig. 7, the top turn in layer 1 is connected to SR1, while the bottom turn in layer 4 is connected to SR2.

A similar configuration is implemented for layers 5 and 8 to form a second center-tapped secondary winding. Finally, the outputs of the two center-tapped structures are connected in parallel to obtain the required 24-V output.

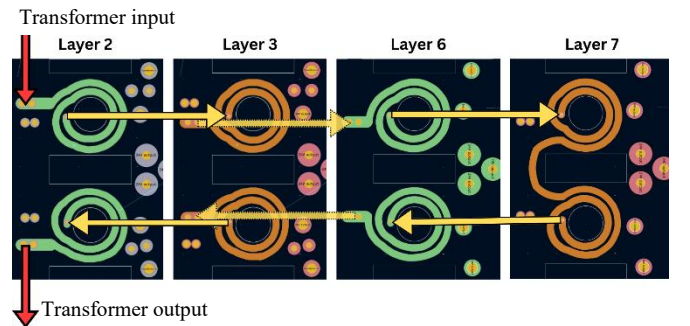


Fig. 6. Designed primary winding.

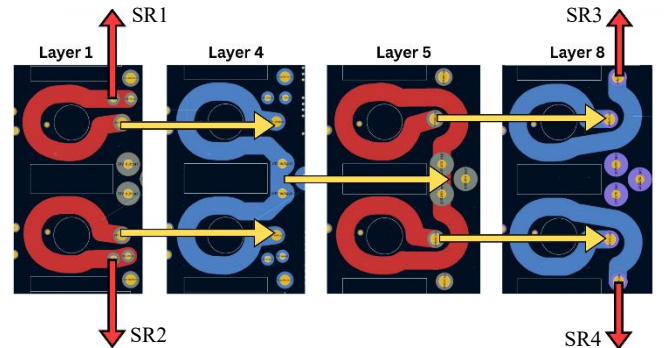


Fig. 7. Designed secondary winding.

III. EXPERIMENTAL RESULTS AND DISCUSSION

The designed LLC converter was tested up to 1.1 kW and operated continuously for 67 min. A photo of the tested prototype is shown in Fig.8. The main components of the power circuit are listed in Tab. II. A forced air cooling was applied using a fan (FFB0912EHE) delivering about 100 CFM. The experimental setup is described as follows.

- 1) **Load type:** A programmable dc electronic load (EA-PSB 9200-210) was used to emulate the output load conditions. The load was configured in constant-current mode.
- 2) **Measurement tools and test conditions:** Voltages and currents were measured with a Tektronix series 6 oscilloscope equipped with differential and current probes (Testec/100 MHz, Keysight N2782B 30 A/50 MHz). Ambient temperature during testing was maintained at 23 ± 2 °C. The temperature is measured using a Fluke Ti9 thermal camera. Efficiency is measured using a high precision power analyzer ZEEMER LMG500.
- 3) **Test duration and consistency:** Each data point was recorded after thermal equilibrium was reached.

Tab. II. Power components of the designed Converter

Component	Model
Inverter switches	GS66508T GaN HEMTs
Rectifier switches	BSC037N08NS5
Resonant Inductor	SER2013-402L
PCB	4 layer, 5oz

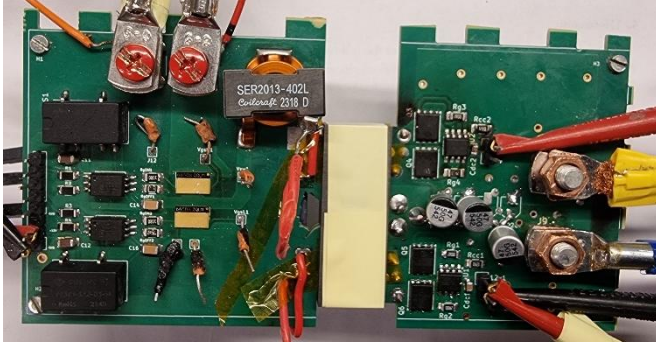


Fig. 8. Photo of the realized LLC converter.

A. Power density

The transformer box volume is 4.78 cm^3 , resulting in a power density of 230 W/cm^3 at the peak power of 1.1 kW. This value is 2.6% higher than the power density reported in [10]. Compared with other MHz-based transformer designs, the achieved power density is $3.83\times$, $3.26\times$, and $3.48\times$ higher than the transformers reported in [2], [3], and [6], respectively, as illustrated in Fig. 9. It should be noted that the comparative analysis is limited to reference designs that provide sufficient and reliable thermal data for their transformer implementations. Some of the limitations related to thermal characterization in existing works are discussed in [9]. The significant increase in power density achieved with the UBFM concept highlights several advantages over conventional MHz-based magnetic designs. First, the UBFM concept introduces a fundamentally different design approach

for high-frequency magnetics, enabling a substantial increase in power density through a simple modification of the magnetic core geometry. Over the past decade, the power density of high-frequency transformers has remained relatively stable, with an average value of approximately 70 W/cm^3 . By modifying the core geometry to reduce magnetic losses and improve thermal performance, the proposed UBFM design increases the power density to 230 W/cm^3 , representing an improvement of more than three times compared with typical designs.

Importantly, this improvement is achieved without relying on ultra-high switching frequencies, specialized magnetic materials, or complex custom magnetic structures. As a result, the proposed approach can simplify the manufacturing process, reduce cost, and mitigate electromagnetic interference (EMI) issues commonly associated with MHz switching operation. With the UBFM concept, high-frequency transformers can reach a footprint comparable to that of the switching devices. This opens the possibility of achieving LLC converter power densities exceeding 3.7 kW/in^3 with optimized system architecture.

The experimental results demonstrate that increasing the magnetic flux density while maintaining acceptable thermal performance is a key factor for improving magnetic power density. This design principle follows directly from the electromagnetic power relation $P \propto fB^2$, which highlights the strong dependence of magnetic power capability on flux density. Future work will investigate the possibility of further improving power density by increasing the switching frequency beyond 300 kHz while maintaining a high operating flux density.

The more important advantage of the UBFM concept is its topological flexibility. Unlike MHz-based magnetic designs, which are typically restricted to soft-switching resonant converters, UBFM can be applied to a wide range of converter topologies. Resonant converters often present limitations related to input voltage range and control complexity. In contrast, the UBFM approach can also be applied to duty-cycle-controlled converters, such as full-bridge and phase-shifted full-bridge converters, enabling higher power density in a broader class of power conversion systems.

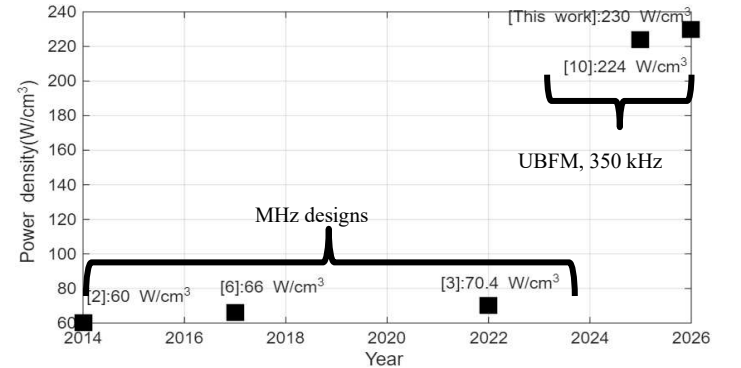


Fig. 9. Advancement of power density in MHz LLC transformers: Demonstrating the impact of UBFM.

B. Resonant current and primary voltage waveforms

The measured resonant current and primary voltage waveforms corresponding to output power levels ranging from 700 W to 1000 W (with a step size of 100 W) are shown in Fig. 10. At 1 kW output power, the measured peak and RMS values of the resonant current are 7.77 A and 5.8 A, respectively. These values are lower than those reported in existing works. For example, at the same output power (1 kW), the reported resonant currents (peak, RMS) are (9.2 A, 5.77 A) and (10 A, 6.69 A), in [2]–[3], respectively.

The quality of the resonant current waveform can be evaluated using the crest factor (CF), defined as the ratio between the peak and RMS values of the current. The crest factors of the different designs are summarized in Table III. As shown, the proposed design achieves the lowest crest factor among the compared works. A lower crest factor indicates reduced current stress on both the switching devices and the magnetic components.

It can also be observed that the converter operates slightly above the resonant frequency f_r . The resonant tank was initially designed to operate at 345 kHz, using a 4.2 μH

resonant inductor and a 50 nF resonant capacitor. However, experimental verification showed that the actual inductance of the resonant inductor is approximately 5 μH , which shifts the operating frequency slightly above the designed resonant frequency.

Future work will focus on the design of a UBFM-based resonant inductor integrated with the transformer, with the aim of further improving both the efficiency and the power density of the converter.

Tab. III. Resonant current characteristics

	This work	[2]	[3]
P_o (kW)	1	1	1
Peak current	7.75	9.2	10
RMS current	5.8	5.77	6.69
CF	1.33	1.59	1.49

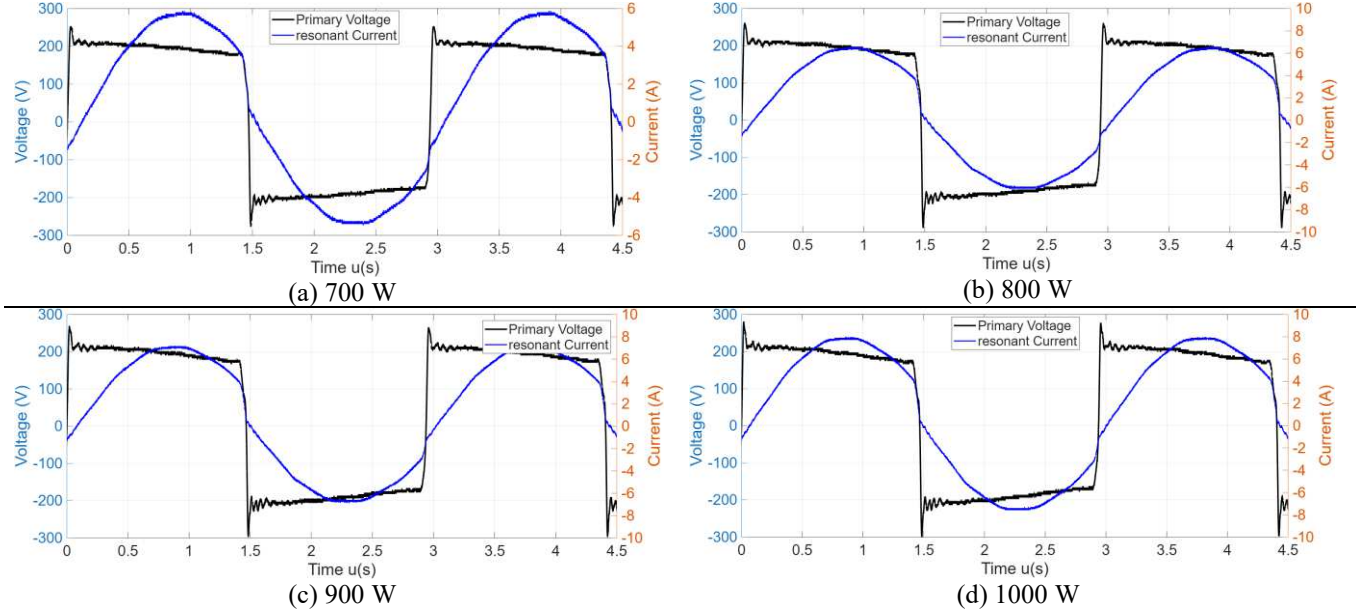


Fig. 10. Measured primary voltage and resonant current.

C. Thermal performance

The transformer temperature, shown in Fig. 11, was recorded at each output power level ranging from 100 W to 1100 W after thermal steady state was reached. For instance, at 1 kW output power, the converter was operated for approximately 14 minutes before the temperature stabilized. The operating time can be read in thermal photographs in Fig. 11. The change from one power level to another is made without stopping the converter.

Compared with the transformer design reported in [10], the present design exhibits improved thermal performance. This improvement can be mainly attributed to the modification of the ratio A_r . According to the fundamental study of the UBFM concept in [9], the optimal value of A_r lies within the range 1–1.25. In the present design, A_r was increased from 0.8, as used in [10], to 1.23. This modification reduces both the core

loss and the thermal resistance, thereby increasing the power capability of the transformer. It is important to note that the magnetic flux density in the central core leg remains unchanged. Instead, the magnetic flux density in the outer region is reduced, which contributes to lowering the core losses and the overall temperature rise.

The variation of transformer temperature with respect to the square of the output current is shown in Fig. 12. As observed, the temperature rise varies approximately linearly with I_o^2 , which is consistent with theoretical expectations. The transformer temperature is influenced by both core losses and winding losses. However, core losses can be assumed nearly constant since they are independent of load current and if we also neglect the small input voltage variation. Consequently, the temperature variation is mainly determined by the winding losses. According to Joule's law, winding losses are

proportional to I_o^2 , which explains the observed linear relationship.

At the rated output power (1 kW), the measured transformer temperature is approximately 83 °C, while at the peak power (1.1 kW) the temperature reaches about 88 °C. Assuming that winding losses are negligible at 100 W, the transformer temperature due to core losses alone is approximately 50 °C. Taking this value as a reference threshold, the temperature rise caused by winding losses is approximately 38.5 °C at the rated load.

The temperature rise associated with winding losses can therefore be expressed as

$$T_r = R_{th} R_{eq} I_o^2 \quad (29)$$

Where R_{th} is the thermal resistance of the magnetic core and R_{eq} is the equivalent resistance reflected to the secondary side. The slope, noted S , of the function $T_r = f(I_o^2)$ is therefore the product:

$$S = R_{th} R_{eq} = 0.0185 \text{ K/A}^2 \quad (30)$$

The parameter S represents an important metric that can be used in future work to evaluate design improvements. For instance, when the magnetic core geometry remains unchanged, changes in S directly reflect modifications in the equivalent winding resistance. Similarly, improvements in core geometry that reduce thermal resistance can also be evaluated using this metric when the winding design is kept unchanged.

In many existing works, the thermal characterization of high-frequency transformers is limited to reporting the temperature at the rated load, which does not allow evaluation of the temperature variation across the operating range. In contrast, the design reported in [10] provides detailed thermal data over a wide range of output power, enabling a more comprehensive analysis of transformer thermal behavior.

For comparison purposes, the present study focuses on the transformer designs reported in [2] and [10]. Although [2] does not provide detailed thermal measurements in the publication, the experimental setup and operation were demonstrated on the CPES research group YouTube channel [11], where the thermal behavior of the transformer can be observed.

Assuming that the cooling fan shown in Fig. 13 has similar performance to the fan used in the present prototype (FFB0912EHE), it can be concluded that the proposed transformer achieves superior thermal performance compared with the design in [2], despite having a 3.5x smaller volume. Under comparable cooling conditions, the transformer temperature of the proposed design is approximately 83 °C, whereas the temperature reported for the design in [2] exceeds 95 °C.

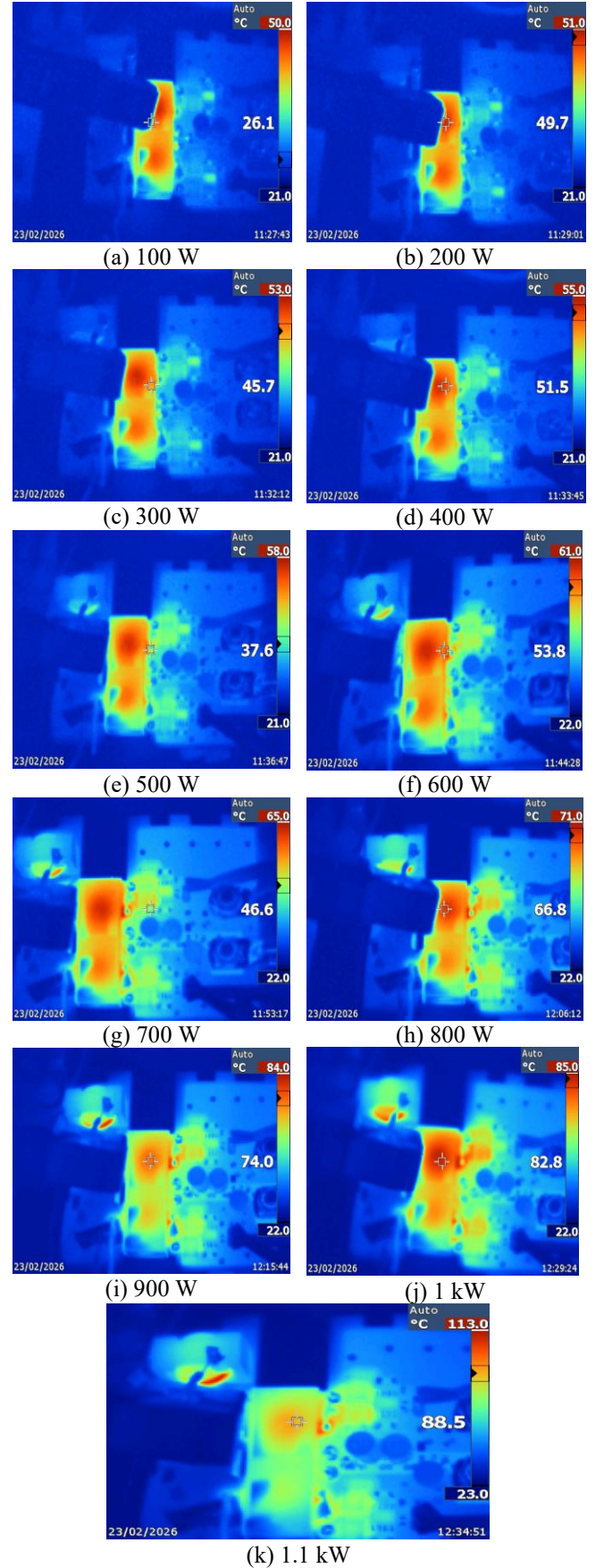


Fig. 11. Thermal images of the tested converter at different power levels

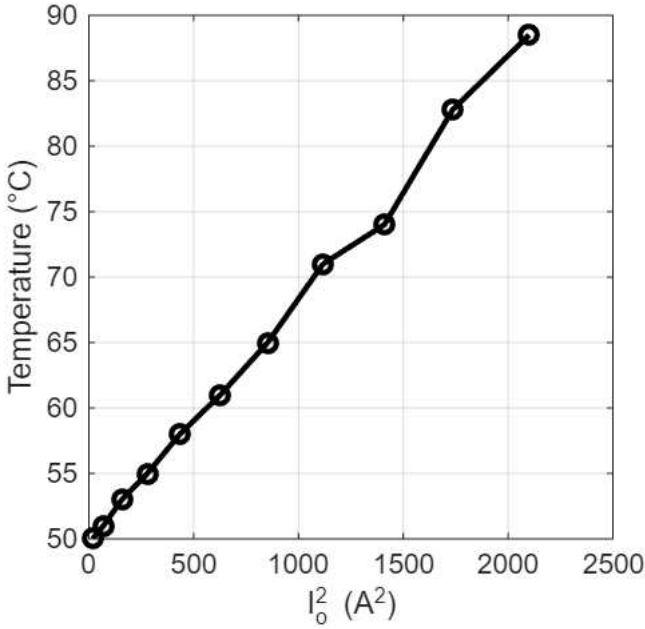


Fig. 12. Variation of the measured temperature with respect to current squared.

With a fan delivering approximately 65 CFM, the temperature of the designed transformer at 1 kW reaches 103 °C. Compared with the transformer design reported in [10], the proposed version exhibits improved thermal performance. At the same output power (700 W), the transformer temperature decreases from 100 °C in [10] to 65 °C in the present design when using a 100 CFM cooling fan. It should be noted, however, that the previous design in [10] was tested using a 65 CFM fan. When the new transformer is tested under the same cooling conditions (65 CFM), the measured temperature is approximately 77 °C, which still represents a significant thermal improvement compared with the previous design.

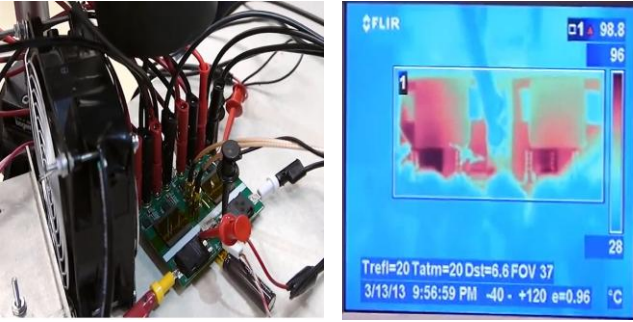


Fig. 13. Measured temperature of LLC converter of [2] shown in [11].

For comparison with the first prototype reported in [10], one additional factor must be considered: the magnetic material used for the transformer core. The first version was implemented using DMR96, which exhibits a lower core loss density compared with 3C96. For example, at 350 kHz, 100°C, and 250 mT, the core loss density of DMR96A is approximately 3000 kW/m³, whereas it reaches about 4000 kW/m³ for 3C96. Despite this disadvantage in core material, the proposed design still demonstrates improved thermal performance.

D. Efficiency

The efficiency of the converter is measured using ZEEMER power analyzer where the input voltage and input current are connected directly to the voltage source. This means the input voltage is the voltage source and the input current is the DC current absorbed by the converter. The output power is measured by the power analyzer. The maximum registered efficiency is 96.07 %, reached within the power range [400-500W]. The maximum efficiency of this design is higher than the efficiency achieved in [10] (94.6 %). The registered efficiency at the rated power and the peak power are 95.4% and 94.9% respectively.

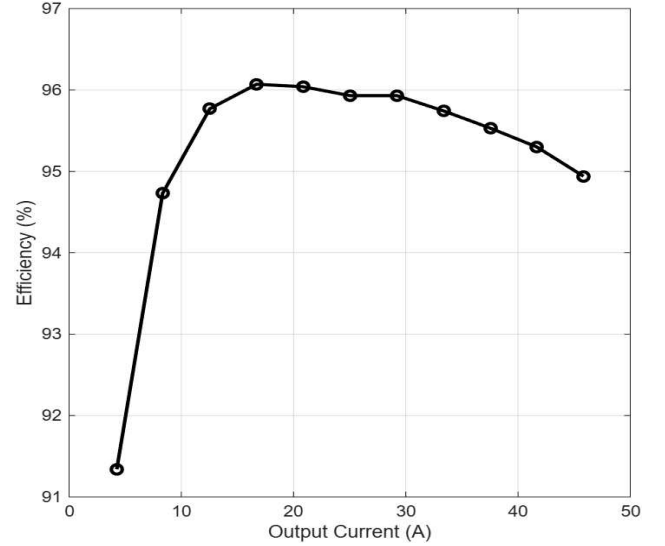


Fig. 14. Measured efficiency from 100 W to 1.1 kW, $V_o=24$ V.

IV. CONCLUSION

This paper presented a second experimental demonstration of the unbalanced-flux magnetics (UBFM) concept applied to a 1-kW LLC transformer. Compared with the first UBFM prototype, the improved design exhibits higher efficiency, improved thermal performance, and increased power capability. The proposed transformer achieves a power density of 230 W/cm³ with a maximum converter efficiency of 96%. These results were obtained at a switching frequency of 350 kHz and an operating magnetic flux density of approximately 0.25 T.

The proposed approach differs fundamentally from the conventional MHz-based magnetic design methodology, which typically increases the switching frequency to the MHz range while reducing the operating magnetic flux density to approximately 50–70 mT. In contrast, the UBFM approach is based on maximizing the magnetic flux density while maintaining acceptable thermal performance, thereby significantly increasing the magnetic power capability.

The experimental results demonstrate several advantages of the UBFM approach over conventional MHz designs. First, the proposed design achieves more than three times higher power density, which can significantly reduce the size and cost of power converters. Second, the improved thermal performance enables the use of simpler and smaller cooling systems. Third, and most importantly, the UBFM concept is

not restricted to resonant converter topologies. Unlike MHz-based designs, which typically rely on soft-switching operation, the UBFM approach can be applied to a wide range of converter topologies, including hard-switching converters. Future work will investigate the application of the UBFM concept to duty-cycle-controlled converters, such as full-bridge and phase-shifted full-bridge converters, with the objective of further increasing converter power density.

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