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Globally Unified ZVS and Quasi-optimal Minimum Conduction Loss Modulation of DAB Converters

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Abstract—Due to the quick development of wide bandgap power devices, the power density of the converters is expected to be much higher. In order to increase the power density and efficiency of the DAB converter, especially the efficiency under light load, a quasi-optimal current root-meansquare (RMS) control strategy to achieve All-ZVS is proposed by taking the inductor RMS current optimization algorithm into account in this paper. First the constraint conditions for realizing ZVS of all switching devices under different operating modes are concluded by analyzing the switching mode and the working principle of the DAB converter. The ZVS domain for All-ZVS operation in each operating mode is further obtained. Then, the GOC strategy is combined with the ZVS condition to derive the novel modulation for all operating modes. Based on this scheme, the switching conditions of all switches can be improved, and the efficiency of the DAB system under the light load is optimized. With ZVS operation, the switching frequency can be raised. Finally, a GaN HEMT based prototype of DAB converter was built and the switching frequency is 200kHz, the effectiveness of the proposed modulation scheme was verified by the experimental results.

Index Terms—Dual Active Bridge converter, Current Control, Conduction Loss, Triple Phase Shift Modulation, ZVS

I. INTRODUCTION

With the rapid growth of renewable energy, electric vehicle, data centers and the development of distribution system or DC micro-grid [1]-[2], the demand for bidirectional AC/DC power converters is also increasing [3]. The isolated bidirectional dc-dc converters, which are applied to adjust DC voltage in AC/DC conversion system, have been an attractive solution for these increasing new requirements [4]-[6]. The dual active bridge (DAB) dc-dc converters have the advantages of high-power density, high voltage conversion ratio, galvanic isolation and easy to realize soft switching. Therefore, these types of converters have continuously attracted more attention in recent years [7]-[9].

The traditional modulation of DAB converter is single phase shift (SPS). It is well known that SPS modulation is able to realize zero voltage switching (ZVS) at heavy-load. However, when the voltage conversion ratio is not equal to one in light load state, the DAB converter cannot realize ZVS operation. Moreover, the increasing of circulating power and RMS current will reduce the efficiency of the converter. To further improve the performance of DAB converters, some novel modulation schemes, such as dual phase shift (DPS) modulation, extended phase shift (EPS) modulation, triple phase shift (TPS) modulation [8] and some other control methods have been proposed [10]. By introducing more control degrees of freedom in these modulations, the loss optimization of the DAB converter has been implemented.

In [10], an asymmetrical pulse-width modulation is proposed to achieve ZVS and ZCS operation for a wide range power and increase the conversion efficiency. When the modulation is used in the case of low output voltage and high-power, the converter has large conduct loss due to the passive rectifier. Besides, the transmission power of this modulation method is limited due to the DCM mode of the primary side. In the modulations of DPS and EPS, the circulating power and current stress can be relatively reduced by optimizing the two control variables. Therefore, the optimal control of expanding the power range of ZVS have been successfully achieved [11]-[14]. Reference [15] has proposed an EPS modulation strategy, which allows DAB converter to operate under the ZVS condition in the whole power range, but the inductor and other components still have large current stress. TPS can control the inner phase shift of two full bridges and the phase shift between them individually, facilitates the most degrees of control freedom. The modulation of SPS, EPS and DPS can be considered as the special cases of TPS, hence the optimizing strategies of DPS and EPS modulation, which can be regarded as the local optimal method, still have limitations.

In order to achieve further power loss optimization for the converter in high frequency application, more work has been done with TPS modulation, such as soft switching operation and conduction loss optimization [16]-[20]. The global optimal...
control (GOC) of the RMS value of the inductor current for the DAB converter in whole power range is proposed by adopting TPS modulation [21], which can minimize the conduction loss for the converter during whole power range. Reference [22]-[23] have investigated the operation of ZVS and reduction of power loss under TPS modulation, proposing a unified PWM control strategy. But his strategy in [22] is complex to be solved and only the numerical solution can be obtained, which is not practical to be widely used. In [23], the control strategy, which uses the fundamental component of the voltage and current signal, is proposed for the case $M=1$ ($M=nV_2/V_1$, $M$ is defined as the voltage conversion ratio between the output voltage ($V_2$) and the input voltage ($V_1$), $n$ is the turn ratio of the transformer). However, the optimization performance of this control strategy is limited. In reference [24], one novel closed form solution of ZVS strategy is proposed for $M<1$. However, for the case of $M>1$, no closed form of optimal modulation scheme has been proposed so far. Besides, the problem of global loss optimization for whole power range by taking both the conduction loss and switching loss optimization into account hasn’t been addressed. Therefore, it is necessary to further study the global optimal analytic solutions to realize ZVS operation and reduce the conduction loss under TPS modulation method [25].

According to the inductor current expression and the constraints of ZVS for selected operating mode, the control domain of realizing All-ZVS control strategy can be obtained. Based on that, combining the previously proposed optimal control strategy of RMS value of the inductor current [21], this paper proposes an All-ZVS control strategy of quasi-optimal RMS current, which not only achieves All-ZVS control, but also optimizes the RMS value of the inductor current. It should be pointed out that a small amount of circulating power is required to realize ZVS control, hence the RMS value of inductor current under this control strategy is larger compared with that under the RMS optimal control strategy [21]. Consequently, a unified analytic method is obtained, which can be well adapted to both cases of $M>1$ and $M<1$. For certain determined transmission power, the control coordinate $D$ can be achieved by solving the analytical function, which is the only solution within the whole control domain for realizing All-ZVS operation, being able to make all switching devices realize ZVS and achieve approximating minimum RMS inductor current. The proposed control strategy not only improves the operating condition of all switches, but also reduce switching and conduction losses, then the converter is able to work in higher frequency, which is useful for improving the power density. Finally, the theoretic analysis and the proposed modulation schemes are verified by the experimental results, which are obtained by an experimental platform working at 200kHz.

II. DERIVATION AND ANALYSIS OF THE GENERAL MODEL OF DAB CONVERTER

Fig. 1 shows the circuit of the DAB converter. The two full bridge circuits (i.e., $H_1$ and $H_2$), which generates AC voltages $v_p(t)$ and $v_s(t)$, are connected by a magnetic tank that includes an inductor $L$ and a high frequency transformer with the turn ratio $n:1$. It is assumed that the power transferred from Port 1 to Port 2 is regarded as positive power. $V_1$ and $V_2$ denote the DC voltages of Port 1 and Port 2, respectively.

The typical waveforms and gate drive signals of the DAB converter controlled by the TPS modulation scheme is depicted in Fig. 2. Note that $T$ is a half of the switching period, and the switching frequency can be expressed as $f_s=1/(2T)$. $D_i$ (i.e., the phase shift ratio between $S_1$ and $S_2$) is the inner phase shift ratio of $H_i$, $D_2$ (i.e., the phase shift ratio between $Q_1$ and $Q_2$) is the inner phase shift ratio of $H_2$, and $D_0$ (i.e., the phase shift ratio between $S_1$ and $Q_1$) is the outer phase shift ratio between $H_1$ and $H_2$.

For TPS modulation scheme, control variables, i.e. $D_0$, $D_1$, and $D_2$ are independent to each other, for convenience, the input control array $D_0$, $D_1$, $D_2$ can be denoted by $D=(D_0, D_1, D_2)$. According to relationship among the three control variables in $D$, TPS modulation scheme can be divided into six operating modes (mode 1 to mode 6), which are listed in Table I. Each control coordinate determines the corresponding transmission power $P_t$ and RMS inductor current $I_{rms}$, and vice versa. According to the calculation formula of $P_t$ and $I_{rms}$, the operation of mode 1, mode 3, mode 4, mode 5 and SPS modulation covers the whole power range, which is shown in Table I. The

<table>
<thead>
<tr>
<th>Mode</th>
<th>Power Range</th>
<th>Field of Definition</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>Lower power</td>
<td>$-1 &lt; D_1 &lt; 0 \land 0 &lt; D_2 &lt; D_0 + D_1 + D_2 &lt; 1$</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>Medium power</td>
<td>$D_1 \leq D_2 \land D_2 \leq D_0 + D_2 \leq 1$</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>Medium power</td>
<td>$D_1 \leq D_2 \land D_2 \leq D_0 + D_2 \leq 1$</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>Medium power</td>
<td>$D_1 \leq D_2 \land D_2 \leq D_0 + D_2 \leq 2$</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>Lower power</td>
<td>$D_0 \leq D_1 \land D_1 \leq D_0 + D_2 \leq 1$</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>Lower power</td>
<td>$D_0 \leq D_1 \land D_1 \leq D_0 + D_2 \leq 1$</td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>SPS High power</td>
<td>$D_0 = D_1 = 0 \land 0 \leq D_2 \leq 0.5$</td>
<td></td>
</tr>
</tbody>
</table>
power ranges of mode 2, mode 3, and mode 6 are already included in those modes, however the RMS inductor current $I_{rms}$ is much higher [21, 23, 26].

It should be pointed out that mode 5 is an alternative to mode 5 to realize the optimal control in low power range of M>1, and the detailed operation of mode 0 will be discussed in section III, part B. Thus this paper will lay emphasis on analyzing mode 1, mode 4, mode 5 and mode 0.

Expressions of transmission power $P_1$ and RMS inductor current $I_{rms}$ for each operating mode are calculated in [26]. For convenience, $P_1$ and $I_{rms}$ need to be normalized. The benchmark value of normalization is defined as,

$$P_{net} = \frac{nV_i^2}{8f_L L}$$

$$I_{net} = \frac{V_i^2}{24f_n^2 L^2}$$

$P_{net}$ denotes normalized value of transmission power, $I_{net}$ denotes normalized value of RMS inductor current square.

Optimization of RMS value of the inductor current is proposed in [21], which realizes minimized conduction losses of the circuit loop. However, the related study hasn’t been implemented on the condition of ZVS operation of all switching devices. To improve the operating condition of all devices of the converter, which is beneficial to increase the switching frequency, efficiency of the converter and power density, this paper proposed an improved control strategy which can facilitate realizing All-ZVS and reducing RMS current of the inductor at the same time in whole power range for both cases of $M>1$ and $M<1$.

### III. ALL-ZVS CONTROL STRATEGY UNDER QUASI-OPTIMAL RMS VALUE OF THE INDUCTOR CURRENT

By controlling the direction of the inductor current appropriately, zero voltage switching (ZVS) can be realized in all switching devices. Taking the waveforms in Fig. 2 into account, the condition for All-ZVS can be summarized as the constraints listed in Table II.

**TABLE II**

<table>
<thead>
<tr>
<th>Switching Devices</th>
<th>Constraints</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_1, S_3, Q_2, Q_1$</td>
<td>$i_i&lt;0$</td>
</tr>
<tr>
<td>$S_2, S_3, Q_2, Q_1$</td>
<td>$i_i&gt;0$</td>
</tr>
</tbody>
</table>

As shown in Fig. 2, the waveform of inductor current of DAB is semi-periodic (0 to $T=T/2$) symmetry in the steady state, therefore when the upper side devices of the half bridge realize ZVS, then the corresponding lower side devices of half bridge can realize ZVS. In other word, the condition of ZVS of switching devices $S_1, S_4, Q_3, Q_4$ is equal to that of switching devices $S_2, S_3, Q_3, Q_1$. There are four favorable operating modes shown in Table I and the corresponding RMS current optimization solution in each mode is different. The control method of quasi-optimal RMS current to achieve All-ZVS for each mode will be discussed. Besides, the voltage gain $M$ affects the current waveform of the inductor. The analysis with $M>1$ and $M<1$ will be implemented individually.

A. The Control Strategy of All-ZVS when $M<1$

1) ZVS control strategy in low power level of mode 4

For the characteristics of semi-periodic symmetry, the operating waveforms and ZVS condition in half of switching period will be taken into consideration. In mode 4, to get the instant inductor current value in half of period, the operation of half period is divided into four intervals according to the applied voltage $U_L$ of the inductor as shown in Fig. 3. The inductor voltage $U_L$ and the changing value of inductor current $Δi_L$ in each interval have been calculated, and further listed in Table III.

**TABLE III**

<table>
<thead>
<tr>
<th>Interval</th>
<th>Inductor voltage $U_L$</th>
<th>Change Amounts of Inductor Current $Δi_L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0−$D_2T$</td>
<td>$nV_2$</td>
<td>$nV_2D_2T/L$</td>
</tr>
<tr>
<td>$D_2T−(D_2+D_3)T$</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>$(D_3+D_4)T−D_2T$</td>
<td>$-nV_1$</td>
<td>$-nV_2D_2T/2L$</td>
</tr>
<tr>
<td>$D_2T−T$</td>
<td>$V_fV_2$</td>
<td>$V_fV_2(1-D_2)T/L$</td>
</tr>
</tbody>
</table>

Because the waveform of inductor current is semi-periodic symmetric, i.e. $i_i(0)=i_i(T)$. Combined with the total change of inductor current in a half of switching period, the value of inductor current at zero time can be calculated. Then according to Table III, the inductor current expression at each interval of switching action can be obtained, listed in Table IV.

**TABLE IV**

<table>
<thead>
<tr>
<th>Switching Moment</th>
<th>Switching Device</th>
<th>Expression of Inductor Current</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>$S_1$</td>
<td>$-T[1−(D_2W_2+(D_2+D_3−1)nV_2)/2L]$</td>
</tr>
<tr>
<td>$D_2T$</td>
<td>$Q_1$</td>
<td>$-T[1−(D_2+D_3−1)nV_2)/2L]$</td>
</tr>
<tr>
<td>$(D_3+D_4)T$</td>
<td>$Q_2$</td>
<td>$-T[1−(D_3−nV_2)/2L]$</td>
</tr>
<tr>
<td>$D_2T−T$</td>
<td>$S_2$</td>
<td>$T(D_2−nV_2)+D_2+2D_3−2D_2nV_2)/2L$</td>
</tr>
<tr>
<td>$T$</td>
<td>$S_2$</td>
<td>$T(D_1−nV_2)+D_2+2D_3−nV_2)/2L$</td>
</tr>
</tbody>
</table>

The constraints of realization for All-ZVS can be concluded from Table IV and Table V. Moreover, when the transmission power is positive, $i_i(0)=i_i(T)$ can always be satisfied, according to the range of control coordinates $D=(D_0, D_1, D_2)$ of this mode.

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Fig. 3. The operating waveforms of Mode 4 and the corresponding definition scope.
So, the constraints for realizing All-ZVS can be simplified as follows,

\[
\begin{align*}
(1-D_1)V'_1 + (D_1-1)nV_2 &< 0 \\
(D_1+1)V'_1 + (1+D_2+2D_0-2D_1)nV_2 &< 0
\end{align*}
\] (2)

Fig. 4(a) illustrates the sets of control coordinates \( \textbf{D}=(D_0, D_1, D_2) \) of mode 4, which satisfies the constraints in (2). When the control coordinate is located in the polyhedron, all devices can realize ZVS operation.

According to Table V, a small amount of circulating power is required to achieve All-ZVS in DAB converter under the operating mode 4, which will increase \( \hat{P}_{n,\text{rms}} \). Combined with the optimal control strategy of RMS current of inductor [21], the quasi-optimal RMS current controller with All-ZVS modulation strategy can be obtained, which realizes All-ZVS and has small \( \hat{P}_{n,\text{rms}} \). The solution for the problem above will be discussed as following.

By analyzing the relationship between \( P_{n,t}\text{,} \hat{P}_{n,\text{rms}} \) and control coordinate \( \textbf{D} \), the optimal control strategy of RMS current of the inductor has been proposed as (3) with mode 4 [21]:

\[
\begin{align*}
D_0 &= \frac{(1-M)(1-D_1)}{M} \\
D_1 &= 1 - \frac{1-D_2}{M}
\end{align*}
\] (3)

The control coordinates given in (3) are exactly on the boundary line of the polyhedron shown in Fig. 4(a). In other words, under the constraint given in (3), the current of the inductor reaches zero at the time \((D_0+D_2)T/2 + D_1T\) and \(D_2\), which does not allow the devices to achieve ZVS. It’s interesting that the inequality in (2) will hold if reducing \( D_2 \) and keeping the relationship of \( D_0 \) and \( D_1 \) to meet the requirement for the optimal control of inductor current. To evaluate the degree of reduction of \( D_2 \), the parameter \( \alpha \), which is chosen from 0 to 1, is added into the second equation in (3). The solution of All-ZVS control and RMS current optimal control strategy of inductor is revised as follows,

\[
\begin{align*}
D_0 &= \frac{(1-M)(1-D_1)}{M} \\
D_1 &= \alpha \left( 1 - \frac{1-D_2}{M} \right)
\end{align*}
\] (4)

\( D_1 \) is determined by the transmission power. The operation of reducing \( D_2 \) will increase a certain amount of circulating power essentially and the condition of All-ZVS is realized. With the above calculation, the control coordinate for All-ZVS control and RMS current optimal control strategy of inductor is extremely close to the RMS current optimal control strategy of inductor, which realizes quasi optimized RMS current control with All-ZVS control for DAB converter.

\( D_2 \) not only determines the current value at the time \((D_0+D_2)T/2 \) and \( D_1T \), but also affects the RMS current of the inductor. \( D_2 \) is determined by \( \alpha \), hence it’s important to select an appropriate parameter \( \alpha \). The value of the current at the time \((D_0+D_2)T/2 \) and \( D_1T \) will increase with \( \alpha \) gradually decreasing, which reduces the time for parallel capacitance of the MOSFET to discharge. It can be concluded that it’s easier for the devices to achieve ZVS with smaller \( \alpha \). As mentioned above, \( \alpha \) needs to be small enough for the reason that the current is large enough at the time \((D_0+D_2)T/2 \) and \( D_1T \) for \( Q_1 \), \( Q_4 \) and \( S_4 \) to achieve ZVS. However, the RMS current of the inductor will increase significantly under the same transmission power. In this paper, the parameter \( \alpha \) is determined according to the experimental experience value.

Taking the case of voltage conversion ratio \( M=0.7 \) for an example, Fig. 4(b) illustrates the waveforms of \( \hat{P}_{n,\text{rms}} \) versus the transmission power \( P_{n,t} \) when \( \alpha \) takes different value. As shown in the figure, when \( \alpha \) is decreasing, \( \hat{P}_{n,\text{rms}} \) significantly increases at the same power. So \( \alpha \) has to be selected as large as possible, within the domain of achieving ZVS. When \( \alpha \) is near 1 and the operating point is close to \( P_{n,t}=2M(1-M) \), the larger dead time is needed to discharge the parasitic capacitance of MOSFET completely, which will make \( S_3 \) and \( S_4 \) more difficult to achieve ZVS. Therefore, \( \alpha \) can be selected according to the detailed parameters of DAB converter. The parameter \( \alpha \) is usually selected to be 0.65.

The expression (5) illustrates the applicable power range for DAB converter operating in this mode. The control scheme of the mode 4 is suitable for low power level:

\[
P_{n,t} \in [0, 2M(1-M)]
\] (5)

2) ZVS control strategy in medium power level

Fig. 5. The operating waveforms of mode 1 and the corresponding definition scope with \( M \leq 1 \).

Fig. 6. (a) The ZVS control coordinate set of mode 1. (b) Curve of \( \hat{P}_{n,\text{rms}} \) versus \( P_{n,t} \) of DAB for different \( D_2 \) when \( M=0.7 \)

The main operating waveforms and the corresponding control domain of mode 1 have been given in Fig. 5. By analyzing Fig. 5 with the same analysis method of mode 4, the instant inductor current at the action time of each switching device and the constraints of realizing All-ZVS can be
concluded separately. The constraints of All-ZVS can be obtained by further analysis as follows,

\[
\begin{align*}
& (D_1 - 1)V_1 + (1 - 2D_0 - D_1 + 2D_1)nV_2 < 0 \\
& (2D_0 - D_1 - 1)V_1 + (1 - D_2)nV_2 > 0
\end{align*}
\]  

(6)

Fig. 6(a) illustrates the sets of control coordinate \( D=(D_0, D_1, D_2) \) that satisfies the constraints in (6). When the control coordinate is located in the polyhedron as shown in Fig. 6(a), the converter can achieve All-ZVS operation in mode 1.

As shown in Fig. 5, the circulating power still exists in mode 1, resulting in the increasing of \( P_{n,\text{rms}} \). The RMS current optimal control strategy of the inductor is proposed in reference [21]:

\[
\begin{align*}
D_0 &= \frac{-1 + D_1 + M + DM + \sqrt{(D_1 - 1)^2 + M^2(D_1^2 - 1)}}{2M} \\
D_2 &= 0
\end{align*}
\]  

(7)

As further analyzed from Fig. 5, within the domain of mode 1, the control variables satisfy the condition of \( D_1 \leq D_0 \), meanwhile, the change value of inductor current is positive during the interval \((D_1T, D_0T)\), so the time for inductor current crossing zero can be selected at \((D_1 + D_0)T/2\), which forces all the switching devices to satisfy the constraints.

It is indicated in optimal control strategy of inductor current that RMS value of the inductor current can be reduced by reducing \( D_2 \) in mode 1 [21], so All-ZVS control strategy can be obtained by selecting \( D_0=0 \).

\[
\begin{align*}
D_1 &= (1 - D_0)\frac{1 - M}{M} \\
D_2 &= 0
\end{align*}
\]  

(8)

Taking the voltage conversion ratio \( M=0.7 \) as an example, Fig. 6(b) illustrates the waveforms of \( P_{n,\text{rms}} \) versus the transmission power when \( D_2 \) takes different value. As shown in the figure, \( P_{n,\text{rms}} \) will decrease significantly when \( D_2 \) decreases at the same power level. So \( D_2 \) should be reduced as small as possible within the definition domain of mode 1.

The expression (9) illustrates the applicable power range for DAB converter operating in this mode. The control scheme of this mode is suitable for medium power levels.

\[
\begin{align*}
P_{n,\text{rms}} &\in \left[ 2M(1 - M), \frac{2(M^2 - 1 + \sqrt{1 - M^2})}{M^2} \right]
\end{align*}
\]  

(9)

When the transmission power is further increased, the traditional SPS control strategy can be applied to realize minimized RMS \( P_{n,\text{rms}} \) in DAB converter. The corresponding range of transmission power is as follows,

\[
\begin{align*}
P_{n,\text{rms}} &\in \left[ \frac{2(M^2 - 1 + \sqrt{1 - M^2})}{M^2}, 1 \right]
\end{align*}
\]  

(10)

Then the range of \( D_0 \) can be obtained as:

\[
D_0 \in \left[ \frac{-1 + M + \sqrt{1 - M^2}}{2M}, 1 \right]
\]  

(11)

For the case of \( M \leq 1 \), the domain of the realization of All-ZVS for each device by SPS control scheme is as follows,

\[
D_0 \geq \frac{M - 1}{2M}
\]  

(12)

It can be verified that equation (11) meets the constraints proposed by (12), so within the transmission power range in (10), called high power level, All-ZVS and optimized \( F_{n,\text{rms}} \) can be achieved by using SPS control strategy at the same time.

Based on the above analysis for the operating condition \( M<1 \), All-ZVS modulation with quasi optimal RMS current control of the inductor can be realized by using the proposed method in the whole power range.

B. The Control Strategy of All-ZVS for DAB when \( M>1 \)

In the existing DAB optimal control strategy literature, the case of reverse power transmission is seldom considered. The reverse power transmission for \( M \leq 1 \) transmitted from PORT 2 to PORT 1 is equivalent to the forward power transmission for \( M>1 \) transmitted from PORT 1 to PORT 2, so the reverse power transmission can be considered and analyzed as the case of \( M>1 \).

The following sections will analyze different power levels in the case of \( M>1 \).

1) ZVS control strategy in low power level

As reference [21] pointing out, mode 5 can be applied to achieve optimal control of RMS inductor current in low power range when \( M>1 \). Employing the same method as \( M \leq 1 \), based on the ZVS condition for the devices, both the constraints of realizing All-ZVS and the instant inductor current at the switching moment of each device can be obtained. The constraints for achieving All-ZVS in this operating mode can be achieved as follows,

\[
\begin{align*}
i_L(0) &< 0 \\
i_L(D_0T) &> 0 \\
i_L(D_1T) &< 0 \\
i_L((D_0 + D_2)T) &> 0
\end{align*}
\]  

(13)

Because \( i_L(D_0T)=i_L(D_1T) \) in this mode, the constraints discussed above cannot be satisfied at the same time, therefore All-ZVS control cannot be realized in mode 5.

Fig. 7. The operating waveforms of Mode 0 and the corresponding definition scope

As shown in Fig. 3 in mode 4, all time periods when \( v_p \) is high is within the time period when \( v_s \) is high. So, all time periods when \( v_p \) is high should be within the time period when \( v_p \) is high in mode 5, which means that \( D_0=0 \) should be satisfied in mode 5, then the appropriate solution can be constructed. \( D_0=0 \) means Q1 of secondary side should be turned on before \( S_1 \) of primary side, which is called mode 0 in this paper, as illustrated in Fig. 7.

According to the constraints of inductor current of All-ZVS, and the condition, \( i_L(0)=i_L(D_1T) \), the in-equation \( i_L((D_2+D_0)T)>0 \) is always satisfied in this mode. The constraints of realization of All-ZVS are as follows,
The critical plane \( D_0 + D_2 = 1 \)

The critical plane \( i_L(T) = 0 \)

The critical plane \( i_L(D_1T + T) = 0 \)

Fig. 8. (a) The ZVS control coordinate set of Mode 0. (b) Curve of \( \dot{P}_{\text{rms}} \) versus \( P_{\text{n,t}} \) of DAB for different \( \alpha \) when \( M=1.2 \)

\[
\begin{align*}
(1 - D_1)V_1 + (D_2 - 1)nV_2 &> 0 \\
(2D_0 - D_1 + 1)V_1 + (D_2 - 1)nV_2 &< 0
\end{align*}
\]  
(14)

Fig 8(a) illustrates the sets of control coordinate \( D=(D_0, D_1, D_2) \), which satisfy constraints (14). When the control coordinate is located in polyhedron as shown in Fig 8(a), all the switching devices are able to achieve ZVS operation.

The solution of optimal control strategy of inductor current in this power level proposed by reference [21] is as follows,

\[
\begin{align*}
D_0 &= 0 \\
D_1 &= 1 + MD_2 - M 
\end{align*}
\]  
(15)

Within the control domain of mode 0, the control variables satisfy the constraints: \( D_1 < D_2 + D_0 \), during the time interval \( (D_0T, 0) \), the change value of inductor current is negative, so the time of inductor current crosses zero is selected at the time of \( D_0T/2 \). Then the constraints of All-ZVS in mode 0 are satisfied. The zero crossing condition of the inductor current at this point can be equal to the condition shown in (16).

\[
1 - D_1 + D_0 = M(1 - D_2) \\
\Rightarrow D_1 - D_0 = 1 + MD_2 - M
\]  
(16)

According to the optimal control strategy of inductor current, when \( D_0=0 \), it is the optimal modulation strategy for the RMS current of the inductor. However, when \( D_0=0 \), some switching devices cannot achieve ZVS. Therefore, ZVS control strategy in this mode should be further optimized as follows,

\[
\begin{align*}
D_1 &= \alpha(1 + MD_2 - M) \\
D_0 &= (\alpha - 1)(1 + MD_2 - M)
\end{align*}
\]  
(17)

\( D_2 \) is determined by the transmission power. For parameter \( \alpha \), it can be selected between zero and one. Taking the voltage conversion ratio \( M=1.2 \) as an example, Fig. 8(b) illustrates the curve of \( P_{\text{rms}} \) versus transmission power when \( \alpha \) is selected from different values. It’s easier for the devices to achieve ZVS with smaller \( \alpha \). However, the large circulating power is generated, namely, the RMS current of inductor current will increase significantly when the same power is transferred by selecting a smaller \( \alpha \). When \( \alpha \) is relatively large, the parasitic capacitor of the several devices cannot be completely discharged at some operating point. Therefore, the selection of parameter \( \alpha \) needs to consider the actual operating condition of converters. As shown in the figure, within the domain of realization of All-ZVS, the value of \( \alpha \) should be selected as large as possible to achieve quasi RMS current optimization. The parameter \( \alpha \) is selected to 0.8 according to the experiment results.

\[
\begin{align*}
D_0 &= 0 \\
D_1 &= 1 - D_2 - M + D_2M + \sqrt{D_2^2 - 1 + M^2(1 - D_2^2)} \\
D_1 &= 0
\end{align*}
\]  
(20)

Within the control domain of mode 1 shown in Fig. 10(b), the control variables satisfy the constraints: \( D_1 < D_0 \). At the interval \( (D_1T, D_0T) \), the change amount of inductor current is positive, so the inductor current at \( (D_1T + D_0T)/2 \) is selected to be zero. The constraints denoted by equation (19) are satisfied naturally. Moreover, according to the optimal control strategy of inductor current in this power level, RMS current of inductor current is minimum when \( D_0=0 \). Therefore, the control strategy of realization of All-ZVS with quasi RMS current optimization in this power range is obtained as follows,
Taking the voltage conversion ratio $M=1.2$ as an example, Fig. 10(b) illustrates the waveforms of $P_{n,ms}$ versus the transmission power when $D_1$ is selected from different value. As shown in the figure, $P_{n,ms}$ will significantly decrease when $D_1$ is decreased when the same power is transferred. The value of $D_0$ should be selected as small as possible at model 1.

Under the condition of $M>1$, the operation of mode 1 is in the medium power level, which is denoted by equation (22).

$$P_{ad} = \left[ \frac{2}{M^2} (M^2 - 1), 2(1 - M^2 + M\sqrt{M^2 - 1}) \right]$$

With the transmission power further increasing, by the traditional SPS control strategy, the DAB converter can realize optimal control of $P_{n,ms}$. The corresponding power range is as follow,

$$P_{ad} = \left[ 2(1 - M^2 + M\sqrt{M^2 - 1}), 1 \right]$$

Then the range of $D_0$ can be solved as in (24).

$$D_0 \in \left[ \frac{1 - M + \sqrt{M^2 - 1}}{2}, \frac{1}{2} \right]$$

For $M>1$, the constraints of the realization of SPS control strategy is shown in (25).

$$D_0 < \frac{1 - M}{2}$$

**Fig. 11. Flowchart of the modulation scheme.**

$$\begin{align*}
D_0 &= 1 - \frac{M}{M - 1} \\
D_1 &= 0
\end{align*}$$

(21)

IV. EXPERIMENTAL RESULT

A prototype of the DAB converter, with the parameters shown in Table VI, was built in the laboratory. In the converter, GaNFET (i.e., GS66508T) and MOSFET (i.e., BSC016N06S) are selected as power devices. The proposed control strategy is implemented in a digital signal processor (i.e., TMS320F28377) that is responsible for sensing of signal, calculating and generating the gate driving signals. In order to test the effectiveness of proposed modulation schemes at different voltage ratio condition for both of $M=1$ and $M \leq 1$, $V_1$ and $V_2$ of the inductor can be realized by using the proposed method in the whole power range. The corresponding solutions for different power range are in (17) and (21).

The flowchart of the modulation scheme for the operating conditions both for $M=1$ and $M>1$ is shown in Fig. 11. The output voltage is controlled by proportional integral (PI) controller and the output of the controller is the normalized value of output power ($P_{o,n}$). Since the control parameters $D_0$, $D_1$ and $D_2$ are related in each mode, therefore $P_{o,n}$ can be represented by one of them, and then the parameter can be solved base on a certain value of $P_{o,n}$, consequently all control parameters can be calculated. The curves of control variables ($D$) versus power and power versus operation condition ($M$) are shown in Fig. 12. It can be seen that the working mode of the converter can be switched as $P_{o,n}$ falls into different power range.

**Fig. 12. Control variables (i.e., $D_0$, $D_1$, $D_2$) versus $P_{o,n}$ and power range versus $M$. (a) $M<1$. (b) $M>1$.**
should be adjustable. Fig. 13 shows the prototype for test. The rating power of the converter is designed to be 1800W.

A. Experimental verification with M≤1

In the experiment of M≤1, the input voltage \( V_1 \) is 300V. For instance, the experimental waveforms with the output voltage \( V_2=20V \) is given to demonstrate the theoretical analysis. Fig. 14(a) and Fig. 15(a) illustrate the inductor current waveforms and the AC waveforms on both sides of the transformer for low power level with mode 4 and medium power level with mode 1, respectively. Fig. 14(b) and Fig. 15(b) illustrate the driving waveforms of switching devices: \( S_2, S_4, Q_2, Q_4 \) and the corresponding waveforms of VDS under different operating power respectively. It is observed that the ZVS operation has been realized.

B. Experimental verification with \( M > 1 \)

For the case of \( M>1 \), the input voltage \( V_1 \) is 200V and the output voltage is \( V_2 \) is 28.8V. Fig. 16(a) and Fig. 17 (a) illustrate the inductor current waveforms and the AC waveforms on both sides of the transformer for low power level with mode 0 and medium power level with mode 1 respectively. Fig. 16(b) and Fig. 17(b) illustrate the driving waveforms of switching devices: \( S_2, S_4, Q_2, Q_4 \) and the corresponding drain to source waveforms VDS, respectively. It is observed that all these devices can realize the operation of ZVS.
Fig. 18 shows the transient experimental results when $V_{in}$ changed from 300V to 200V and $V_o=20V$. During the transition, the converter switches from mode 4 to mode 1. The experimental result verifies the effectiveness of the implemented controller during mode switching.

Fig. 19 and Fig. 20 illustrates the RMS current and efficiency curves which employs the proposed quasi RMS current optimization with All-ZVS control method and GOC (global optimization control) control strategy proposed in [21] at different voltage ratio condition for both of M>1 and M≤1. The RMS current curves for different selected operating points, which is determined by the transferred power $P_t$, and output voltage $V_o$, are depicted in Fig. 19(a) and Fig. 20(a), the corresponding efficiency curves are depicted in Fig. 19(b) and Fig. 20(b). The blue lines mark the corresponding value of the proposed quasi RMS current optimization All-ZVS control strategy, and the red lines mark the corresponding value of the GOC control strategy. It should be noted that the power range of the efficiency comparison given in Fig. 19 and Fig. 20 belongs to low and medium power range. Whereas, when the converter operates in high power range, SPS modulation can help to achieve the control of GOC and ZVS, the conversion efficiency comparison does not need to be given.

As observed from the experimental results, the RMS current of proposed quasi RMS current optimization with All-ZVS control strategy is slightly higher than the GOC control strategy in low and medium power range. As shown in Fig. 19(a), for instance, at the transferred power point of 200W and 909W, the RMS current increases from 1.67A and 4.45A by the GOC control strategy to 1.92A and 4.53A by the proposed quasi optimization with All-ZVS control respectively. However, from Fig. 19(b), the total efficiency of the proposed strategy is higher than the GOC control strategy due to the realization of All-ZVS for the converter which reduces the switching loss. The conversion efficiency increases from 85.23% and 92.18% by the GOC control strategy to 86.37% and 92.27% by the proposed quasi optimization with All-ZVS control respectively. The experiment result in Fig. 20 is similar to that of Fig. 19. It can be concluded that the proposed method which can realize quasi RMS current optimization and All-ZVS at the same time significantly improves the efficiency of the converter in light load range, compared to the experimental results in other literature. Hence the modulation method proposed in this paper is an effective solution with high frequency application, especially when the converter operates in wide range.

ZVS operation of tested power points in this paper has been achieved. However, when the converter operates in extremely lighter load, the inductor current at the switching point might not be large enough to fully discharge the capacitor. Therefore, ZVS operation of the converter might be lost at light load.

V. CONCLUSION

A control strategy that combines quasi inductor current optimization with All-ZVS for DAB converter under TPS modulation is proposed in this paper. Based on the mathematical model of the converter, the operating mode of converter has been divided, and the constraints of realizing All-ZVS under different operating modes has been analyzed and obtained for both $M>1$ and $M≤1$. Combining optimal control strategy of inductor current, a quasi-optimal All-ZVS control strategy is proposed, which not only forces all the switching devices to realize ZVS, but also almost minimizes the conduction losses under the condition of achieving quasi-optimal RMS inductor current. Furthermore, the operating condition of the switching devices is improved. The proposed strategy has analytic expression, which is convenient to be used in implementation.

In summary, this paper proposes a practical and effective control method that can be used to improve the operating condition of switching, to increase the efficiency, and to push the switching frequency to much higher range for DAB converter.

REFERENCES


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