Hybrid SPS Control for ISOP Dual-Active-Bridge Converter Based on Modulated Coupled Inductor With Full Load Range ZVS and RMS Current Optimization in DC Transformer Applications

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ABSTRACT The Input-serial-output-parallel (ISOP) dual-active-bridge (DAB) converter is an attractive solution to be used as a DC transformer (DCT) in the DC power distribution system. However, due to zero-voltage-switching (ZVS) limitation and large circulating current, the efficiency of the converter will be reduced. To improve the efficiency, an ISOP DAB converter based on a modulated coupled inductor, and a corresponding hybrid SPS control are proposed in this paper. With the hybrid SPS control strategy, the equivalent leakage inductance of modulated coupled inductor can be controllable. The large inductance is adopted at light load and the small inductance for the transformer is applied at heavy load to reduce circulating current and extend the ZVS range. By optimizing the root-mean-square (RMS) current of two operation modes of the converter, a control boundary of the hybrid SPS control can be obtained to modulate the coupled inductor. Meanwhile, the additional magnetizing inductance is utilized to achieve ZVS for all switches. The analysis based on the RMS current optimization and ZVS condition is given. Based on the conditions of the full load range ZVS and small RMS current, the parameters of the modulated coupled inductor, the turn ratio of the transformer, and magnetizing inductance are designed. Finally, a 250kHz, 800/400V, 2000W ISOP DAB prototype is built to verify the effectiveness of the proposed method. Meanwhile, a comparison experiment with the conventional individual leakage inductors solution is given, which shows performance improvements of the proposed solution.

INDEX TERMS DC transformer, ISOP, DAB, modulated coupled inductor, ZVS.

I. INTRODUCTION
Isolated bidirectional dc-dc converter (IBDC) is widely used as a DC transformer (DCT) in the DC power distribution system to achieve power flow balance between two DC buses [1]. In view of the topology, the dual-active-bridge (DAB) converter is popularly regarded as the constituent unit of DCT for its high reliability, control flexibility, soft switching, and easy control [2].

To meet voltage differences between different DC buses and achieve high power features of DCT, the modular structure is a popular solution to increase the voltage and current levels [3]. For high input voltage and large output current applications, the input-serial-output-parallel (ISOP) structure is often employed in DCT [4], [5], [6]. Because the voltage variations of DC buses are small, and the voltage gain of DCT is relatively fixed, the single-phase-shift (SPS) control strategy is widely applied to each DAB module to obtain simple implementation and high performance in industrial applications [7]. However, under SPS control, ZVS range limitation will occur at light load for the DAB converter. Moreover, a high circulating current will exist when the phase-shift ratio is large, leading to extra undesirable power loss [8].
To improve the performance of the DAB converter, many researchers have proposed lots of solutions to extend the ZVS range and reduce the circulating current in other applications, which can be roughly divided into two categories: control improvement and topology modifications. In the aspect of control improvement, the multiple control degree of freedoms strategies, such as extend-phase-shift (EPS) control strategies [9], [10], dual-phase-shift (DPS) control strategy [11], [12], triple-phase-shift (TPS) control strategy [13], [14] can be adopted aimed at different performance improvement goals. However, these methods will lead to a complex control strategy due to multiple working modes, compared to the SPS control strategy. In addition, for some DCT applications, its basic unit is a half-bridge DAB converter, which will lose the ability to adopt the increased control degree of freedom rather than SPS control. A hybrid-mode-modulated strategy with natural boundary transition and inherent dynamic control was proposed for the DAB converter [15]. According to the output current, the control strategy is changed among trapezoidal continuous conduction mode, triangular discontinuous conduction mode, and SPS control, which can improve the efficiency and transient response. In [16], a variable switching modulation was proposed to reduce the circulating current under SPS control, which can be used for a half-bridge DAB converter. However, the switching frequency will be extremely high to guarantee soft switching at light load.

In the aspect of topology modifications, some additional components can be combined with conventional DAB converters. A DC blocking capacitor for the DAB converter with hybrid modulation was proposed to optimize RMS current [17]. In [18], to enlarge the ZVS range and reduce current stress, a hybrid EPS control strategy was proposed for the DAB converter with a DC-blocking capacitor. A center-tapped transformer was applied to improve the performance of the DAB converter [19], which significantly minimizes the switching losses and alleviates electromagnetic interference. To enhance the efficiency of the converter, a T-type DAB converter was proposed in [20], which utilizes four-level voltage at one port of the transformer to obtain matched voltage waveforms. To extend the ZVS range, adding the auxiliary inductor is also an effective method, which can inject extra current to help the switches charge or discharge the junction capacitor sufficiently during the dead time [21]. However, adding an auxiliary inductor will increase the volume of the converter. To reduce the auxiliary circuit for the converter, the magnetizing inductance of the transformer can be adopted [9], [22], [23], which uses the magnetizing current to achieve ZVS with consideration of the switch junction capacitors.

For the DAB converter in the DCT application, there is a contradiction between ZVS at light load and circulating current at heavy load. For example, the small leakage inductance will reduce circulating current, but the ZVS range will be limited at light load. If large leakage inductance is adopted, the ZVS range will be extended, but the circulating current will be increased. However, for these aforementioned modulation strategies and topology modification methods [9], [10], [11], [12], [13], [14], [15], [16], [17], [18], [19], [20], [21], [22], [23], the value of the inductive element (leakage inductance or magnetizing inductance) is fixed during the whole operation range. To deal with this, in [24] and [25], the authors use a variable inductor to reduce circulating current and extend the ZVS range. The large equivalent leakage inductance is employed at light load to extend the ZVS range, and the small equivalent leakage inductance is used to reduce the circulating current at heavy load. An adjustable-tap high-frequency transformer was proposed in [26]. By changing the tap of the transformer, the DAB converter can boost efficiency over wider voltage gains and provide adjustable power-transfer capability compared to a normal DAB. In [27], the switch-controlled magnetizing inductor was proposed to extend the ZVS range. However, these methods, in [24], [25], [26], and [27], need an extra auxiliary circuit to achieve variable inductance. To eliminate the extra auxiliary circuit and achieve variable leakage inductance, the authors proposed a modulated coupled inductor to limit start-up inrush current and reduce the circulating current without any other auxiliary circuit [28]. By controlling the operating phase of the coupled inductor, the equivalent leakage inductance of the converter can be changeable. The large equivalent inductance is applied in the start-up interval to limit inrush current, and the small equivalent inductance is applied in a steady state to reduce the circulating current. However, the equivalent inductance does not exchange in steady state, and the small equivalent inductance causes the ZVS limitation for the converter, leading to low efficiency at light load.

In this paper, to achieve low RMS current and full load range ZVS, an ISOP DAB converter based on a modulated coupled inductor, and a corresponding hybrid SPS control are proposed. And the contributions can be summarized as follows.

Firstly, by utilizing the modulated coupled inductor, the equivalent leakage inductance can be adjusted by modulating the PWM signals. The larger equivalent leakage inductance is applied at light load to reduce the RMS current and extend the ZVS range when the two port voltages of the proposed converter are mismatched. And the smaller equivalent leakage inductance is applied at heavy load to reduce the RMS current and improve the conversion efficiency.

Secondly, a corresponding hybrid SPS control scheme is proposed for the converter, which can control the equivalent leakage inductance of the modulated coupled inductor according to different operation conditions, leading to optimization for the RMS current.

Finally, a design method for the modulated coupled inductor and the transformer is introduced to achieve full load range ZVS. By designing the equivalent leakage inductance, turn ratio of the transformer, and magnetizing inductance, the full-load ZVS operation can be obtained.

This paper is organized as follows. The proposed modified ISOP DAB converter and its working principle are
described in Section II. The RMS current optimization and ZVS range considering magnetizing inductance are analyzed in Section III. Section IV is the parameters design consideration of the converter. The experimental results are shown in Section V. Finally, Section VI provides the conclusions.

**FIGURE 1.** The circuit of the proposed converter.

**II. THE TOPOLOGY AND WORKING PRINCIPLE OF THE PROPOSED CONVERTER**

**A. CIRCUIT CONFIGURATION**

The schematic circuit of the proposed converter is shown in Fig.1. As seen, the converter consists of two half-bridge DAB modules connected with the ISOP structure, and the leakage inductors of two DAB modules are coupled. And the magnetizing inductance is specially addressed instead of ignoring it, since it is utilized to help achieve ZVS in this paper. The parameters of the two DAB modules, including the isolation transformer, the half-bridge capacitors, the leakage inductor, and the switch device, are designed to be the same to ensure power balance. The two ports of the converter are connected to two DC buses.

**B. WORKING PRINCIPLE OF THE PROPOSED CONVERTER**

The hybrid SPS control method is adopted in the proposed converter, whose key waveforms are illustrated in Fig.2 and Fig.3. As seen, the phase-shift ratio between primary side switches and secondary side switches is defined as $\varphi$. The phase between two DAB modules, namely the phase between the voltages of two windings of the coupled inductor, is defined as $\varphi_{ab}$.

There are two working modes for the proposed converter. When the converter works in Mode I, $\varphi_{ab}$ is operated at 0 degree, whose key operating waveforms are shown in Fig.2. In this mode, the phase between two DAB modules is controlled to be 0 degree by modulating the PWM signals. The driving signal of $S_{1a}$ and $S_{2b}$ is the same, and $S_{3a}$ and $S_{4b}$ is the same. And the phase between $v_{La}$ and $v_{Lb}$ is 0 degree, namely $v_{La} = v_{Lb}$.

When the converter works in Mode II, $\varphi_{ab}$ is operated at 180 degrees, whose key operating waveforms are shown in Fig.3. In this mode, the phase between two DAB modules is controlled to be 180 degrees by modulating the PWM signals. The driving signal of $S_{1a}$ and $S_{2b}$ is the same, and $S_{3a}$ and $S_{4b}$ is the same. And the phase between $v_{La}$ and $v_{Lb}$ is 180 degrees, namely $v_{Lb} = -v_{La}$.

**C. MODEL ANALYSIS OF THE MODULATED COUPLED INDUCTOR**

As for the coupled inductor shown in Fig.1, its mathematical model can be derived by the following equation according to the basic circuit theory.

\[
\begin{align*}
\frac{dv_{La}}{dt} &= L_a \frac{di_{pa}}{dt} + M \frac{di_{pb}}{dt} \\
\frac{dv_{Lb}}{dt} &= M \frac{di_{pa}}{dt} + L_b \frac{di_{pb}}{dt}
\end{align*}
\]

(1)
Equation (1) can be rearranged as

\[
\begin{align*}
(1 - k_c^2)L_a \frac{di_{pa}}{dt} &= v_{La} - k_c v_{Lb} \\
(1 - k_c^2)L_b \frac{di_{pb}}{dt} &= v_{Lb} - k_c v_{La}
\end{align*}
\]  

(2)

where \(k_c\) is the coupling coefficient, which is defined as

\[
k_c = \frac{M}{\sqrt{L_a L_b}}
\]  

(3)

When the converter operates in Mode I, the relation of the two voltages across the two windings of the coupled inductor can be expressed as

\[
v_{La} = v_{Lb}
\]  

(4)

To achieve the power balance of two DAB modules, the parameters of two DAB converters are designed to be the same. Then, the self-inductance of the coupled inductor is designed to be the same that \(L_k = L_a = L_b\). Therefore, according to (4), (2) can be simplified as

\[
\begin{align*}
(1 + k_c)L_k \frac{di_{pa}}{dt} &= v_{La} \\
(1 + k_c)L_k \frac{di_{pb}}{dt} &= v_{Lb}
\end{align*}
\]  

(5)

Then, the equivalent leakage inductance when \(\varphi_{ab}\) is 0 degree can be obtained as

\[
L_{eq0} = (1 + k_c)L_k
\]  

(6)

When the converter operates in Mode II, the relation of the two voltages across the two windings of the coupled inductor can be expressed as

\[
v_{La} = -v_{Lb}
\]  

(7)

According to (7), (2) can be simplified as

\[
\begin{align*}
(1 - k_c)L_k \frac{di_{pa}}{dt} &= v_{La} \\
(1 - k_c)L_k \frac{di_{pb}}{dt} &= v_{Lb}
\end{align*}
\]  

(8)

Then, the equivalent leakage inductance can be obtained as

\[
L_{eq180} = (1 - k_c)L_k
\]  

(9)

As analyzed above, when the converter operates in different modes, which can be controlled by modulating the PWM signals of the converter, the equivalent leakage inductance can be exchanged between \(L_{eq0}\) and \(L_{eq180}\).

**D. POWER TRANSFER CHARACTERISTICS OF THE PROPOSED CONVERTER**

Because of the symmetrical current characteristics of two DAB modules, only module A is analyzed in detail. When the converter works in Mode I, the instantaneous current of the coupled inductor can be calculated as

\[
\begin{align*}
I_{pa}(t_1) &= I_{pa}(t_0) + \frac{V_1 + 2nV_2}{4(1 + k_c)L_k f_s} k_c \left(1 - \varphi\right) \\
I_{pa}(t_2) &= I_{pa}(t_1) + \frac{V_1 - 2nV_2}{4(1 + k_c)L_k} \left(1 - 2\varphi\right)
\end{align*}
\]  

(10)

where \(f_s\) is the switching frequency of the converter, \(\varphi\) is the phase-shift ratio of each DAB module. According to the volts-second theory, the relation between \(I_{pa}(t_0)\) and \(I_{pa}(t_2)\) is

\[
I_{pa}(t_0) = -I_{pa}(t_2)
\]  

(11)

Then, according to (10) and (11), \(I_{pa}(t_0)\) can be solved as

\[
I_{pa}(t_0) = -\frac{V_1 - 2nV_2 + 8n\varphi V_2}{16(1 + k_c)L_k f_s}
\]  

(12)

After that, the transferred power of the converter can be calculated as

\[
P_I(\varphi) = 2f_s \int_{t_0}^{t_2} V_I I_{pa}(t) \, dt = \frac{nV_1 V_2 \varphi(1 - 2\varphi)}{4(1 + k_c)L_k f_s}
\]  

(13)

To normalize the transferred power, the base power can be set when the converter works with \(\varphi = 0.25\), \(k_c = 0\), and \(V_1 = 2nV_2\), which can be expressed as

\[
P_{base} = P(0.25)|_{V_1=2nV_2,k_c=0} = \frac{V_1^2}{64L_k f_s}
\]  

(14)

Then the normalized power under Mode I can be obtained as

\[
P_I^*(\varphi) = \frac{P_I(\varphi)}{P_{base}} = \frac{8G\varphi(1 - 2\varphi)}{1 + k_c}
\]  

(15)

where \(G\) is the voltage ratio of the converter, which is defined as

\[
G = \frac{2nV_2}{V_1}
\]  

(16)

When the converter operates in Mode II, the transferred power can be calculated with the same approach, which is expressed as

\[
P_{II}(\varphi) = \frac{nV_1 V_2 \varphi(1 - 2\varphi)}{4(1 - k_c)L_k f_s}
\]  

(17)

And the normalized power under Mode II can be obtained as

\[
P_{II}^*(\varphi) = \frac{P_{II}(\varphi)}{P_{base}} = \frac{8G\varphi(1 - 2\varphi)}{1 - k_c}
\]  

(18)

By comparing (13) and (18), it can be concluded that the transferred power when the converter operates in Mode II is larger than that when the converter operates in Mode I with the same \(\varphi\), since the range of \(k_c\) is \(0 < k_c < 1\).
III. ZVS CONDITIONS ANALYSIS AND CONTROL BOUNDARY OF THE HYBRID SPS CONTROL

A. ZVS ANALYSIS CONSIDERING JUNCTION CAPACITANCE AND MAGNETIZING CURRENT INJECTION

As for the proposed converter, the working principles of the two DAB modules are symmetrical. Therefore, in this section, only module A is analyzed in detail. To achieve ZVS for switches, the energy stored in the inductive components should be high enough to charge/discharge the switch junction capacitors sufficiently.

For the primary side switches, the equivalent ZVS circuit considering the junction capacitors is illustrated in Fig.4(a). When \( S_{1a} \) is turned off, whose waveforms are shown in Fig.2 and Fig.3, \( i_{pa} \) will approach its maximum value. Then, \( i_{pa} \) will charge \( C_{s1} \) and \( C_{s2} \), and discharge \( C_{1a} \) and \( C_{1b} \). During the charging/discharging process, there are three parts of energy including the energy stored in \( L_{eq} \), energy in all capacitors, and the energy flowing into \( V_2 \). According to the law of energy conservation, the sum of all energy stored in the capacitors is a constant value, while the energy stored in \( L_{eq} \) is decreased and fed back to \( V_2 \). The relation among these energies can be calculated as

\[
E_{Leq}(t_{2-1}) = E_{Leq}(t_{2-2}) + E_{feedback} = \frac{1}{2} L_{eq} i_{pa}(t_{2-1})^2 \quad (19)
\]

where \( t_{2-1} \) and \( t_{2-2} \) are the beginning and end of the charging/discharging process, respectively.

According to Kirchhoff’s current law, the relation between the leakage inductor current and the capacitor voltage can be expressed as

\[
i_{pa}(t) = C_{s1} \frac{du_{s1}(t)}{dt} - C_{s2} \frac{du_{s2}(t)}{dt} = 2C_{s1} \frac{du_{s1}(t)}{dt} \quad (20)
\]

where \( u_{s1}(t) \) and \( u_{s2}(t) \) are the voltage across the switch junction capacitors for \( S_{1a} \) and \( S_{2a} \), respectively.

When the charging/discharging process is finished, the energy stored in \( L_{eq} \) is fed back to \( V_2 \), which is calculated as

\[
E_{feedback} = \int_{t_{2-1}}^{t_{2-2}} \left[ \frac{1}{2} n V_2 i_{pa}(t) \right] dt \quad (21)
\]

After the charging/discharging process, if the switch junction capacitors are charged/discharged sufficiently, the remaining energy stored in \( L_{eq} \) should be greater than zero. Therefore, the constraint to realizing ZVS for the primary side switches can be obtained as

\[
E_{Leq}(t_{2-1}) - E_{feedback} \geq 0 \quad (22)
\]

According to (19), (20), and (21), (22) can be rearranged as

\[
i_{pa}(t_{2-1}) \geq \sqrt{\frac{n V_1 V_2 C_{s1}}{L_{eq}}} \quad (23)
\]

which is the constraint for primary side switches to achieve ZVS.

For ZVS realization of the secondary side switches, different from the conventional DAB converter, the magnetizing current participates in the charge/discharging process after the switches are turned off. The equivalent ZVS circuit considering the junction capacitors and the magnetizing current after \( S_{1a} \) is turned-off is illustrated in Fig.4(b). As seen, the magnetizing current will benefit to charge/discharge the junction capacitors. The magnetizing current \( i_{ma} \) will cancel the leakage inductor current \( i_{pa} \) and change the shape of \( i_{sa} \), which will help secondary side switches to achieve ZVS. During the charging/discharging process, \( i_{sa} \) should always be greater than zero, which can be expressed as

\[
i_{s}(t_3) = n \left[ i_{ma}(t_3) - i_{pa}(t_3) \right] \geq 0 \quad (24)
\]

According to (23) and III-B, Fig.5 plots the ZVS range versus \( P^* \) and \( G \) with different \( k_c \) and \( k_L \). Noted that, \( k_L \) is the ratio between the leakage inductance and magnetizing inductance, which is defined as

\[
k_L = \frac{L_k}{L_m} \quad (25)
\]

\[
FIGURE 4. Equivalent ZVS circuit. (a) The equivalent ZVS circuit after \( S_{1a} \) is turned off. (b) The equivalent ZVS circuit after \( S_{1a} \) is turned off.
\]

\[
FIGURE 5. ZVS range versus \( P^* \) and \( G \). (a) ZVS range comparison with different \( k_c \). (b) ZVS comparison with different \( k_L \).
\]
converter. Fig. 5(b) shows the ZVS range comparison with different \( k_L \). As seen, when \( k_L = 0 \), which means the magnetizing inductance is ignored, none ZVS region will occur at light load. When \( k_L \) is changed from 0 to 0.15, which means the magnetizing current is injected, the ZVS region is expanded, and the full load range ZVS can be realized. Therefore, with the proper design of \( G \) and \( k_L \), full load range ZVS can be achieved for the converter under SPS control.

**B. RMS CURRENT DERIVATION**

Because the RMS current of the two DAB modules is the same, only module A is analyzed. When the converter operates in Mode I, the normalized RMS current can be derived as

\[
I_{p_{\text{rms}, I}}(\varphi) = \frac{1}{I_{\text{base}}} \left[ \frac{1}{I_{\text{zvs}, I}} \right] \left[ \int_{t_0}^{t_2} i_{\text{pa}}(t) \, dt \right] = \frac{4\sqrt{16G\varphi^2(3 - 4\varphi) + (G - 1)^2}}{\sqrt{3}(1 + k_c)} \tag{26}
\]

where \( I_{\text{base}} \) is the base current, which is defined as \((P_{\text{base}}/V_1)\). When the converter operates in Mode II, the normalized RMS current can be expressed as

\[
I_{p_{\text{rms}, II}}(\varphi) = \frac{4\sqrt{16G\varphi^2(3 - 4\varphi) + (G - 1)^2}}{\sqrt{3}(1 - k_c)} \tag{27}
\]

Substituting (15) and (18) into (26) and (27) respectively, Fig. 6 plots the normalized RMS current versus normalized power \( P^* \) with different coupling coefficient \( k_c \) and voltage ratio \( G \) under different operation modes in the 3D plot. As seen, when the converter operates at the voltage matching point, namely \( G = 1 \), the RMS current in Mode II is always smaller than that in Mode I under the full load range. However, when the converter does not operate at the voltage matching point, the RMS current in Mode II is smaller than that in Mode I at light load. Therefore, Mode I is more suitable for the converter working at light load, and Mode II is more suitable for the converter working at heavy load to reduce RMS current.

**C. CONTROL BOUNDARY BETWEEN MODE I AND MODE II**

To achieve a low RMS current for the converter, the different operating modes can be adopted in different operating conditions. It can be observed from Fig. 6 that the converter is more suitable to operate in Mode I at light load and Mode II at heavy load. Therefore, it is necessary to propose a hybrid SPS control strategy, of which the key point is how to determine the control boundary of different operating modes. For example, Fig. 7(a) illustrates the normalized RMS current when \( k_c = 0.5 \). And the control boundary can be obtained by solving the equations of (15), (18), (26) and (27). However, the calculation results are complicated, resulting in a large amount of resources consumed by the MCU controller.

**TABLE 1.** The fitted control boundary with different \( k_c \).

| \( k_c \) | \( P^* \) |
| --- | --- |
| 0.3 | \(-77.732G^2 + 265.06G^2 + 303.69G^2 - 168.81G + 48.637 \) |
| 0.5 | \(-89.228G^2 + 226.97G^2 - 193.13G + 55.422 \) |
| 0.7 | \(-86.01G^2 - 294.58G^2 + 338.22G - 129.63 \) |

Therefore, to simplify the expressions, the fitted boundary lines are used to approximate the original control boundary lines, shown as the dotted lines \( a \) and \( b \) in Fig. 7(b). Meanwhile, the control boundary with different \( k_c \) is expressed in Table 1. Then, the hybrid SPS control strategy can be obtained according to the control boundary in Table 1.

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Fig. 8 shows the comparisons of normalized RMS current between the proposed solution and the conventional solution. Noted that, the conventional solution is that the leakage...
inductors are not coupled, and the coupled coefficient $k_c$ is 0. As seen, under the proposed hybrid SPS control strategy, the RMS current of the converter can be reduced in some operating conditions. But, if the coupled coefficient $k_c$ is too large, the RMS current will be increased at the middle power point. Therefore, a suitable coupled coefficient is needed to optimize the RMS current of the converter, which will be discussed in the next Section. The control block of the proposed solution is illustrated in Fig.9, of which the boundary law can be obtained from Table 1. As Fig.9 shows, $V_{in1}$ and $V_{in2}$ are the input voltages of two DAB modules, respectively. $\psi_a$ and $\psi_b$ are the phase shift ratio of two DAB modules, respectively. $V_2$ is the output voltage, and $V_{2, ref}$ is the reference value of the output voltage. To control the output voltage, the error between $V_2$ and $V_{2, ref}$ is calculated through the PI controller. Then, the phase shift ratio can be obtained for the two DAB modules. To achieve voltage balance between two DAB modules, the error between $V_{in1}$ and $V_{in2}$ is calculated through the PI controller. Then, $\phi_{ab}$ can be obtained, which is added to $\phi_a$. By slightly adjusting the phase shift ratio of module B, voltage balance between the two modules can be achieved. Meanwhile, the phase $\phi_{ab}$ between two DAB modules can be calculated by the control boundary law.

![Image](image.png)

**FIGURE 9.** The control block of the proposed hybrid SPS control.

IV. PARAMETERS DESIGN CONSIDERATION

As previously analyzed, when the proposed ISOP DAB converter operates under different modes, the equivalent leakage inductance can be changeable by utilizing the modulated coupled inductor, which can reduce the RMS current. However, a suitable coupling coefficient $k_c$ of the modulated coupled inductor should be properly designed to optimize the RMS current. Meanwhile, to achieve full load range ZVS for primary side switches, the voltage ratio $G$ should be designed.

In this paper, a design example is given with the parameters of 250kHz, 2000W, $V_{in} = 800V$, and $V_o = 400V$. And the voltage is considered with ±5% fluctuation under rated operating conditions. With the given parameters of the converter, the leakage inductance can be calculated according to (13) and shown as

$$L_k = \frac{V_1 V_2}{32f_s P_{rated}}$$

**A. VOLTAGE RATIO G AND COUPLED COEFFICIENT kC DESIGN TO ACHIEVE FULL LOAD RANGE ZVS FOR PRIMARY SIDE SWITCHES WITH SMALL RMS CURRENT**

To achieve full load range ZVS for primary side switches, the voltage ratio $G$ should be designed. Fig.10(a), Fig.10(b) and Fig.10(c) plot the normalized RMS current contour lines and the ZVS range of primary side switches with different coupling coefficient $k_c$. As seen, to achieve full load range ZVS for primary side switches, $G$ should be designed in the full load ZVS range region. Meanwhile, $G$ should

![Image](image.png)

**FIGURE 10.** The normalized RMS current contour lines and the ZVS range of primary side switches with different coupling coefficient $k_c$. (a) $k_c = 0.3$. (b) $k_c = 0.5$. (c) $k_c = 0.7$. (d) The operation range for the converter.
also be designed to achieve a relatively small RMS current. As Fig. 10(a), Fig. 10(b), and Fig. 10(c) show, for a certain output power \( P^* \), the RMS current is decreased as \( G \) increases. Therefore, to achieve low RMS current, \( G \) should be designed as large as possible in the full load ZVS range.

For the design example of this paper, considering the voltage fluctuation, the range of \( G \) is designed with \( \pm 5\% \) fluctuation. Fig. 10(d) shows the optimized operation range of \( G \). As seen, to achieve low RMS current, \( G \) should be designed as large as possible in the full load ZVS range.

Fig. 11(a) plots the operation range versus \( k_c \). As seen, \( G \) decreases monotonically with \( k_c \). To find a suitable \( G \), \( k_c \) should be designed first.

Fig. 11(b) plots the normalized RMS current versus \( P^* \) with different \( k_c \) at the rated voltage point. As seen, the normalized RMS current will decrease as \( k_c \) increases. However, when \( k_c \) is designed too large, the normalized RMS current will increase at the middle power point. Therefore, to make a tradeoff for the normalized RMS current over the entire load, \( k_c \) can be designed to be 0.5.

Then, the voltage ratio \( G \) can be designed according to Fig. 11(a), which is 0.857. Afterward, the turns ratio of the transformer can be calculated as

\[
n = \frac{V_1 G}{2V_2} = 0.857
\]  

(29)

B. MAGNETIZING INDUCTANCE \( L_m \) DESIGN TO ACHIEVE FULL LOAD RANGE ZVS FOR SECONDARY SIDE SWITCHES

To achieve full load range ZVS for secondary side switches, the magnetizing inductance of the transformer should be properly designed. However, if the magnetizing inductance is too small, the RMS current of secondary side switches will be increased. Therefore, under the condition of ensuring ZVS, \( L_m \) should be designed as large as possible. Namely, \( K_L \) should be designed as small as possible.

Fig. 12 plots the operation range and the ZVS boundary of the converter. The secondary side ZVS boundary should be designed close to the operating range. As seen in Fig. 12, \( k_L \) should be designed larger than 0.15 to achieve ZVS for secondary side switches in the operation range. Then, the range of magnetizing inductance \( L_m \) can be calculated as

\[
L_m < \frac{L_k}{K_L} = 6.67L_k
\]  

(30)

V. EXPERIMENTAL RESULTS

A 2000W prototype of the proposed ISOP DAB converter is built to verify the effectiveness of the proposed solution,
which is shown in Fig.13. The detailed parameters of the converter are shown in Table 2. To balance the transferred power of two modules, the parameters of two DAB modules are designed to be the same, including the turns ratio of the transformer, magnetizing inductance, and leakage inductance. Besides, when producing the transformers and the coupled inductor, the tolerance of these components should be controlled as small as possible to facilitate the accuracy of power control for the hybrid SPS control scheme. Meanwhile, a comparison between the proposed solution and the conventional individual inductor solution is given. The parameters of the conventional individual inductor solution are also included in Table 2.

The middle point voltages of the H-bridge when the converter operates at steady state are illustrated in Fig.14. As seen, when the converter operates in different modes,
the operating phase between two DAB modules is different. When the converter operates at Mode I, the operating phase between two DAB modules is 0 degree. As Fig.14(a) shows, the phase between $v_{AB}$ and $v_{EF}$ is 0 degree. When the converter operates at Mode II, the operating phase between two DAB modules is 180 degrees. As Fig.14(b) shows, the phase between $v_{AB}$ and $v_{EF}$ is 180 degrees.

Fig.15 illustrates the steady-state waveforms of the converter when $V_2 = 400V$. As seen in Fig.15(a) and Fig.15(b), when the converter operates at light load, the polarity of the primary side and secondary side current is opposite, which helps the converter to achieve full load range ZVS. As Fig.15(c) and Fig.15(d) show the steady-state waveforms of Module A and Module B during heavy load operation.

As seen, the converter can be operated in typical DAB waveforms. Fig.16 shows the ZVS waveforms of the converter when $V_2 = 380V$. As seen in Fig.16(a) and Fig.16(b), when the converter operates at light load, all switches can achieve ZVS. As Fig.16(a) shows, when $S_{1a}$ is turned on, the polarity of $i_{pa}$ is negative, which is helpful for $S_{1a}$ to achieve ZVS. As Fig.16(b) shows, when $S_{3a}$ is turned on, the polarity of $i_{sa}$ is positive, which is helpful for $S_{3a}$ to achieve ZVS. When the converter operates at heavy load, all switches can also achieve ZVS, which can be seen in Fig.16(c) and Fig.16(d). Because the operation condition of the two DAB modes is the same, ZVS for all switches of module B can also be achieved.

Fig.17 shows the ZVS waveforms of the converter when $V_2 = 400V$. As seen in Fig.17(a) and Fig.17(b), the
drain-source voltage of \(S_1a\) and \(S_3a\) have been reduced to zero before the driving signals are coming. Therefore, \(S_1a\) and \(S_3a\) can be achieved ZVS at light load. As Fig.17(c) and Fig.17(d) show, when the converter operates at heavy load, ZVS can also be achieved for \(S_1a\) and \(S_3a\).

Fig.18 shows the ZVS waveforms of the converter when \(V_2 = 420\) V. As Fig.18(a) and Fig.18(b) show, the polarity of \(i_{pa}\) and \(i_{sa}\) is helpful for \(S_1a\) and \(S_3a\) to achieve ZVS at light load. When the converter operated at heavy load, whose waveforms can be seen in Fig.18(a) and Fig.18(b), \(S_1a\) and \(S_3a\) can achieve ZVS.

Fig.19 plots the efficiency comparisons with different solutions of the ISOP DAB converter. As seen, with the proposed solution, the efficiency is improved compared to the conventional individual inductor solution. Especially, the efficiency is significantly improved at light load. When the converter operates when the power is around 700~1100W, the efficiency of the conventional solution is a little higher than that of the proposed method. It can be observed that the proposed solution has more advantages when the converter operates at light load or heavy load.

The qualitative comparison between the proposed method and the different state-of-art control methods is summarized in Table 3. Compared to conventional SPS control [7] and SPS control with the Tapped Transformer [26], the proposed method has two control variables to achieve full load range ZVS and relative optimal RMS current. Because of the magnetizing current injection for the converter, compared to [28], the proposed method can achieve ZVS at light load and improve the light load efficiency. Compared to the EPS control [9] and EPS control with DC-blocking capacitors [18], the optimization process of the proposed method is easier. Although the full load range soft switching achievement and easy optimization for the control method can be realized in [15], it has one more control variable than the proposed method.

| Mode                                      | Number of variables | Soft-sw. range       | Type of soft-sw. | Optimization complexity |
|-------------------------------------------|---------------------|----------------------|------------------|-------------------------|
| SPS [7]                                   | 1                   | Partial              | ZVS              | None                    |
| SPS with Tapped transformer [26]          | 1                   | Partial (improved)   | ZVS              | Low                     |
| SPS with modulated coupled inductor [28]  | 2                   | Partial              | ZVS              | Low                     |
| EPS [9]                                   | 2                   | Partial (improved)   | ZVS              | Medium                  |
| Hybrid EPS with DC-blocking capacitor [18]| 2                   | Full                 | ZVS              | High                    |
| Hybrid trapezoidal, triangular and SPS modulation [15]| 3      | Full                 | ZVS/ZCS          | None                    |
| Proposed method                           | 2                   | Full                 | ZVS              | Low                     |

**VI. CONCLUSION**

This paper proposes an ISOP DAB converter based on the modulated coupled inductor for the DC transformer application, and a corresponding hybrid SPS control is also proposed. To optimize the RMS current, the modulated coupled inductor is utilized for the converter. Meanwhile, a corresponding control boundary is introduced to modulate the coupled inductor and control the equivalent leakage inductance, leading to optimization for the RMS current. To achieve full load range ZVS, the design process is studied for the parameters of the transformers, including the turns ratio and the magnetizing inductance. From the experimental results of a 2000W prototype, it can be verified that the full load range ZVS can be achieved for all switches. Besides, by comparing with the conventional individual leakage inductors solution, the efficiency of the proposed method is significantly improved at light load and heavy load. Meanwhile, when the ISOP system has more DAB modules, the proposed ISOP DAB converter with the modulated coupled inductor can be regarded as a sub-module. Then, each sub-module can be designed and controlled with the proposed method.

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