

Efficiency Improvement and High-performance Control of Dual-active-bridge DC-DC Converter with Triple-phase-shift Modulation

Qiang Ren, Fei Xiao, Jilong Liu, Peng Chen, and Zhichao Zhu

Abstract—This paper analyzes the optimal switching modes of the dual-active-bridge (DAB) DC-DC converter to minimize losses in DC-DC converters that exhibit a wide voltage range, high efficiency, and high performance. By solving the multi-objective constrained conditional extremums with the Karush-KuhnTucker (KKT) method, a triple-phase-shift (TPS) optimal modulation with full range of operation, global continuity, and simple calculation is obtained. By adopting TPS modulation, the DAB can decrease reactive power when dc voltages are mismatched and obtain global optimization of conduction loss and soft-switching in the full range of operation. Generation of dc-offset current in the ac link is analyzed. Static and dynamic dc-offset currents are suppressed based on TPS modulation. With direct-power feed-forward, high-dynamic control is used to improve system response to external disturbances. The digital control strategy of the whole system is designed, and optimal modulation and high-performance control are verified on the experimental prototype.

Index Terms—DC-DC converter, DC-offset, dual-active-bridge, feedforward control, optimal control, phase-shift control, soft switching.

I. INTRODUCTION

WITH rapid development of new clear energy, applications such as electric vehicles and energy stations have demand for high-performance isolated bi-directional DC-DC converters [1]. The dual-active-bridge (DAB) DC-DC converter with high-frequency isolation has the advantages of high-power density and good environmental adaptability, which is widely used in medium-low power DC-DC conversions [2].

The conventional single-phase-shift (SPS) modulation realizes power transmission by adjusting the phase-shift angle between the output square-wave voltages of the primary and secondary H-bridges. When battery voltage changes in a wide range, dc voltages of the primary and secondary sides do not

match any more, resulting in large reactive power, high current stress, large loss, and small soft-switching range [3].

To reduce reactive power and expand the soft-switching range, multi-phase-shift modulation becomes an effective way. References [4]–[6] study double-phase-shift (DPS) modulation with the mathematical relationship obtained, and the soft-switching range expanded. Switching modes of extend-phase-shift (EPS) modulation are analyzed in [7]–[9]. A good compromise between complexity and effectiveness is achieved to reduce reactive power and expand the soft-switching range. What's more, triple-phase-shift (TPS) modulations with segmented analysis, reinforcement learning, neural network, particle swarm optimization, etc., are explored to make full use of the control freedoms in [10]–[17]. Optimal switching modes are selected from numerous modes. Current stress optimization and soft-switching are realized in almost the full range of operation. In summary, for multi-phase-shift modulations, reactive power, maximum current or its effective value, etc., are selected to be the optimal goals and then control strategies are obtained by solving conditional extremes. Compared to SPS modulation, multi-phase-shift modulations have lower current stress, wider soft-switching range, and lower loss. However, problems such as numerous switching modes, cumbersome analysis and solving, discontinuous control strategies, and complex expressions still limit application of the multi-phase-shift modulations. Exploring a modulation and control strategy with simple and continuous expression will facilitate use of DAB in engineering.

Due to factors such as non-ideal devices and external disturbances, an ac link will generate DC-offset current. The DC-offset current not only increases current stress and loss, but also saturates the magnetic elements, resulting in peak current and device damage [18]. Therefore, DC-offset current must be suppressed in practical applications. The most commonly used ways are the hardware method in series with a capacitor connected in the ac link and the software method with control strategies [19], [20]. Limited by ac-link current and aging of the capacitor, the hardware method is only suitable for small power occasions. The software method is suitable for medium-high power occasions. In [21]–[23], a low-pass filter is used to detect the average current of the ac link, and then the static DC-offset current is suppressed. A low-pass filter has slow dynamic response, which cannot be used for suppression of dynamic DC-offset current. To quickly suppress the dynamic DC-offset current, phase-shift angle is adjusted based on the

Manuscript received April 13, 2021; revised August 19, 2021; accepted November 19, 2021. Date of online publication September 5, 2022; date of current version November 30, 2022. This work was supported by the National Natural Science Foundation of China (51807200).

Q. Ren (corresponding author, email: im_qiangren@126.com), F. Xiao, J. L. Liu, P. Chen, and Z. H. Zhu are with National Key Laboratory of Electromagnetic Energy, Naval University of Engineering, Hubei 430033, China.

DOI: 10.17775/CSEEJPES.2021.02880

prediction of the offset current [24]–[26]. However, predictive control is mostly used for SPS modulation, which is difficult to extend to multi-phase-shift modulations.

For battery applications, fast dynamic response becomes a basic requirement of DAB. For example, a power supply for a battery test requires settling time less than 3 ms with a step change in the current. Generally, settling time of the conventional single-voltage or -current closed-loop is long. The range of commands of the voltage-current double-closed-loop is wide, resulting in big differences between light and heavy loads. To improve dynamic response, a feed-forward control based on the load-current is proposed, which is only suitable for SPS modulation [27]. An EPS modulation with high dynamic is analyzed based on model prediction, whose control commands depend on system parameters [28], [29].

To meet the requirements of isolated bi-directional DC-DC converters for a wide range of voltage, high efficiency and high-dynamic response, this paper presents a comprehensive analysis of optimal modulation and high-performance control of DAB. TPS modulation is analyzed, and then DC-offset current suppression and high-dynamic control are proposed based on the optimal modulation. This paper is arranged as follows: Section II presents TPS modulation; Section III analyzes switching characteristics; Section IV studies DC-offset current suppression; Section V presents high-dynamic control; Section VI is the experiment and analysis; Section VII concludes the paper.

II. TPS OPTIMAL MODULATION

A. TPS Optimal Process

A typical DAB with primary and secondary H-bridges, an isolation transformer and an external inductor is shown in Fig. 1. For DAB operating in a wide range of voltages, TPS modulation becomes a feasible way to obtain high efficiency. In TPS modulation, the primary and secondary duty ratios and phase-shift angle are adjusted in real-time according to the dc voltages and the output power. Current stress minimization and soft-switching range expansion are the objectives to decrease loss.

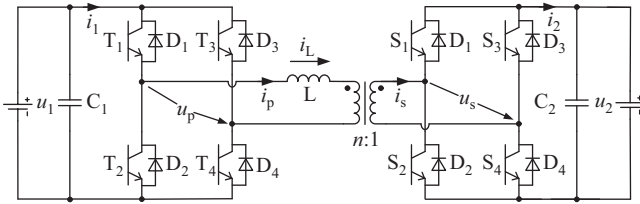


Fig. 1. Topology of DAB converter.

In Fig. 1, the transformer ratio is n and the dc voltages of the primary-secondary side are u_1 and u_2 , respectively. L is the sum of the leakage inductance and external inductance. The output rectangular voltages of the primary-secondary H-bridges are u_p and u_s . The switching frequency is f_s and switching cycle is T_s .

According to the voltage ratio $d = nu_2/u_1$, the full range of operation is divided into three situations, $d > 1$, $d = 1$ and

$d < 1$. For $d > 1$ and $d < 1$ situations, the internal phase-shifts of the primary and secondary H-bridges are the duty ratios D_1 and D_2 , and the phase-shift between the fundamental components of the two bridges is D_f . $D_1, D_2 \in [0, 1]$ and $D_f \in [-0.5, 0.5]$. $D_1(D_2)$ equals 1 when the output voltage of the H-bridge is a square wave. D_f equals 0.5 when the phase shift angle is 90° . Power is transmitted from primary side to secondary side when $D_f > 0$, and power transmits in the opposite direction when $D_f < 0$.

According to the power transmission of DAB, the proportion of power directly transmitted from primary side to secondary side or vice versa should be as large as possible, and the proportion of power indirectly transmitted through the leakage inductor should be as small as possible, to reduce reactive power [28]. Therefore, the waveforms of the rectangular-wave voltage u_p and u_s (three-level or two-level) should overlap, and their volt-second products should be as equal as possible. For this purpose, optimal switching modes with minimum current stress under $d > 1$ and $d < 1$ situations are achieved, as shown in Figs. 2 and 3. Each situation has two optimal switching modes.

For mode 1 of $d > 1$ situation, current at each switching moment is calculated, and the range of phase-shift angle, the transmission power and the maximum current are obtained, as shown in (1)~(3). Power is normalized to $nu_1u_2/(8f_sL)$ and

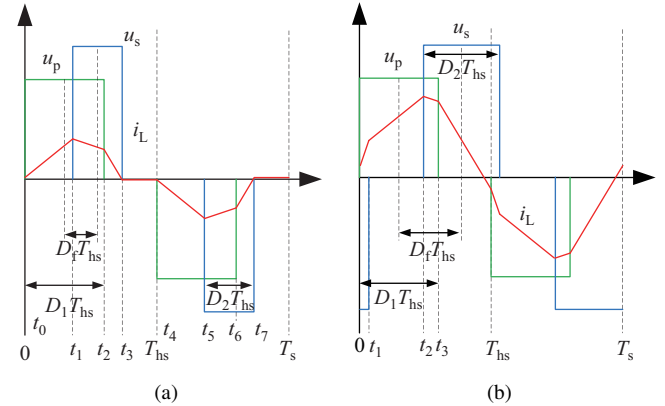


Fig. 2. Optimal switching modes under $d > 1$ situation. (a) Mode 1 (u_p overlaps u_s once). (b) Mode 2 (u_p overlaps u_s twice).

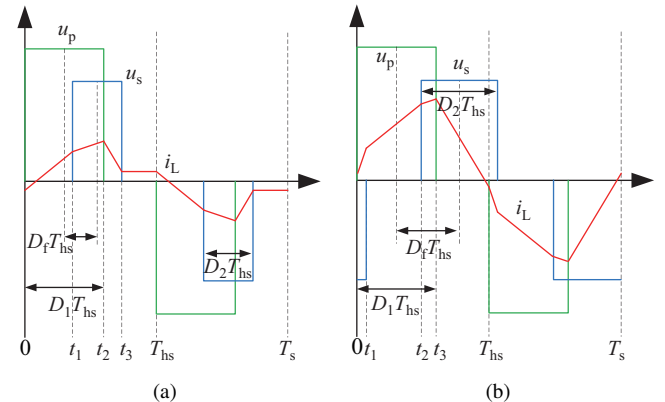


Fig. 3. Optimal switching modes under $d < 1$ situation. (a) Mode 1 (u_p overlaps u_s once). (b) Mode 2 (u_p overlaps u_s twice).

current is normalized to $nu_2/(4f_sL)$.

$$|D_1 - D_2|/2 \leq D_f \leq \min \left\{ \frac{D_1}{2} + \frac{D_2}{2}, 1 - \frac{D_1}{2} - \frac{D_2}{2} \right\} \quad (1)$$

$$P_o = 8D_f^2/(d-1) \quad (2)$$

$$I_{\max} = 4D_f/d \quad (3)$$

Maximum current $I_{\max}$ is determined by the phase-shift angle D_f (inequality constraint) and the output power P_o (equality constraint). Thus, finding the minimum current that satisfies the constraints and obtaining the mathematical relationships among the control variables (D_1, D_2, D_f) become the goals of TPS optimal modulation.

Referring to the KKT method [12] and [30], maximum current is set as the objective function, and phase-shift angle and output power are set to be the constraints. The Lagrange polynomial $L(X)$ is constructed and equations with local optimal solution of mode 1 are obtained.

$$\begin{aligned} L(D_1, D_2, D_f) = & \frac{4D_f}{d} + \lambda \left(\frac{8D_f^2}{d-1} - P_{on} \right) \\ & + \mu_1 \left(D_f - \frac{D_1}{2} + \frac{D_2}{2} \right) \\ & + \mu_2(-D_f) + \mu_3(-D_2) + \mu_4(D_1 - 1) \end{aligned} \quad (4)$$

$$\begin{cases} \frac{dL}{dD_1} = \frac{dL}{dD_2} = \frac{dL}{dD_f} = 0 \\ P_{on} = 8D_f^2/(d-1) \\ \mu_1 \left(D_f - \frac{D_1}{2} + \frac{D_2}{2} \right) = 0, \mu_2(-D_f) = 0 \\ \mu_3(-D_2) = 0, \mu_4(D_1 - 1) = 0, \mu_1, \mu_2, \mu_3, \mu_4 \geq 0 \\ 0 \leq D_f \leq \frac{D_1}{2} - \frac{D_2}{2}, D_1 \leq 1, D_2 \geq 0, \lambda \neq 0 \end{cases} \quad (5)$$

Then, the local optimal solution of mode 1 is obtained.

$$D_1 = \frac{2d}{d-1}D_f, D_2 = \frac{2}{d-1}D_f, 0 \leq D_f \leq \frac{d-1}{2d} \quad (6)$$

Similarly, the local optimal solution of mode 2 with minimum current stress under the constraints can be achieved.

$$D_1 = 1, D_2 = 1 - (d-1)(1-2D_f), \frac{d-1}{2d} < D_f \leq \frac{1}{2} \quad (7)$$

Based on the local optimal solutions of the two modes, the unified optimal strategy is achieved for $d > 1$ situation.

$$\begin{cases} D_1 = \frac{2d}{d-1}D_f, D_2 = \frac{2}{d-1}D_f, D_f \in \left[0, \frac{d-1}{2d}\right] \\ D_1 = 1, D_2 = 1 - (d-1)(1-2D_f), D_f \in \left[\frac{d-1}{2d}, \frac{1}{2}\right] \end{cases} \quad (8)$$

With the same way, the unified optimal strategy for $d < 1$ situation can be had.

$$\begin{cases} D_1 = \frac{2d}{1-d}D_f, D_2 = \frac{2}{1-d}D_f, D_f \in \left[0, \frac{1-d}{2}\right] \\ D_1 = 2 - \frac{1}{d} + \frac{2(1-d)}{d}D_f, D_2 = 1, D_f \in \left[\frac{1-d}{2}, \frac{1}{2}\right] \end{cases} \quad (9)$$

For $d = 1$ situation, the optimal modulation is SPS, whose duty ratios and phase-shift angle are as follows.

$$D_1 = D_2 = 1, D_f \in \left[0, \frac{1}{2}\right] \quad (10)$$

According to the symmetry of DAB, the global optimal strategy for reverse power conversion satisfies the relationship among D_1, D_2 and D_f remains the same, only the sign of D_f is opposed.

B. Effects of TPS Modulation

According to the TPS modulation in (8), (9) and (10), the mathematical relationship between duty ratios and phase-shift angle is only related to the voltage ratio, which can be determined by detecting the primary and secondary dc voltages. With TPS modulation, the control variable is reduced from three (D_1, D_2 and D_f) to one (D_f), that a single regulator can be adopted to realize system optimization.

First, in TPS modulation, electrical parameters such as current and power are only related to voltage ratio d , phase-shift angle D_f and the hardware parameters f_s, L , and n . Fig. 4 shows the RMS of inductor current under different conditions. It is shown that current near-linearly increases as output power increases for a given voltage-ratio d . Current increases little as the increasing of voltage mismatch (d deviates from 1) for a given output power. Figures show that TPS modulation can effectively reduce the impact of the voltage mismatch.

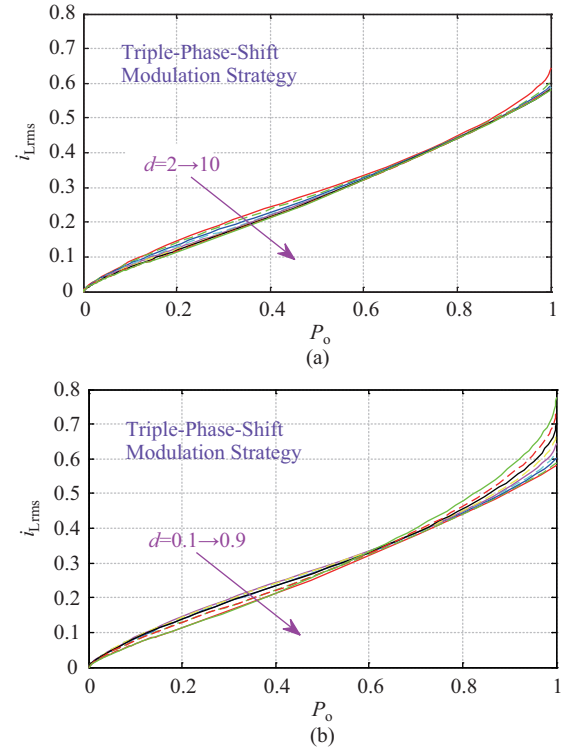


Fig. 4. RMS of inductor current with TPS modulation. (a) $d > 1$. (b) $d < 1$.

Secondly, with TPS modulation, active power can be calculated in the whole range of operation, as shown in Fig. 5. Both $d > 1$ and $d < 1$ situations are composed of two switching modes with continuous transition. Mode 1 works in a low

power range and mode 2 works in high power range. For each voltage ratio d , active power P_o monotonically increases with phase-shift angle D_f . Gains of the two modes are different. When the mismatch of dc voltages increases, power decreases. Maximum power can be had when the phase-shift angle is 90° ($D_f = 0.5$).

To reveal the effectiveness of TPS modulation in reducing reactive power, the power factor of the high-frequency transformer is shown in Fig. 6. What's more, the power factor with SPS modulation is given for comparative analysis, as shown in Fig. 7.

With TPS modulation, mode 1 obtains a constant power factor, while the power factor of mode 2 changes with the output power. TPS can obtain a higher power factor than SPS when operating in a wide range of voltages. That is, TPS can

achieve higher efficiency than SPS.

In summary, for DAB operating in a wide range of voltages, the proposed TPS modulation has simple calculation, a uniform form and a continuously adjustable power. When the dc voltages are mismatched, conduction loss is reduced and efficiency is improved by reducing current stress.

III. BEHAVIOR OF POWER SWITCHING

To reduce switching loss, soft-switching has become a key element to DAB modulation. By transmitting power through the inductor, the inductive current creates conditions for soft-switching. TPS modulation can expand soft-switching range and reduce switching loss by reducing reactive power and the current stress.

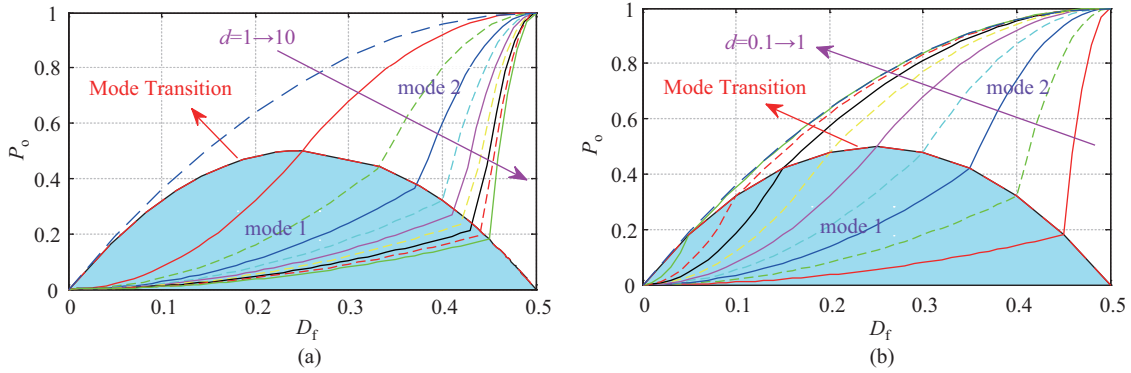


Fig. 5. Active power under $d > 1$ and $d < 1$ conditions. (a) $d > 1$. (b) $d < 1$.

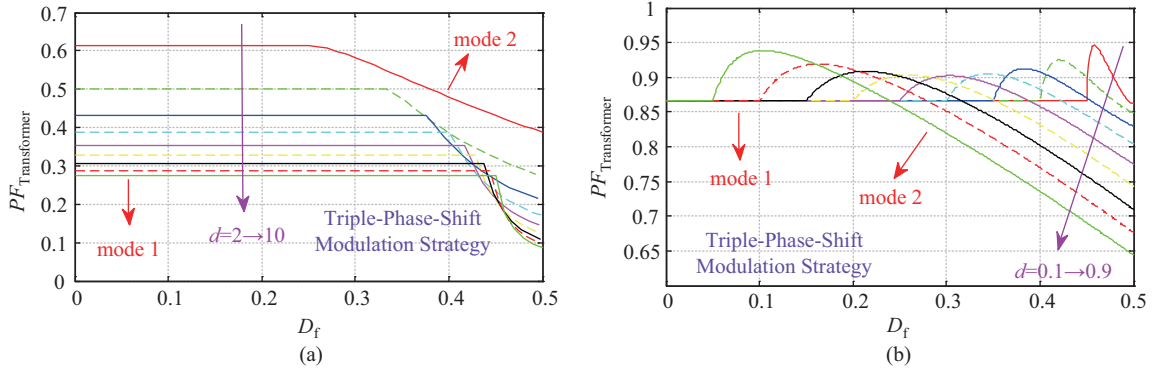


Fig. 6. Transformer power factor with TPS modulation. (a) $d > 1$. (b) $d < 1$.

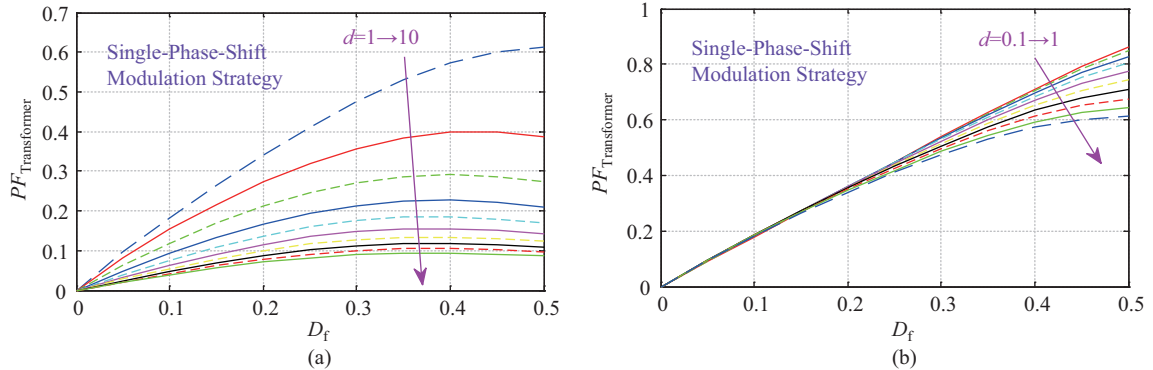


Fig. 7. Transformer power factor with SPS modulation. (a) $d > 1$. (b) $d < 1$.

A. Switching Characteristic of Mode 1 Under $d < 1$ Situation

To simplify analysis, dead-time is not considered. Current at each switching time of every device in mode 1 is calculated with TPS modulation.

$$\begin{cases} I_{T_{1_off}} = -I_{T_{2_off}} = -\frac{2nu_2D_f + (-u_1 + nu_2)D_1}{2L}T_{hs} = 0 \\ I_{T_{3_off}} = -I_{T_{4_off}} = -\frac{(u_1 - nu_2)D_1 + 2nu_2D_f}{2L}T_{hs} < 0 \\ I_{S_{1_off}} = -I_{S_{2_off}} = n\frac{u_1D_1 - nu_2D_2}{2L}T_{hs} = 0 \\ I_{S_{3_off}} = -I_{S_{4_off}} = -n\frac{u_1D_1 - nu_2D_2}{2L}T_{hs} = 0 \end{cases} \quad (11)$$

All devices except for the primary T_3/T_4 can achieve zero-current switching, that is ZCS can be had. For T_3/T_4 , the commutation at the moment when T_3 turns off and T_4 turns on is shown in Fig. 8.

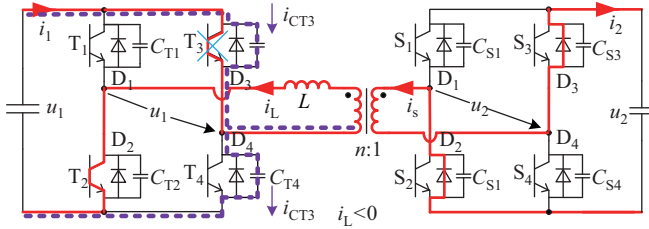


Fig. 8. Commutation when T_3 turns off and T_4 turns on in mode 1.

Before T_3 turns on, inductor current is negative and T_3 conducts the current (the red line). After T_3 turns off, the parasitic capacitors C_{T3}/C_{T4} and inductor L start to resonate (the dotted line). C_{T3} is charged to u_1 and C_{T4} is discharged to zero. Then, current flows through the anti-parallel diode D_4 , resulting in T_4 turning on with zero voltage.

B. Switching Characteristic of Mode 2 Under $d < 1$ Situation

Similar to the above analysis, currents at the turn-on/off moments of each switching device in mode 2 are calculated.

$$\begin{cases} I_{T_{1_off}} = I_{T_{4_off}} = -I_{T_{2_off}} = -I_{T_{3_off}} \\ = \frac{u_1 + nu_2(2D_f - 1)}{2L}T_{hs} > 0 \\ I_{S_{1_off}} = -I_{S_{2_off}} = -n\frac{(-u_1 + nu_2)D_1 + 2u_1D_f}{2L}T_{hs} < 0 \\ I_{S_{3_off}} = -I_{S_{4_off}} \\ = \frac{(-2(-1 + D_f)nu_2 + d(-1 + 2D_f)(u_1 + nu_2))}{2L}T_{hs} > 0 \end{cases} \quad (12)$$

In mode 2, each switching device conducts current before turning off, resulting in hard switching off. However, when the device is turning on, current flows through the anti-parallel diode and zero-voltage turn-on can be had.

According to symmetry of DAB, the switching characteristics of $d < 1$ situation can be expanded to $d > 1$. Undoubtedly, both modes 1 and mode 2 of $d > 1$ situations can realize soft-switching in the full range of operation.

It can be concluded that all switching devices can obtain soft-switching in the whole range of operation with TPS modulation. That is, there is no need to attach additional

constraints to control parameters (D_1 , D_2 and D_f) for soft-switching when the proposed TPS modulation is used.

However, in practical applications, dead-time will be added in each bridge and switching behaviors will be changed. For mode 1, inductor current will have a slight change during the dead-time, and devices no longer switch in zero current but a small current. For mode 2, current direction remains unchanged in the dead-time with no effect to switching behavior. Considering that dead-time is very short, the change of inductor current of mode 1 has little influence on switching behaviors and loss. That is, by adopting TPS modulation, soft-switching can be achieved in almost the whole range of operation when dead-time is under consideration.

IV. SUPPRESSION OF DC-OFFSET CURRENT

In DAB, a high-frequency ac link is used to achieve electrical isolation. DC-offset current in the ac link caused by non-ideal devices and external disturbances will make the flux of the transformer and inductor saturation, result in a surge current in severe case. Therefore, suppression of DC-offset current is necessary in practical applications.

A. Static DC-offset Current Suppression

Non-ideal factors, such as transmission delay and pulse-width distortion of control pulse, differences in switching devices and loop impedances, etc., will cause a static DC-offset current in the ac link. Static DC-offset current usually increases system loss without causing deep flux saturation of the magnetic elements for its large randomness, low destructiveness and continuous existence.

Since it is difficult to accurately estimate static DC-offset current, suppression based on offset current detection becomes an effective way. As shown in Fig. 9, a low-bandwidth circuit is used to extract the offset current and then compensation of the modulation ratio is generated by a proportional regulator K_{p_offset} . By compensating the duty ratio of the primary (or secondary) half-bridge, the volt-second product of the rectangular-wave voltage output by the primary (or secondary) H-bridge is changed, and DC-offset current can be eliminated.

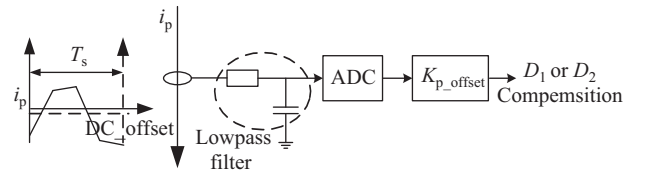


Fig. 9. Diagram of static DC-offset current suppression.

B. Dynamic DC-offset Current Suppression

Affected by external factors such as sudden changes of load and voltages, the closed-loop regulator will quickly change the control commands, generating a dynamic DC-offset current in the ac link. The dynamic offset current will increase loss and make deep flux saturation of the magnetic elements for its large amplitude, strong destructiveness and fast response.

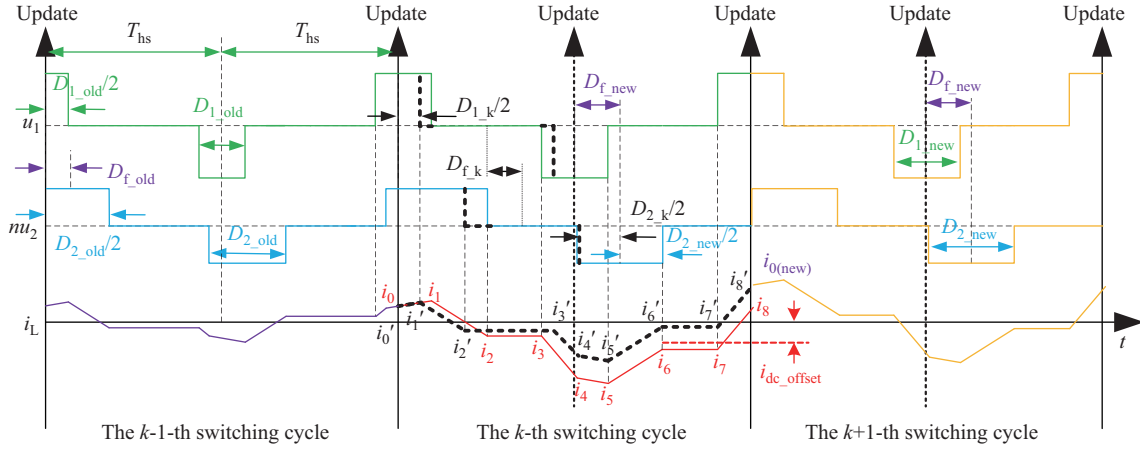


Fig. 10. Current trajectory of switching mode 1 under $d < 1$ situation.

Since dynamic DC-offset current is caused by the changes of control commands, an offset current is predictable. Adopting current trajectory planning, suppression of dynamic DC-offset current is shown in Fig. 10.

Under normal operation, control commands are updated only once per switching cycle. That is, commands change from D_{1_old} , D_{2_old} , and D_{f_old} to D_{1_new} , D_{2_new} , and D_{f_new} when the k -th cycle begins, causing a DC-offset current (red solid line) in the ac link at the end of the k -th cycle.

With current trajectory planning, each switching cycle is divided into upper and lower half cycles for control respectively, and commands are updated every half cycle. For the k -th cycle, commands in the first half cycle are D_{1_k} , D_{2_k} , and D_{f_k} , and are D_{1_new} , D_{2_new} , and D_{f_new} in the second half cycle. D_{1_k} , D_{2_k} , D_{f_k} are determined by the control commands of the PI regulator in the $k-1$ and k -th cycles. To eliminate the DC-offset beginning from the $k+1$ -th cycle, current after trajectory planning should meet the constraint that current i'_8 of the k -th cycle should be equal to current $i_{0(new)}$ of the $k+1$ -th cycle without DC-offset current. Based on TPS modulation, the control commands for dynamic DC-offset current suppression are as follows.

$$\begin{cases} D_{1_k} = (D_{1_new} + D_{1_old})/2 \\ D_{2_k} = (D_{2_new} + D_{2_old})/2 \\ D_{f_k} = (D_{f_new} + D_{f_old})/2 \end{cases} \quad (13)$$

By adopting current trajectory planning, dynamic DC-offset current can be completely eliminated within one cycle, which can effectively improve dynamic response and prevent flux saturation of magnetic components.

V. HIGH DYNAMIC CONTROL

Fast dynamic response has become a basic requirement of DAB for battery applications. For external disturbances, adjustment time of the conventional single closed-loop control is long. The range of commands of the voltage-current double closed-loop control is wide. It is not possible to obtain good performance in the full range of conditions.

Based on direct-power feed-forward (DPF) control, output

power is real-time estimated online, and open-loop commands can be calculated [29]. What's more, a voltage closed-loop regulator can be used to compensate for the static error caused by switching loss and line resistance.

With TPS modulation, the mathematical relationship between output power P_o and phase-shift angle D_f is obtained for $d > 1$ and $d < 1$ situations.

$$D_f = \begin{cases} \sqrt{\frac{P_o(d-1)}{8}} & 0 \leq P_o \leq \frac{2(d-1)}{d^2} \text{ \& } d > 1 \\ \frac{1}{2} - \frac{1}{2} \sqrt{\frac{1-P_o}{d^2-2d+2}} & \frac{2(d-1)}{d^2} < P_o \leq 1 \text{ \& } d > 1 \end{cases} \quad (14)$$

$$D_f = \begin{cases} \sqrt{\frac{P_o(1-d)}{8d}} & 0 \leq P_o \leq 2d(1-d) \text{ \& } d < 1 \\ \frac{1}{2} - \frac{1}{2} \sqrt{\frac{d^2-d^2P_o}{2d^2-2d+1}} & 2d(1-d) < P_o \leq 1 \text{ \& } d < 1 \end{cases} \quad (15)$$

Suppose the reference output voltage, current and resistance are u_2^* , i_2^* and R_{Load}^* , respectively. When there is a disturbance in the load, actual resistance R_{Load} changes, resulting in an error in output voltage. To quickly compensate for the error, real load is estimated. R_{Load}^* is set to track R_{Load} , and then the expression of i_2^* can be obtained.

$$R_{Load}^* = R_{Load} \Rightarrow \begin{cases} R_{Load} = u_2/i_2 \\ R_{Load}^* = u_2^*/i_2^* \end{cases} \Rightarrow i_2^* = \frac{u_2^*}{u_2} i_2 \quad (16)$$

Based on i_2^* , the output reference power P_o^* is obtained.

$$P_o^* = \frac{u_2^* i_2^*}{P_N} = \frac{u_2^* u_2^* i_2}{u_2} \frac{8f_s L}{nu_1 u_2} \quad (17)$$

Substituting (17) into (14) and (15), the feed-forward phase-shift angle D_f^* is obtained.

$$D_f^* = \begin{cases} \sqrt{\frac{P_o^*(d-1)}{8}} & 0 \leq P_o^* \leq \frac{2(d-1)}{d^2} \text{ \& } d > 1 \\ \frac{1}{2} - \frac{1}{2} \sqrt{\frac{1-P_o^*}{d^2-2d+2}} & \frac{2(d-1)}{d^2} < P_o^* \leq 1 \text{ \& } d > 1 \end{cases} \quad (18)$$

$$D_f^* = \begin{cases} \sqrt{\frac{(1-d)P_o^*}{8d}} & 0 \leq P_o^* \leq 2d(1-d) \text{ \& } d < 1 \\ \frac{1}{2} - \frac{1}{2} \sqrt{\frac{d^2-d^2P_o^*}{2d^2-2d+1}} & 2d(1-d) < P_o^* \leq 1 \text{ \& } d < 1 \end{cases} \quad (19)$$

To eliminate the static error, compensation is directly added onto output voltage u_2^* . The control diagram with DPF is shown in Fig. 11.

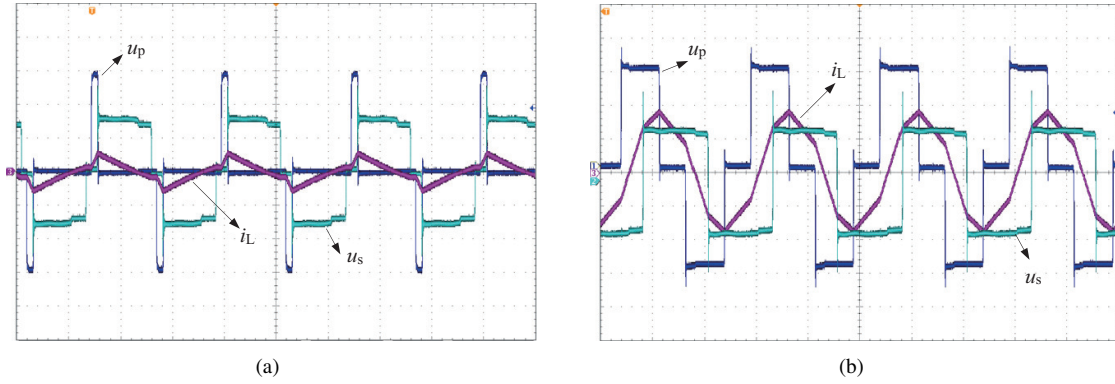


Fig. 14. Experiment waveforms of TPS modulation when $d < 1$. (a) $d = 0.25$. (b) $d = 0.5$.

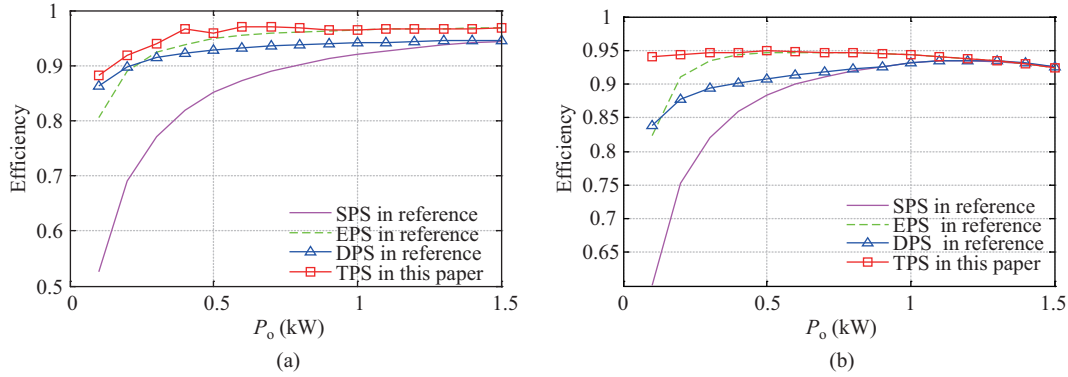


Fig. 15. Efficiency with different modulations. (a) $d = 1.5$. (b) $d = 0.5$.

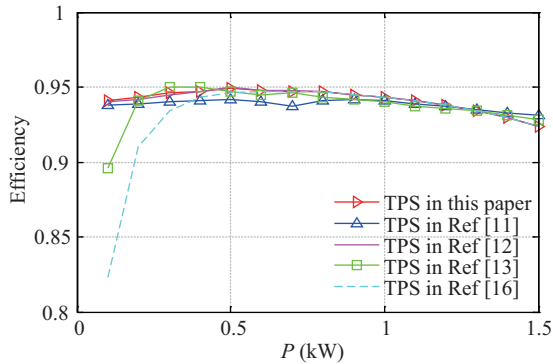


Fig. 16. Efficiencies of five different TPS modulations.

can almost achieve high efficiency under full load conditions.

B. Switching Behavior

In forward power conversion, switching behaviors of the devices in mode 1 and mode 2 under $d < 1$ situations are shown in Figs. 17 and 18.

In mode 1, output power and inductor current are small. The actual switching current is near zero when dead time is under considerations. All switching devices except for T_3/T_4 are approximately zero-current turn-on and -off. T_3 and T_4 are hard turn-off and zero-voltage turn-on.

In mode 2, output power is large, and the inductor is fully energized. The stored energy ensures the commutation process goes on smoothly. Primary and secondary devices are all hard

turn-off and zero-voltage turn-on. Experimental results agree with the theoretical analysis.

C. DC-offset Current Suppression

For static DC-offset current caused by non-ideal factors, current can be easily eliminated based on the detection circuit. For dynamic DC-offset current caused by the change of control commands, current trajectory planning is adopted. Both static and dynamic suppression methods are used in the experiment. When there is a sudden change in load, the current waveform is shown in Fig. 19. Without suppression of DC-offset current, offset current is about 26% of the peak-to-peak current. After suppression is adopted, the current adjusts to steady-state in one cycle and current is symmetrical without offset, as shown in Fig. 19(b).

D. High-dynamic Control

In forward power conversion, dynamic waveforms using DPF control and conventional voltage closed-loop (CVL) control are shown in Fig. 20.

When there is a step change in output voltage, dynamic adjustment time of DPF is 0.34 ms, while the adjustment time of CVL is 0.97 ms, as shown in Fig. 20(a) and 20(b). When load is suddenly added, the operation will change from switching mode 1 to mode 2. Dynamic adjustment time of DPF is 0.56 ms, while adjustment time of CVL is 1.2 ms with a drop in output voltage, as shown in Figs (c) and (d). Based on experimental results, it can be concluded that DPF can effectively improve the dynamics of the DAB.

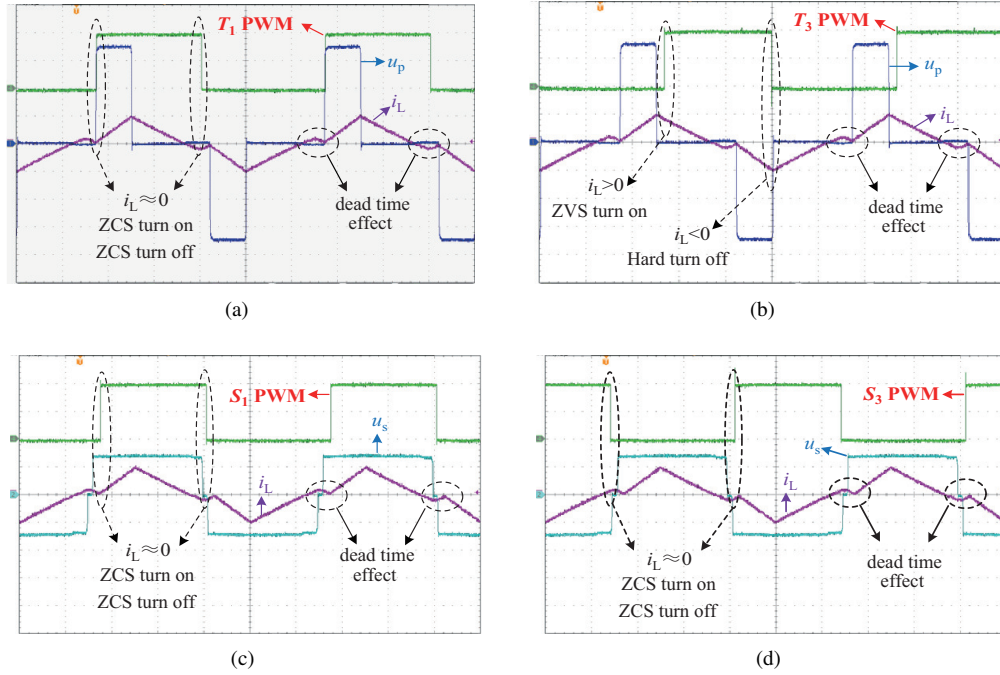


Fig. 17. Experiment waveforms of mode 1 under $d < 1$ condition. (a) T_1 approximately zero-current turns on and off. (b) T_3 hard turns off and zero-voltage turns on. (c) S_1 approximate zero-current turn-on and -off. (d) S_3 approximate zero-current turn-on and -off.

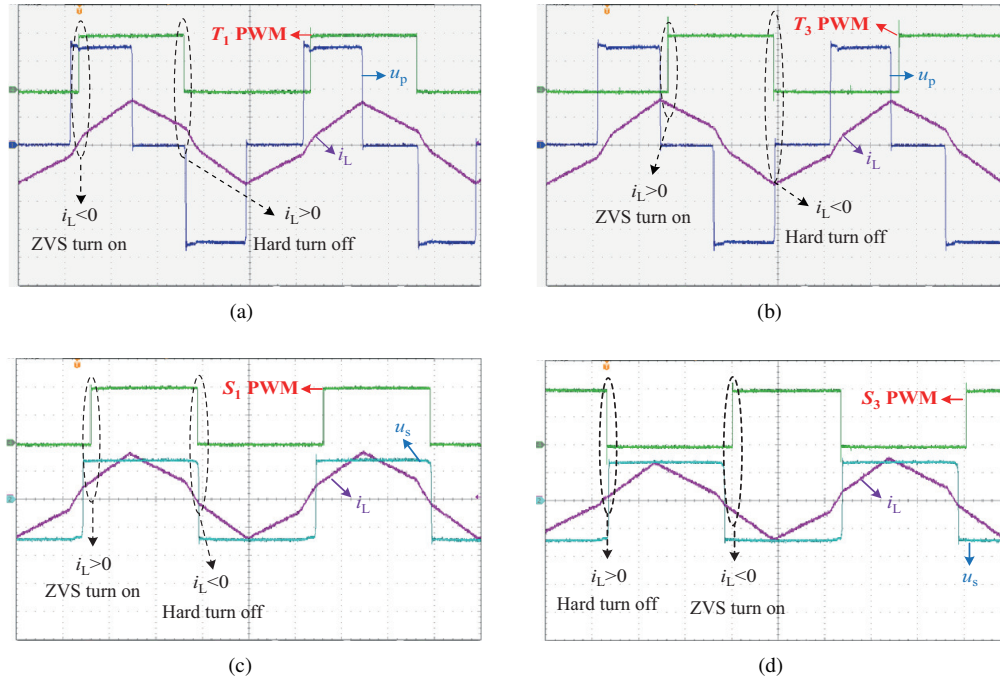


Fig. 18. Experiment waveforms of mode 2 under $d < 1$ condition. (a) T_1 hard turns off and zero-voltage turns on. (b) T_3 hard turns off and zero-voltage turns on. (c) S_1 hard turns off and zero-voltage turns on. (d) S_3 hard turns off and zero-voltage turns on.

VII. CONCLUSION

For DAB applications with wide-voltage range, high-dynamic response and high efficiency, this paper studies TPS optimal modulation, DC-offset current suppression and high-dynamic control. TPS modulation can achieve optimization of current stress for the full range of operation. While reducing conduction loss of switching devices, switching loss is also

decreased by expanding the soft-switching range. Then, static DC-offset current can be suppressed by detecting the offset current, and dynamic DC-offset current can be suppressed based on current trajectory planning. Dynamic response to external disturbances can be improved by direct-power feed-forward control. Finally, effectiveness of optimal modulation and high-performance control is verified by the experimental results.

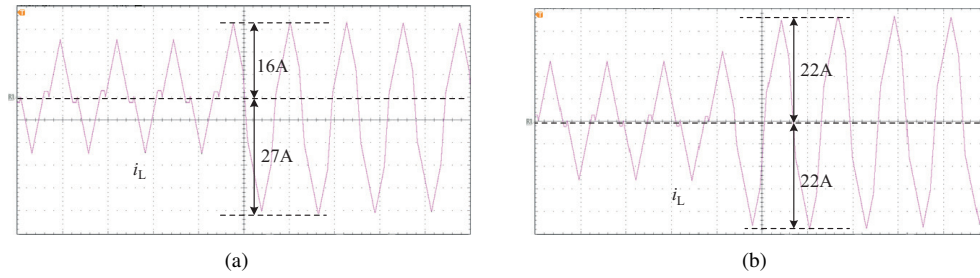


Fig. 19. Experiment waveforms of dc-offset current control with load sudden change. (a) Without dc-offset current suppression. (b) With dc-offset current suppression.

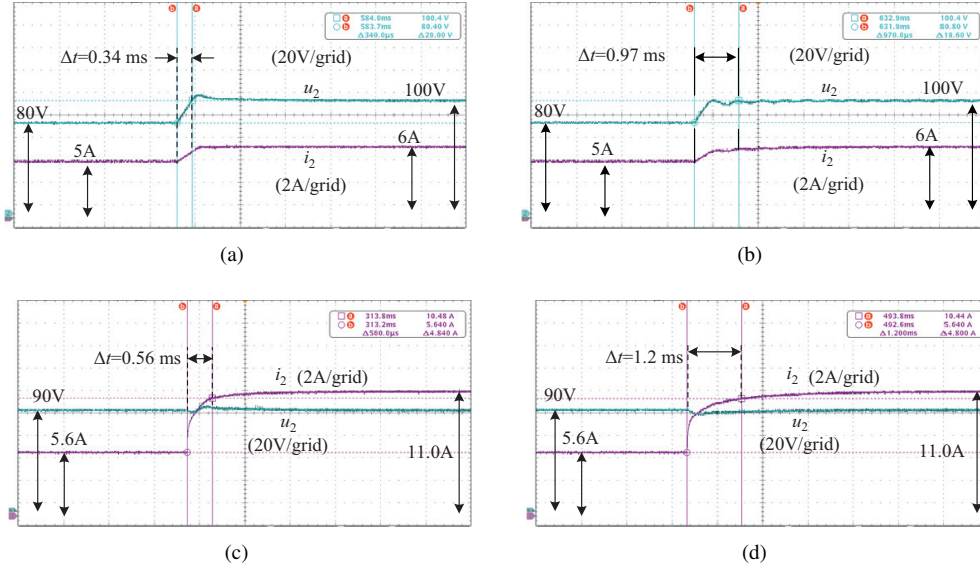


Fig. 20. Dynamic response experiment waveforms. (a) Dynamic response of DPF when output voltage changes. (b) Dynamic response of CVL when output voltage changes. (c) Dynamic response of DPF when load adds with a Li-battery load. (d) Dynamic response of CVL when load adds with a Li-battery load.

APPENDIX

Differences between the proposed TPS and the newly TPS modulations are listed in Table AI.

TABLE AI
DIFFERENCES BETWEEN THE PROPOSED TPS AND THE NEW TPS MODULATIONS

TPS modulation	Optimal target	Optimal strategy	Soft-switching	Global continuity	Calculation resource used
Proposed TPS modulation	Current stress	Step-up Conversion ($d = nu_2/u_1 > 1$): $\begin{cases} D_1 = \frac{2d}{d-1} D_f, D_2 = \frac{2}{d-1} D_f, P \in [0, \frac{2(d-1)}{d^2}] \\ D_1 = 1, D_2 = 1 - (d-1)(1-2D_f), P \in [\frac{2(d-1)}{d^2}, 1] \end{cases}$ Step-down Conversion ($d = nu_2/u_1 \leq 1$): $\begin{cases} D_1 = \frac{2d}{1-d} D_f, D_2 = \frac{2}{1-d} D_f, P \in [0, 2d(1-d)] \\ D_1 = 2 - \frac{1}{d} + \frac{2(1-d)}{d} D_f, D_2 = 1, P \in [2d(1-d), 1] \end{cases}$	Full working range soft-switching range	Continuous	Simple calculation with one PI regulator
Ref [11] TPS modulation	Minimize RMS Current	Step-up Conversion ($d = nu_2/u_1 > 1$): $\begin{cases} D_1 = 1 - d + dD_2, D_f = 0, 0 \leq P < \frac{2(d-1)}{d^2} \\ D_1 = 0, D_f = \frac{(d-1)(D_2-1)}{2} + \frac{\sqrt{d^2(1-D_2)^2 + D_2^2 - 1}}{2}, \\ P \in [\frac{2(d-1)}{d^2}, 2(1-d^2 + d\sqrt{d^2-1})] \\ D_1 = 0, D_2 = 0, P \in [2(1-d^2 + d\sqrt{d^2-1}), 1] \end{cases}$ Step-down Conversion ($d = nu_2/u_1 \leq 1$): $\begin{cases} D_2 = 1 - \frac{1-D_1}{d}, D_f = \frac{(1-d)(1-D_1)}{d}, P \in [0, 2d(1-d)] \\ D_2 = 0, D_f = \frac{(1+d)D_1-1+d}{2d} + \frac{\sqrt{(D_1-1)^2 + d^2(D_1^2-1)}}{2d}, \\ P \in [2d(1-d), \frac{2(d^2-1+\sqrt{1-d^2})}{d^2}] \\ D_1 = 0, D_2 = 0, P \in [2(d^2-1+\sqrt{1-d^2})/d^2, 1] \end{cases}$	NULL	Continuous	Complex calculation with one PI regulator

TABLE AI
DIFFERENCES BETWEEN THE PROPOSED TPS AND THE NEW TPS MODULATIONS (CONTINUED)

TPS modulation	Optimal target	Optimal strategy	Soft-switching	Global continuity	Calculation resource used
Ref. [12] TPS modulation	Current Stress	Step-up Conversion ($d = nu_2/u_1 > 1$): NULL. Step-down Conversion ($d = nu_2/u_1 \leq 1$): $\begin{cases} D_1 = \sqrt{\frac{2P}{\pi(1-d)}}, D_2 = \frac{1}{d} \sqrt{\frac{2P}{\pi(1-d)}}, P \in [0, \frac{\pi d^2(1-d)}{2}] \\ D_1 = 1 - (1-d) \sqrt{\frac{\pi d - 4P}{\pi d(1-2d+2d^2)}}, D_2 = 1, P \in [\frac{\pi d^2(1-d)}{2}, \frac{\pi d}{4}] \end{cases}$	Full working range soft-switching range	Only step-down conversion; Continuous	Complex calculation with one PI regulator
Ref. [13] TPS modulation	Minimize RMS Current	Step-up Conversion ($d = nu_2/u_1 > 1$): NULL. Step-down Conversion ($d = nu_2/u_1 \leq 1$): $\begin{cases} D_2 = D_1/dD_f = 0.5 \left(D_1 - D_2 + \frac{P}{2dD_1} \right) P \in [0, 0.5d] \\ D_2 = 1, D_f = \frac{(1-D_2+D_1) - \sqrt{2D_1+2D_2-D_1^2-D_2^2-P/d-1}}{2}, \\ P \in [0.5d, 1] \end{cases}$	NULL	Only step-down conversion; Discontinuous	Complex calculation with one PI regulator
Ref. [14] TPS modulation	Conduction loss	Step-up Conversion ($d = nu_2/u_1 > 1$): $\begin{cases} D_1 = \frac{\varphi}{\pi} \frac{d}{d-1}, D_2 = \frac{\varphi}{\pi} \frac{d}{d-1}, P \in [0, \frac{\pi(d-1)}{2d}] \\ D_1 = \frac{1}{2}, D_2 = \frac{1}{4} (2\sqrt{e_7} - 2\sqrt{e_8} - e_1), \\ e_8 = \frac{1}{4} \left(\frac{-e_1^3 - e_6}{\sqrt{e_7}} + 3e_1^2 - 8e_2 - 4e_7 \right), P \in [\frac{\pi(d-1)}{2d}, P_{opt}] \\ D_1 = \frac{1}{2}, D_2 = \frac{1}{2}, P \in [P_{opt}, 1] \end{cases}$ where e_1, e_7, e_8 are the polynomial expressions. Step-down Conversion ($d = nu_2/u_1 \leq 1$): NULL.	NULL	Only step-down conversion; Discontinuous	Complex calculation with one PI regulator
Ref. [15] TPS modulation	Efficiency Optimization	Step-up Conversion ($d = nu_2/u_1 > 1$): NULL. Step-down Conversion ($d = nu_2/u_1 < 1$): $\begin{cases} \text{offline PSO LUT}, P \in [0, d-d^2] \\ D_1 = 2\varphi - \varphi\sqrt{d} + \sqrt{d} - 1, D_2 = \frac{3}{2}\varphi - \frac{\varphi-1}{\sqrt{d}} - 1, \\ P \in [d-d^2, 3d-d^2] \\ \text{offline PSO LUT}, P \in [3d-d^2, 1] \end{cases}$	NULL	Only step-down conversion; Discontinuous	Complex calculation with two techniques
Ref. [16] TPS modulation	Minimize Current	Step-up Conversion ($d = nu_2/u_1 > 1$): NULL. Step-down Conversion ($d = nu_2/u_1 \leq 1$): $\begin{cases} D_1 = \frac{(1+d)D_3}{1-d}, D_2 = D_1 + D_3, P \in [0, \frac{\pi d(3d+1)(1-d)}{8}] \\ D_1 = \frac{1}{2} - \frac{1-d}{d} (\frac{1}{4} - D_3), D_2 = \frac{1}{2}, P \in [\frac{\pi d(3d+1)(1-d)}{8}, \frac{\pi d}{4}] \end{cases}$	Full working range soft-switching range	Only step-down conversion; Continuous	Simple calculation with one PI regulator

REFERENCES

- [1] W. Deng, W. Pei, N. Li, X. Zhang, Y. Yi and L. Kong, "Operational Control of Low-voltage MTDC Systems in a Cyber-physical Environment," *CSEE Journal of Power and Energy Systems*, vol. 8, no. 6, pp. 1569–1582, Nov. 2022.
- [2] Y. Xu, G. Tang, W. Li, G. Zhao and Z. Deng, "A three-phase integrated power electronic transformer with the capability of DC fault clearance," *CSEE Journal of Power and Energy Systems*, vol. 7, no. 4, pp. 785–796, July 2021.
- [3] H. Bi and C. Mi, "Eliminate reactive power and increase system efficiency of isolated bidirectional dual-active-bridge DC–DC converters using novel dual-phase-shift control," *IEEE Transactions on Power Electronics*, vol. 23, no. 6, pp. 2905–2914, Dec. 2008.
- [4] Y. C. Wang, Y. C. Wu, and T. L. Lee, "Design and implementation of a bidirectional isolated dual-active-bridge-based DC/DC converter with dual-phase-shift control for electric vehicle battery," in *2013 IEEE Energy Conversion Congress and Exposition*, Denver, CO, USA, 2013, pp. 5468–5475.
- [5] M. Kim, M. Rosekeit, S. K. Sul, and R. W. A. A. De Doncker, "A dual-phase-shift control strategy for dual-active-bridge DC–DC converter in wide voltage range," in *8th International Conference on Power Electronics*, Jeju, Korea (South), 2011, pp. 364–371.
- [6] B. Zhao, Q. Song, and W. H. Liu, "Efficiency characterization and optimization of isolated bidirectional DC–DC converter based on dual-phase-shift control for DC distribution application," *IEEE Transactions on Power Electronics*, vol. 28, no. 4, pp. 1711–1727, Apr. 2013.
- [7] B. Zhao, Q. G. Yu, and W. X. Sun, "Extended-phase-shift control of isolated bidirectional DC–DC converter for power distribution in microgrid," *IEEE Transactions on Power Electronics*, vol. 27, no. 11, pp. 4667–4680, Nov. 2012.
- [8] F. An, W. S. Song, and K. X. Yang, "Optimised power control with extended phase shift in dual-active-bridge DC–DC converters," *Electronics Letters*, vol. 54, no. 10, pp. 651–653, May 2018.
- [9] B. Zhao, Q. Song, W. H. Liu, G. W. Liu, and Y. M. Zhao, "Universal high-frequency-link characterization and practical fundamental-optimal strategy for dual-active-bridge DC–DC converter under PWM plus phase-shift control," *IEEE Transactions on Power Electronics*, vol. 30, no. 12, pp. 6488–6494, Dec. 2015.
- [10] N. Hou, W. S. Song, Y. W. Li, Y. N. Zhu, and Y. T. Zhu, "A comprehensive optimization control of dual-active-bridge DC–DC converters based on unified-phase-shift and power-balancing scheme," *IEEE Transactions on Power Electronics*, vol. 34, no. 1, pp. 826–839, Jan. 2019.
- [11] A. P. Tong, L. J. Hang, and G. J. Li, "Modeling and analysis of a dual-active-bridge-isolated bidirectional DC/DC converter to minimize RMS current with whole operating range," *IEEE Transactions on Power Electronics*, vol. 33, no. 6, pp. 5302–5316, Jun. 2018.
- [12] J. Huang, Y. Wang, Z. Q. Li, and W. J. Lei, "Unified triple-phase-shift control to minimize current stress and achieve full soft-switching of isolated bidirectional DC–DC converter," *IEEE Transactions on Industrial Electronics*, vol. 63, no. 7, pp. 4169–4179, Jul. 2016.
- [13] O. M. Hebal, A. A. Aboushady, K. H. Ahmed, and I. Abdelsalam, "Generic closed-loop controller for power regulation in dual active bridge DC–DC converter with current stress minimization," *IEEE Transactions on Industrial Electronics*, vol. 66, no. 6, pp. 4468–4478, Jun. 2019.
- [14] F. Krismer and J. W. Kolar, "Closed form solution for minimum conduction loss modulation of DAB converters," *IEEE Transactions on Power Electronics*, vol. 27, no. 1, pp. 174–188, Jan. 2012.
- [15] G. Jean-Pierre, N. Altin, A. El Shafei, and A. Nasiri, "Efficiency optimization of dual active bridge DC–DC converter with triple phase-shift control," in *2020 IEEE Energy Conversion Congress and Exposition*, Detroit, MI, USA, 2020, pp. 1217–1222, doi: 10.1109/ECCE44975.2020.9235620.
- [16] D. Mou, Q. M. Luo, Z. Q. Wang, J. Li, Y. Q. Wei, H. C. Shi, and X. Du, "Optimal asymmetric duty modulation to minimize inductor peak-to-peak current for dual active bridge DC–DC converter," *IEEE Transactions on Power Electronics*, vol. 36, no. 4, pp. 4572–4584, Apr. 2021.
- [17] Y. H. Tang, W. H. Hu, J. Xiao, Z. Y. Chen, Q. Huang, Z. Chen, and F. Blaabjerg, "Reinforcement learning based efficiency optimization scheme for the DAB DC–DC converter with triple-phase-shift modu-

- lation," *IEEE Transactions on Industrial Electronics*, vol. 68, no. 8, pp. 7350–7361, Aug. 2021, doi: 10.1109/TIE.2020.3007113.
- [18] S. Han, I. Munuswamy, and D. Divan, "Preventing transformer saturation in bi-directional dual active bridge buck-boost DC/DC converters," in *2010 IEEE Energy Conversion Congress and Exposition*, Atlanta, GA, USA, 2010, pp. 1450–1457.
- [19] N. M. L. Tan, T. Abe, and H. Akagi, "Design and performance of a bidirectional isolated DC-DC converter for a battery energy storage system," *IEEE Transactions on Power Electronics*, vol. 27, no. 3, pp. 1237–1248, Mar. 2012.
- [20] M. Stojadinović, E. Kalkounis, F. Jauch, and J. Biela, "Generalized PWM generator with transformer flux balancing for dual active bridge converter," in *2017 19th European Conference on Power Electronics and Applications*, Warsaw, Poland, 2017, pp. 1–10.
- [21] Y. Panov, M. M. Jovanović, and B. T. Irving, "Novel transformer-flux-balancing control of dual-active-bridge bidirectional converters," in *2015 IEEE Applied Power Electronics Conference and Exposition*, Charlotte, NC, USA, 2015, pp. 42–49.
- [22] Y. Panov, M. M. Jovanović, G. Liu, and Y. Y. Meng, "Transformer-flux-balancing control in isolated bidirectional DC-DC converters," in *2014 IEEE Applied Power Electronics Conference and Exposition*, Fort Worth, TX, USA, 2014, pp. 49–56.
- [23] J. A. Claassens and I. W. Hofsafer, "A flux balancer for phase shift ZVS DC-DC converters under transient conditions," in *21st Annual IEEE Applied Power Electronics Conference and Exposition*, Dallas, TX, USA, 2006, pp. 523–527.
- [24] K. Takagi and H. Fujita, "Dynamic control and performance of a dual-active-bridge DC-DC converter," *IEEE Transactions on Power Electronics*, vol. 33, no. 9, pp. 7858–7866, Sep. 2018.
- [25] B. Zhao, Q. Song, W. H. Liu, and Y. M. Zhao, "Transient DC bias and current impact effects of high-frequency-isolated bidirectional DC-DC converter in practice," *IEEE Transactions on Power Electronics*, vol. 31, no. 4, pp. 3203–3216, Apr. 2016.
- [26] S. M. Kaviri, H. Hajebrahimi, S. Eren, and M. Pahlevani, "A digital active DC-eliminating method for DC/DC converters," *IEEE Transactions on Power Electronics*, vol. 34, no. 4, pp. 3014–3019, Apr. 2019.
- [27] F. Xiong, J. Y. Wu, Z. C. Liu, and L. L. Hao, "Current sensorless control for dual active bridge DC-DC converter with estimated load-current feedforward," *IEEE Transactions on Power Electronics*, vol. 33, no. 4, pp. 3552–3566, Apr. 2018.
- [28] C. Sun and X. D. Li, "Fast transient modulation for a step load change in a dual-active-bridge converter with extended-phase-shift control," *Energies*, vol. 11, no. 6, pp. 1569, May 2018.
- [29] W. S. Song, N. Hou, and M. Y. Wu, "Virtual direct power control scheme of dual active bridge DC-DC converters for fast dynamic response," *IEEE Transactions on Power Electronics*, vol. 33, no. 2, pp. 1750–1759, Feb. 2018.
- [30] Q. Ren and S. Ai, "A three degree freedom optimal control strategy of dual-active-bridge converters for full range operations," *Proceedings of the CSEE*, vol. 40, no. 11, pp. 3613–3621, Jun. 2020.



Qiang Ren received a B.S. degree in Electronic Information Engineering from Ocean University of China, Shandong, China, in 2008, and a Ph.D. degree in Electrical Engineering from Naval University of Engineering, Hubei, China. He is a Lecturer at Naval University of Engineering, currently. His research interests include high-power density power conversion and its combined converters.



Fei Xiao received his B.S. and M.S. degrees in Electrical Engineering from Naval University of Engineering, Hubei, China; and his Ph.D. degree in Electrical Engineering from Zhejiang University, Zhejiang, China, in 2012. He is currently a Professor at Naval University of Engineering, Hubei, China. His main research interests include wind power generation and DC area distribution.



Jilong Liu received B.S. and Ph.D. degrees from Xi'an Jiaotong University, Xi'an, China, in 2010 and 2015, respectively. He is currently a Professor at Naval University of Engineering, Hubei, China. His main research interests include modular multilevel converter and AC motor control.



Peng Chen received B.S. degree in Electrical Engineering from Southeast University, Nanjing, China, in 2016, where he is currently pursuing the Ph.D. degree in Electrical Engineering. He has been working as a joint training student in the National Key Laboratory for Vessel Integrated Power System Technology, Wuhan, China, since 2017. His research interests include modular multilevel DC-DC converter, and control strategies for multi-port converter.



Zhichao Zhu received his B.S. degree in Naval University of Engineering Wuhan, China, in 2017. He is currently working toward the Ph.D. degree in Naval University of Engineering. His current research interests include modular multilevel converter and large-capacity energy storage converter.