An Integrated Transformer Design With a Center-Core Air-Gap for DAB Converters

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ABSTRACT In this paper, an integrated transformer design methodology for dual-active-bridge (DAB) converters, which basically utilize leakage inductance so as to substitute a cumbersome inductor in the conventional DAB converters, is newly proposed. The proposed integrated transformer with integrated leakage inductance inherently includes leakage inductance and magnetizing inductance, whose values can be appropriately selected by modulating the number of turns and the center-core air-gap. Based on the theoretical analysis of the DAB converters, the major parameters, i.e., primary and secondary current levels and its phase difference, can be derived, and these values are utilized for magnetic analysis of the proposed integrated transformer. Considering core loss and copper loss of the proposed integrated transformer in the DAB converters, optimal values of the center-core air-gap \(l_g\) and the optimal number of turns \(N_{1,\text{op}}\) \& \(N_{2,\text{op}}\) can be determined so that high power efficiency is achieved at a normal operating point of the DAB converters. The 20W prototypes of the DAB converter with EER2834 and EER3019 of the ferrite cores were fabricated and verified by simulation and experiment. The design results showed that the optimal number of turns at \(l_g = 2.0\)mm are found to be 11 and 10, for the EER2834 and EER3019, respectively. Compared to 94.5\% and 96.3\% of AC efficiencies for the conventional transformer and inductor set, the proposed integrated transformer achieved 95.8\% and 96.9\% of AC efficiencies at maximum power transfer point of \(\theta_p = \pi/2\) for EER2834 and EER3019, respectively.

INDEX TERMS Transformer design, integrated transformer, transformer leakage, leakage inductance, dual-active-bridge (DAB) converter, DC/DC converter, magnetic analysis, ferrite core air-gap.

I. INTRODUCTION

DC/DC power conversion implemented by high-frequency switching devices is essential to various electronic devices. For safety of human and electronic equipment, isolation-type DC/DC converters are required and plenty of DC/DC converter topologies have been developed thus far [1]–[20]. Dual-active-bridge (DAB) converters are the most popular topology as an isolation-type bidirectional DC/DC converter due to using less component, simple circuit structure, low switching losses, bidirectional power flow, low sensitivity to parasitic components, and relatively wide-range of zero-voltage switching (ZVS) operations [3]–[20]. By virtue of such advantages, the DAB converters are widely used for various power applications, e.g., DC smart grid [3], [4], photovoltaic applications [5], solid-state-transformer (SST) [6], [7], battery storage systems (BSS) [8], [9], and automotive applications [10], [11]. As a typical method to control the load power through a transformer, a simple control method, e.g., single phase shift (SPS) modulation, is generally adopted in the DAB converters. The SPS modulation methods, however, result in high-circulating currents, hard-switching issues at light load conditions, and limited ZVS range when source and load voltages are mismatched [1], [2]. Therefore, novel control methods to overcome such disadvantages of the SPS modulation methods have been widely suggested thus far [12]–[20]. Extending of voltage gains and ZVS operation ranges have been studied for high efficient operation of the DAB converters [12], [13]. Furthermore, there are plenty of novel modulation methods, e.g., dual phase...
shift (DPS) methods [14], [15], front and rear end switch (FRS) methods [16], extended phase shift (EPS) methods [17], [18], triple phase shift (TPS) methods [19], [20]. Such previous studies mentioned above eventually decrease circulating current, expand ZVS operation ranges, and reduce non-active power of the DAB converters for high efficiency operation [12]–[20]. From recent studies of the DAB converters, high-frequency transformers and additional inductors are essentially adopted in the DAB converters. Although the transformers are manufactured as small as possible, the transformer loss may become severe due to the core and copper losses, which are not deeply considered for a minimum transformer loss [1]–[23]; hence, a practical solution to design an inductor and a transformer is necessary for high power efficiency and high power density of the DAB converters.

There are plenty of issues to design high performance transformers [23]–[32]. As the representative methods of designing transformers, various transformer design algorithms have been widely studied for transformer design optimization (TDO) [24]–[26]. Above TDO methodologies, however, are not suitable to directly apply in DAB converters because the current waveforms of the transformer in DAB converters are not sinusoidal, and the transformer efficiency of the DAB converters depends on load condition and phase difference between primary and secondary voltages of DAB converters [1]–[4]. As an optimal designed compact transformer, variety of transformer shapes, e.g., adjustable-tap transformers, foil winding-based transformers, and compact planar transformers, could be applicable to DAB converters [23], [27], [28]. However, an inductor should be additionally adopted in DAB converters, and furthermore winding turns in the transformers should be appropriately designed for a minimum transformer loss. To make better utilization of the core material, stacking the magnetic components of transformers or inductors may be preferred for compact magnetic components [23], [29], [30], [31]. Such stacking the magnetic components are only applicable to specially designed transformer and inductor set, and may increase the total size of magnetic components due to additional inductors [23], [29]. Integrated LCLC resonant transformers and integrated multiphase transformers have been used in resonant converters for high power density and low core loss by current balancing method [30], [31]. Although these types of the integrated transformers can be utilized in special cases of the resonant converters, e.g., LLC resonant converters and multiphase resonant converters, they cannot be applied to the DAB converters due to different characteristics between DAB converters and resonant converters [30], [31]. As an integrated inductor and transformer set that has been utilized to DAB converters, the gap between primary and secondary windings wound in a center core can be modulated to obtain small leakage inductance [32]. Due to a strong magnetic coupling between primary and secondary windings, large leakage inductance cannot be obtained by this way and eventually the total size of the transformer core may become large due to the large gap between primary and secondary windings. To sufficiently integrate an inductor and a transformer for DAB converters, various integrated transformers have been examined to obtain leakage inductance [33]. However, they focused on comparative magnetic loss characteristics for various magnetic materials, and only core loss has been considered for transformer loss. Gapped E-I ferrite cores-based LLC converter has been proposed for integrated transformers [34]. Two side legs and center leg are the variable parameters to design the integrated transformer for the LLC converters. This research, however, focused on the LLC converter, which operating principle is different from the DAB converters. In addition, plenty of parameters, e.g., side and center legs gaps and four windings, make the integrated transformer design be so complicated; hence, a simple and practical design guideline to select an appropriate leakage inductance for the DAB converters is highly necessary and furthermore a minimum transformer loss analysis should be carried out for high power efficiency operation of the integrated transformer in the DAB converters.

In this paper, an integrated transformer design methodology applied to DAB converters, which excludes an additional inductor for high power density and provides a low transformer loss for high efficiency operation of the proposed integrated transformer, is newly proposed, as shown in Fig. 1. By modulating an air-gap \( l_g \) in center-core of the transformer, the leakage inductance utilized for magnetic power storage component can be appropriately selected for design goal of the DAB converters. The optimal number of turns in accordance with \( l_g \) can be determined, and furthermore optimum value of \( l_g \) having a minimum transformer loss can be derived by the proposed design methodology; hence, contrary to the previous integrated transformer design methods [30]–[34], the proposed design procedure can provide an intuitive and practical solution to design high efficiency and high power density transformers. A design example of 20W DAB converter was built and evaluated by two ferrite core shapes. The validity of the proposed transformer design methodology for DAB converters is verified by simulation and experiment, providing a practical solution to design high power density and high power efficiency of the DAB converters.

II. STATIC ANALYSIS OF THE INTEGRATED TRANSFORMER BASED ON DAB CONVERTERS

In this section, the static analysis of the DAB converter including the proposed integrated transformer will be described to derive the characteristics of the proposed integrated transformer. The DAB converter is composed of primary and secondary inverters, and the proposed integrated transformer, as shown in Fig. 1. The amplitude of \( V_1 \) and \( V_2 \) are determined by the DC voltage \( V_d \) and the load voltage \( V_L \), respectively, and the duty cycle of the gate signals for primary and secondary sides is assumed to be fixed as 0.5 in this section. Contrary to the conventional DAB converters, no lumped inductor is used in the DAB converters; instead, leakage inductance in the primary and secondary sides of the transformer is utilized for a magnetic energy storage...
component. To implement the DAB converter in Fig. 1, a simple control method, i.e., phase shift modulation, is adopted in this paper [1], [2]. Thus, switching loss in the DAB converters is not considered and only transformer loss is considered to focus on verifying the proposed integrated transformer.

To analyze the proposed transformer including two leakage inductance, an equivalent circuit of the proposed transformer is described in Fig. 2. Parasitic resistance or capacitance may be significant when the number of winding becomes so large [31]. However, in this case, because primary and secondary winding for the proposed transformer are far enough separated and the turn ratio is only 1.0, then the parasitic components are assumed to be ignored for simplicity of analysis. Then, from (1)-(2), magnetizing current \( i_m \) can be derived as follow:

\[
\begin{align*}
\dot{i}_m(t) &= \frac{v_m(t)}{j\omega_s L_m}, \\
\dot{v}_m(t) &= \frac{L_m}{L_{d1} + \frac{L_2}{n^2}} v_1(t) + \frac{L_m}{L_{d1} + L_{d2}} v_2(t), \\
\dot{v}_n(t) &= \frac{L_2}{n^2 L_{d1} + L_{d2}} v_2(t).
\end{align*}
\]

where \( \omega_s \) is switching angular frequency (\( \omega_s = 2\pi f_s \)). Based on (3) and Fig. 2, primary current \( i_1 \) can be found if secondary current \( i_2 \) is determined as follows:

\[
i_1(t) = i_m(t) + n i_2(t).
\]

As shown in Fig. 3, due to same operating principle for positive and negative polarities of the proposed DAB converters, two operating modes are only specified for positive period of \( v_1 \). The operating modes begin when \( v_1 \) is positive. The operating modes are as follows:

**Mode 1 \([t_0, t_1]\):** At \( t_0 \), \( v_1 \) changes to \( + V_s (= V_1) \) and \( v_2 \) maintains \( -V_{L/n} (= -V_2) \), as shown in Fig. 3. Then, magnetizing voltage \( v_m \) in this mode is determined as follow:

\[
V_m = \frac{L_{d2} V_1 - n L_{d1} V_2}{n^2 L_{d1} + L_{d2}}.
\]

From (5), the secondary current \( i_2 \) can be derived as follow:

\[
i_2(t) = \frac{1}{L_{d2}} n V_m - V_2 - V_{L/n}(t - t_0) = \frac{n V_1}{n^2 L_{d1} + L_{d2}}(t - t_0).
\]
Mode 2 \([t_1, t_2]\): At \(t_1\), \(v_1\) maintains \(+V_s\) and \(v_2\) changes to \(+V_L/n\), as shown in Fig. 3. Then, magnetizing voltage \(V_m\) in this mode is determined as follow:

\[
V_m = \frac{L_{12}V_1 + nL_{11}V_2}{n^2L_{11} + L_{12}}. \tag{7}
\]

From (7), the secondary current \(i_2\) can be derived as follow:

\[
i_2(t) = i_2(t_1) + \frac{1}{L_{12}}(nV_m-V_2)(t-t_1) = i_2(t_1) + \frac{nV_1-V_2}{n^2L_{11}+L_{12}}(t-t_1). \tag{8}
\]

This mode ends when \(v_1\) changes to negative at \(t_2\).

As shown in Fig. 3, the phase difference between \(V_1\) and \(V_2\) is defined as \(\theta_p\) in this paper. To specify the secondary current, \(i_2\) at \(t = t_1\) and \(t_2\) can be determined from (6) and (8), considering \(i_2(t_0) = -i_2(t_2)\), as follows:

\[
i_2(t_1) = \frac{T_s}{4\pi(n^2L_{11}+L_{12})}n(2\theta_p-1)\left(V_1+\pi V_2\right) \tag{9a}
\]

\[
i_2(t_2) = \frac{T_s}{4\pi(n^2L_{11}+L_{12})}\left[n\pi V_1 + (2\theta_p - \pi )V_2\right] \tag{9b}
\]

\[
\therefore t_1-t_0 = \frac{T_s}{2\pi}\theta_p, \ t_2-t_1 = \frac{T_s}{2\pi}(\pi - \theta_p), \quad 0 \leq \theta_p \leq \pi, \tag{9c}
\]

where \(T_s\) is the switching period, i.e., \(T_s = 1/f_s\).

From (6) and (8)-(9), the load current \(I_L\) in Fig. 1, which is average value of the output current \(I_{o,avg}\), can be calculated as follows:

\[
I_{o,avg} = I_L = \frac{2}{T_s} \int_0^{T_s} i_2(t)dt = \frac{2}{T_s} \left( \int_{t_0}^{t_1} i_2(t)dt + \int_{t_1}^{t_2} i_2(t)dt \right) = \frac{nV_1\theta_p(\pi - \theta_p)}{2\pi^2(n^2L_{11}+L_{12})f_s}. \tag{10}
\]

Therefore, load power \(P_L\), primary and secondary currents \(I_1\) and \(I_2\) can be straightforwardly derived as follows:

\[
P_L = \frac{2}{T_s} \int_0^{T_s} v_2(t) \cdot i_2(t)dt = \frac{nV_1V_2\theta_p(\pi - \theta_p)}{2\pi^2(n^2L_{11}+L_{12})f_s} \tag{11a}
\]

\[
I_1 = \sqrt{\frac{1}{T_s} \int_0^{T_s} i_1^2(t)dt}, \quad I_2 = \sqrt{\frac{1}{T_s} \int_0^{T_s} i_2^2(t)dt}. \tag{11b}
\]

For magnetic analysis of the proposed integrated transformer, it is necessary to identify not only the primary and secondary current levels but also phase difference between primary and secondary currents \(\theta_d\), as shown in Fig. 3. To identify \(\theta_d\), zero-crossing points of \(i_1\) and \(i_2\), i.e., \(t_a\) and \(t_b\) during Mode 1 in Fig. 3, can be derived from (3)-(6) and (9b).
as follows:

\[
\begin{align*}
\theta_d &= \frac{2\pi}{T_s} \left( t_a - t_b \right) \\

\theta_d &= \frac{2\pi}{T_s} \left( t_a - t_b \right) \quad (13)
\end{align*}
\]

From the theoretical results of (11)-(13) described in this section, core and copper losses, which are major source losses of the proposed transformer, can be analyzed by a finite-element-method (FEM) based simulation tool, which will be described in the next section.

To specifically design the DAB converters throughout this paper, a design condition, e.g., source voltage, load conditions, and switching frequency should be preliminarily provided; the design parameters for the DAB converter are summarized in Table 1, whose values will be used to design the proposed integrated transformer throughout this paper. Considering the control slope to modulate load power \( P_L \) w.r.t. \( \theta_p \) for a wide control range, a nominal operating point of \( \theta_p \) is assumed to be set as \( \pi/4 \) in this paper. The load voltage and load power are 20 V and 20 W at \( \theta_p = \pi/4 \), respectively, as one of design examples. The load voltage and load power are 20 V and 20 W at \( \theta_p = \pi/4 \), respectively, as one of design examples. Although the load power delivery condition is only 20 W in this paper, the proposed integrated transformer design can be applicable to the various DAB converters, regardless of the transformer power delivery level. From the result of (11a) and Table 1, the remaining design parameters are \( n^2 L_{12} = L_{12} & 9.35 \mu H \), i.e., \( L_{11} = L_{12} = 9.35 \mu H \), which is a design parameter of this paper to satisfy design goal of the proposed DAB converters. Based on those parameter results, the load power w.r.t. \( \theta_p \) is shown in Fig. 4, where the simulation results were implemented by a PSIM simulation tool. The result of Fig. 4 shows \( P_L \) at \( \theta_p = \theta_p = \pi/4 \), which is a nominal operating point to design the proposed integrated transformer.

### III. THE DESIGN PROCEDURE OF THE PROPOSED INTEGRATED TRANSFORMER

#### A. MAGNETIC MODELING OF THE INTEGRATED TRANSFORMER

To describe the geometrical structure of the transformer core, the physical dimension of transformer core shape is shown in Fig. 5. Magnetic flux path through the core is defined as \( l_1 \) and \( l_2 \) and a center-core air-gap is \( l_g \). Magnetic flux area is \( A_{ct} \) in the center core and such magnetic flux area is divided in half at left and right sides. When primary and secondary currents are applied to the left and right sides of the transformer, respectively, the magnetic flux is generated, as shown in Fig. 6. Then, total magnetic flux of the primary, secondary windings and center core \( \phi_{1l}, \phi_{2l}, \) and \( \phi_{ct} \) are defined as follows:

\[
\begin{align*}
\phi_{1l} &= \phi_{1c} + \phi_{1g1} - \phi_{21} \\
\phi_{2l} &= \phi_{2c} + \phi_{2g2} - \phi_{12} \\
\phi_{ct} &= \phi_{1c} + \phi_{2c}
\end{align*}
\]

where magnetic vectors of \( \phi_{12} \) and \( \phi_{21} \) are generally opposite directions.

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**Table 1. Design goal of the DAB converter.**

| Parameters   | Values | Parameters   | Values |
|--------------|--------|--------------|--------|
| \( V_s(V) \) | 20 V   | \( C_L \)   | 10 \cdot F |
| \( V_r(V) \) | 20 V   | \( P_L \)   | 20 W   |
| \( R_L \)    | 20 \Omega | \( f \) | 100 kHz |
| \( I_L \)    | 1 A    | \( \theta_{p, nom} \) | \pi/4 |

---
To establish magnetic characteristics of the proposed integrated transformer, an equivalent magnetic circuit of Fig. 6 is described in Fig. 7. Reluctance of $N_1$, $N_2$, and $L_0$ correspond to $l_1$, $l_2$, and $l_g$, respectively. As described in Figs. 6-7, it is almost impossible to analyze self-leakage magnetic flux $\phi_{lk1}$ and $\phi_{lk2}$ due to non-linear characteristics of the magnetic field distortion through the core and unexpected air path; hence, such self-leakage terms of $\phi_{lk1}$ and $\phi_{lk2}$ are not considered for theoretical analysis in this section. Then, self-inductance of the primary and secondary windings $L_1$ and $L_2$ can be derived from Fig. 7 if primary and secondary windings are symmetrically designed as follows:

$$L_1 = N_1\phi_1 \rightarrow L_1 = L_2 = \frac{N_1^2}{\mathcal{R}_{th}} = \frac{N_2^2}{\mathcal{R}_{th}} \quad (n = 1) \quad (15a)$$

where $\mathcal{R}_{th}$ is the equivalent magnetic reluctance of the transformer core.

From (15), magnetizing inductance $L_m$ and leakage inductance $L_{l1}(= L_{l2})$ can be straightforwardly derived based on the static magnetic circuit of Fig. 7, as follows:

$$L_m = \frac{g_{c1} + (1 - 1/\mu_r) g_{a}}{3g_{c1} + 4g_{c2} + (1 - 1/\mu_r) g_{a}} L_1 = k L_1 \quad (16a)$$

$$L_{l1} = \frac{g_{c1} + (1 - 1/\mu_r) g_{a}}{3g_{c1} + 4g_{c2} + (1 - 1/\mu_r) g_{a}} L_1 = (1 - k)L_1 \quad (16b)$$

$$k \equiv \frac{g_{c1} + (1 - 1/\mu_r) g_{a}}{3g_{c1} + 4g_{c2} + (1 - 1/\mu_r) g_{a}}. \quad (16c)$$

Based on the results of (15)-(16), $L_m$, $L_{l1}$, and $L_{l2}$ can be modulated by a center-core air gap $l_g$ and the number of turns $N_1$ and $N_2$. In Fig. 6, different coil winding positions may change the magnetic characteristics, i.e., leakage inductance. In this paper, it is assumed that the coil winding is positioned in the center of each core leg, as one of design examples. Although different coil winding positions are applied in the proposed transformer structure, the leakage inductance can be found by 3D FEM simulation analysis.

### B. SIMULATION VERIFICATION

To evaluate the proposed integrated transformer design, a Mn-Zn-type ferrite core made by TODAISU is used and material of the ferrite core adopted in this paper is PM7 [37]. Two shapes of the ferrite core, i.e., EER2834 and EER3019, are selected. According to the datasheet [37], for the magnetic flux area $A_{ct}$, EER3019 is 66% larger than EER2834. On the other hand, for the magnetic flux path $2(l_1 + l_2)$, EER2834 is 43% larger than EER3019. Thus, those two ferrite cores are so distinguished to prove the universality of the proposed transformer design methodology. Physical dimensions are summarized in Table 2. It is noted that the proposed design methodology for the integrated transformer can be utilized, regardless of the shapes of the transformer core. To identify magnetic characteristics of the proposed integrated transformer, a FEM simulation model is established, as shown in Fig. 8. Considering that AWG 17 wire ($= 1.04 \text{ mm}^2$) have 8.0A of ampacity, according to NFPA 70, 0.1mm/60EA lit wire is selected for primary and secondary coil windings, which are sufficient to carry 1 ~ 3A of coil current [38], [39]. The other wire thickness selection is possible if the selected thickness is enough to have ampacity for targeting load condition. This wire selection is fixed as one of design baseline, and the other wire selection cases are not considered due to plenty of design parameters in this paper. Therefore, the thickness of the primary and secondary coils is calculated as 1.0 mm.
for the shape of the square copper in the simulation model. To apply litz wire model in FEM 3D simulation condition as accurately as possible, Stranded-based-copper wire model, which modelled evenly distributed current density in current path area, is built, as shown in Fig. 8. Based on the simulation models of Fig. 8, the magnetic coupling coefficient between primary and secondary windings w.r.t. \( l_g \) is simulated and compared to the calculated results of (16), as shown in Fig. 9. As \( l_g \) increases, the coupling coefficient of simulation results becomes lower than that of calculation results. This means that the unexpected self-leakage inductance, i.e., \( \phi_{lk1} \) and \( \phi_{lk2} \) in Figs. 6-7, becomes dominant, and these values cannot be evaluated by theoretical analysis, as described in Fig. 6. Based on the results of Fig. 9, magnetizing and leakage inductance are shown in Fig. 10. A discrepancy between calculation and simulation results for magnetizing inductance is not a critical factor due to \( L_m \gg L_{l1} \) and \( L_{l2} \). Because a little discrepancy between calculation and simulation results for leakage inductance becomes severe to determine the performance of the DAB converters, effect of \( \phi_{lk1} \) and \( \phi_{lk2} \) can no longer be ignored in case of designing leakage inductance; hence, it is assumed that effects of \( \phi_{lk1} \) and \( \phi_{lk2} \) in Figs 6-7 are found by the FEM simulation analysis and reflected to the calculation results in order to determine leakage inductance for the proposed integrated transformer design throughout this paper.

Based on the simulation results of Figs. 9-10, the optimal number of turns, satisfying \( L_{lk1} = L_{lk2} = 9.35 \, \mu H \) for \( P_L = 20W \), for various center-core air gap can be calculated from (15)-(16), as shown in Fig. 11; hence, the optimum number of turns for primary and secondary winding exist in accordance with \( l_g \) for \( P_L = 20W \). The simulation results of magnetizing inductance w.r.t. \( l_g \) when \( N_1 = N_2 = 10 \) turns is shown Fig. 12, where \( L_{lk1} = L_{lk2} = 9.35 \, \mu H \) in this case. It is noteworthy that the value of \( L_m \) in Fig. 12 determines core loss of the proposed integrated transformer.
The core loss and copper loss for the proposed integrated transformer should be specified when the number of turns for primary and secondary windings \(N_1\) and \(N_2\) are decided. Hysteresis loss, which is the major source of core loss, can be represented in watt per unit volume by the following Steinmetz equation [35], [36]:

\[
P_{cv} = C_m x^y B_1^z \quad [W/m^3].
\] (17)

where \(C_m, x,\) and \(y\) are the coefficients, which are generally provided by the characteristics of the ferrite core, these values can be found by the power loss versus frequency and magnetic flux, according to the datasheet information [37]. Because transient flux change is not considered, steady-state condition based Steinmetz equation (17) is used in this paper.

Quantitative core loss can be assessed for the transformer core when the amount of core in the transformer is provided. Then, the core loss of the transformer can be analytically calculated from (17) and Fig. 6 as follows:

\[
P_{co} = \iiint P_{cv} \, dx \, dy \, dz = C_m x^y \left( \iiint_{U_1} B_{1r}^2 \, dx \, dy \, dz + \iiint_{U_2} B_{2r}^2 \, dx \, dy \, dz \right)
\] + \iiint_{U_{ct}} B_{c}^2 \, dx \, dy \, dz \quad \text{(18a)}

\[
U_{1r} = U_{2r} \equiv \frac{(l_{c1} + 2l_{c2})}{2} A_{cr}, \quad U_{ct} \equiv (l_{c1} - l_{g}) A_{cr}, \quad \text{(18b)}
\]

where \(U_{1r}, U_{2r}\) and \(U_{ct}\) are the core volumes that \(B_{1r}, B_{2r},\) and \(B_{c}\) are passing through the ferrite core, respectively, as described in Figs. 5-6 and (14). The values of \(B_{1r}, B_{2r},\) and \(B_{c}\) can be found by the FEM simulation analysis.

On the other hand, copper loss \(P_{cp}\) can be determined by \(I_1, I_2,\) and copper resistance \(r_{cp1}\) and \(r_{cp2}\) as follows:

\[
P_{cp} = I_1^2 r_{cp1} + I_2^2 r_{cp2}. \quad \text{(19)}
\]

where \(r_{cp1}\) and \(r_{cp2}\) include only copper loss terms and can be found by FEM simulation analysis, and \(I_1\) and \(I_2\) in (19) are rms values. It is noted that the copper type in the simulation condition is selected as a stranded type, which is closer to the real characteristics of the litz copper wire.

To evaluate \(P_{co}\) and \(P_{cp}\) in (18)-(19), the values of \(I_1, I_2,\) and \(\theta_d\) were simulated and compared to the calculation results, as shown in Fig. 13. Calculation results can be obtained by (11)-(13) and simulation results have been obtained by a PSIM simulation tool. When \(I_1\) becomes larger, magnitude difference between \(I_1\) and \(I_2\) in Fig. 13(a) and phase difference \(\theta_d\) in Fig. 13(b) become smaller due to the increment of \(L_m\) in Fig. 12. Based on the results of Fig. 13, core and copper losses of the proposed integrated transformer w.r.t. \(l_g\) are calculated and simulated, as shown in Fig. 14. The calculated results of the coil currents and phase difference have been compared with the PSIM simulation results in Fig. 13(a), and such calculated results have been reflected to the 3D FEM simulation results in Figs. 14-15 for core and copper loss analysis. In order to evaluate the core loss, core loss parameters, i.e., \(C_m, x,\) and \(y\) in (17)-(18), are selected: \(C_m = 8.5, x = 1.25,\) and \(y = 2.02\) for PM7 ferrite core material at 20°C of a core temperature, according to the datasheet [37]. The core loss and copper loss can be found by (18)-(19) based on the FEM simulation results of magnetic field \(B_{1r}, B_{2r}, B_{ct}\) and copper loss resistance \(r_{cp1}\) and \(r_{cp2}\).
When \( l_g \) increases, \( P_{co} \) decreases and \( P_{cp} \) increases, as shown in Fig. 14. From the results of Fig. 14, it is recommended to select large \( l_g \) for both ferrite core cases to achieve a minimum loss of the proposed integrated transformer.

To identify characteristic of core loss in the proposed integrated transformer, core loss resistance, including only core loss term \( r_{co} \) in Fig. 2, can be defined from Fig. 14 as follows:

\[
P_{co} = \frac{V_m^2}{r_{co}} \quad \Rightarrow \quad r_{co} = \frac{V_m^2}{P_{co}}
\]  

(20)

The calculation results of core loss resistance \( r_{co} \) based on the simulation results of Fig. 14 are shown in Fig. 15, whose results will be used to verify power efficiency of the proposed integrated transformer in the experimental verifications. The proposed design procedure of the integrated transformer described in this section is summarized in Fig. 16. After the optimal air-gap is determined, magnetic field level of the transformer core should be examined by the FEM simulation analysis. Because the saturation level of the magnetic field \( B_{sat} \) is usually around 0.3T [37], the average value of the magnetic field inside the ferrite core should be enough lower than 0.3T. To guarantee the stable operation of the proposed integrated transformer, the core temperature should be monitored when the proposed transformer is operated by experiment process. If the core temperature is over 40°C, then the volume of the transformer core should be increased so that the core temperature is enough low for thermal stable of the transformer core.
IV. EXPERIMENTAL VERIFICATIONS
The proposed integrated transformer design discussed in the previous sections has been evaluated by two experimental setups, as shown in Figs. 17-18. To compare efficiencies between conventional and proposed transformers, a conventional DAB converter including an inductor and a transformer was fabricated, as shown in Fig. 17. To fabricate an additional inductor in Fig. 17(a), the number of winding turns and air-gap of ferrite core are selected as 8 and 0.2mm, respectively, so that series inductance is designed to be 18.7 $\mu$H, which is equivalently the same value as the sum of leakage inductance in the proposed integrated transformers ($L_{l1} + L_{l2} = 18.7 \, \mu$H). In case of selecting the transformers, the EER2834 and EER3019 ferrite cores have been selected as one of design examples for both the conventional transformers and the proposed integrated transformers. As shown in Fig. 17(a), the primary and secondary coils are wound in center core, as conventional DAB converters are usually designed. On the other hand, the proposed integrated transformers have a center-core air gap, as shown in Fig. 18(a). The physical dimensions and material of the ferrite core adopted in the proposed integrated transformer are the same as the conventional transformers, except for a center-core air gap. The semiconductor switching devices for primary and secondary sides are selected as PRM3R7N10CTB, which are maximum 100V drain-source voltage and 110A drain current rating with 3.7$m\Omega$ turn-on resistance. The window utilizations of the conventional transformers for the EER2834 and EER3019 ferrite cores having the optimal number of turns are calculated as 38.6% and 72.5%, respectively. On the other hand, the window utilizations of the proposed integrated transformers for the EER2834 and EER3019 ferrite cores having the optimal number of turns are calculated as 12.3% and 26.1%, respectively, whose values are lower than the conventional one in the case of the window utilization of the transformer. The window utilization can be increased if the larger thickness of the copper wire is adopted. To provide appropriate DC voltage $V_s$ and load resistance $R_L$ in Fig. 1, a DC power supply in the source side and a DC electronic load in the load side were used. The duty cycle of the gate signals for primary and secondary inverters is 0.48 for ZVS operation of the switching devices, and the phase difference between primary and secondary inverters $\theta_p$ is modulated by MCU. 0.1mm/60EA litz wires for primary and secondary windings were used, as identified from previous sections.

To confirm electrical characteristics of the proposed integrated transformer, magnetizing inductance $L_m$ and copper loss resistance $r_{cp}$ were measured and compared to the simulation results, as shown in Figs. 19-20. Concerning to obtain the magnetizing inductance in Fig. 19, measuring inductance in two primary winding nodes when shorting the secondary nodes is ($L_{l1} + L_{l2}$); hence, $L_m$ is calculated by $L_1 - (L_{l1} + L_{l2})$. The copper loss resistance $r_{cp}$ in Fig. 20 has been obtained by measuring resistance at $f_s = 100$ kHz without the transformer ferrite core. It is noteworthy that all the parameters derived from this simulation analysis include the eddy loss effect depending on operating frequency. A little discrepancy of $L_m$ and $r_{cp}$ between simulation and experiment results in Figs. 19-20 may come from additional parasitic components in the experimental setup and imperfect simulation modeling.

To verify efficiencies of conventional DAB converters for two ferrite cores of EER2834 and EER3019 in Fig. 17, DC and AC efficiencies in Fig. 1 have been measured.
TABLE 3. Measured and selected parameters for 20W load power.

| EER2834 ferrite core | EER3019 ferrite core |
|----------------------|----------------------|
| $I_p$ [mm] | Calculated $N_1$ | Selected $N_1$ | Measured $k$ | Measured $L_m$ [$\mu$H] | $I_p$ [mm] | Calculated $N_1$ | Selected $N_1$ | Measured $k$ | Measured $L_m$ [$\mu$H] |
| 0.5 | 8.5 | 9 | 0.876 | 73.2 | 0.5 | 6.8 | 7 | 0.892 | 95.6 |
| 1.0 | 9.8 | 10 | 0.903 | 89.7 | 1.0 | 8.2 | 8 | 0.928 | 118.5 |
| 2.0 | 11.0 | 11 | 0.921 | 109.6 | 2.0 | 9.8 | 10 | 0.947 | 185.9 |
| 4.0 | 11.9 | 12 | 0.930 | 138.8 | 4.0 | 11.2 | 11 | 0.955 | 238.1 |
| 10.0 | 12.9 | 13 | 0.934 | 162.5 | 10.0 | 12.0 | 12 | 0.960 | 286.3 |

as follows:

$$\eta_{dc} \equiv \frac{P_L}{P_s}, \quad \eta_{ac} \equiv \frac{P_2}{P_1}.$$  \hspace{1cm} (21)

where DC source power $P_s$ and DC load power $P_L$ were measured by a DC power supply and a DC electronic load, respectively. AC powers of $P_1$ and $P_2$ were measured by YOKOGAWA WT310 AC power meters and Lecroy 610Zi oscilloscope: the power measurement results of AC power meters and oscilloscopes were compared to surely confirm the power efficiency for this low power delivery condition. Those measurement methods provide the number to two decimal places for obtaining exact efficiency results, and those experimental results are so reliable to verify the proposed transformer.

Experimental results of the measured DC and AC efficiencies for the EER2834 case are shown in Fig. 21. Only three different experimental results, e.g., $N_1 = 8, 16,$ and 28 turns, have been shown as one of examples, and the experimental results for the EER3019 are omitted due to the similar tendency between the EER2834 and EER3019 cases. As identified from Fig. 21, there is an optimum $N_1$ for the highest efficiency, i.e., $N_1 = 16$ turns in this case. Although the design process to find $N_1 = N_{1,op} = 16$ for the high efficiency is not described and only verified by experiments for the conventional transformers in this paper, the experimental results of Fig. 21 imply that lower values of $N_1$ causes a large core loss and larger values of $N_1$ causes a large copper loss. As the same principle of Fig. 21, optimal number of turns for EER3019 case $N_{1,op}$ is found to be 12, whose value will be used to compared with the efficiency of the proposed integrated transformer.

To verify efficiencies of the proposed integrated transformer, the optimal number of turns $N_1$ ($= N_{2,op}$) for various air-gap cases, satisfying $L_{l1} = L_{l2} = 9.35 \, \mu$H, have been already calculated, as identified from Fig. 11; hence, integer values of $N_1$ is selected based on the results of Fig. 11. Major electrical characteristics of the proposed integrated transformers are summarized in Table 3, where the measured $k$ and $L_m$ are in good agreements with simulation results of Figs. 9 and 12. The load power $P_L$, primary and secondary currents $I_1$ and $I_2$ were measured and compared to calculation of the number to two decimal places for obtaining exact efficiency results, and those experimental results are so reliable to verify the proposed transformer.

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and simulation results, as shown in Fig. 22, where only \( l_g = 2\) mm case is shown due to same tendency of those characteristics for the other \( l_g \) cases. All the parasitic components are not included in the calculation results and the simulation results were performed by a PSIM simulation tool in Fig. 22. The measured load power is 18.5W, which is slightly lower than simulation result of 19.1W. A little discrepancy between simulation and experiment results in Fig. 22 may come from unexpected parasitic components in the experimental setup.

The tendencies of \( P_L, I_1 \) and \( I_2 \) w.r.t. \( \theta_p \) are the same for the other \( l_g \) cases and two ferrite cores, although not described in this paper.

Measured DC and AC efficiencies for the EER2834 and EER3019 cases are shown in Figs. 23-24, where \( l_g = 2\) mm case are described as one of examples. The integrated transformer model can be represented by inductance and resistance, as described in Fig. 2, and reflected to the PSIM simulation results based on the parameters of \( L_{m}, L_{11}, L_{12}, r_{cp1}, r_{cp2}, \) and \( r_{co} \), whose values were already specified in the previous sections. The efficiencies of conventional one are higher than those of \( l_g = 0.5\) mm, whose results are in good agreement with Fig. 14. At the nominal operating point of \( \theta_p \) in Figs. 23-24, inductor loss and transformer loss for the conventional DAB magnetic components in Fig. 17(b) occupy roughly 15% \sim 20\% and 80% \sim 85\%, respectively. The waveforms of primary and secondary voltages and currents are shown in Fig. 25, which correspond to Fig. 3 for the operating principle of the proposed DAB converters. Measured DC and AC efficiencies at \( \theta_p = \pi/4 (=45\)°) and \( \theta_p = \pi/2 (=90\)°) are summarized in Table 4. It is noteworthy that the larger the air-gap \( l_g \) is, the larger the efficiency \( \eta_{dc} \) is, as shown in Table 4, whose results are well matched with the simulation results of Fig. 14. From the efficiency results in Table 4, optimal \( l_g \) can be selected as 20mm and 10mm for the EER2834 and EER3019, respectively.

After optimal \( l_g \) and optimal number of turns \( N_{1,op} \) & \( N_{2,op} \) are determined, the magnetic field inside the transformer core should be identified, according to the proposed design procedure, as shown in Fig. 26. To evaluate the magnetic flux density inside the transformer core, the FEM simulation analysis has been carried out, as shown in Fig. 27.
The average values of the magnetic field for the EER2834 and EER3019, operating at an optimal point were 75.9 mT and 58.4 mT, respectively, whose values are well below the saturation level of 300 mT. The highest temperatures of the transformer cores for both the EER2834 and EER3019 in the steady state were measured below 35 °C at the ambient temperature of 25 °C. These temperature results imply that the proposed integrated transformers are well designed for the thermal stable operation by the proposed design methodology.

To comparatively evaluate the transformer loss, the FEM simulation, the PSIM simulation, and experimental results of the transformer loss have been compared w.r.t. $l_g$, as shown in Fig. 28. The FEM simulation results are based on the Steinmetz equation for core loss, i.e., Fig. 14. On the other hand, according to the proposed design procedure, the core loss and copper loss resistance can be found from (19)-(20). From these obtained calculation results, the transformer loss can be simulated based on the equivalent circuit of Fig. 2. A little discrepancy between the experiment results and the PSIM simulation results may come from additional copper wires in order to connect each circuit components and to measure the efficiency by measurement devices. A little discrepancy between the PSIM simulation and the FEM simulation may come from the core loss modeling methods. In case of $r_{co}$ in Fig. 2 for the PSIM simulation, $r_{co}$ is assumed to be constant if the transformer windings are decided; in this case, the core loss is represented by $\frac{V^2_{in}}{r_{co}}$, as identified from (20). On the other hand, the Steinmetz equation model for the FEM simulation depends on the core temperature, the switching frequency, and the magnetic field. In particular, exponential values of $x$ and $y$ in (17)-(18) varies at different operating
FIGURE 24. Measured efficiencies w.r.t. $\theta_p$ for the EER3019 case to compare with the conventional DAB converter having $N_1 = 12$ turns.

FIGURE 25. Experimental waveforms of $v_1$, $v_2$, $i_1$ and $i_2$ when $l_g = 2\text{mm}$.

FIGURE 26. Magnetic flux density of the proposed integrated transformers to confirm the core saturation level and core heating.

conditions. Nevertheless, the tendencies of these three cases w.r.t. $l_g$ in Fig. 28 are the same, which guarantees that optimal number of turns $N_{1,\text{op}}$ and optimal value of $l_g$ can be found. From the simulation and experimental results in Fig. 28, it is found that the larger value of $l_g$ is recommended for the minimum transformer loss in case of the presented design goal of the DAB converter in Table 1 in this paper. By the same design principle of the proposed integrated transformer, the optimal $l_g$ and $N_{1,\text{op}}$ can be found for various power design requirements. Therefore, the high power density for
magnetic components in the DAB converter can be possible by the proposed integrated transformer structures and design procedures.

V. CONCLUSION

The proposed integrated transformer design methodology presented this paper has been proven to be a practical solution to design compact DAB converters. The leakage inductance in the proposed integrated transformer can be appropriately modulated by selecting the optimal number of turns and the center-core air gap. To determine the optimum number of turns, satisfying the design goal of the DAB converter, the FEM simulation analysis based transformer loss analysis is carried out to analyze core and copper loss of the proposed integrated transformer. Contrary to the previous literature that uses both side legs and center leg [34], the proposed idea modulates only a center leg so that the leakage inductance can be intuitively controlled. Based on this design approach, appropriate leakage inductance and optimal winding turns w.r.t. the air-gap can be simply and easily obtained. The efficiencies of the DAB converters for 20W load power were evaluated and compared to the conventional DAB converters. The total volumes of the proposed integrated transformer are 10.9 cm$^3$ and 12.0 cm$^3$ for the EER2834 and EER3019, respectively, which are the 40% volume reduction compared to the conventional inductor and transformer set in this paper. Although 20 W power rating case has been verified in this paper, the proposed integrated transformer can be also applicable to the high power applications, according to the design procedure in Fig. 16. By virtue of the proposed design procedure of the integrated transformer, the DAB converters can exclude a bulky and cumbersome inductor and guarantee the thermal stable operation of the proposed integrated transformer.

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