

Synchronous PWM Modulation to Eliminate Circulating Power in Bidirectional Dual-active-bridge Converter

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Abstract—To solve the problem of circulating power of dual active bridge (DAB) DC-DC converter over a wide voltage conversion ratio, this paper proposes a novel synchronous PWM (S-PWM) modulation. Existence of circulating power increases current stress of devices and decreases efficiency, especially under light load conditions. Several modulation methods have been proposed to overcome the problem. They can reduce or eliminate either input or output side circulating power. In contrast, S-PWM not only eliminates both sides circulating power and reduces current stress, but also achieves zero-current-switching (ZCS) turn-on for all switches and ZCS turn-off for most across the full power range. No auxiliary or snubber circuits are increased. In addition, the control can be simplified so the transmitted power is related to only one variable. The S-PWM has four cases under different gain and power conditions. The detailed operation principle and modes of DAB under S-PWM are analyzed in the paper. In addition, four modulations in literature are discussed, and corresponding comparative analyses with S-PWM are given. Finally, a laboratory prototype is built to verify advantages and effectiveness of the proposed modulation.

Index Terms—Bidirectional converter, circulating power, DC-DC converter, dual active bridge (DAB), soft switching.

I. INTRODUCTION

THE dual active bridge (DAB) dc-dc converter has gained extensive attention since it was proposed, which consists of two active full bridges and a high-frequency transformer [1]. It has the advantages of simple and symmetrical structure, galvanic isolation, bidirectional power flow, and soft-switching ability [2]–[4]. Therefore, DAB plays an important role in power conversion applications, such as energy sources [5], energy router [6], hybrid electric vehicles [7], uninterruptible power supply [8], etc. However, DAB suffers problems of circulating power and limited soft-switching range under conventional modulations. Circulating power represents the

reverse power that flows into the input source and out of the output source. Circulating power is not transmitted to the load in reality but increases current stress and conduct loss of the converter. To solve the problem of circulating power, several modulation methods and techniques have been proposed in literature [9]–[11].

Single-phase-shift (SPS) modulation is a simple and widely used modulation [12]. It controls power transfer by the phase shift ratio between output voltages of primary and secondary full bridges. SPS has best performance when gain is united [13]. But when gain is far from unified, the circulating power will be huge and increase current stress [14]. Especially when operating in light load conditions, circulating power will occupy a considerable proportion, and soft-switching ability may be lost. In literature, extended-phase-shift (EPS) [15], dual-phase-shift (DPS) [16], triple-phase-shift (TPS) [17] and other improved modulation methods are gradually proposed [18]–[20]. They can reduce circulating power through more flexible control. EPS modulation introduces an inner phase shift to the primary full bridge. It extends control freedom and power transfer range. In addition, it can reduce backflow power sent back to the input source [21]. DPS further introduces an inner phase shift to the secondary full bridge. Three phase shift ratios are further obtained in TPS. Therefore, control flexibility is further improved, and more complex control can be realized [22]–[24]. All EPS, DPS and TPS can reduce circulating power and current stress through control. However, the proportion of circulating power is still non-negligible, and control can be very complicated. To minimize circulating power, an optimal three-level phase-shift control strategy is proposed in [25]. It unifies all operation modes of DAB by harmonic analysis and builds a complex mathematic model. However, the analysis and control process is too complex.

Modulations mentioned above can reduce circulating power, but cannot be employed to eliminate it. In [26], [27], a masked drive modulation (MDM) is proposed to overcome the problem of switching surge under light load conditions, which is caused by loss of soft-switching. MDM masks drive signals at specific intervals to prevent reverse of leakage inductor current. It can eliminate both switching surges and output side circulating power. However, it only works under light load conditions, and SPS is employed when the power boundary is crossed. In [28]–[30], novel semi-DAB converters are proposed. By replacing some switches of DAB with diodes, semi-DAB can eliminate

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input or output side circulating power. But it can only be used for unidirectional power flow. In [31], a novel front and rear end switch DAB is proposed. It connects additional switches in input and output side ports compared to conventional DAB, and both input and output sides circulating power can be eliminated through control of auxiliary switches. However, two additional active switches are employed, which means high cost and complex control methods.

This paper proposed a synchronous PWM (S-PWM) modulation with no inner and outer phase shifts in the primary and secondary full bridges. It can eliminate both input and output sides circulating power over a wide voltage conversion ratio without adding auxiliary and snubber circuits. Therefore, conduct loss and current stress can be reduced. Besides, all switches can achieve zero-current-switching (ZCS) turn-on and most switches can achieve ZCS turn-off within full power range, which means there is no worry for switching surges. Therefore, losses are minimized, and efficiency can be improved. The proposed S-PWM modulation has four cases according to the voltage gain and transmission power. Control can be simplified since transmitted power is related to only one variable.

This paper is organized as follows. In Section II, the problem of circulating power and ideal operating waveforms of SPS, EPS, and MDM are discussed. Besides, the operation principle and modes analysis of the S-PWM is introduced. Then, performance analysis of the S-PWM, and comparison analysis with SPS, EPS, TPS, and MDM are given in Section III. In Section IV, experimental results are given, which verify advantages and effectiveness of the proposed S-PWM modulation. Finally, Section V concludes this paper.

II. OPERATE PRINCIPLE OF DAB AND PROBLEM OF CIRCULATING POWER

A. Topology Description and Definition of Circulating Power

A circuit diagram of a basic DAB dc-dc converter is shown in Fig. 1. Input and output side active full bridges are linked by a high-frequency transformer. L_k is leakage inductance of the

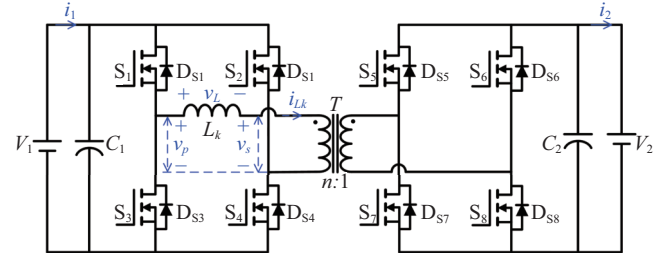


Fig. 1. Circuit diagram of the DAB dc-dc converter.

transformer, n is turns ratio, S_1 – S_8 are active switches, and C_1 & C_2 are filter capacitors. Furthermore, v_p is output voltage of the primary full bridge and v_s is output voltage of the secondary full bridge referred to the primary side, v_L and i_{Lk} are voltage and current of the leakage inductor, respectively, i_1 and i_2 are output and input current of V_1 and V_2 , respectively.

There are three flow forms of power in the DAB, power flows from the input source to output load is defined as active power, and power flows back to the input source and out of the output source is defined as circulating power. In this paper, V_1 and V_2 are set as input and output sources, respectively, and the opposite case is not discussed since DAB is symmetrical in the view of topology and control. When power flows from V_1 to V_2 , i_1 and i_2 should be positive to transfer active power. Otherwise, circulating power will be produced when the direction of i_1 or i_2 is reversed.

B. Modulation Methods in the Literature

1) SPS Modulation

For SPS, 50% of the duty cycle is employed for active switches of both primary and secondary full bridges. In addition, power transfer is controlled by the outer phase shift between output voltages of primary and secondary full bridges. Ideal operating waveforms of the DAB under SPS are shown in Fig. 2(a). D is outer phase shift ratio. As can be seen, in intervals t_0 – t_2 and t_3 – t_5 , value of i_1 or i_2 is negative. It means the blue shaded areas of current in Fig. 2(a) represent circulating power. For required transmission power, transmitted

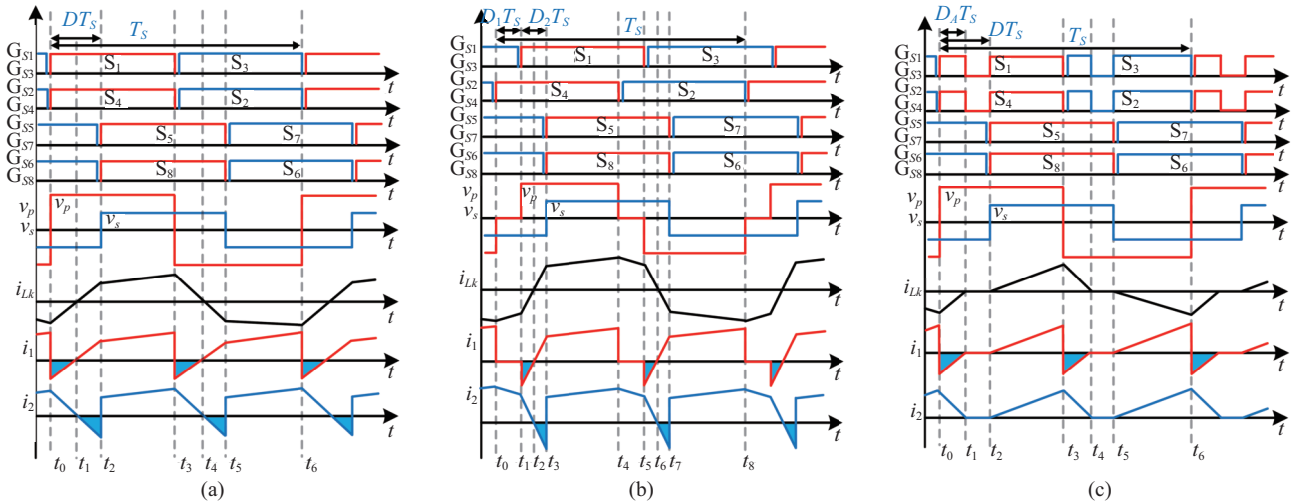


Fig. 2. Ideal waveforms of DAB under different modulations in literature. (a) SPS modulation. (b) EPS modulation. (c) MDM modulation.

active power has to be increased to compensate for circulating power, which increases current stress and conduction loss of the converter. In addition, performance of DAB under SPS will deteriorate further when operating in light load mode.

To visually observe impact of circulating power, Fig. 3 shows peak current curve of the converter with different transmission power. It is drawn according to the equation (14). Orange and yellow sections represent the increment of peak current with increasing transmitted power. As can be seen, peak current does not change very much as output power increases exponentially. That means there is a considerable portion of current used to produce circulating power. It does not contribute to power transmission but increases conduct loss and current stress.

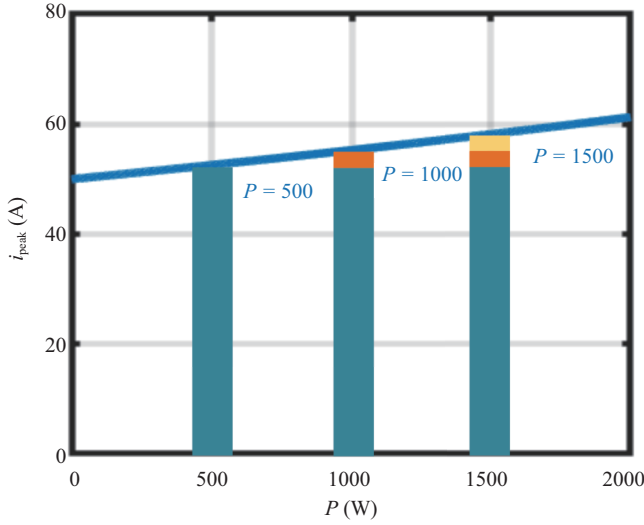


Fig. 3. Peak current of DAB with different transmitted power under SPS modulation.

2) EPS Modulation Proposed in [21]

Compared with SPS, EPS increases the inner phase shift ratio of the primary bridge to make output voltage of the primary bridge a three-level voltage. Ideal operation waveforms of DAB under EPS are shown in Fig. 2(b). D_1 is inner phase shift ratio between drive signals of S_1 & S_4 or S_2 & S_3 , and D_2 is the phase shift ratio between v_p and v_s . Blue shaded areas displayed represent input and output side circulating power. To reduce backflow power, D_1 and D_2 are restricted by $0 < D_1 < 1$, $0 < D_2 < 1$, and $0 < D_1 + D_2 < 1$. From Fig. 2(b), due to three-level of v_p , negative interval (t_0-t_1) of i_1 in SPS is divided into two intervals (t_0-t_1 and t_1-t_2) in EPS. i_1 keeps zero in the interval t_0-t_1 , which means circulating power is zero. Therefore, overall circulating power can be reduced for given transmitted power. Besides, input side circulating power can be eliminated when $t_1 = t_2$. But

the negative interval of i_2 always exists so that output side circulating power cannot be eliminated.

3) MDM Proposed in [27]

MDM only works under light load conditions, the boundary of light and heavy loads for MDM has been given in [27]. Fig. 2(c) shows ideal operation waveforms of DAB under MDM modulation. As can be seen, it masks drive signals of specific switches in intervals t_1-t_2 and t_4-t_5 to prevent reverse of leakage inductor current i_{Lk} and output current i_2 . MDM eliminates output side circulating power and overcomes the problem of switching surges under light load conditions. However, input side circulating power still exists, which is represented by the blue shaded areas in Fig. 2(c).

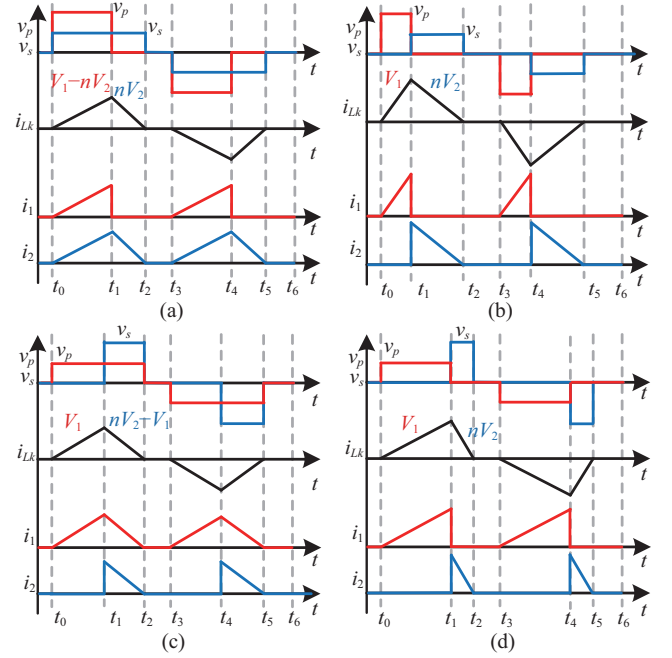


Fig. 4. Four cases of S-PWM modulation. (a) Case A. (b) Case B. (c) Case C. (d) Case D.

C. Operation Principle of the Proposed S-PWM Modulation

When power is forward transmitted, S-PWM can be divided into four cases according to voltage gain $G = nV_2/V_1$ and maximum transmitted power range. As shown in Table I, different cases have different rising or falling slopes of i_{Lk} . It should be noted $G = 1$ corresponds to case B and D. $P_{\max_A}$, $P_{\max_B}$, $P_{\max_C}$, and $P_{\max_D}$ are maximum transmission power of cases A, B, C, and D, respectively. All cases can operate from zero to their maximum transmission power, and their suitable operating load ranges are given in Table I. Fig. 4 shows the waveforms of DAB under four cases. As can be seen, they all can eliminate both input and output sides

TABLE I
FOUR CASES OF S-PWM MODULATION

Cases	Condition	Rising slope of i_{Lk}	Falling slope of i_{Lk}	Suitable operating load range
A	$G < 1$	$(V_1 - nV_2)/L_k$	nV_2/L_k	$0 < P < P_{\max_A}$
B		V_1/L_k	nV_2/L_k	$P_{\max_A} < P < P_{\max_B}$
C		V_1/L_k	$(nV_2 - V_1)/L_k$	$0 < P < P_{\max_C}$
D	$G > 1$	V_1/L_k	nV_2/L_k	$P_{\max_C} < P < P_{\max_D}$

circulating power. Therefore, current stress of DAB can be reduced compared with the above modulations under the same transmitted power.

On the other hand, peak current of i_{Lk} in case A is lower than B under the same transmitted power, and $P_{\max_A}$ is smaller than $P_{\max_B}$, too. Similarly, case C has a lower peak current of i_{Lk} and smaller maximum transmission power than D. Therefore, it is suitable to operate preferentially in cases A and C. Cases B and D are suitable when load is beyond the maximum transmission power of cases A and C. To simplify analysis, case A is an example for detailed analysis, and other cases have similar characteristics. In the following, S-PWM represents case A of S-PWM modulation.

Figure 5 shows ideal operation waveforms of the DAB converter under S-PWM. D and D^* are duty cycle ratio when i_{lk} decreases and increases, respectively. Drive signals of up and down switches in a bridge leg have the same duty cycle ratio and 180° phase shift. From Fig. 5, current of leakage inductor i_{lk} is restricted to not reverse during the half-cycle. Thanks to this restriction, input and output currents i_1 & i_2 will not flow in a reverse direction, which eliminate circulating power. In addition, all switches can achieve ZCS turn-on, and all switches can achieve ZCS turn-off except S_1 and S_3 . Detailed analysis and mathematical expressions are illustrated as follows.

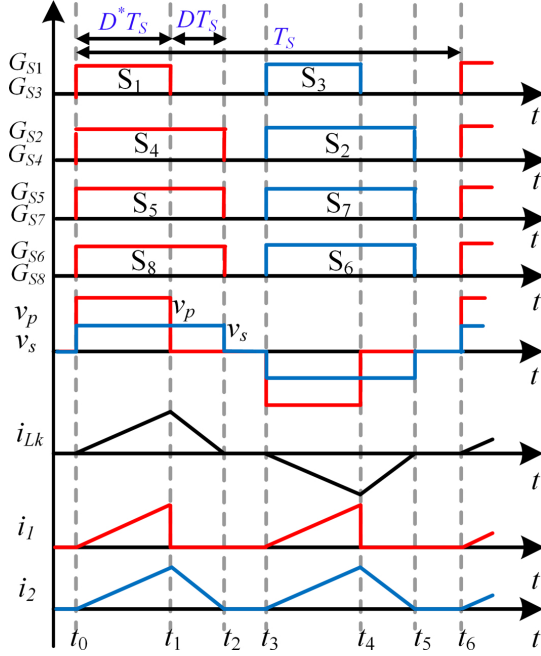


Fig. 5. Ideal operation waveforms of DAB under S-PWM modulation.

From Fig. 5, a switching cycle can be divided into six operation modes. To simplify analysis, we assume that filter capacitors are big enough to constantly input and output voltages, and parasitic parameters of devices can be neglected. Due to symmetry of the first and second half-cycles, only three modes of the first half cycle (t_0 – t_3) are analyzed.

Mode 1 (t_0 – t_1): Equivalent circuits of this mode is shown in Fig. 6(a). Before t_0 , all switches are turned off, and leakage inductor current is zero. Therefore, S_1 , S_4 , S_5 , and S_8 are

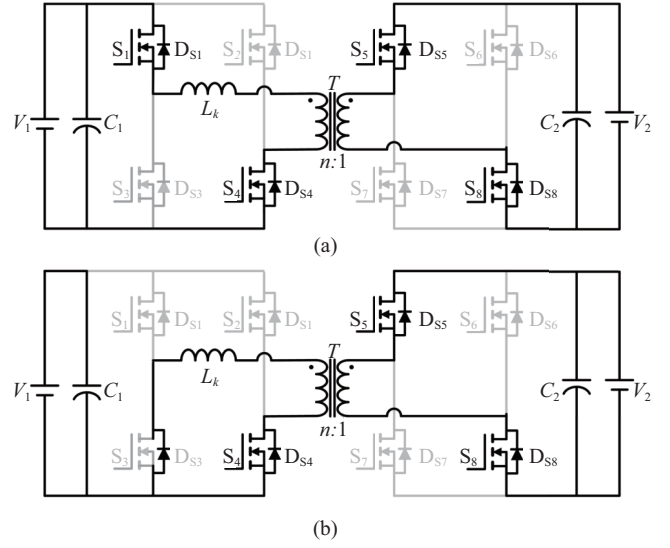


Fig. 6. Equivalent circuit of DAB under S-PWM modulation. (a) Mode 1. (b) Mode 2.

turned on with ZCS at t_0 . In this interval, voltage of the leakage inductor is clamped by $v_L = (V_1 - nV_2)$, and i_{lk} increases linearly. Input current i_1 is equal to i_{lk} in this interval. Related mathematic expressions are derived as:

$$i_1(t) = \frac{i_2(t)}{n} = i_{Lk}(t) = \frac{V_1 - nV_2}{L_k}(t - t_0) \quad (1)$$

Mode 2 (t_1 – t_2): Equivalent circuit of this mode is shown in Fig. 6(b). At t_1 , S_1 is turned off, and then the primary current flows through the body diode of S_3 . In this interval, the leakage inductor is clamped by voltage $v_L = nV_2$ and i_{lk} decreases linearly to zero. Related mathematic expressions are derived as:

$$\frac{i_2(t)}{n} = i_{Lk}(t) = i_{Lk}(t_1) - \frac{nV_2}{L_k}(t - t_1) \quad (2)$$

Mode 3 (t_2 – t_3): At t_2 , leakage inductor current drops to zero, and switches (S_4 , S_5 , S_8) are turned off with ZCS. All switches remain turned-off during this interval. In addition, this operation mode will be lost when the converter operates at the maximum transmission power point.

III. PERFORMANCE ANALYSIS AND COMPARISON

A. Power Transfer Analysis of S-PWM Modulation

From Fig. 5, assuming $t_0 = 0$ and switching cycle is T_s , we can get $t_1 = D^*T_s$, $t_2 = (D^* + D)T_s$, $t_3 = 0.5T_s$, $t_4 = (0.5 + D^*)T_s$, $t_5 = (0.5 + D^* + D)T_s$, $t_6 = T_s$. According to analysis in the previous section, from (1) and (2), the relationship between D^* and D is

$$D^* = \frac{nV_2}{V_1 - nV_2}D \quad (3)$$

and leakage inductor current at t_1 and t_4 can be derived as:

$$i_{Lk}(t_1) = \frac{nV_2}{L_k f_s}D \quad (4)$$

$$i_{Lk}(t_4) = -\frac{nV_2}{L_k f_s}D \quad (5)$$

where $f_s = 1/T_s$ is the switching frequency. Then peak current and average input current can be derived as:

$$i_{p_S-PWM} = i_{Lk}(t_1) = \frac{nV_2}{L_k f_s} D \quad (6)$$

$$I_{1_S-PWM} = \frac{2}{T_s} \int_{t_0}^{t_1} i_{Lk}(t) dt = \frac{(nV_2)^2 D^2}{(V_1 - nV_2) L_k f_s} \quad (7)$$

then the transmitted power is

$$P_{S-PWM} = I_1 V_1 = \frac{(V_1 - nV_2) V_1 D^{*2}}{L_k f_s} = \frac{(nV_2)^2 V_1 D^2}{(V_1 - nV_2) L_k f_s} \quad (8)$$

As can be seen, transmitted power can be adjusted by D . Maximum power transfer point is reached when $D^* + D = 0.5$. From (3) and (8), maximum transmission power and corresponding duty cycle ratio can be derived

$$P_{\max_S-PWM} = \frac{(nV_2)^2 (V_1 - nV_2)}{4V_1 L_k f_s} \quad (9)$$

$$D_{\max_S-PWM} = \frac{V_1 - nV_2}{2V_1} \quad (10)$$

The curve of normalized transmission power with respect to duty ratio D^* is plotted in Fig. 7, which is normalized by $P_{\text{base}} = (nV_2)^2 V_1 / [16(V_1 - nV_2) L_k f_s]$. As can be seen, transmitted power is increasing monotonically with respect to D^* . This helps simplify the control algorithm. Fig. 8 shows

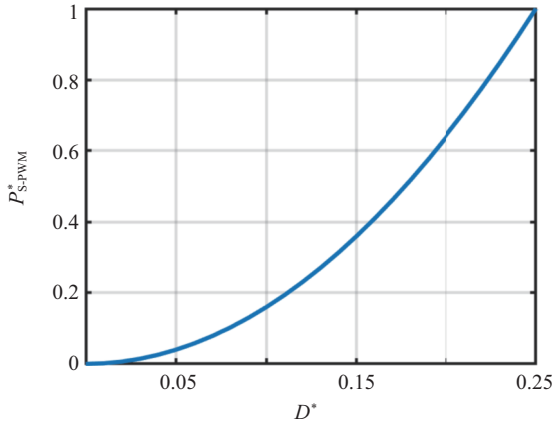


Fig. 7. Normalized power transfer curve with duty cycle ratio D^* under S-PWM modulation.

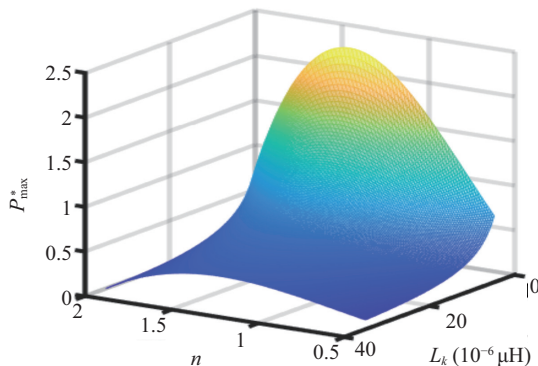


Fig. 8. Curve of the normalized maximum transmission power varies with turn ratio n and leakage inductance L_k .

the curve of unified maximum transmission power varies with n and L_k . As illustrated, for given input and output voltages, maximum transmission power can be adjusted by n and L_k to meet design requirements.

B. Comparative Analysis

1) Comparison of Peak Current Stress with SPS, MDM, and S-PWM

Peak current stress comparison of DAB under different modulations should be based on the same transmitted power. From (6) and (8), peak current of S-PWM with respect to transmitted power can be derived

$$i_{p_S-PWM_P} = \sqrt{\frac{(V_1 - nV_2)P}{V_1 L_k f_s}} \quad (11)$$

For SPS, transmission power and peak current are [16]

$$P_{\text{SPS}} = \frac{nV_1 V_2}{L_k f_s} D(1 - 2D) \quad (12)$$

$$i_{p_SPS} = \frac{V_1 + nV_2(4D - 1)}{4L_k f_s} \quad (13)$$

Using (12) and (13), peak current with respect to transmitted power can be derived

$$i_{p_SPS_P} = \frac{V_1^2 - \sqrt{nV_1 V_2(nV_1 V_2 - 8L_k f_s P)}}{4V_1 L_k f_s} \quad (14)$$

For MDM, transmission power and peak current are [27]

$$P_{\text{MDM}} = \frac{nV_1 V_2}{2L_k f_s} \cdot \frac{V_1 - nV_2}{V_1 + nV_2} \cdot (1 - 2D)^2 \quad (15)$$

$$i_{p_MDM} = \frac{(V_1 - nV_2)}{L_k f_s} \cdot \left(\frac{1}{2} - D\right) \quad (16)$$

Using (15) and (16), peak current with respect to transmitted power can be derived

$$i_{p_MDM_P} = \sqrt{\frac{(V_1^2 - n^2 V_2^2)P}{2nV_1 V_2 L_k f_s}} \quad (17)$$

Using (11), (14), and (17), comparative curves of normalized peak current stress varied with transmitted power under different n can be shown in Fig. 9. Currents are normalized by $i_{\text{base}} = nV_2 / L_k f_s$. As can be seen, peak current stress of SPS is largest, especially when transmitted power is low. MDM successfully reduces current stress during light load conditions and gradually approaches SPS as power increases. In contrast, the proposed S-PWM reduces peak current stress significantly, which is due to elimination of circulating power.

2) Comparison of Peak Current Stress with EPS and S-PWM

To compare the EPS and S-PWM, the relationship of D and D_1 & D_2 under the same transmitted power should be derived first. For EPS, transmission power and peak current are [21]

$$P_{\text{EPS}} = \frac{nV_1 V_2}{2f_s L_k} \left[D_2(1 - D_2) + \frac{1}{2} D_1(1 - D_1 - 2D_2) \right] \quad (18)$$

$$i_{p_EPS} = \frac{nV_2}{4f_s L_k} \left[\frac{V_1}{n \cdot V_2} (1 - D_1) + (2 \cdot D_1 + 2D_2 - 1) \right] \quad (19)$$

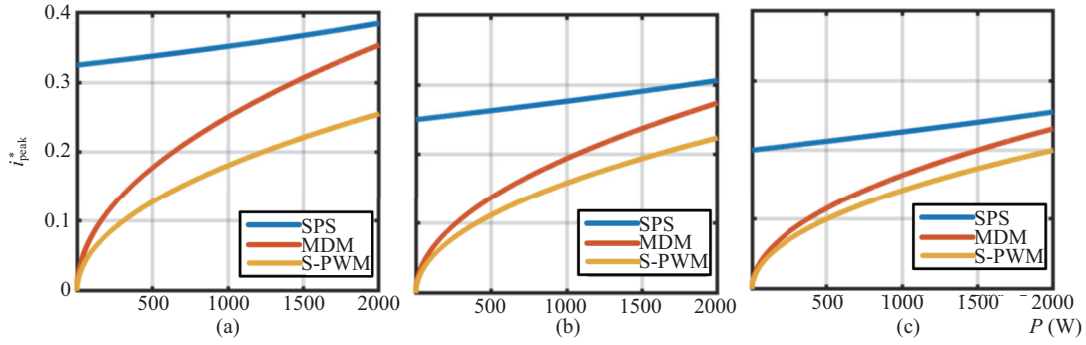


Fig. 9. Comparative curves of normalized peak current stress i_{peak}^* varied with transmitted power P . (a) $n = 0.7$. (b) $n = 1$. (c) $n = 1.2$.

Using (8) and (18), with specific values of D_1 and D_2 in EPS, phase shift ratio in S-PWM can be derived

$$D = \sqrt{\frac{(V_1 - nV_2)(2D_2 + D_1 - 2D_2^2 - D_1^2 - 2D_1D_2)}{4nV_2}} \quad (20)$$

Then, peak current of S-PWM with respect to D_1 and D_2 can be derived from (6) and (20)

$$i_{\text{p_S-PWM_D12}} = \frac{\sqrt{nV_2(V_1 - nV_2)(D_1 - 2D_1D_2 - D_1^2 - 2D_2^2 + 2D_1D_2)}}{2L_k f_s} \quad (21)$$

From (19) and (21), 3-D curves of normalized peak current stress under EPS and S-PWM can be shown in Fig. 10. Parameters of EPS are designed according to [18]. It is clear from Fig. 10 the curve of EPS is always above S-PWM, which means higher peak current stress.

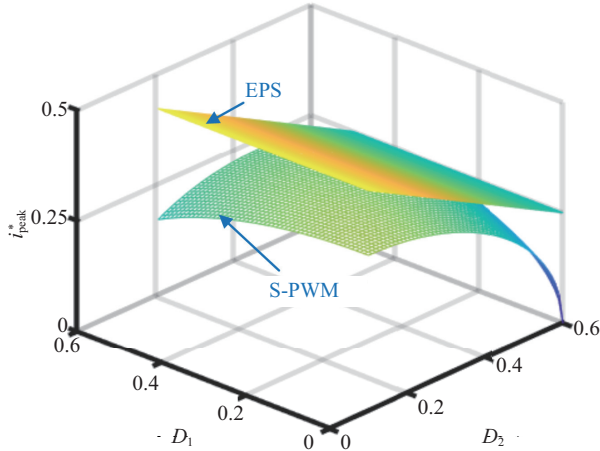


Fig. 10. 3-D curves of normalized peak current stress of EPS and S-PWM modulation.

In conclusion, though the S-PWM tends to have a higher peak current due to discontinuous current, elimination of circulating power and a reasonable duty cycle help to reduce peak current significantly. Therefore, S-PWM can always keep a lower peak current compared to continuous current modulations with circulating power. Besides, the S-PWM also has an advantage in input and output ripple, which helps save filters.

3) Comparison of Circulating Power with SPS, EPS, MDM, and S-PWM

Current areas that produce circulating power are shown in Fig. 2(a), (b), and (c) with blue shadows. For SPS, circulating power appears during intervals t_0-t_2 and t_3-t_5 in Fig. 2(a). Then, circulating power expressions of V_1 and V_2 sides can be derived

$$P_{\text{cir_SPS_1}} = \frac{2}{T_s} \int_{t_0}^{t_1} -V_1 i_{Lk} dt = \frac{V_1 [V_1 + nV_2(4D - 1)]^2}{16L_k f_s (V_1 + nV_2)} \quad (22)$$

$$P_{\text{cir_SPS_2}} = \frac{2}{T_s} \int_{t_1}^{t_2} nV_2 i_{Lk} dt = \frac{nV_2 [V_1(4D - 1) + V_2 n]^2}{16L_k f_s (V_1 + nV_2)} \quad (23)$$

From (22) and (23), total circulating power of SPS is

$$P_{\text{cir_SPS}} = \frac{V_1 [V_1 + nV_2(4D - 1)]^2 + nV_2 [V_1(4D - 1) + V_2 n]^2}{16L_k f_s (V_1 + nV_2)} \quad (24)$$

For MDM, only V_1 side circulating power is produced during interval t_0-t_1 in Fig. 2(c). Similarly, circulating power of MDM can be derived

$$P_{\text{cir_MDM}} = \frac{2}{T_s} \int_{t_0}^{t_1} -V_1 i_{Lk} dt = \frac{V_1 (V_1 - nV_2)^2 (2D - 1)^2}{4L_k f_s (V_1 + nV_2)} \quad (25)$$

For EPS, circulating power appears during intervals t_1-t_3 and t_5-t_7 in Fig. 2(b). Then, circulating power expressions of V_1 and V_2 sides can be derived

$$P_{\text{cir_EPS_1}} = \frac{2}{T_s} \int_{t_1}^{t_2} -V_1 i_{Lk} dt = \frac{V_1 [(1 - D_1)V_1 - nV_2(1 - 2D_2)]^2}{16L_k f_s (V_1 + nV_2)} \quad (26)$$

$$P_{\text{cir_EPS_2}} = \frac{2}{T_s} \int_{t_2}^{t_3} -nV_2 i_{Lk} dt = \frac{\left(nV_2 [(D_1 - 1 + 2D_2)V_1 + nV_2][(D_1 - 1) + 4D_2]V_1 + (1 + 2D_2)nV_2 \right)}{16L_k f_s (V_1 + nV_2)} \quad (27)$$

From (26) and (27), total circulating power of EPS is

$$P_{\text{cir_EPS}} = \frac{V_1 [(1 - D_1)V_1 - nV_2(1 - 2D_2)]^2}{16L_k f_s (V_1 + nV_2)}$$

$$+ \frac{\left(\frac{nV_2[(D_1 - 1 + 2D_2)V_1 + nV_2][(D_1 - 1 + 4D_2)V_1 + (1 + 2D_2)nV_2]}{16L_k f_s (V_1 + nV_2)} \right)}{(28)$$

To compare SPS and MDM with EPS, variable D in SPS and MDM should be mapped to D_1 and D_2 . Similarly, based on the same transmitted power, mapping relationships can be derived

$$D_{\text{SPS}} = \frac{1 + \sqrt{(D_1 + 2D_2)^2 + D_1^2 - 2D_1 - 4D_2 + 1}}{4} \quad (29)$$

$$D_{\text{MDM}} = \frac{1}{2} + \frac{1}{4} \sqrt{\frac{\left(\frac{2(V_1 + nV_2)[D_1 + 2D_2]}{V_1 - nV_2} - D_2^2 - (D_1 + D_2)^2 \right)}{V_1 - nV_2}} \quad (30)$$

From (24), (25), (29), and (30), circulating power of SPS and MDM with respect to D_1 and D_2 can be derived

$$P_{\text{cir_SPS_D12}} = \frac{1}{16L_k f_s} [(D_1 - 1)^2 + (2D_2 + D_1)^2 + 2\sqrt{2D_1(D_1 + 4D_2 - 1) + (2D_2 - 1)^2 - 4D_2 - 1}] \quad (31)$$

$$P_{\text{cir_MDM_D12}} = \frac{\left(\frac{V_1(V_1 - nV_2)[D_1 + 2D_2]}{-D_2^2 - (D_1 + D_2)^2} \right)}{8L_k f_s} \quad (32)$$

Using (28), (31), and (32), the 3-D and 2-D curves of circulating power under SPS, MDM, and EPS are drawn in Figs. 11 and 12, respectively. As can be seen, the circulating power of SPS is larger than MDM and EPS with the same transmitted power, which results in a larger peak current. MDM and EPS can reduce circulating power, but considerable values remain. In contrast, the proposed S-PWM modulation eliminates circulating power by restricting current flow. Transmission of invalid power can be avoided. Therefore, current stress and conduction loss can be reduced, and efficiency can be increased.

4) Comparison of Peak Current and Circulating Power with TPS and S-PWM

TPS can achieve all kinds of control algorithms with respect to various operating conditions and different coordination of

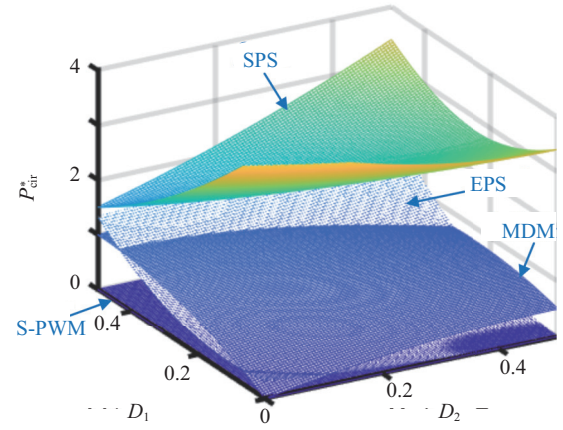


Fig. 11. 3-D curves of normalized circulating power under SPS, EPS, MDM, and S-PWM modulations.

three phase shift ratios. Due to the complex control three coupled variables have to be calculated, the common way to employ TPS control in the microcontroller is to save optimization results in a look-up table in advance [32]. Take optimal operating mode TPS_III in [17] as an example. When minimum peak inductor current strategy is employed to design coordination of three phase shift ratios, expressions of peak current and circulating power with respect to transmitted power can be given as [17]

$$i_{p_TPS_P} = \frac{\sqrt{L_k f_s G(G-1)P}}{GL_k f_s} \quad (33)$$

$$P_{\text{cir_TPS_P}} = \frac{8(G-1)^2 L_k f_s P}{GV_1} \cdot \sqrt{\frac{2}{V_1} \sqrt{\frac{G}{(G-1)L_k f_s P}} - \frac{8}{3V_1^2}} \quad (34)$$

From (11), peak current expression of S-PWM with respect to transmitted power can be given as

$$i_{p_S\text{-PWM_P}} = \sqrt{\frac{(G-1)V_1 P}{GV_1 L_k f_s}} \quad (35)$$

From (33) and (35), comparative curves of peak current stress under TPS and S-PWM can be shown in Fig. 13(a).

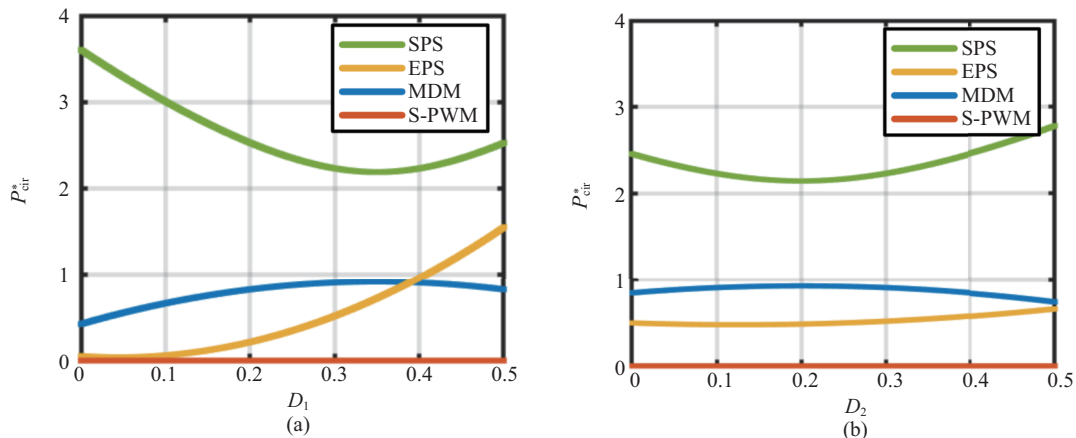


Fig. 12. 2-D curves of normalized circulating power under SPS, EPS, MDM, and S-PWM modulations. (a) $D_2 = 0.3$. (b) $D_1 = 0.3$.

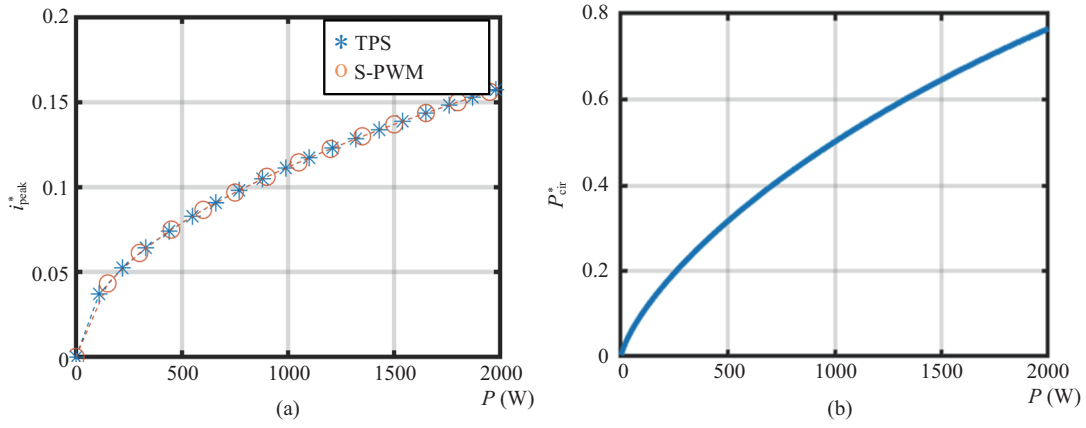


Fig. 13. Curves of normalized parameters under TPS. (a) Comparative curves of normalized peak current stress with S-PWM. (b) Curve of normalized circulating power

TABLE II
PERFORMANCE COMPARISONS

Components	Circulating power	Peak current stress	Auxiliary devices	Operation range	Number of control variables	Soft-switching	
						Type	Range
S-PWM modulation	Eliminate	Low	No	Full operating power	1	S_1 – S_8 : ZCS turn-on S_2 & S_4 – S_8 : ZCS turn-off	Full operating power range
SPS modulation	High	High	No	Full operating power	1	S_1 – S_8 : ZVS turn-on	Except light load
EPS modulation	Low	Middle	No	Full operating power	2	S_1 – S_8 : ZVS turn-on	Except light load
TPS Modulation	Low	Low	No	Full operating power	3	S_1 – S_8 : ZVS turn-on	Large operating power range
MDM modulation	Low	Middle	No	Light load	1	S_1 – S_4 : ZVS turn-on S_5 – S_8 : ZCS turn-on(off)	Light load
Ref. [30]	Eliminate	Low	Yes	Full operating power	1 or 2	S_1 – S_8 : ZVS turn-on	Except light load

From (34), the curve of normalized circulating power under TPS can be shown in Fig. 13(b). As illustrated, peak current of TPS is similar to S-PWM. However, circulating power still exist in TPS modulation and increases with transmitted power. It means there is always some additional active power to compensate for circulating power, which will increase the conduction loss. Additionally, the phase-shift trajectory of TPS has to be obtained using look-up tables, considering it is extremely complex to calculate online [17]. In contrast, S-PWM eliminates circulating power, and modulation can be easily achieved where only one phase shift variable is employed.

Performance comparisons between SPS, EPS, TPS, MDM, FRS-DAB presented in [31], and S-PWM are summarized in Table II. As can be seen, the proposed S-PWM modulation eliminates circulating power without adding auxiliary devices, and better soft-switching characteristic can be kept during full operating power range. However, SPS, EPS, TPS, and MDM cause circulating power during the full operating power range and soft-switching range is limited. Additionally, TPS can significantly reduce circulating power and current stress by reasonably coordinating three phase shift ratios. However, the phase-shift trajectory is extremely complex and influences maximum soft switch range control. The FRS-DAB eliminates circulating power by increasing two switches, which means complex control and increased cost.

C. Loss Analysis of S-PWM Modulation

The equivalent circuit considering parasitic parameters is

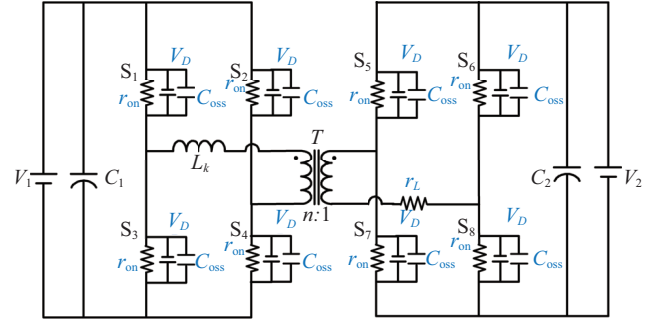


Fig. 14. Equivalent circuit considering parasitic parameters.

shown in Fig. 14, where r_{on} is on-resistance of switches, V_D is forward voltage drop of body diode of switches, C_{oss} is parasitic capacitance of switches, and r_L represents line resistance.

Referring to the Fig. 5, Fig. 6, and operation principle analysis of the S-PWM, conduction loss of S_1 & S_2 , S_3 & S_4 , and S_5 – S_8 are equivalent. Their conduction loss can be derived as

$$P_{loss_on_S1_S2} = \frac{1}{T_s} \int_0^{t_1} i_{S1}^2 r_{on} dt = \frac{D^3 n^3 V_2^3 r_{on}}{3L_k^2 f_s^2 (V_1 - nV_2)} \quad (36)$$

$$P_{loss_on_S3_S4} = \frac{1}{T_s} \int_0^{t_2} i_{S4}^2 r_{on} dt = \frac{r_{on} D^3 V_2^2 n^2 (7V_1^2 - 22V_1 nV_2 + 18(nV_2)^2)}{3L_k^2 f_s^2 (V_1 - nV_2)^2} \quad (37)$$

$$\begin{aligned}
P_{\text{loss_on_S5-S8}} &= \frac{1}{T_s} \int_0^{t_2} i_{S5}^2 r_{\text{on}} dt \\
&= \frac{r_{\text{on}} D^3 n^4 V_2^2 (7V_1^2 - 22V_1 n V_2 + 18n^2 V_2^2)}{3L_k^2 f_s^2 (V_1 - nV_2)^2} \quad (38)
\end{aligned}$$

Similarly, discharge loss of parasitic capacitance of S_1 – S_4 and S_5 – S_8 are equivalent, respectively. Their discharge loss can be derived as

$$P_{\text{loss_Coss_S1-S4}} = \frac{1}{T_s} \frac{1}{2} C_{\text{oss}} V_1^2 = \frac{f_s C_{\text{oss}} V_1^2}{2} \quad (39)$$

$$P_{\text{loss_Coss_S5-S8}} = \frac{1}{T_s} \frac{1}{2} C_{\text{oss}} V_2^2 = \frac{f_s C_{\text{oss}} V_2^2}{2} \quad (40)$$

As for body diodes and line resistance, conduction loss can be calculated as

$$\begin{aligned}
P_{\text{loss_D_S1-S3}} &= \frac{1}{T_s} \int_{t_1}^{t_2} -i_{S3} V_D dt \\
&= \frac{D^2 V_{D3} n V_2 (3V_1 - 5nV_2)}{2L_k f_s (V_1 - nV_2)} \quad (41)
\end{aligned}$$

$$\begin{aligned}
P_{\text{loss_Line}} &= \frac{1}{T_s} \int_0^{t_6} i_{Lk}^2 r_L dt \\
&= \frac{2r_L D^3 n^2 V_2^2 (7V_1^2 - 22V_1 n V_2 + 18n^2 V_2^2)}{3L_k^2 f_s^2 (V_1 - nV_2)^2} \quad (42)
\end{aligned}$$

Based on (36)–(42) and Table III, Fig. 15 gives the loss ratio of output power at 500, 1125, and 1500 W when $r_{\text{on}} = 8 \text{ m}\Omega$, $V_D = 1.3 \text{ V}$, $C_{\text{oss}} = 810 \text{ pF}$, and $r_L = 1 \text{ m}\Omega$. It can be found the loss and percentage of switching conduction, body diode conduction, and line conduction increase as output power increases. Discharge loss of parasitic capacitance is constant with constant switching frequency and input/output voltage. Therefore, percentage of parasitic capacitor discharge loss decreases as output power increases. It should be noted discharge loss of parasitic capacitors is small for setup of the experimental prototype, but it requires some attention for super high switching frequency applications.

IV. EXPERIMENTAL RESULTS

With main parameter specifications listed in Table III, an experimental prototype has been built. Peak load of the experimental prototype is 3.125 kW according to (9), and a 1 kW rated load is employed to verify effectiveness of the proposed S-PWM modulation.

Steady-state experimental waveforms under small load (200 W), half load (500 W), and rated load (1 kW) are shown

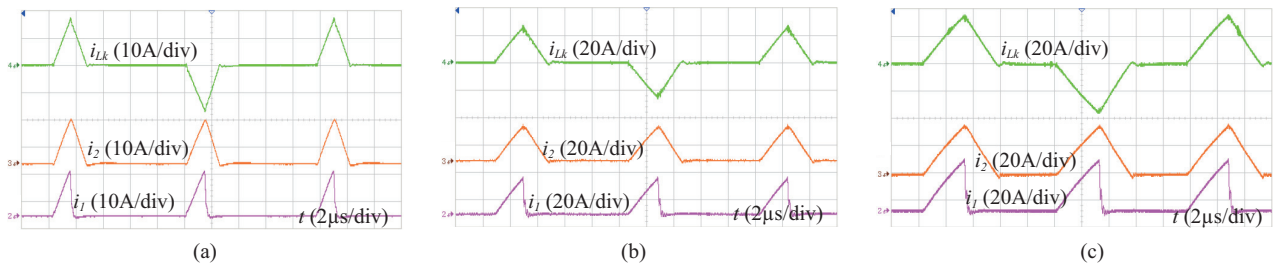


Fig. 16. Steady state experimental waveforms. (a) Small load (200 W). (b) Half load (500 W). (c) Rated load (1 kW).

TABLE III
SYSTEM SPECIFICATION

Parameters	Values
Input voltage (V_1)	200 V
Output voltage (V_2)	100 V
Rated Power (P)	1 kW
Magnetization inductance (L_m)	1.3 mH
Leakage inductor (L_k)	8 μ H
Turn ratio (n)	16:16
Switching frequency (f_s)	50 kHz

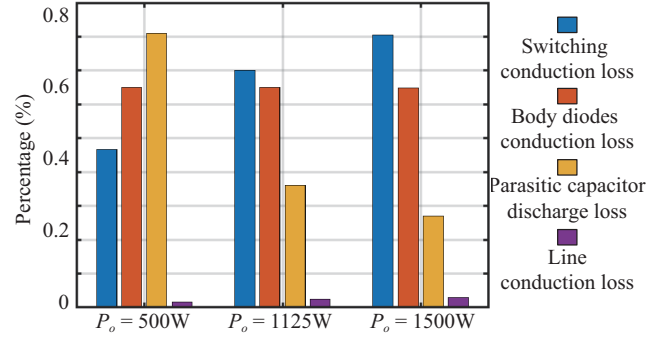


Fig. 15. Loss ratio of output power at 500, 1125, and 1500 W.

in Fig. 16(a), (b), and (c). Input current i_1 , output current i_2 , and leakage inductor current i_{Lk} are illustrated. Actually, waveforms of i_1 and i_2 are parts of i_{Lk} . As can be seen, there is no negative value portion in i_1 and i_2 under different load conditions, which means there is no circulating power on V_1 and V_2 sides.

Specific soft-switching waveforms at small load conditions are shown in Fig. 17 to verify ZCS soft-switching ability. Fig. 17(a) shows the leakage inductor current and driver signals waveforms of switches S_1 and S_4 . As can be seen, S_1 and S_4 are turned on before the increase of i_{Lk} , and S_4 is turned off after i_{Lk} decreases to zero. So, the ZCS-off of S_4 and ZCS-on of S_1 and S_4 are achieved. Similarly, it can be observed from Fig. 17(b) the ZCS-off of S_2 and ZCS-on of S_2 and S_3 are achieved. Fig. 17(c) and (d) show leakage inductor current and driver signal waveforms of switches S_5 & S_8 and S_6 & S_7 . It can be seen the rising edges of driver signals are before the i_{Lk} and falling edges are after the i_{Lk} . So, the ZCS-on and ZCS-off of S_5 , S_6 , S_7 , and S_8 can be achieved. ZCS soft-switching characteristics can be achieved under different load conditions.

Figure 18 shows the dynamical response waveforms of the experimental prototype. Transmitted power is switched

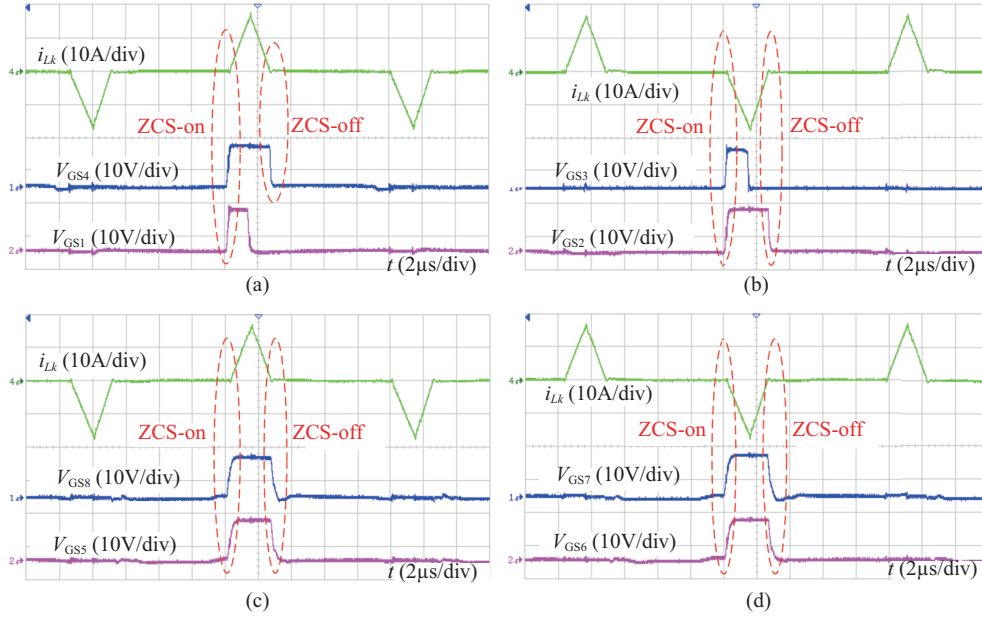


Fig. 17. Experimental waveforms of soft switching under S-PWM modulation at small load condition. (a) Leakage inductor current and driver signals waveforms of switches S_1 and S_4 . (b) Leakage inductor current and driver signals waveforms of switches S_2 and S_3 . (c) Leakage inductor current and driver signals waveforms of switches S_5 and S_8 . (d) Leakage inductor current and driver signals waveforms of switches S_6 and S_7 .

from 200 to 500 W and then back from 500 to 200 W. Leakage inductor current, input and output current, and output voltage of the primary bridge are captured. As can be seen, the converter under S-PWM modulation can switch to a new steady-state fast after load condition changes. Experimental results verify effectiveness of the S-PWM modulation during mode switching.

Figure 19 shows peak current curves of SPS, EPS, and

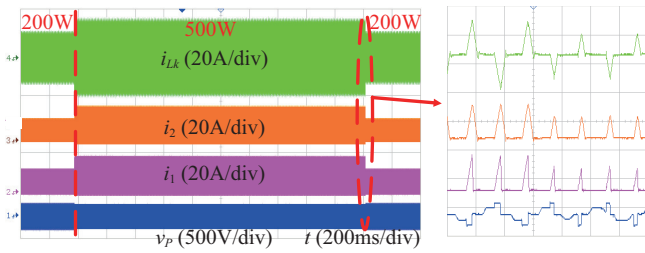


Fig. 18. Dynamical response waveforms with different load conditions.

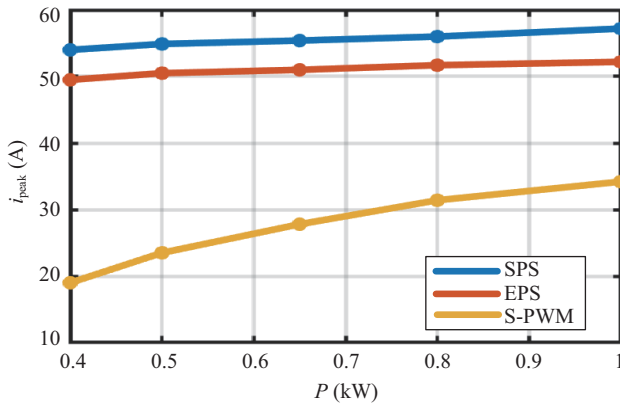


Fig. 19. Peak current stress curves of SPS, EPS, and S-PWM modulations.

S-PWM modulations under different transmission power. As can be seen, peak currents of SPS and EPS are much bigger than S-PWM with the same transmitted power. This is due to elimination of circulating power. Therefore, current stress and conduct loss of S-PWM can be minimized. All switches can achieve ZCS, which further reduces switching loss and increases efficiency. Besides, the rising speed of the peak current of S-PWM decreases with increasing output power from Fig. 19, which means it is suitable for high-power applications. Efficiency curves of the converter under SPS, EPS, and S-PWM for different transmitted power are illustrated in Fig. 20. As can be seen, overall conversion efficiency of the proposed S-PWM modulation is higher than SPS and EPS modulations, especially when transmitted power is low.

V. CONCLUSION

This paper proposes a novel S-PWM modulation to eliminate both input and output sides circulating power of DAB

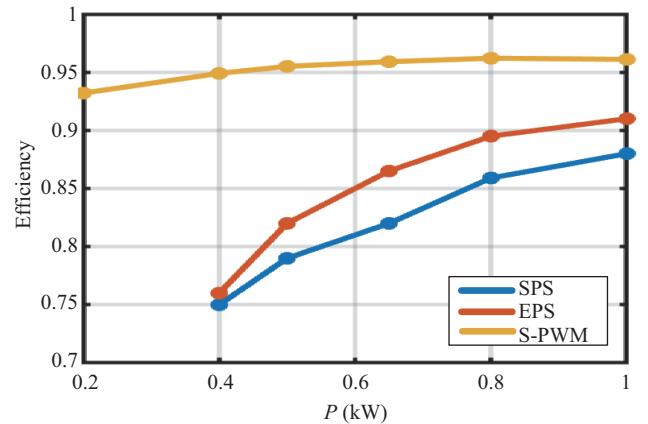


Fig. 20. Efficiency curves of SPS, EPS, and S-PWM modulations.

over a wide voltage gain. The proposed S-PWM has four cases according to voltage gain and transmission power. The operation principle, modes, and performance of DAB under one case are analyzed in detail. In addition, comparison analysis of peak current stress and circulating power with SPS, EPS, TPS, and MDM modulations are presented. From theoretical analysis and experimental results, it can be found the DAB converter under the S-PWM modulation has the following advantages: 1) S-PWM can eliminate circulating power. Current stress of devices and conduction loss can be reduced for the same transmitted power, and efficiency can be increased. 2) Soft-switching characteristics can be achieved within the full power range for the S-PWM. 3) S-PWM has a wide power transmission range, and control can be simplified because power transfer is related to only one variable.

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