A New Isolated Multi-port Converter Using Interleaving and Magnetic Coupling Inductor Technologies

Masanori Ishigaki, Kenichi Ito, Shuji Tomura and Takaji Umeno

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ABSTRACT This paper proposes a new multi-port converter with interleaved magnetic coupling technologies. With the focus on the impedance behavior of coupled magnetic components and the control parameters, this multi-port converter integrates one isolated dc-dc converter and two multi-phase boost converters, and controls these topologies independently. The proposed converter has the capability to connect four dc applications or sources in total, and so brings significant cost reduction and improves the total loss consumption of power systems. In addition, magnetic isolation brings high reliability of complex power systems such as electric vehicles or grid-connected industries. In this paper, the basic circuit behavior is verified by theoretical considerations and experiments. An example of the proposed circuit as a multifunctional dc power system is demonstrated with the simulation results. Such an approach could bring about variable application fields to dc electric power systems.

KEYWORDS Power Electronics, De-De Converter, Multi-port, Power Semiconductor, Magnetic Devices, Electrical Vehicle, Grid Connected Industries

1. Introduction

With the rapid development of energy storage device technologies, dc power systems are becoming widely used in various applications. In electricity infrastructures, large energy storage devices such as Li-ion and NaS batteries promote distributed renewable energy sources, which include photovoltaic power systems, wind generator systems, and fuel cell systems in micro grids. In vehicles, gas-electric hybrid vehicles are becoming mainstream all over the world, and with further development of the power/energy densities of batteries, plug-in or purely electric vehicles can also become mainstream in future transportation systems. As energy storage device technologies progress, the electric systems become more complex. Therefore, cost reduction and downsizing of dc-dc converter circuits is actively sought as a common goal for power conversion circuits. To achieve this goal, power electronics technologies are developing new semiconductor devices such as SiC and GaN. In other approaches, inventions of circuit topologies have been used with some success for cost reduction and weight saving of complex dc power systems.

This paper proposes a new multi-port converter capable of controlling four dc components. With magnetic coupling technologies, the proposed circuit achieves integration of two multi-phase boost converters and one isolated dc-dc converter in one circuit. By controlling the duty and phase of each switch, it is possible to control voltages and transfer isolated dc power independently. In this paper, the basic functions and design of magnetic components are shown with a theoretical analysis. Then, the static characteristics and conversion efficiency are verified with experimental results. The simulation results demonstrate the application to an electric vehicle.

2. Electrical Vehicle System and Issues

2.1 Basic System Configuration

Figure 1 shows an example of the basic power system of an electric vehicle. This electric vehicle system mainly consists of a motor MG, a large capacity and high-voltage main battery BATT_m for driving, and a low voltage system battery BATT_s for ensuring a stable supply of power to computer elements and other important subsystems. Generally, a conventional lead battery is the system battery BATT_s, and the voltage is typically $V_{bs} = 12 – 14$ V. Depending on the vehicle...
system specification, the voltage of the main battery $V_{Bm}$ is set as $V_{Bm} = 100 – 500$ V. Often, the system battery $V_{Bm}$ is set as $100 – 200$ V, with a connect boost “CONVERTER A” between the battery $BATT_m$ and the motor MG to improve the motor efficiency and the controlled performance.\(^{(14-17)}\) For the same reason, electrical connections using “CONVERTER B” are used between the system battery $BATT_S$ and the load of high-power actuators $LOAD_b$.\(^{(4)}\) Additionally, the main battery $BATT_m$ and the system battery $BATT_S$ are interconnected by an isolated dc-dc converter “CONVERTER C” for hardware redundancy and fail-safe design. The main battery $BATT_m$ needs to be charged from an external AC or DC power supply, $EXT_{AC}$ or $EXT_{DC}$, through the battery charger circuit “CHARGER”.

### 2. 2 Standard Circuit Schematics and Functions

Figures 2 and 3 show conventional converter circuits and ideal waveforms.\(^{(18-20)}\) Figure 2(a) is an interleaved multi-phase converter circuit using “CONVERTER A” and “CONVERTER B”, shown in Fig. 1. The two inductors $L_{b1}$ and $L_{b2}$ can be directly or inversely coupled by constructing directional selections between the two windings. In this case, $L_{b1}$ and $L_{b2}$ are an inverse coupling with coupling coefficient $k_b$. Figure 3(b) shows the ideal waveforms of the interleaved multi-phase converter. The phase between $V_u$ and $V_v$ is set as $\theta$, and the duty ratio is set as $D = \delta/2\pi$. In this configuration, the voltage relationship between input voltage $V_i$ and output voltage $V_o$ is given by

$$V_o = \frac{1}{1 - D} V_i \quad (1)$$

The output voltage $V_o$ is controlled by adjusting the duty ratio $D$. In each period, the derivative value of the input current $i'_{in}$ and the inductor current $i'_{Lb1}$ can be described in

$$i'_{in}[\theta ≤ \theta ≤ \theta_2] = -\frac{1}{(1 + k_b)L_b}(V_o - 2V_i) \quad (2)$$

$$i'_{in}[\theta_2 ≤ \theta ≤ \theta_3] = \frac{2}{(1 + k_b)L_b}V_i \quad (3)$$

$$i'_{Lb1}[\theta_1 ≤ \theta ≤ \theta_2] = \frac{1}{1 - k_b^2 L_b} V_o + \frac{1}{1 + k_b L_b} V_i \quad (4)$$

$$i'_{Lb1}[\theta_2 ≤ \theta ≤ \theta_3] = \frac{1}{(1 + k_b)L_b}V_i \quad (5)$$

Because of the inversely coupled effect, the current ripple of $i_{Lb1}$ is a monotonically increasing function of $k_b$. Thus, the core and winding losses of the inductors increase with increasing $k_b$. In contrast, the switching losses decrease drastically if the boost converter operates in the discontinuous current mode, when $i_{Lb1}$ across zero is $\theta = \theta_2$. Therefore, the total efficiency can be optimized to select the parameters $L_b$ and $k_b$ by considering the steady-state duty, the power ratio, and characteristics of the magnetic components and the semiconductors.

Figure 2(b) is an isolated dc-dc converter circuit using “CONVERTER C”, shown in Fig. 1.\(^{(21-23)}\) The primary bridge $S_{d1} – S_{d4}$ and the secondary bridge $S_{a1} – S_{a4}$ are interconnected by a transformer. The inductor $L_{b1}$ is connected between the two bridges. The voltage $V_o$ is controlled by adjusting the duty ratio $D$. In each period, the derivative value of the input current $i'_{in}$ and the inductor current $i'_{Lb1}$ can be described in

$$i'_{in}[\theta ≤ \theta ≤ \theta_2] = -\frac{1}{(1 + k_b)L_b}(V_o - 2V_i) \quad (2)$$

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\(S_{d1} - S_{d8}\) are connected through inductors \(L_{d1}, L_{d2}\), and transformer Tr with turn ratio 1: \(N\). Figure 3(b) shows the ideal waveforms of the bi-directional isolated dc-dc converter. The adjusted phase angle between the primary and the secondary inductor currents is \(\phi_d\), and the duty ratio of each bridge is set as \(D = 1/2\). In the period \(\theta_1 \leq \theta \leq \theta_2\), the derivative value of the primary inductor current \(i'_d\) is as follows:

\[
\begin{align*}
    i'_d &= \frac{V_A + V_B}{2L_{DD}} \\
    L_{DD} &= L_{d1} + L_{d2}/N^2 + L_i
\end{align*}
\]  

(6)  

(7)

where \(L_i\) is the primary equivalent value of the leakage inductance of the transformer. The transmitted power \(P_{DCDC}\) is given by

\[
P_{DCDC} = \frac{V_AV_B}{\omega_{\alpha}L_{DD}} \left( \frac{\phi_d - \phi_c^2}{\pi} \right)
\]

(8)

According to Eq. (8), the \(P_{DCDC}\) applies control by adjusting the phase angle \(\phi_d\).

In the case of Fig. 1, the number of semiconductors and magnetic components is equal to 21, and these thus lead to issues of cost, volume, and productivity of converters.

3. Proposed Multi-port Converter

Figure 4 shows the proposed multi-port converter. Figure 4(a) is the circuit schematic, which shows two full-bridge inverters and three magnetic components constructing four dc ports (port A – port D). Figure 4(b) shows the structure of the magnetic components of Fig. 4(a). The magnetic components consist of three cores: one is for the transformer, and the other two are for the inductors \(L_x - L_y\). The parameters of the magnetic components must be symmetric, the number of windings is \(1:n:n\), and the inductance is \(L_u = L_v, L_a = L_b\). To simplify the theoretical analysis, the magnetic coefficients are set as \(k_1 = k_2 = k\). Each winding of the transformer has a center tap that makes port C and port D such that primary ports A and C, and secondary ports B and D are connected by the multi-phase boost converter, as shown in Fig. 5(a), and ports A and B are connected by the isolated dc-dc converter, as shown in Fig. 5(b). Thus, all four ports A – D are connected bi-directionally. The primary inductors \(L_x, L_y\) and the secondary inductors \(L_a, L_b\) are strongly inversely coupled inductors connected to each winding of the transformer. Consequently, the number of magnetic components and semiconductors is reduced by more than half of that of the conventional system example described in Section 2.

In the theoretical analysis of this converter, it is

Fig. 3 Ideal waveforms of conventional circuits. The boost converter controlled duty angle is \(\delta b\), and the dc-dc converter is controlled by phase angle \(\phi_d\)

Fig. 4 The proposed multi-port converter. Major parts of the proposed converter are the same as the isolated dc-dc converter, but center-tapped \(C_1, C_2\) of the transformer adds two more ports.
important to understand the behavior of the magnetic components. Table 1 shows the impedance behavior of the magnetic components, the impedance of the inversely coupled inductors, the increase against the common mode current, and the decrease against the differential mode current. As indicated in this table, the inductors work as large inductance $L_b$ in Eqs. (4) and (5) for the boost converter function, and also work as small inductance $L_{DD}$ in Eq. (8) for the isolated dc-dc converter function. The exiting inductance of the transformer is canceled against the common mode current, and so the transformer only works as a differential current, so that no interference exists between the two functions.

Figures 6 and 7 show examples of the ideal waveforms and equivalent circuits of the proposed converter. To control multi-phase boost operation, the duty ratio of all lower switches is adjusted as $D = \delta/2\pi$, and the phase angles between $V_U$ and $V_V$, $V_A$ and $V_B$ are set as $\pi$. By using the same principle of the isolated dc-dc converter illustrated in Fig. 2(b), the phase angle between primary voltage $v_{uv}$ and secondary voltage $v_{ab}$ is the adjusted $\phi$. The behavior of the proposed converter becomes different with each change in the relationship between the duty angle $\delta$ and the phase angle $\phi$. Figures 6(a) and 7(a) show the ideal waveforms and equivalent circuits in the case of $\delta \geq \pi + \phi$. In Fig. 7, all descriptions are shown as the primary side, and the arrows indicate the current direction of each period. For example, in periods I and II, the input current $i_C$ is increased and the common current is through the inversely coupled inductors $L_u$ and $L_v$. Aside from this, the differential current is through $L_u$ and $L_v$ by secondary voltage $v_B / N$, and the transformer current $i_{DD}$ increases in period I. In period II, the common current is through the transformer from $V_C$ but its own exciting is canceled, so that no interference or magnetic saturation occurs in this converter. Figures 6(b) and 7(b) show the ideal waveforms and equivalent circuits in the case of $\delta \leq \pi + \phi$. In Fig. 6(b), the transformer exited as $\pm (V_A + V_B / N)$ in period 2 ($\theta_2 \leq \theta \leq \theta_3$) and period 6 ($\theta_6 \leq \theta \leq \theta_7$), so the shape of transformer current $i_{DD}$ is a different waveform in Fig. 6(a).

Tables 2 and 3 show the derivative values of $i_u$ and $i_v$.

![Fig. 5 Illustration of the basic functions of the proposed converter. Due to the directions of the windings, the flux of the magnetic components in each function is distinct.](image)

![Fig. 6 Ideal waveforms of the proposed circuit.](image)

Table 1  Impedance behavior of inversely coupled inductor and transformer.

<table>
<thead>
<tr>
<th>Current direction in Fig. 3</th>
<th>Primary inductance of coupled inductor</th>
<th>Secondary inductance of coupled inductor</th>
<th>Exciting inductance of transformer</th>
</tr>
</thead>
<tbody>
<tr>
<td>Common</td>
<td>$2L_u(1 + k_1)$</td>
<td>$2L_v(1 + k_1)$</td>
<td>zero (canceled)</td>
</tr>
<tr>
<td>Differential</td>
<td>$2L_u(1 - k_1)$</td>
<td>$2L_v(1 - k_1)$</td>
<td>large (excited)</td>
</tr>
</tbody>
</table>
in each period of Fig. 7(a) and 7(b). Due to symmetry, these tables describe only a half period of Fig. 7, and for simplicity, the voltage is set as $V_A = V_B / N$. Here $L_s$ is comparable to $L_{DD}$ in Fig. 2(b) and is given by

$$L_s = L_u + \frac{L_a}{N^2} \quad (9)$$

According to Table 2 and 3, the common mode current $i_{cm} = i_L$ has the same value as that of the conventional boost converter in Eqs. (2) and (3). Therefore, the proposed converter makes it possible to control voltages by adjusting the duty ratio $D$. However, due to the differential mode current $i_{df}$ through the transformer, the isolated dc-dc power $P_{DD}$ is transferred between port A and port B. The transfer power $P_{DD}$ can be calculated by

$$P_{DD} = \frac{V_A V_B}{\omega \{2(1-k)L_u + L_i\} N \pi} \times f(\delta, \phi) \quad (10)$$

$$f(\delta, \phi) = \delta(\pi - \delta) \quad : [0 \leq \delta \leq |\phi|] \quad (11)$$

$$\phi(\delta - \phi) \quad : [|\phi| \leq \delta \leq \pi] \quad (12)$$

$$\phi(\pi - \delta - \phi) \quad : [\pi \leq \delta \leq |\phi| + \phi] \quad (13)$$

$$\phi(2\pi - \delta) \quad : [\pi + |\phi| \leq \delta] \quad (14)$$

By using Eq. (10), $P_{DD}$ is expressed as a function of $\phi$ and $\delta$, and variables $L_u$, $L_v$, $L_a$, $L_b$, and $k$.

### 4. Experimental Results

#### 4.1 Experimental Verifications

The experimental configuration is shown in Fig. 8 and Table 4. The voltage of the input dc source of port D is set as $V_D = 100$ V, and load $R_v$ is connected to port B. Port A and port C have contactors SW$_A$ and SW$_B$ for choosing either primary power source $V_A$ or $V_C$. All switching devices are MOSFET (IXFN130N30, 300 V, 130 A, IXYS), and the core material of the transformer and the inductors is Finemet (F3CC0025, Hitachi). The turn ratio of the transformer is set as $N = 5$, and the number of primary windings is $n_1 = n_2 = 2$.

#### Table 2 Derivative values of $i_u$ and $i_v$ in Fig. 7(a).

<table>
<thead>
<tr>
<th></th>
<th>$i'_u$</th>
<th>$i'_v$</th>
<th>$i'_{cm}$</th>
<th>$i'_{df}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
<tr>
<td>II</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
<tr>
<td>III</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
<tr>
<td>IV</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
</tbody>
</table>

#### Table 3 Derivative values of $i_u$ and $i_v$ in Fig. 7(b).

<table>
<thead>
<tr>
<th></th>
<th>$i'_u$</th>
<th>$i'_v$</th>
<th>$i'_{cm}$</th>
<th>$i'_{df}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
<tr>
<td>II</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
<tr>
<td>III</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
<tr>
<td>IV</td>
<td>-1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>1 &amp; 0</td>
<td>0</td>
</tr>
</tbody>
</table>
The primary inductance is set as $L_u = L_v = 10 \, \mu H$, secondary inductance is set as $L_a = L_b = 50 \, \mu H$, and both coupling coefficients are $k_1 = k_2 = 0.8$. Given this specification, the switching frequency is set as 20 kHz.

**Figures 9 and 10** illustrate the result of steady-state operations. In this condition, dc power is transferred form port C to port A by using the boost converter. This power is transferred to port B by using the isolated dc-dc converter, and finally reaches port D by using the down conversion. Thus, the transferred power is described as $P_{DD} = i_C V_C$. In Fig. 9, the experimental waveforms in three cases of duty ratio $D$ are shown. The voltages $V_a$ and $V_b$ are estimated from the output voltages of each bridge, $v_{uv}$ and $v_{ab}$: $V_a = 32 \, V$, $V_b = 165 \, V$ at $D = 0.4$, $V_a = 40 \, V$, $V_b = 200 \, V$ at $D = 0.5$, $V_a = 50 \, V$, $V_b = 250 \, V$ at $D = 0.6$, and the result shows the boost converter function follows Eq. (1). The inductor current $i_u$ is shaped like traperized, biased half of $i_C$. The input current $i_c$ shape is nearly flat due to the effect of multiphase boost conversion, and the average value is increased monotonously by duty ratio $D$. In Fig. 10, the experimental results of transferred power $P_{DD}$ are compared with estimated values from Eq. (10) and Table 4. The experimental values correspond to the estimated lines, controlled by phase angle $\phi$ and duty ratio $D$.

![Fig. 8 Setup of the experimental circuit.](image)

**Table 4** Circuit parameter of the experimental circuit.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated power</td>
<td>$P_{DD}$</td>
<td>±800 W</td>
</tr>
<tr>
<td>Rated input voltage</td>
<td>$V_A$, $V_C$, $V_D$</td>
<td>50 V, 20 V, 100 V</td>
</tr>
<tr>
<td>Rated Duty</td>
<td>$D$</td>
<td>0.2 ~ 0.8</td>
</tr>
<tr>
<td>Turn ratio of transformer</td>
<td>$N$</td>
<td>1:5</td>
</tr>
<tr>
<td>Turn number of transformer</td>
<td>$n_t$</td>
<td>2</td>
</tr>
<tr>
<td>Magnetizing inductance (primary)</td>
<td>$L_{m}$</td>
<td>300 $\mu H$</td>
</tr>
<tr>
<td>Coupling coefficient of transformer</td>
<td>$k_{c}$</td>
<td>0.98</td>
</tr>
<tr>
<td>Coupled inductor (primary)</td>
<td>$L_{uc}$, $L_{ub}$</td>
<td>10 $\mu H$</td>
</tr>
<tr>
<td>Coupling inductor (secondary)</td>
<td>$L_a$, $L_b$</td>
<td>50 $\mu H$</td>
</tr>
<tr>
<td>Coupling coefficient of inductor</td>
<td>$k$</td>
<td>0.8</td>
</tr>
<tr>
<td>Core material of inductor and transformer</td>
<td>Hitachi material</td>
<td></td>
</tr>
<tr>
<td>dc capacitor</td>
<td>$C_f$</td>
<td>1400 $\mu F$</td>
</tr>
<tr>
<td>Dead time</td>
<td>$t_d$</td>
<td>1 $\mu s$</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>$f_{sw}$</td>
<td>20 kHz</td>
</tr>
</tbody>
</table>

![Fig. 9 Experimental waveforms of the proposed circuit, port C and port D are dc sources set as $V_C = 20 \, V$, $V_D = 100 \, V$; port B and port A are opened.](image)

![Fig. 10 Power-transferred value $P_{DD}$ from port C to port D vs. phase shift angle $\phi$. The three dark symbols are the result of Fig. 9.](image)
Figure 11 plots the measured converter efficiency. Figure 11(a) shows the conversion efficiency of each function, and these values are above 90% over a wide range of output power. The efficiency of the boost converter is 90.3% at $P_B = 200$ W, 96.2% at $P_B = 600$ W, and that of the isolated dc-dc converter is 92.7% at $P_B = 200$ W, 91.2% at $P_B = 600$ W. Figure 11(b) shows the total efficiency of the hybrid mode. In this mode, a part of $P_B$ is provided from port A, and the other part is provided from port D, under the constant load configuration $P_L = 400$ W. The distribution factor $k$ is defined as $k = P_{DD} / P_B \times 100$, and rated at $0 - 100\%$. The total efficiency $\eta$ is equal to $P_B / (P_C + P_D) \times 100$ and reaches a maximum 97% at $k = 50\%$. Because the proposed converter reduces the number of components, the total loss and consumption are improved drastically. If the same power is given to port B by using the conventional boost converter and the isolated dc-dc converter, the total efficiency is lower than that of the hybrid mode. For example, in the estimate from the result of Fig. 11(a), the total efficiency is only 91% for the same distribution factor.

4.2 System Example and Simulation Result

As one example of using the proposed converter, an electric vehicle power system and its controller are shown in Fig. 12. The motor MG connects to port B, the main battery $\text{BATT}_s$ connects to port B, and the system battery $\text{BATT}_m$ connects to port C. A solar panel is placed at port A with a contactor SW for charging two batteries. In Fig. 12, there are two feedback controllers in one system. To drive the vehicle, contactor SW turns off, the secondary boost converter controls the voltage of port B to command $V_{B*}$ from the motor MG by adjusting $\delta$ with the PI controller. Depending on the usage of the batteries, the transferred power between the primary and the secondary $P_{DD}$ is controlled by another PI controller. In a parking situation, the contactor SW is turned on, and $\text{BATT}_m$ and $\text{BATT}_s$ are changed by controlling $P_{DD}$. The feedforward values $d_{ff}$ and $f_{ff}$, estimated by Eqs. (1) and (10), have a role in reducing the interference between $P_{DD}$ and $V_{B*}$.

Figure 13 shows the simulation waveforms of Fig. 12. The parameters are the same as those for the measured efficiency of each function. When measuring the isolated dc-dc converter, the input power source is connected to port A and $V_A = 50$ V, $V_B = 250$ V, and port C and port D are opened. When measuring the boost converter, the input power source is connected to port D and $V_B = 250$ V, $V_D = 100$ V, and port A and port D are opened. When measuring the hybrid mode, the input power source is connected to ports A and D and $V_A = 50$ V, $V_B = 250$ V, $V_D = 100$ V, and port C is opened.
experimental circuit configuration, as listed in Table 4. In the period \( t_1 \leq t \leq t_2 \), \( P_{DD}^* \) is 0 W to 250 W, transfers power from port C to port D. The feedback controller adjusts \( \phi \) by approximately 20 degrees and keeps \( P_{DD} \) as the command value in \( t_2 \leq t \leq t_3 \). In the next period \( t_3 \leq t \leq t_4 \), load \( P_B \) increases from 0 W to 600 W, and the port D supply is \( P_D = 350 \) W in accordance with the relationship, \( P_B = P_{DD}^* + P_D \). In the last period \( t_5 \leq t \leq t_6 \), the command of output voltage \( V_B^* \) increases from 150 V to 300 V, then the feedback controller adjusts \( \delta \) from 110 degrees to 250 degrees by following Eq. (1). In this case, the controller decreases phase angle \( f \) to keep \( P_{DD} = 250 \) W under Eq. (10), that is, non-interacting control is achieved in this converter.

5. Conclusion
In this paper, a multi-port converter formed by inversely coupled inductors and center-tapped was presented. For adjusting the duty ratio and the phase angle of full-bridged switches, it is possible to integrate boost conversion and isolated dc-dc converter function and connect four dc sources/load in one circuit. The theoretical analysis and the experimental results verified the multi-functionality of the magnetic components and the converter operation, and it was also shown that the proposed converter has the ability to downsize and improve energy consumption of complex dc power systems. In addition, an example of system utilization was shown, and the simulation result confirmed that it is able to replace discrete converters with one proposed converter without any difficulty in terms of control performance. This converter is suitable for multi-input multi-output systems, such as electric vehicle systems and renewable energy systems.

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