Optimized Control for Modified Push-Pull Dual Active Bridge Converter to Achieve Wide ZVS Range and Low Current Stress

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ABSTRACT

Compared with the traditional dual active bridge (DAB) converter, the push-pull DAB converter can reduce the number of switching devices and is suited for low voltage applications. In order to extend the zero voltage switching (ZVS) range and reduce the current stress within wide voltage range, this paper modifies output rectifier circuit, and proposes an optimized control strategies for push-pull DAB converters. Unlike conventional modulation that only utilizing the primary duty cycle and phase shift angle to adjust the power, the duty cycle for the secondary side is also utilized for the PWM modulation. With the modified PWM modulation, a control law for achieving minimum root mean square (RMS) current within the precondition of achieving wide range ZVS is proposed. The working modes under the modified PWM modulation, current and power analysis and optimized control law are analyzed in depth. Finally, the experimental results using a 1 kW prototype verify the effectiveness of the proposed converter.

INDEX TERMS

Modified push-pull dual active bridge converter, optimized control strategy, minimum RMS current.

I. INTRODUCTION

With the increasing penetration of distributed renewable generations, railway traction systems and new energy electric vehicle charging systems, energy router is one of the most promising technologies which has multi DC buses [1]–[4]. Therefore, it needs the converter to flexibly connect different voltage buses to achieve bidirectional power transmission. Due to the advantages of electrical isolation, high power density, flexible bidirectional power regulation, wide load range zero voltage switching (ZVS) operation, dual active bridge (DAB) converter gets more and more attractive in industry applications [5], [6]. However, when the load is light, it is difficult to realize soft switching, and the high turn-off loss exists at rated load, resulting in lower efficiency. For overcoming the shortages, decoupled extended phase shift (EPS) control utilizing magnetizing current is proposed in [7] to achieve full load range ZVS for DAB converter. Even though there are many methods for improving DAB, due to the large number of switching devices, in low voltage and high current applications, it will bring more conduction losses. At the same time, the drive circuit design is more complicated, which brings further challenges to reliability and safety.

In order to overcome the above shortcomings, scholars have proposed and studied many alternative topologies from the perspective of topology and reducing the number of switching devices. In low power application, like small and medium sized energy storage systems, the bidirectional flyback converter is suitable, which has simple structure, simple control and requires less power components. However, the switching loss exists due to the leakage inductor of the transformer. The output diode also has reverse recovery problem. Thus, the power conversion efficiency is significantly affected. To reduce the influence of leakage inductor, an active clamp circuit is added to the flyback converter in [8], which recovers the energy stored in the leakage inductor and magnetizing inductor of the transformer. It keeps the voltage on the power switch at a specific voltage level. Then, low-rated voltage switches with low on-resistance can be used to improve the power conversion efficiency. In high-power transmission applications, the push-pull circuit is an effective
alternative topology with a small number of switches and is suitable. But when the switches are turned-off, the energy stored in the leakage inductor can cause voltage spikes. The clamp circuit is generally used to absorb this stored energy. In [9], one auxiliary switch and a clamping capacitor are added in the primary side of the push-pull converter to recycle the energy stored in the leakage inductors and clamp the voltage spike. Therefore, for the rectifier diodes in the converter, the reverse recovery of the diodes needs to be suppressed or even eliminated, rather than forced commutation [10], [11]. Jiang et al [12] proposes a new push-pull DC/DC converter topology with complementary active clamped. Based on this design, the switches actively clamp each other, and the natural soft-switching can be achieved. And the secondary side diodes can also achieve zero current switching (ZCS). However, these structures are unidirectional, which cannot be used in the bidirectional applications. In addition, many converters adopt active clamp circuits [13]–[15] or auxiliary inductor-capacitor resonant tank [16] to absorb the energy stored in leakage inductor or achieve the ZVS operation.

In bidirectional applications, a full-bridge push–pull bidirectional DC/DC converter is proposed in [17] with the drawback of voltage spike. For avoiding it, [18] added an active clamp capacitor to improve the performance. For making full use of the advantages of push-pull converter and DAB converter, many studies have combined these two converters. In [19] a novel bidirectional DC/DC converter is proposed, which applies a three switches topology substituting for the full-bridge structure with seven switches in all. In [20], a four-phase current-fed push–pull dual active bridge converter is proposed for wide voltage range applications, which can cancel the dc-flux offset. However, the voltage spike is still existing and the number of magnetized components and switches is also large. In [21], a push-pull forward half-bridge bidirectional DC/DC converter is proposed, which adopts the single phase shift modulation to achieve the power regulation. All the switches operate under ZVS condition at rated output power. Obviously, the disadvantages of this structure are similar to DAB converters, which is that the ZVS will lose in light load conditions. And, because of the half bridge structure in secondary side, the converter cannot extend the phase shift modulation. Thus, when it applied in wide input voltage, it also needs to use variable switching frequency control, which brings control complexity. In [22], a bidirectional push-pull resonant converter is proposed, which uses a bidirectional switch in the secondary side. This topology and corresponding modulation naturally yield almost ZVS turn-off of switches. However, it still has eight switches, which is same with DAB converter, and the ZVS operation cannot be fully realized within full load range.

To avoid these problems, this paper modifies the half bridge circuit in secondary side of the converter in [21] to be full bridge circuit to increase the control degree freedom, and then proposed an optimized control strategy. Different from the conventional modulation which only utilizes the primary duty cycle and phase shift angle to adjust the power, the modified PWM modulation strategy also utilizes the duty cycle of the secondary side. Through the analysis of different working modes under the modified PWM modulation, a control law for achieving the minimum root mean square (RMS) current within the precondition of achieving wide ZVS range is proposed, which is obtained by calculated the partial differential of the RMS current expression. Finally, a 1 kW prototype was built up, and the experimental results compared with those from traditional modulation verify the effectiveness of the proposed optimized control strategy.

The rest of this paper are organized as follows. Section II introduces the topology and analyzes the different operation modes and power characteristics. The analysis of ZVS range under different operation modes for all the switches is presented in Section III, Section IV gives minimum RMS current control strategy, and Section V illustrates the experimental results from a 1 kW prototype, which verify the effectiveness. Finally, the conclusions are made in Section VI.

II. OPERATION MECHANISM ANALYSIS OF PUSH-PULL DAB CONVERTER

The topology of modified push-pull dual active bridge converter is shown in Fig. 1. The push-pull side is composed of two power switching devices, \( S_1 \) and \( S_2 \), a clamp capacitor \( C_c \) and two leakage inductors, \( L_1 \) and \( L_2 \). \( T \) represents the three winds high frequency transformers, where \( N_1, N_2 \) and \( N_3 \) represent their number of coils. Compared with [21], the secondary side is a full bridge structure composed of four power switching devices. The principle of bidirectional power flow control of the converter is the same as the traditional DAB converter, that is, the power transmission is realized by controlling the phase difference between the push-pull side and the secondary side voltage. Because the principle of the backward operation is similar to that of the forward operation, this paper will focus on the forward operation, and the forward direction is defined as from the push-pull side to the secondary side.

Unlike the conventional modulation for push-pull DAB converter [21], this paper utilizes the duty cycle of secondary side (\( D_2 \) shown in Fig. 2) to improve the performance. The key waveforms with the modified PWM modulation are shown in Fig. 2, which are divided into three modes. In Fig. 2, \( D_1, D_2 \) and \( D_3 \) mean duty cycle of phase shift angle between
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FIGURE 2. Key waveforms of different modes. (a) Mode 1. (b) Mode 2. (c) Mode 3. $V_{ab}$ and $V_{cd}$, duty cycle of $V_{cd}$ and duty cycle of phase shift angle between driving signals of $S_1$ and $Q_1$, respectively. And, $T_s$ means switching period. The relationship between $D_1$, $D_2$ and $D_\alpha$ is as follows:

\[
D_\alpha = \frac{1}{2}(1 - D_2 - 2D_1)
\] (1)

Then, the relationship between $D_1$ and $D_2$ under different modes, including Mode 1, Mode 2 and Mode 3, can be obtained as

\[
\begin{align*}
0 & \leq D_2 \leq 1 - 2D_1 \quad \text{Mode 1} \\
1 - 2D_1 & \leq D_2 \leq D_1, \quad D_1 \in [0, 0.5] \\
2D_1 - 1 & \leq D_2 \leq 1, \quad D_1 \in [0.5, 1] \\
0 & \leq D_2 \leq 2D_1 - 1 \quad \text{Mode 3}
\end{align*}
\] (2)

\[\text{Stage 1 [t}_0, \ t_1] \text{ [see Fig. 3(a)]: At } t_0, \ Q_3 \text{ is turned off. At this time, the secondary current } i_3 \text{ changes/discharges the junction capacitors of } Q_3 \text{ and } Q_4 \text{ respectively in the dead time.}
\]

\[\text{Stage 2 [t}_1, \ t_2] \text{ [see Fig. 3(b)]: At } t_1, \ Q_1 \text{ is turned off. For achieving the ZVS operation for } Q_2, \text{ the secondary current } i_3 \text{ needs to be positive. When } Q_2 \text{ is turned on, the voltages across the leakage inductors can be expressed as}
\]

\[
\begin{align*}
V_{L1} &= V_{in} - \frac{V_o}{n} \\
V_{L2} &= \frac{V_o}{n} - V_{CC}
\end{align*}
\] (3)

\[\text{Stage 3 [t}_2, \ t_3] \text{ [see Fig. 3(c)]: At } t_2, \ S_1 \text{ is turned off. For achieving the ZVS operation for } S_2, \text{ the current } i_1 \text{ and } i_2 \text{ need to be positive and negative for charging/discharging the junction capacitors. Then, } S_2 \text{ is turned on. The voltages across the leakage inductors can be expressed as}
\]

\[
\begin{align*}
V_{L1} &= -V_{CC} \\
V_{L2} &= V_{in}
\end{align*}
\] (4)

\[\text{Stage 4 [t}_3, \ t_4] \text{ [see Fig. 3(d)]: At } t_3, \ Q_4 \text{ is turned off, and } i_3 \text{ is decreased towards to zero. Then, it begins increase. } i_2 \text{ and } i_3 \text{ change similarly. The voltages across the leakage inductors are expressed as}
\]

\[
\begin{align*}
V_{L1} &= -V_{CC} \\
V_{L2} &= V_{in}
\end{align*}
\] (5)

\[\text{FIGURE 3. Equivalent circuit in mode 1. (a) } [t_0, t_1]. \ (b) \ [t_1, t_2]. \ (c) \ [t_2, t_3]. \ (d) \ [t_3, t_4].\]
equal to the voltages in Stage 3, as expressed in (5). As the waveforms are symmetrical, the analysis of the other half period is similar with the previous stages.

**B. MODE 2 ANALYSIS**

In the proposed converter, the change rate of current $i_1$ and $i_2$ are opposite, and the current $i_3$ equals $(i_1 - i_2)/n$, which means the waveform shape of $i_1$ is the same as $i_3$ with only one proportional factor difference. Therefore, the subsequent analysis will only focus on the change law of $i_1$. Fig. 2(b) shows the key waveforms in mode 2. The equivalent circuits during each time interval are shown in Fig. 4. Before $t_5$, the process is same as the mode 1.

**Stage 1** [$t_5$, $t_6$] [see Fig. 4(a)]: At $t_5$, $Q_4$ is turned on. The current change rate of $i_1$ can be expressed as

$$\frac{di_1}{dt} = \frac{V_{in} - V_n}{L_1} \tag{6}$$

**Stage 2** [$t_6$, $t_7$] [see Fig. 4(b)]: At $t_6$, $S_1$ is turned off. During the dead time, if the junction capacitor of $S_2$ is completely discharged by $i_2$, then $S_2$ can be turned on with ZVS. When $S_2$ is turned on, the current change rate of $i_1$ can be expressed as

$$\frac{di_1}{dt} = -\frac{V_{Cc} - V_n}{L_1} \tag{7}$$

**Stage 3** [$t_7$, $t_8$] [see Fig. 4(c)]: At $t_7$, $i_3$ is decreased towards to zero. The current change rate of $i_1$ is same as Stage 2.

**Stage 4** [$t_8$, $t_9$] [see Fig. 4(d)]: At $t_8$, $Q_1$ is turned off. During the dead time, $i_3$ is negative, which can charge /discharge the junction capacitors of $Q_1$ and $Q_2$. The current change rate of $i_1$ can be expressed as

$$\frac{di_1}{dt} = -\frac{V_{Cc}}{L_1} \tag{8}$$

As the waveforms are symmetrical, the analysis of the other half period is similar with the previous stages.

According to the analysis in [5], a smaller $D_1$ can make the effective value and peak value of the current of the converter smaller, thereby reducing losses and improving the efficiency. Therefore, in practical, $D_1$ is designed from $-0.5$ to $0.5$. Thus, mode 3 will not be analyzed. According to above analysis, the normalized average output power of the proposed converter can be calculated as follows:

$$P^* = \frac{P}{P_{base}} = \begin{cases} 
-k[1 - (1 + 2D_1)^2 - (1 - D_2)^2], & \frac{1 - D_2}{2} \leq D_1 < -\frac{1 - D_2}{2} \\
4kD_1D_2, & \frac{1 - D_2}{2} \leq D_1 \leq \frac{1 - D_2}{2} \\
k[1 - (1 + 2D_1)^2 - (1 - D_2)^2], & \frac{1 - D_2}{2} < D_1 \leq \frac{1 + D_2}{2}
\end{cases} \tag{9}$$

where $P$ is the average output power, and $P_{base}$ is the base of $P$, which is expressed as $P_{base} = (V_{in})^2/4L_1f_s$. And the gain $k = V_o/nV_{in}$. Then, the power characteristic of the proposed converter can be drawn in Fig. 5.

It can be seen that the forward and the backward power transmission are symmetric with respect to the origin, in which the sign of $D_1$ determine the direction of power transmission. The power characteristic in the same direction is symmetrical about $|D_1| = 0.5$. The value of power is determined by $D_1$ and $D_2$. Under the same $D_1$, the larger the $D_2$ is, the greater the power is.

![FIGURE 4. Equivalent circuit in mode 2. (a) [t5, t6]. (b) [t6, t7]. (c) [t7, t8]. (d) [t8, t9].](image-url)
III. ZVS RANGE ANALYSIS

In order to reduce the loss of the push-pull dual active bridge converter and improve efficiency, it is necessary to realize ZVS within the wide load range under wide input voltage range. This section will analyze the conditions of ZVS operation in different modes.

A. ZVS ANALYSIS IN MODE 1

It can be seen from Fig. 3(a) that when $Q_1$ is turned off at $t_0$, the current $i_3$ should be greater than or equal to 0 to achieve ZVS operation of $Q_1$. And it can be seen from Fig. 3(b) that when $Q_1$ is turned off at $t_1$, the current $i_3$ should be less than or equal to 0 to ZVS operation of $Q_2$. Then, the analysis is similarly when $Q_4$ and $Q_2$ are turned off. According to the relationship between $i_1$, $i_2$ and $i_3$, the conditions of ZVS operation on each switching device are listed in the Table 1.

| Switches | Corresponding to the value range of $D_2$ with respect to $D_1$ |
|----------|---------------------------------------------------------------|
| $S_1$, $S_2$ | $D_2 \leq 1/k$ |
| $Q_1$, $Q_4$ | $D_2 \geq (2D_1)/(1-k)$, $D_2 < 0$ |
| $Q_2$, $Q_3$ | $D_2 \geq (2D_1)/(1-k)$, $D_2 \geq (2D_1)/(1-k)$ |

As seen from Table 1 that when $k < 1$, $Q_1$ and $Q_2$ cannot achieve ZVS. Therefore, $k$ should be designed as greater than or equal to 1 in order to achieve the ZVS operation of all power switching. And, when the $k > 1$, the range of $D_2$ when all switches achieve ZVS can be obtained as

$$\frac{2D_1}{k-1} \leq D_2 \leq \frac{1}{k}$$  

(10)

B. ZVS ANALYSIS IN MODE 2

Consistent with analysis of ZVS conditions in Mode 1, the conditions for each switch to achieve ZVS in Mode 2 can also be derived.

It can be seen from Fig. 4(a) that when $Q_3$ is turned off at $t_5$, the current $i_3$ should be greater than or equal to 0 to achieve ZVS operation of $Q_3$. And it can be seen from Fig. 4(b) that when $S_1$ is turned off at $t_6$, the current $i_3$ should be greater than or equal to 0 to ZVS operation of $Q_2$. Then, the analysis is similarly when $Q_4$ and $Q_2$ are turned off. The conditions of ZVS operation on each switching device are listed in the Table 2.

| Switches | Corresponding to the value range of $D_2$ with respect to $D_1$ |
|----------|---------------------------------------------------------------|
| $S_1$, $S_2$ | $D_2 \geq (k-1)/(2k)$ |
| $Q_1$, $Q_2$ | $D_2 \geq (1-D_1)/(1-k)$ |
| $Q_3$, $Q_4$ | $D_2 \leq (2D_1)/(1-k)$, $D_2 \geq 0$ |

It can be seen from Table 2 that when $k < 1$, the range of $D_2$ when all switches achieve ZVS can be obtained as

$$\frac{2(1-D_1)}{1+k} \leq D_2 \leq \frac{2D_1}{1-k}$$  

(12)

Therefore, according to the above analysis, in practical applications, $k$ should be reasonably designed constantly greater than or equal to 1 where the input and output voltages are variable, so as to obtain the wide range of ZVS operation. Then, the relationship of value of $D_2$ with respect to $D_1$ when ZVS operation is realized is shown in Fig. 6.

Substituting (11) and (12) into the normalized average output power expression (9), the power characteristics when all switches achieve the ZVS operation can be obtained as

$$\frac{8kD_1^2}{k-1} \leq P^* \leq 4D_1,$$  

Mode 1

$$P^* \geq k \left[ (1-2D_1)^2 - \frac{2(1-D_1)}{1+k} \right]$$,  

Mode 2

(13)

IV. MINIMUM RMS CURRENT CONTROL STRATEGY WITH PRECONDITION OF ACHIEVING WIDE ZVS RANGE

As above analyzed, the proposed converter can achieve ZVS operation of all switches under a wide input voltage range within wide load range. In order to further increase the efficiency, the control strategy can be modified to reduce the RMS current, thereby reducing the conduction loss.

When the transmitted power is the same, the combination of $D_1$ and $D_2$ are not unique, and the corresponding effective current values are not the same. The working waveforms of Mode 1 are shown in Fig. 2(a), the value of RMS current in half a switching period can be calculated as

$$I_{i(rms)} = \sqrt{3V_{in} \left[ 12kD_1^2D_2 - 2k^2D_1^3 + kD_1^3 + 3k^2D_2^2 - 3kD_1 + 1 \right]}$$  

(14)

In order to facilitate calculation and analysis, the normalization is also used to define the reference value of the RMS current, which is defined as

$$I_{sb} = \frac{P_b}{V_o} = \frac{V_{in}}{4kD_1L_{eqf}s}$$  

(15)
Therefore, divide the RMS current by (15) to get the normalized value as shown follow.

\[
I_{1\text{rms}*} = \frac{2\sqrt{3}}{3} k \sqrt{12kD_1^2D_2 - 2k^2D_2^3 + kD_2^2 + 3kD_2 - 3D_2^2 + 3D_2 - 3D_2^2 + 3D_2 - 3D_2^2 + 3D_2 - 3D_2^2 + 3D_2 - 3}
\]

(16)

To analyze the relationship between \( I_{1\text{rms}} \) and \( D_2 \) when the transmitted power is same, it is necessary to replace \( D_1 \) in (17), which can be obtained by (9).

\[
D_1 = \frac{P*}{4kD_2}
\]

(17)

Substituting (17) into (16) can derive the following.

\[
I_{1\text{rms}*} = \frac{2\sqrt{3}}{3} k \sqrt{ \frac{3(P*)^2}{4kD_2^2} - 2k^2D_2^3 + kD_2^2 + 3kD_2^2 - 3kD_2 + 3D_2^2 - 3}
\]

(18)

Then, the normalized value of the leakage RMS current in Mode 2 can also be calculated as

\[
I_{1\text{rms}}^2 = \frac{2\sqrt{3}}{3} k \sqrt{ \frac{2kD_1(-4D_2^2 + 6D_1 - 3D_2^2 + 6D_2 - 3)}{\sqrt{1 + kD_2(-2kD_2^2 + 3kD_2^2 - 3kD_2 + 3D_2^2 - 3kD_2 + 3D_2 - 3)}} + k + 1
\]

(19)

In order to obtain the minimum RMS current, take the partial derivative of (18) and (19) with respect to \( D_2 \) to find the \( D_2 \) corresponding to the minimum value. The derivation
results are as follows, where the first term is the partial derivative of (18).

\[
\frac{\partial I_{1\text{(rms)}}}{\partial D_2} = \left\{ \begin{array}{ll}
\frac{\sqrt{3} k \left(3k - 6k^2D_2 - 3kD_1^2 + 6k^2D_1^2 + 12kD_1^3 \right)}{3 \sqrt{kD_1^2 - 3kD_2 + 3k^2D_1^2 - 2k^2D_1^2 + 12kD_1^2 + 1}} \\
\sqrt{3} k^2 \left((1 - 2D_1) \left[2D_2^2 + [2k(1 - 2D_1) - 2]D_2 - (1 - 2D_1)^2 \right] \right)
\end{array} \right.
\]

\[= \left(1 - 2D_1 \right) \left\{ \frac{k^2(1 + 3k^2D_2^2 - 2k^2D_1^2)}{+3k^3(D_2^2 - 2D_2)(1 - D_1) + k^3(1 - D_1)^3} \right\}
\]

(20)

Let (20) equals zero to find the value of \(D_2\), as follows shown, where (21) and (22) are the results of \(D_2\) in Mode 1 and Mode 2, respectively.

\[D_2 = \frac{k - \sqrt{-8kD_1^2 + 4D_1^2 + (k - 1)^2}}{2k - 1}
\]

(21)

\[D_2 = \begin{cases} 1 - k(1 - 2D_1) + \sqrt{(1 - k(1 - 2D_1))^2 + (1 - 2D_1)^2} & \text{for } D_2 = 1, D_1 \geq 1 + \sqrt{k^2 - 1 - k} \\
1 & \text{for } D_2 = 1, D_1 \leq 1 + \sqrt{k^2 - 1 - k} 
\end{cases}
\]

(22)

Fig. 7 shows the relationship between \(D_2\) and \(D_1\) when the value of the RMS current is the smallest in the area of Mode 1 and Mode 2. It can be seen that the curves corresponding to the minimum RMS current are all within the range where all switched can achieve ZVS operation. Substituting (21) and (22) into (9), and draw the power characteristic of the proposed converter with optimized control strategy, as shown in Fig. 8.

In order to better understand the control method proposed in this paper, Fig. 9 shows the control block diagram. By sampling the input voltage \(V_{in}\) and the output voltage \(V_o\), the module of proposed control strategy calculate the value of \(D_2\) and \(D_4\). Then, according to Fig. 2, the PWM generator generates the drive signals by judging the operation mode.

V. EXPERIMENT RESULTS

A 1 kW push-pull dual active bridge converter prototype is built up with 28-50 V input and 200 V output voltage. The system circuit parameters are shown in Table 3. MOSFET (Manufacturer number: IPP110N20NA) is used for the pull-push side power devices. The control strategy is implemented in a digital signal processor TMS320F28069. The laboratory prototype is shown in Fig. 10.

A. STEADY STATE PERFORMANCE

The steady state performance of the proposed converter under full load in different input voltage is shown in Fig. 11, which includes secondary side bridge arm midpoint voltage \(V_{ab}\), leakage current \(i_1\), pull-push side bridge arm midpoint voltage \(V_{cd}\), respectively. Fig. 11(a) and (b) show the waveforms when the input voltages are 50V and 36V, respectively.

Fig. 12 shows the steady state waveforms when input voltage is 50V under full load and light load, which includes pull-push side bridge arm midpoint voltage \(V_{ab}\), secondary side bridge arm midpoint voltage \(V_{cd}\), leakage current \(i_1\) and \(i_2\), respectively. In order to see the relationship between the pull-push side currents, the current \(i_2\) is reversed. It can be

![Figure 13](image-url)
seen that $i_1$ is larger than $i_2$, and their difference can transmit the power.

**B. SOFT SWITCHING PERFORMANCE AND COMPARISON**

In order to verify ZVS performance at different input voltages under different loads, the soft switching performance of the proposed converter is shown in Fig. 13 to Fig. 15, which include the secondary side bridge arm midpoint voltage $V_{cd}$, leakage current $i_1$, drain to source voltage and drive voltage, respectively. It can be seen from the measured waveforms, the ZVS operation on all switches can be realized under the full load (1 kW) and light load (100 W), when the input voltages are 28V, 36V and 50V. Therefore, the proposed converter can achieve ZVS operation with wide load range within wide input voltage range.

To demonstrate the superiority of the proposed minimum RMS current control strategy in terms of ZVS operation, comparison experiments are carried out on the traditional method in [21], which use the same circuit parameters, shown in Table 3. Fig. 16 shows its soft switching performance at full...
load and light load condition when the input voltages are 36V and 50V. It can be seen that the ZVS operation is lost under light load and full load conditions.

C. DYNAMIC PERFORMANCE

In order to verify the voltage regulation capability, the load is step changed from 200 W to 1kW. The key waveforms are shown in Fig. 17. It can be seen that the output voltage remains stable at 200 V under the load step change when the input voltages are 50V and 36V, respectively.

D. PERFORMANCE UNDER BACKWARD POWER TRANSMISSION

Since the proposed push-pull DAB converter can realize bidirectional power transmission, its performances under backward power transmission are shown in Fig. 18 and Fig. 19. Fig. 18 shows the steady-state performance under full load when the input voltages are 50V and 36V, respectively. And, it can also be seen from Fig. 19 that a good dynamic
performance with a load stepped change under different input voltages is still maintained.

**E. EFFICIENCY COMPARISONS**

The comparisons of the measured efficiency under different power transmission direction in different input voltages between the proposed optimized control strategy and the traditional strategy in [21] are shown in Fig. 20. It can be seen that the peak efficiency of the proposed control strategy are 96.2% and 95.9% under forward and backward power transmission, respectively. And, the efficiency of the traditional strategy is lower, especially under light loads.

**VI. CONCLUSION**

In order to extend the ZVS operation range of the push-pull DAB converter in a wide input voltage range and reduce the current stress of the switches, this paper modified the output circuit to be full bridge to increase the control degree of freedom, and proposed an optimized control strategy, which utilizes the duty cycle of the secondary side. Through the analysis of the working modes, the realization conditions of ZVS operation in the wide input voltage range are calculated. Based on the preconditions, this paper further proposed the control law of the minimum leakage RMS current based on the calculation of the current stress to reduce the conduction loss. Finally, by comparing its performance with the traditional strategy, the experimental results verify the effectiveness of the proposed converter.

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