Soft-Switching Modulation Strategy of High-Frequency Dual Active Bridge Converter Based on Triple Phase-Shifting Control

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Abstract. Dual active bridge (DAB) converters are widely used in work scenarios that require energy efficient bidirectional flow. In order to reduce the on-state loss of the converter while reducing the switching loss to improve the transmission efficiency of the converter, this paper analyzes the waveforms of the DAB converter in different modes, and derive the constraint conditions to ensure the zero voltage switching (ZVS) of the low-voltage side under different working modes. On this basis, combined with the optimal control algorithm of inductor current stress, a quasi-optimal low-voltage side ZVS modulation strategy of inductor current stress is proposed. This strategy improves the working conditions of the device in the full power range, further eliminates the switching loss under the premise of reducing the conduction loss, and greatly improves the efficiency. The efficiency improvement is particularly obvious in the low power section, which is beneficial to further improve the switching of the converter's Frequency and power density. Finally, an experimental platform is built for verification, and the experimental results verify the effectiveness of the proposed modulation strategy.

1. Introduction
In recent years, DC microgrids, new energy vehicles, and energy storage systems have developed rapidly. As a key device among them, dual active bridge (DAB) converters have also received extensive attention and applications[1-2]. DAB converters have the advantages of high efficiency, high power density, wide-range soft switching, and bidirectional energy flow. Therefore, DAB converters have high research value. The loss of the DAB converter is mainly composed of the on-state loss of the switching device and the switching loss. Considering that SIC and GaN devices have gradually been used in actual production in recent years, the application requirements of DAB converters have gradually developed to high frequency. While reducing the current stress, try to ensure that the converter's full power range zero voltage Switching (ZVS) is the key to its efficient operation.

Literature [3] proposed a full power range ZVS control strategy based on extended phase shift (EPS), and optimized the inductor current stress, but because EPS can be regarded as triple phase shift (TPS) Is a special case, the result is not the optimal solution. Literature [4] proposed an optimal strategy for inductor current stress based on TPS, which can achieve the minimum current stress of DAB in the full...
power range and greatly reduce the on-state loss. However, due to the influence of dead zone and junction capacitance, this strategy can only achieve ZVS with fewer switching devices, which is not applicable in high-switching frequency scenarios.

Based on optimizing the TPS operating mode, this paper uses the low-voltage side ZVS as a constraint to solve the feasible region of the phase shift angle, and achieves a substantial increase in the number of low-power switching devices ZVS. Further combining with the existing optimal control method of inductor current stress, a ZVS modulation strategy for the low-voltage side of dual active bridges with optimized inductor stress is proposed. Finally, an experimental platform is built to verify the correctness of the theoretical analysis and the effectiveness of the control strategy in this paper.

2. DAB topology and its working principle
The typical topology of the DAB converter is shown in Figure 1. In the figure: \( U_{in} \) and \( U_{out} \) are the DC side voltages of the full bridge \( H_1 \) and \( H_2 \) respectively; \( U_{ab} \) is the primary side full bridge \( H_1 \) AC voltage measurement, and \( U_{cd} \) is the secondary side full bridge \( H_2 \) AC measurement voltage; \( L \) is the sum of the auxiliary inductance and the leakage inductance of the transformer equivalent to the primary side; \( TF \) is the high-frequency transformer with a transformation ratio of \( n:1 \); \( C_1 \) and \( C_2 \) are the DC side filter capacitors. In order to simplify the process of analyzing the problem, it is assumed that the power is transmitted from the primary side to the secondary side and \( K \geq 1 \), where \( K \) is defined as \( \frac{U_{ab}}{nU_{cd}} \).

When the DAB converter adopts TPS modulation, there are three control variables \( D_1, D_2, \) and \( D_3 \). Among them, \( D_1, D_3 \) are the internal phase shift angles of the two full bridges, \( D_2 \) is the external phase shift angle between the two full bridges, the power transmission of the converter is controlled by controlling \( D_1, D_2, \) and \( D_3 \), and \( D_1, D_2, D_3 \) The value range of is \([0,1]\), \( T_{hs} \) is half a switching period, \( f_s \) is the switching frequency, and \( f_s = \frac{2}{T_{hs}} \) is satisfied.
3. DAB low-voltage side ZVS modulation strategy

3.1 Low power section control strategy

![Figure 2. Low-power operating waveform](image)

Figure 2 shows the steady-state operating waveforms of the converter in the low-power section under low-side ZVS strategy modulation. It can be seen from the figure that the working waveform of the DAB converter is symmetrical in the half cycle, so the following analysis is carried out in the half cycle. When the inductor current satisfies $i_L(t) = -i_L(t + T_{hs})$ during steady-state operation, the time domain expression of $i_L(t_0) - i_L(t_3)$ can be obtained, and combined with the low-voltage side ZVS condition, the switching device action time in the half cycle can be obtained.

Constraints of inductor current:

$$
0 < t_0 < t_1 < t_2 < t_3
$$

$$
I_{zvs} < I_L(t_0) < I_L(t_3) < 0
$$

Since in this mode $K \geq 1$ and power forward transmission, $i_L(t_1) \geq i_L(t_0)$ is always satisfied, so formula (1) can be simplified to:

$$
\frac{nU_{cd}T_{hs}}{2L} [1 - D_3 - K(1 - D_2 - D_3)] = I_{zvs}
$$

In the formula: $I_{zvs}$ is the minimum inductor current to achieve low-voltage side ZVS, $I_{zvs} > 0$

The expression of the transmission power of the DAB converter is:

$$
P = \frac{1}{T_{hs}} \int_0^{T_{hs}} U_{ab} i_L(t) dt
$$

Take the power reference value as:

$$
P_N = nU_{ab} U_{cd} / (8f_s L)
$$

The power expression after unitization of the low power section can be obtained as:

$$
P_o = P / P_N = 2D_2(1 - D_3)
$$

Combining formula (2)(5), we can get the expressions of $D_1$ and $D_2$ with respect to $P$:...
Where: $a = 2L_{ms}/(nU_{cd}T_{hs})$.

Combining with literature [2] the optimal control strategy of inductor current stress, it can be known that the phase shift angle of the low power section satisfies $D_2 + D_3 = D_1$, and the expression of $D_3$ with respect to $p$ can be obtained:

$$D_3 = 1 - \frac{(2 - K)(a + \sqrt{a^2 + 2(K-1)P}) - aK}{2(K-1)}$$  \hspace{1cm} (7)

3.2 Mid-power section control strategy

Figure 3 shows the steady-state operating waveforms of the mid-power section. As mentioned in the previous analysis, the low-voltage side ZVs constraint conditions and the expression after the standardization of the power of the mid-power section can be obtained:

$$D_1 = 1 - \frac{\sqrt{a^2 + 2(K-1)P - a}}{2(K-1)}$$
$$D_2 = \frac{\sqrt{a^2 + 2(K-1)P + a}}{2}$$

$$D_3 = \left\{ \begin{array}{ll} 
K & \text{if } D_2 - D_1 \\
1 + a & \text{if } D_3 \\
0 & \text{else} \end{array} \right.$$  \hspace{1cm} (6)

Combining equations (8)(9) can obtain the expressions of $D_1$ and $D_2$ with respect to $P$:

$$D_1 = \frac{\sqrt{K^2(1-P^2) - (1-a)^2}}{K}$$
$$D_2 = \frac{K + \sqrt{K^2(1-P^2) - (1-a)^2} - 1 + a}{2K}$$
$$D_3 = \begin{cases} 
K & \text{if } D_2 - D_1 \\
1 + a & \text{if } D_3 \\
0 & \text{else} \end{cases}$$  \hspace{1cm} (10)
3.3 High-power section control strategy

When the transmission power is further increased, the DAB converter adopts the control method of optimizing TPS mode 2, as shown in formula (11), it can realize ZVS of all devices with minimum current stress, and its steady-state operating waveform is shown in Figure 4.

$$\begin{align*}
D_1 &= (K - 1) \sqrt{\frac{1 - P}{K^2 - 2K + 2}} \\
D_2 &= \frac{K - 2}{2} \sqrt{\frac{1 - P}{K^2 - 2K + 2}} + \frac{1}{2} \\
D_3 &= 0
\end{align*}$$

(11)

4. Experiment

In order to verify the low-voltage side ZVS modulation strategy proposed in this article, this article uses TI's TMS320F28069 as the control chip to build an experimental platform for dual active bridge DC converters in the laboratory. The specific parameters are shown in Table 1.

| Quantity                      | Symbol | Values |
|-------------------------------|--------|--------|
| Input voltage                 | $U_{in}$ | 600V   |
| Output voltage                | $U_{out}$ | 70V   |
| Switching frequency           | $f_s$  | 100 kHz |
| Leakage inductance reflected to the primary side | $L$ | 30 uH |
| Transformer turn ratio        | $n$ | 4:1 |

In the experiment, the input voltage is 600V and the output voltage is 70V. Figures 5(a) and 6(a) are the voltage waveforms on both sides of the transformer and the primary inductor current waveforms of the transformer working in the low and medium power sections, respectively. Figures 5(b), (c) and, Figure 6(b), (C) The $V_{ds}$ and $V_{gs}$ waveforms of the switching devices $P_1$, $S_1$ and $S_4$ in the low and medium power stages, respectively. It can be seen from the figure that the $V_{ds}$ of the corresponding switching device has been reduced to 0V before the rising edge of its driving signal comes, that is, the corresponding switching device has realized ZVS.
Because the DAB converter works in a steady state, the upper and lower tubes of the same bridge arm are symmetrical, so it can be inferred from Figures 5 and 6 that DAB achieves ZVS for all switching devices except \( P_2 \) and \( P_3 \) in the low and medium power range. Since the high-power stage converter continues to use the control method of optimizing TPS mode 2 and can realize the ZVs of all switching devices, repeated verification will not be performed here.

**Figure 5.** Experimental waveform when \( U_{\text{out}} = 70\, \text{V}, \ P_o = 700\, \text{W} \).

**Figure 6.** Experimental waveform when \( U_{\text{out}} = 70\, \text{V}, \ P_o = 3.5\, \text{KW} \).

5. **Conclusion**

Based on the existing optimal control algorithm for inductor current stress, this paper proposes a quasi-optimal low-voltage side ZVS modulation strategy with low-voltage side ZVS as the constraint condition. Compared with the inductor current stress optimization algorithm, this strategy optimizes the inductor current stress and at the same time, by ensuring the low-voltage side ZVS, the number of
switching devices that can achieve ZVS in the low-power section of the converter is increased from two to six. Which greatly reduces the switching loss of the converter in the low power section and improves the transmission efficiency of the converter. The proposed strategy gives the analytical expressions of phase shift angle and power, which is convenient for use in practical engineering.

Finally, the effectiveness of the modulation strategy proposed in this paper is verified through experiments, which can effectively improve the switching conditions of the converter in the low-power section. In order to further improve the transmission efficiency, how to ensure the ZVS of the high-voltage side while ensuring the ZVS of the low-voltage side can be studied later.

References
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