Integrated Magnetic Component of a Transformer and a Magnetically Coupled Inductor for a Three-Port DC-DC Converter

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This paper discusses a design method for a proposed integrated magnetic component for an isolated bidirectional three-port DC-DC converter (TPC). TPC comprises a dual active bridge converter (DAB) and a non-isolated bidirectional DC-DC converter (NBC); each converter is independently controlled with a transformer and a magnetically coupled inductor. To reduce the size of the magnetic components, an integrated magnetic component that can integrate a magnetically coupled inductor and a transformer is implemented. A 750-W magnetically integrated TPC prototype was constructed and tested to validate the operation. The experimental results show that the efficiency of the integrated TPC is above 90% for the entire output power range, which is nearly equal to that of the conventional magnetic component. As a result, the proposed component was 10% smaller than the conventional magnetic components, and the overall size of the integrated TPC was 33% smaller than that of the conventional one.

Keywords: magnetic integration, multiport, bi-directional, galvanic isolation, DC48V, electric vehicle

1. Background

Due to the strengthening of CO₂ emission regulations in each country, the mainstream in the powertrain of vehicles is expected to shift from gasoline-powered vehicles to electrically driven vehicles such as hybrid vehicles (HV) and electric vehicles (EV) (1). As a consequence, the introduction of on-board DC-DC converters and AC-DC converters is progressing. Non-isolated DC-DC converter for inverter voltage boosting (2), isolated AC-DC converters for vehicle charging circuits (18,19), and isolated DC-DC converter for 12 V auxiliary systems (20) can be given as examples. On the other hand, in Europe, the introduction of 48 V auxiliary equipment is being studied as one of the measures towards CO₂ emission control (10). If a system requiring multiple inputs and multiple outputs, such as a 12 V/48 V mixing system, is realized using a conventional power conversion circuit, the number of conversion circuits, their size and cost will increase. However, multiport converter technology is being developed to reduce the size and cost by sharing circuit components (9)-(14). In this paper, we propose a technology (15)-(16) that integrates the two magnetic components used in the three-port DC-DC converter (TPC) discussed in Reference (13) and (14). Instances of the usage of magnetic component integration technology in the inductor for filters (17) can be seen as early as in the 1930s. In the power conversion circuits, the magnetic component integration technology has been addressed starting from the technology to integrate the transformer and reactor of a forward converter (18) reported in the 1970s, to the Ćuk converters (19), push-pull converters (20), high-boost tapped inductor circuits (21), LLC circuits (22), and various other circuits (23)-(27). In this paper, we have used the above knowledge to study the technology for integrating magnetic components in conventional TPCs. In Chapter 2, we describe the operating principle of the conventional TPC, and in Chapter 3, we describe the operating principle and design method of the proposed integrated magnetic component.

In Chapter 4, we compare the experimental results of a circuit using the integrated magnetic component and the conventional circuit, and show that for the same efficiency, the core volume of the magnetic component can be reduced by 10%, and the circuit volume can be reduced by 33%. In Chapter 5, we present the conclusion.

2. Three-Port DC-DC Converter

Figure 1 shows a conventional TPC that can realize two circuit operations with a single circuit, using a magnetic coupling component.

Conventional TPCs have a center tap transformer and a magnetic coupling reactor as the magnetic coupling components, which are connected to the center points M<sub>α</sub>, M<sub>β</sub>, M<sub>γ</sub>,...
and \( M_{ph} \) of a full-bridge circuit. Taking the DC voltage parts of the two full-bridge circuits as Port A and Port B, and the DC voltage part obtained from the center-tap of the transformer as Port C, one circuit is made to have three DC ports, by which it becomes a power conversion circuit capable of bidirectional power transmission between the ports. When focusing on Port A-B, the conventional TPC can be considered as a bidirectional isolated DC-DC converter (Dual active bridge converter: DAB) as shown in Fig. 2. The power transmitted \( P_O \) due to DAB operation is expressed as the following equation, where \( N \) is the turns ratio of the transformer, \( f_{sw} \) is the switching frequency, \( \psi \) is the phase difference between the rectangular wave voltages generated from the primary and secondary side full-bridge circuits, and \( \delta \) is the ON ratio of the lower arm of the full-bridge circuit. The transmitted power between ports A and B can be controlled by the phase difference \( \psi \). \(^{(18\text{-}40)} \)

\[
P_O = \frac{2N V_A V_B}{f_{sw} L_{eq}} \left( 2\pi - \delta - \left| \frac{\psi}{2} \right| \right) \quad \text{................. (1)}
\]

Here, \( L_{eq} \) is the total leakage inductance when the magnetic coupling component is converted to a T-type equivalent circuit. If \( k_L \) and \( k_T \) denote the coupling coefficients of the magnetic coupling reactor and the center tap transformer respectively, and \( L_n \) and \( L_1 \) denote the self-inductance of the magnetic coupling reactor and the primary of the transformer respectively, \( L_{eq} \) is given by the following equation.

\[
L_{eq} = 2N^2 \left[ (1 - k_L)L_n + (1 - k_T)L_1 \right] \quad \text{................. (2)}
\]

When focusing between the ports A and C, conventional TPC can be regarded as a non-isolated bidirectional DC-DC converter (NBC) as shown in Fig. 3. The voltage ratio between \( V_A \) and \( V_C \) can be expressed in terms of switching ratio \( \delta \) by the following equation.

\[
V_C = \left( 1 - \frac{\delta}{2\pi} \right) V_A \quad \text{......................... (3)}
\]

Therefore, the voltage ratio between ports A and C can be controlled by the duty ratio \( \delta \).

Figure 4 shows the direction of change in magnetic flux in the core of each magnetic coupling component according to the direction of change in current.

Figure 4(a) shows the change in the reverse-phase sequence current \( i_{D0} \) (= \( i_n - i_s \))/2 related to the DAB operation mode. The magnetic coupling reactor and the center tap transformer operate as a small inductance \( L_{eq} \) shown in Equation (2). Since the transmitted power \( P_O \) due to the DAB operation shown in Equation (1) is inversely proportional to the total leakage inductance \( L_{eq} \), the transmitted power of DAB operation can be designed to have a large value using the small inductance of the magnetic coupling reactor. Further, at this time, due to the excitation of the center tap transformer, isolated power is transmitted through the transformer winding. Figure 4(b) shows the change in the in-phase current \( i_c \) (= \( i_n + i_s \)) related to the NBC operation mode. The center tap transformer does not function, as the magnetic flux inside the core is cancelled. At this time, the magnetic flux of the magnetic coupling reactor reinforces and operates as the necessary reactor component \( L_n \) in the NBC operation mode represented by the following equation.

\[
L_n = \frac{(1 + k_L)L_{eq}}{2} \quad \text{................. (4)}
\]

Since the inductance of the magnetic coupling reactor due to the strengthening of the magnetic flux can be designed to be large, the ripple current flowing through the capacitor connected to Port C can be decreased, and hence, the capacitor can be designed to be small.

Figure 5 shows the ideal operation waveform when the power is transmitted from Port B to Port C. Since the reactor currents \( i_n \) and \( i_s \) have a DC offset component due to NBC operation, a DC offset component occurs in the reactor core flux \( \phi_{Tr} \). However, since the DC offset component of the reactor current flows only from the center tap to Port C, no DC offset component occurs in the transformer core magnetic flux \( \phi_{Tr} \).

Although conventional TPCs can reduce the number of components compared to a system configured with DAB and NBC by circuit integration, the problem is that since two magnetic components are required, the reduction in circuit size is hindered by the component size and the dead space between the magnetic components.

### 3. Magnetic Component Integrated TPC

#### 3.1 Operating Principle

Figure 6 shows the circuit diagram of a TPC using the proposed integrated magnetic component. The primary winding \( n_1 \), the secondary winding \( n_2 \), and the primary winding \( n_3 \), coming out of the center tap are magnetically coupled, and the coupling coefficient
between the windings \( n_1 \) and \( n_2 \) is \( k_X \).

Figure 7 shows the change in magnetic flux of the integrated magnetic component for each operation mode, and Fig. 8 shows the equivalent circuit for each operation mode. In the DAB operation mode shown in Fig. 7(a), the change in magnetic flux occurs in the outer peripheral path of the core. A self-induced electromotive force is generated in the primary winding \( n_1 \) wound on both ends of the three-legged core, and a mutually induced electromotive force is generated in the secondary winding \( n_2 \). Since the magnetic flux is cancelled in the center leg, no mutually induced electromotive force is generated in the center leg winding \( n_L \), and the center leg does not operate as a reactor. Therefore, for the change in the reverse-phase current, the integrated magnetic component operates as the small inductances \( N_2(1-k_X)L_1 \) and \( (1-k_X)L_2 \), and the transformer excitation inductance \( k_XL_2 \) necessary for the DAB operation shown in Fig. 8(a). In the NBC operation mode shown in Fig. 7(b), the change in magnetic flux occurs along the path through the center leg. At this time, although a mutually induced electromotive force is generated in the secondary winding \( n_2 \) wound around both legs, since the direction is reversed between the two secondary windings \( n_2 \), it does not operate as a transformer. On the other hand, since a self-induced electromotive force is generated in the primary windings \( n_1 \) and \( n_L \), they act as reactors. Therefore, when the in-phase current changes, the integrated magnetic component operates as the reactor component necessary for NBC operation shown in Fig. 8(b) by the primary windings \( n_1 \) and \( n_L \). Thus, for the two operating modes, the integrated magnetic component functions as a transformer and a reactor without interfering with each other.

While the trapezoidal current due to the DAB operation mode flows through both the transformer winding and the magnetically coupled reactor winding in the conventional circuit, in the integrated magnetic component, since the winding \( n_2 \) functions as both the transformer and reactor winding, and \( n_L \) functions as a reactor winding in a path through which the trapezoidal current does not flow, the winding loss in the DAB operation mode can be reduced. Further, the dead space between the magnetic components can be reduced compared to separate magnetic components.

Figure 9 shows the analysis model of the integrated magnetic component. Figure 9(a) shows the parameter positions of the integrated magnetic component, and Fig. 9(b) shows the equivalent magnetic circuit model of the integrated magnetic component. \( A_s, A_c, R_s \), and \( R_c \) represent the core cross-sectional area and magnetoresistance of the outer legs and the center leg, respectively.

If the magnetic fluxes \( \phi_1 \) and \( \phi_2 \) are defined by the loop shown in Fig. 9(b), the magnetic circuit equation is as
First, we derive a theoretical formula relating to the transformer operation using Equation (5). By taking the sum of the first and second rows and differentiating both sides, Equation (5) can be rearranged as follows:

$$\frac{d\phi_1}{dt} + \frac{d\phi_2}{dt} = n_1 \frac{d}{dt}(i_a - i_b) - n_2 \frac{d}{dt}(i_a - i_b)$$

Applying Faraday’s law of electromagnetic induction to the terminals $M_a$-$M_b$ and $M_a$-$M_c$, respectively, it can be expressed as

$$\nu_a - \nu_b = -n_1 \frac{d\phi_1}{dt} - n_2 \frac{d\phi_2}{dt}$$

$$\nu_b - \nu_c = -n_1 \frac{d\phi_1}{dt} - n_2 \frac{d\phi_2}{dt}$$

where $\nu_a$, $\nu_b$, $\nu_c$, and $\nu_d$ are the electrical potential at $M_a$, $M_b$, $M_c$, and $M_d$, respectively.

In order to focus only on the inductance component of the reverse-phase current change, we assume $(i_a - i_b)/2 = i_1$ and $(i_a - i_b)/2 = i_2$. Substituting Equation (6) into Equations (7) and (8), we get,

$$\nu_a - \nu_b = -n_1 \frac{2i_1^2}{L_1} \frac{dt}{dt} + n_1 \frac{2n_1 n_2 i_2}{L_1} \frac{dt}{dt}$$

$$\nu_b - \nu_c = -n_1 \frac{2i_1^2}{L_1} \frac{dt}{dt} - n_2 \frac{2i_2^2}{L_2} \frac{dt}{dt}$$

According to Kirchhoff’s second law, Equations (9) and (10) can also be written as follows.

$$\nu_a - \nu_c = -L_1 \frac{di_1}{dt} + k_X \sqrt{L_1 L_2} \frac{di_2}{dt}$$

$$\nu_b - \nu_c = -k_X \sqrt{L_1 L_2} \frac{di_1}{dt} + L_2 \frac{di_2}{dt}$$

Here, $L_1$ and $L_2$ denote the self-inductance of the primary winding $n_1$ and secondary winding $n_2$, respectively. Comparing Equations (9) and (10) with Equations (11) and (12), $L_1$ and $L_2$ are obtained as follows.

$$L_1 = \frac{2m_1^2}{R_s}$$

$$L_2 = \frac{2m_2^2}{R_s}$$

In the proposed integrated magnetic component, since the leakage inductance component $2(1-k_X)$ generated in the conventional magnetic coupling reactor, is eliminated, the leakage inductance component $L_{eq}$ used in the DAB operation shown in Equation (2) becomes as follows.

$$L_{eq} = 2(1-k_X)N^2 L_1$$

Next, we derive the theoretical expressions relating to the reactor operation. Applying Faraday’s law of electromagnetic induction between the terminals Mu-C and terminals Mv-C respectively, it can be expressed as,

$$\nu_a - \nu_c = n_1 \frac{d\phi_1}{dt} - n_2 \frac{d\phi_2}{dt}$$

$$\nu_b - \nu_c = -n_1 \frac{d\phi_1}{dt} - n_2 \frac{d\phi_2}{dt}$$

where $\nu_c$ is the electrical potential at Port C.

To focus only on the inductance component of the change in the in-phase current, we take

$$\nu_a = \nu_b = 0$$

$$i_a = i_b = i_c$$

Hence, it can be consolidated as,

$$\nu_c = \frac{(n_1 + 2n_2)^2}{2(R_s + 2R_c)} \frac{di_c}{dt}$$

By Kirchhoff’s second law, Equation (18) can also be written as follows.

$$\nu_c = \frac{L_b \frac{di_c}{dt}}{dt}$$

$L_b$ denotes the inductance that functions as a reactor for NBC operation. By comparing Equations (18) and (19), $L_b$ can be expressed as follows.

$$L_b = \frac{(n_1 + 2n_2)^2}{2(R_s + 2R_c)}$$

Since the numerator of Equation (20) includes the windings of the two outer legs $n_1$ and the winding of the center leg $n_2$, it can be seen that, while the separate magnetic component creates $L_b$ only with the winding of the magnetic coupling reactor, the integrated magnetic component creates $L_b$ with both $n_1$ and $n_2$.

Hence, it can be designed to have a reduced core volume compared to the conventional separate core.

Further, under the conditions of Equation (17), from Equation (11), we get,

$$\frac{di_1}{dt} = \frac{di_2}{dt} = 0$$

Substituting Equation (21) in Equation (12), we get

$$\nu_a - \nu_c = 0$$

Thus, in the NBC operation mode, no induced electromotive force occurs.
The voltage is generated in the secondary winding, indicating that the integrated magnetic component operates only as a reactor. From the above, since $k_X$ does not influence $L_b$, but contributes only to the change in $L_{eq}$, it is possible to design $L_{eq}$ and $L_b$ independently using $k_X$.

Next, we derive a theoretical formula for the magnetic flux density generated inside the core, which is required for designing the integrated magnetic component. Using Equation (5), the magnetic flux density generated at the center leg of the three-legged core is obtained as follows.

$$\phi_1 - \phi_2 = \frac{n_1 (L_1 + L_2) + 2n_2 I_c}{R_s + 2R_c} \cdots (23)$$

Therefore, the saturation magnetic flux density of the core material used is $B_{SAT}$, the maximum magnetic flux density of the center leg is $B_{c,\text{MAX}}$, and the maximum current value of Port C is $I_{c,\text{MAX}}$, the following equation must be satisfied inside the core of the center leg.

$$A_c B_{SAT} < A_c B_{c,\text{MAX}} = \frac{(n_1 + 2n_2 I_{c,\text{MAX}})}{R_s + 2R_c} \cdots (24)$$

On the other hand, the condition for $B_{SAT}$ and $I_{c,\text{MAX}}$ at the two outer legs of the three-legged core can be expressed as follows, where the maximum flux density at both the outer legs of the core is $B_{s,\text{MAX}}$.

$$A_b B_{SAT} < A_b B_{s,\text{MAX}} = \frac{T_{\text{ON}}}{4n_1} V_b + \frac{n_1 + 2n_2}{2(R_s + 2R_c)} I_{c,\text{MAX}} \cdots (25)$$

When $\delta < \pi$, it becomes,

$$T_{\text{ON}} = \frac{2\pi - \delta}{2\pi f_s} \cdots (26)$$

$$T_{\text{ON}} = \frac{\delta}{2\pi f_s} \cdots (27)$$

Figure 10 shows the relationship between $L_b$ and $L_1$ with respect to $A_c$, plotted using Equations (13) and (20). It shows the $L_b$ values for three types of $A_c$ values when the number of turns is set to $n_1 = 2$ and $n_2 = 1$. $L_b$ increases linearly with $A_c$, increases sharply with $A_c$, in the range of $A_c < A_c / 2$, and has almost no sensitivity in the range of $A_c > A_c / 2$. On the other hand, $L_1$ increases linearly with $A_c$, but has no sensitivity with respect to $A_c$, as seen from Equation (13). Therefore, $L_b$ and $L_1$ can be independently designed by adjusting $A_c$ and $A_c$.

![Figure 10. Relationship curves of $L_b$, $L_1$, and $A_c$.](image)

### 3.2 Design of the Core Shape of the Magnetic Component

This section shows a core shape design method when the winding is a rectangular wire and a copper wire. It should be noted that, for simplicity, only the DAB operation mode current is considered for the winding copper loss.

Figure 11 shows the definition of each dimension of the integrated magnetic component. Figure 11(a) is the dimensional drawing of the core part of the integrated magnetic component. The length of the center leg is taken as $l_c$, the length of the two outer legs is $l_t$, the depth of the core is $d$, the gap length of the center leg is $l_g$, width of the core window is $w$, and its height is $h$. As the distance between the primary and secondary windings increases, the core window height $h$ increases, and as the winding width increases, the core window width $w$ increases. The core depth $d$ is determined from the viewpoint of the core volume $V_{\text{integ}}$ and copper loss of the primary winding $P_{\text{integ}}$. Using the above variables, the core size $V_{\text{integ}}$ is expressed by the following equation.

$$V_{\text{integ}} = d \left( h + \frac{2A_c}{d} \right) \left( A_c + \frac{2A_c}{d} + 2w \right) - 2d \cdot w \cdot h - A_c l_g \cdots (28)$$

![Figure 11. Dimensions of the integrated magnetic component](image)

Figure 11(b) is the dimensional diagram of the winding part of the integrated magnetic component. Assuming the winding width of the integrated magnetic components as $w_{\text{integ}}$, the winding lengths of conventional components as $l_{\text{conv}}$, the winding width of the integrated magnetic components as $w_{\text{conv}}$, and the sum of the AC resistance losses of the transformer winding and reactor winding on the primary side of the conventional TPC for trapezoidal current as $P_{\text{conv}}$, the sum of the AC resistance loss in the primary winding $n_1$ of the integrated magnetic component $P_{\text{integ}}$ is expressed by the following equation.

$$P_{\text{integ}} = 4n_1 l_{\text{conv}} w_{\text{conv}} \left( \frac{d + A_c}{d} + 2w_{\text{integ}} \right) P_{\text{conv}} \cdots (29)$$

However, $P_{\text{integ}}$ considers only the change in the AC resistance with respect to the change in the winding length from the winding loss of the conventional separate magnetic components $P_{\text{conv}}$, and the influence of the proximity effect between the windings is ignored.
Figure 12 shows the variation of the core volume \( V_{\text{integ}} \) and the primary winding copper loss \( P_{\text{c, integ}} \) with respect to the core depth \( d \). The magnetic component parameters used in the plot are shown in Table 1. Although \( V_{\text{integ}} \) decreases as the core depth \( d \) increases, since the cross-sectional shape of the core becomes elongated, winding length increases, and the winding copper loss increases. Figure 12(a) is a relational diagram of \( V_{\text{integ}} \) to \( d \), plotted using Equation (28). Since the size of the core window is fixed, \( V_{\text{integ}} \) decreases as \( d \) increases, as the core cross-sectional area is kept constant. The total core volume \( V_{\text{conv}} \) for a separate core designed with the same maximum magnetic flux density and number of turns is also shown.

If the depth \( d \) of the integrated magnetic component is more than 30 mm, a core volume reduction effect can be obtained. Figure 12(b) is a plot of the winding loss \( P_{\text{c, integ}} \) versus core depth \( d \) plotted using Equation (29). As the core depth \( d \) is increased from 0, the winding loss \( P_{\text{c, integ}} \) decreases sharply, reaches a minimum at a point where the winding becomes a square, and then gradually increases. The figure also shows the primary-side conventional winding loss \( P_{\text{c, conv}} \). If the depth \( d \) of the integrated magnetic component is between 5 mm and 65 mm, the effect of winding copper loss reduction can be obtained. From Fig. 12, it can be seen that if the depth \( d \) of the magnetic component is between about 30 mm and 65 mm, the winding loss can be suppressed while reducing the volume, as compared to the conventional magnetic component. In the prototype of this paper, depth \( d \approx 37 \) mm was selected as the depth that can reduce the winding loss while reducing \( V_{\text{integ}} \), compared to the conventional core volume.

### Table 1. Parameters of magnetic component for calculated curves

<table>
<thead>
<tr>
<th>Description</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Height of core window</td>
<td>( h )</td>
<td>5.0 mm</td>
</tr>
<tr>
<td>Width of core window</td>
<td>( w )</td>
<td>6.4 mm</td>
</tr>
<tr>
<td>Length of core gap</td>
<td>( l_g )</td>
<td>0.6 mm</td>
</tr>
<tr>
<td>Cross-sectional area of core</td>
<td>( A_c )</td>
<td>97 mm²</td>
</tr>
<tr>
<td>Cross-sectional area of core</td>
<td>( A_1 )</td>
<td>223 mm²</td>
</tr>
<tr>
<td>Total size of conventional</td>
<td>( V_{\text{conv}} )</td>
<td>16.6 cm³</td>
</tr>
<tr>
<td>Chosen total size of proposed core</td>
<td>( V_{\text{integ}} )</td>
<td>14.9 cm³</td>
</tr>
<tr>
<td>Total resistance of conventional components at 175 kHz</td>
<td>( R_{\text{conv}} )</td>
<td>16.9 mΩ</td>
</tr>
<tr>
<td>Winding length of conventional component</td>
<td>( l_{\text{conv}} )</td>
<td>437 mm</td>
</tr>
<tr>
<td>Width of conventional windings</td>
<td>( w_{\text{conv}} )</td>
<td>7.0 mm</td>
</tr>
<tr>
<td>Width of proposed windings</td>
<td>( w_{\text{integ}} )</td>
<td>5.0 mm</td>
</tr>
</tbody>
</table>

### 3.3 Examination of Winding Configuration

In the proposed integrated TPC, the leakage inductance component \( L_{\text{eq}} \) required for DAB operation, which was adjusted using the coupling coefficient \( k_{\text{c}} \) of the magnetic coupling reactor previously, must be adjusted with the coupling coefficient \( k_X \) of the primary winding \( n_1 \) and the secondary winding \( n_2 \). An effective means to obtain the desired \( L_{\text{eq}} \) value by reducing \( k_X \) is to increase the distance between the primary winding \( n_1 \) and the secondary winding \( n_2 \). However, if the distance between the windings is simply increased, the height of the core will increase, which will result in the decrease of the size reduction effect of magnetic component integration.

Therefore, in this paper, secondary winding \( n_2 \) of the integrated magnetic component was wound perpendicular to the primary winding \( n_1 \) to reduce \( k_X \), and by simulation, it was confirmed that the required \( L_{\text{eq}} \) value was obtained.

Figure 13 shows the cross-sectional view of the three-dimensional model of the integrated magnetic component and the separate magnetic component. Table 2 shows each parameter. Using this model, the leakage inductance \( L_{\text{eq}} \) of the integrated magnetic component and separate magnetic component was compared by three-dimensional electromagnetic field simulation. The total leakage inductance of the integrated magnetic component shown in Fig. 13(a) is 3.2 μH, compared to the total leakage inductance of the separate magnetic component shown in Fig. 13(b) and (c). Table 2 shows the parameters used in the simulation.

### Table 2. Simulation parameters of magnetic components

<table>
<thead>
<tr>
<th>Description</th>
<th>Symbol</th>
<th>Conventional</th>
<th>Integrated</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transformer turn number</td>
<td>( n_1, n_2 )</td>
<td>1, 4</td>
<td>-</td>
</tr>
<tr>
<td>Inductor turn number</td>
<td>( n_{\text{mut}}, n_{\text{coil}} )</td>
<td>2, 2</td>
<td>-</td>
</tr>
<tr>
<td>Integrated magnetic component turn number</td>
<td>( n_1, n_2, n_{\text{mu}} )</td>
<td>-</td>
<td>1, 4, 2</td>
</tr>
<tr>
<td>Effective cross section</td>
<td>( A_c, A_1 )</td>
<td>170 mm²</td>
<td>97 mm², 223 mm²</td>
</tr>
<tr>
<td>Gap length</td>
<td>( l_{\text{gap}} )</td>
<td>0.5 mm</td>
<td>0.6 mm</td>
</tr>
<tr>
<td>Window size of core</td>
<td>( w \times h )</td>
<td>-</td>
<td>6.4 \times 5.0 mm</td>
</tr>
<tr>
<td>Depth of core</td>
<td>( d )</td>
<td>37.2 mm</td>
<td>-</td>
</tr>
<tr>
<td>Core size of transformer</td>
<td>-</td>
<td>8.3 cm²</td>
<td>-</td>
</tr>
<tr>
<td>Maximum flux density</td>
<td>( B_{\text{MAX}} )</td>
<td>118 mT</td>
<td>318 mT</td>
</tr>
<tr>
<td>Inductor : 318 mT</td>
<td>-</td>
<td>109 mT</td>
<td>309 mT</td>
</tr>
<tr>
<td>Core size of inductor</td>
<td>-</td>
<td>8.3 cm²</td>
<td>-</td>
</tr>
<tr>
<td>Total core size</td>
<td>( V_{\text{conv}}, V_{\text{integ}} )</td>
<td>16.6 cm³</td>
<td>14.9 cm³</td>
</tr>
</tbody>
</table>
which is found to be equivalent to the sum (3.0 µH) of the leakage inductances of the magnetic coupling reactor model shown in Fig. 13(b) and the center tap transformer model shown in Fig. 13(c). Hence, in the prototype, we adopted a winding configuration where the secondary winding was wound perpendicular to the primary winding. It should be noted that if a leakage inductance larger than the value obtained by the vertical winding is required, it is necessary to further increase the core height of the integrated magnetic component and increase the distance between the primary and the secondary windings.

3.4 Trial Calculation of the Items in the Magnetic Component Losses Using Three-dimensional Model In the above, a design method considering the winding loss during DAB operation of the integrated magnetic component was shown. However, accurate loss estimation of the entire magnetic component requires analysis considering the iron loss, proximity effect of the windings, and the DAB and NBC operations. Therefore, the loss composition of the magnetic component was analyzed by three-dimensional electromagnetic simulation using JMAG FEM. In the analysis, the loss of the proposed integrated magnetic component shown in Fig. 13(a) was compared with that of the conventional transformer/magnetic coupling reactor set shown in Figs. 13(b), (c).

Figure 14 shows the results of the analysis of the loss composition of the magnetic component. Figure 14(a) shows the loss itemization of \( P_A \) single load (transmission from Port B to Port A only) mode. By using the integrated magnetic component, the total value of the magnetic component loss is reduced by 3.4 W from 12.1 W to 8.7 W. This is because, the windings \( n_1 \) of the two outer legs of the integrated magnetic component functions as a transformer winding and a reactor winding as shown in Equation (20), leading to a reduction in the winding loss generated by the trapezoidal current in DAB operation. Figure 14(b) shows the loss itemization of \( P_C \) single load (transmission from Port B to Port C only) mode. By using the integrated magnetic component, the total value of the magnetic component loss increases 1.9 W from 18.2 W to 20.1 W. This is because, the reactor windings, which were conventionally two parallel windings, are consolidated into the center leg winding. From the above, if the integrated magnetic component is designed to be equal to the sum of the iron losses of the separate magnetic components as shown in Fig. 14, a reduction in the copper loss in \( P_A \) single load mode can be expected. Furthermore, when the thickness of copper in the center winding is designed such that the copper volumes of the reactor winding and center winding of the integrated magnetic component are equal, an improvement in efficiency in the \( P_C \) single load mode can be expected. However, increasing the center winding thickness increases the core height and reduces the volume reduction effect.

4. Experimental Results and Discussion

4.1 Experimental Circuit Configuration Figure 15 shows the configuration of the experimental circuit, and Table 3 shows the list of circuit parameters. A DC power supply was connected to Port B, an electronic load was connected to Port A and Port C, and the voltages of Port A, Port B, and Port C were set to \( V_{A,in} = 48 \) V, \( V_{B,in} = 200 \) V, and \( V_{C,in} = 12 \) V, respectively. Figure 16 shows the appearance of two prototypes designed at 750 W and 175 kHz, and Table 4 shows the measured magnetic characteristics of the prototype of the
integrated magnetic component and the separate magnetic component.

The prototype on the left side of Fig. 16 is the magnetic component integrated-type TPC. As the saturation magnetic flux density of ferrite core material at 120°C is 380 mT, the design value of the maximum flux density $B_{\text{MAX}}$ of the center leg of the integrated magnetic component was set to 309 mT, and the design value of the maximum flux density $B_{\text{MAX}}$ of its two outer legs was set to 109 mT. The low-voltage side switching element used a Si MOS (Toshiba, TPW4R50ANH) with a withstand voltage of 100 V, and the high-voltage side switching element used a GaN FET (Transphorm, TPH3006LD) with a withstand voltage of 600 V, a next-generation semiconductor component, which was adopted with the purpose of increasing the efficiency of the circuit. The core material of the magnetic component was Mn-Zn ferrite (TDK, PC95). Although $k_X$ takes a very high value of 0.997, since it is possible to change only the leakage inductance $L_{\text{eq}}$ while keeping the self-inductance ($L_1, L_2$, and $L_b$) of the integrated magnetic component almost constant by changing the distance between the primary and secondary windings, fine adjustment of $k_X$ according to the design requirements is possible. The prototype on the right side of Fig. 16 is conventional TPC created to verify the application effects of the integrated magnetic component technique.

The circuit parameters other than the magnetic components were same as those of the magnetic component integrated-type TPC, and PQ core (PC95PQ32/20Z-12) was used for the magnetic core of the transformer and the magnetic coupling reactor. Each magnetic component parameter is shown in Tables 2 and 4. The total core volume was 16.6 cm$^3$ for the conventional type but 14.9 cm$^3$ for the integrated type, showing a 10% reduction. Further, since the wiring space between the magnetic components can also be reduced compared to the conventional type, the overall circuit size is reduced by 33%, from 300 cm$^3$ (2.50 W/cm$^3$) to 200 cm$^3$ (3.75 W/cm$^3$).

### 4.2 Experimental Results

Figure 17 shows the measured waveforms of $V_{\text{uv}}$, $V_{\text{igf}}$, and $i_g$, in the combined load mode when 375 W of power was simultaneously transmitted from Port B to Port A and Port C. Similar to the ideal operation waveform of Fig. 5, the phase-U current waveform was such that the DC current with the ripple due to NBC operation overlapped with the trapezoidal current due to DAB operation.

Figure 18 shows the comparison of the efficiencies of the magnetic component integrated-type TPC and the conventional TPC in the single load mode. Figure 18(a) shows the efficiency curves of the proposed circuit and the conventional circuit in $P_A$ single load mode, and it can be seen that at the rated output of 750 W, the efficiency improved by 0.3% (93.4%) in the proposed circuit compared to the conventional circuit. Figure 18(b) shows the efficiency curves of the proposed circuit and the conventional circuit in $P_C$ single load mode, and it can be seen that at the rated output of 750 W, the efficiency of the proposed circuit was reduced by 0.4% (90.8%) compared to the conventional circuit.

Table 5 shows the comparison between the experimental results and the simulation results. The experimental results of the difference in circuit loss $\Delta \text{Loss}$ when changing from the conventional magnetic component to the integrated magnetic component were $-2.3$ W for $P_A$ single load mode and $+3.5$ W for $P_C$ single load mode, which was consistent with the increase/decrease trend of loss obtained by simulation ($-3.4$ W in $P_A$ single mode and $+1.9$ W in $P_C$ single mode), thus confirming the validity of the simulation results. Hence, it can be applied in future applications and design. From the above

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### Table 4. Measured parameters of magnetic components

<table>
<thead>
<tr>
<th>Description</th>
<th>Symbol</th>
<th>Conventional</th>
<th>Integrated</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductance of transformer</td>
<td>$L_1$, $L_2$</td>
<td>45 μH, 723 μH</td>
<td>48 μH, 767 μH</td>
</tr>
<tr>
<td>Coupling rate of Integrated core</td>
<td>$k_X$</td>
<td>0.999</td>
<td>0.997</td>
</tr>
<tr>
<td>Coupling rate of inductor</td>
<td>$k_L$</td>
<td>0.97</td>
<td></td>
</tr>
<tr>
<td>Inductance for NBC function</td>
<td>$L_a$</td>
<td>1.70 μH</td>
<td>1.89 μH</td>
</tr>
<tr>
<td>Inductance for DAB function</td>
<td>$L_{\text{eq}}$</td>
<td>4.33 μH</td>
<td>3.85 μH</td>
</tr>
</tbody>
</table>

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![Fig. 16. Prototypes of the integrated TPC and the conventional TPC](image-url)
results, it was confirmed that the magnetic component integrated TPC can reduce the core volume by 10% and the circuit size by 33%, while maintaining the conversion efficiency at the same level as the conventional type.

5. Conclusion

In this paper, we have proposed a transformer/reactor integrated TPC. The proposed circuit realizes the integration of the transformer and magnetic coupling reactor used in the conventional TPCs, using a three-legged core and a vertical winding structure. We created the prototypes of the proposed circuit and the conventional circuit at 750 W, 175 kHz, and measured their efficiencies to verify the validity of the proposed circuit. From the measurement results, it can be seen that the efficiency is 90% or more over a wide output range, and the same efficiency as the conventional type is obtained. Further, with the proposed circuit, we were able to demonstrate the possibility of reducing the core volume by 10% and the circuit volume by 33%, while maintaining the same efficiency. Since the 33% volume reduction of TPC will make it easy to replace the conventional single-port output power supply unit with a multiport power supply unit, it is expected to promote the spread of 12 V and 48 V vehicle accessory systems aiming at higher efficiency. In the future, we aim to further reduce the size of the integrated magnetic component by increasing the frequency of the circuit.

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