# DAB Converter Based on Unified High-Frequency Bipolar Buck-Boost Theory for Low Current Stress 

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#### Abstract

This paper proposes a unified high-frequency bipolar buck-boost (UHFBB) control strategy for a dual-active-bridge (DAB), which is derived from the classical buck and boost $\mathrm{DC} / \mathrm{DC}$ converter. It can achieve optimized current stress of the switches and soft switching in wider range. The UHFBB control strategy includes multi-control-variables, which can be achieved according to an algorithm derived from an accurate mathematical model. The design method for the parameters, such as the transformer turns ratio and the inductance, are shown. The current stress of the switches is analyzed for selecting an optimal inductor. The analysis is verified by the experimental results within a 500 W prototype.


Key words: BCM, Current stress, DAB, DCM, High-frequency bipolar buck-boost, Multi-parameters solving

## I. InTRODUCTION

Bidirectional DC/DC converters (BDCs) are widely used in aerospace applications [1], electrical vehicle chargers [2], distributed generation (DG) systems and AC or DC microgrids [3] due to their bidirectional power flow, low cost, small volume and low weight. Therefore, they have attracted a lot of attention.

BDCs can be divided into two class, non-isolated and isolated [4], [5]. The non-isolated BDC is hard to achieve a high conversion efficiency if a high step-up or step-down voltage ratio is necessary. Meanwhile, the isolated BDC can match the input and output voltage by using a high frequency transformer and it can easily obtain a high efficiency. Generally, BDCs include current-fed and voltage-fed converters [6], [7]. In the current-fed BDC, the spike voltage is usually across the switches and a clamped circuit should be included

[^0]in the converter, which increases the complexity and cost of the converter. The switches in the voltage-fed BDC are all clamped by filter capacitors. The advantages of the voltage-fed BDC have led to its wide adoption. The voltage-fed BDC can be achieved by reorganizing the classical DC/DC converter. The reorganized BDCs are symmetrical about the transformer, and include the flyback BDC [8], the flyback-forward hybrid BDC [9], the half-bridge BDC [10], the full-bridge BDC [11], etc. In large or medium power applications, the dual-activebridge ( DAB ) is usually adopted due to its zero-voltage switching (ZVS) and its bidirectional power flow by a simple phase-shift control strategy [7]. In the DAB, the phase and pulse width of the AC square-wave voltages on both sides of the transformer can all be adjusted to achieve optimal efficiency [12], [13]. In recent years, many research results have been achieved in an effort to decrease the circulating energy, the current stress of the switches and the total loss [14].
The authors of [15] proposed a definition of extended phase shift to decrease the circulating energy and the switches conduction loss, which achieved a high efficiency. However, there are three independent control variables, which are the primary-side full-bridge duty ratio D 1 , the secondary-side fullbridge duty ratio D2 and their phase shift angle $\phi$. The most optimized of the three variables method is not proposed in the paper and the circulating energy is not at the minimum value
because D1 is a fixed value in the control strategy. A double phase-shift control strategy is proposed in an effort to achieve the minimum peak current [16]. However, D1 = D2 is designed for convenient control. Therefore, this method can be further optimized. According to the output power, a segment variablefrequency strategy is proposed to guarantee the optimized efficiency [17]. On the one hand, the realization of the segment variable-frequency is difficult to achieve. On the other hand, the adopted triangle current control results in more current stress. The full power loss model is built and the three variables mentioned above, D1, D2 and $\phi$, are determined according to the principle of the minimum power loss [18]. However, the operation condition depends on the accuracy of the loss model and the modeling process is complicated. Moreover, ZVS in the DAB is somewhat determined by the workload [12].

The conventional control strategy for a DAB comprises of a voltage outer-loop and a current inner-loop [19], [20]. The feedback variable of the current inner-loop is the inductor current, which is generally filtered to obtain a smooth variable. The added low-pass filter affects the dynamic performance of the DAB. Hence, predictive current control is widely investigated [21]. However, the control effect of the predictive current control is determined by the accuracy of the measurement.

This paper proposes a method based on the unified highfrequency bipolar buck-boost theory for the DAB to overcome the shortcomings mentioned in [22]. The inductor current is designed in the discontinuous conduction mode (DCM) or the boundary conduction mode (BCM) in half of a switching cycle and the switching frequency is fixed, which guarantees that the switches realize ZVS or ZCS. Moreover, an algorithm for solving multi-variables is proposed for optimizing the current stress. The current sensor is removed and the dynamic performance is enhanced. Experimental results verify the high performance of the proposed method.

## II. Motivation for Introducing the Unified Bipolar Buck-Boost Theory into a DAB

Fig. 1 shows the topology of a DAB which comprises of a low-voltage-side (LVS) full-bridge formed by S1-S4, a high-voltage-side (HVS) full-bridge formed by S5-S8, an inductor $L$, and a high-frequency transformer T with a turns ratio of 1:n. $C_{\mathrm{r} 1}-C_{\mathrm{r} 8}$ and D1-D8 are the parasitic capacitors and body diodes of the switches S1-S8, respectively. $U_{\mathrm{in}}$ and $U_{\mathrm{o}}$ are the input and output voltages, respectively. $C_{1}$ and $C_{2}$ are the filter capacitors in the LVS and HVS. respectively. $i_{\mathrm{L}}$ is the current through the buffering inductor $L$. $i_{\mathrm{S}}$ and $u_{\mathrm{S}}$ are the secondaryside current and voltage of the transformer, respectively. $u_{\mathrm{L} 1}$ and $u_{\mathrm{L} 2}$ are the input-terminal and output-terminal voltages of the buffering inductor, respectively. $I_{0}$ is the output current of the DAB , and $u_{\mathrm{L} 2}$ is input voltage of the transformer T .

In the existing control strategy for the DAB , some shortcomings, such as switching surges, circulating current


Fig. 1. Topology of a DAB.


Fig. 2. Operational waveforms of a DAB controlled by the dual-phase-shift control strategy.
and high current stress of the switches, overly complicated control strategy, etc., decrease the efficiency and block the widespread use of the DAB. These shortcomings are expected to be overcome after using the method proposed in this paper.

## A. Hard Switching

Under most conditions, the switches can achieve ZVS in the existing control strategy. However, there are some special conditions. The reason for the high diode reverse recovery loss under a light load with the conventional single phase-shift control strategy is analyzed in [23]. Moreover, there is also switching loss in some of the improved control strategies [15], [16], [23]. Fig. 2 shows an example of a DAB controlled by the dual-phase-shift strategy [16].

There are 10 modes in a switching cycle in Fig. 2. There is no surge voltage across the switches in the LVS because $i_{\mathrm{L}}$ lags behind $u_{\mathrm{L} 1}$. However, a surge voltage occurs repeatedly in the HVS switches in the transition from mode 1 (6) to mode 2 (7). The diode D7 and the switch S5 are conducting in state 1. Then the switch S7 is turned off and S8 is turned on when the state changes from mode 1 to mode 2 . At this time, the diode D7 is immediately switched from a forward bias condition to a reverse bias condition, and the switch S7 is turned off with a large reverse recovery loss. Therefore, there is high switching loss condition as shown in Fig. 2. The cyan and black boxes shown in Fig. 2 show that there is circulating energy to the fullbridge in the LVS and HVS in the corresponding intervals because of the opposite polarities of the voltage and current.

## B. Circulating Current and High Current Stress

Many methods have been proposed to decrease or eliminate


Fig. 3. Waveforms of the control strategy in [14].
circulating current [14], [15]. Moreover, a lot of studies have tried to achieve the minimum RMS value or peak value of $i_{\mathrm{L}}$. These two objections should be concurrent, which means that the RMS value of $i_{\mathrm{L}}$ is decreased if the circulating current is well restrained [16], [24]. However, the existing control strategies cannot a good match for these two objectives. For example, main waveforms of the control strategy in [14] are shown in Fig. 3. It completely eliminates the circulating current. However, the RMS value of $i_{\mathrm{L}}$ is still high. The duty ratios of the two bridges in both sides of the transformer are designed too short. Then the zero-time intervals of $u_{\mathrm{L} 1}$ and $u_{\mathrm{L} 2}$ are very long. There is current flowing though the switches in these time intervals. In half a switching cycle, the current in modes b -d cannot produce active power to the bridge in the LVS and the current in modes $\mathrm{a}, \mathrm{b}$ and d cannot produce active power to the bridge in the HVS.

## C. Complexity of Control Strategy

In order to achieve a high efficiency, some control strategies adopt composite modulation methods according to the power boundary [24]-[27]. Although the composite modulation methods can obtain an optimized switching current stress, the calculation process is very complicated and it requires a highperformance digital chip to perform the calculations. On the other hand, the power boundary may have no closed-form solution and a numerical solver has to be used to determine the boundary power [24].

Therefore, the objective of this paper is to propose a simple modulation so the DAB can achieve soft switching, no circulating current, low current stress and high efficiency.

## III. Principle of High-Frequency Bipolar Buck-Boost Theory

There is a large circulating current when using the conventional phase-shift strategy. If the stored energy in the buffering inductor $L$ is entirely released in a positive or negative half switching cycle, that is the current $i_{\mathrm{L}}=0$ at the beginning of a positive or negative half switching cycle, the problem of circulating current can be solved.

A reasonable turns ratio of the transformer ' $n$ ' can match the unbalance between the input and output voltages. However, the fluctuation of $U_{\text {in }}$ and $U_{\mathrm{o}}$ is very large and the amplitude

TABLE I
Switch State of a DC Converter

|  | Buck |  |  | Boost |  |  |
| :--- | :--- | :--- | :--- | :--- | :--- | :--- |
| A | 1 | 0 | 0 | 1 | 1 | 0 |
| B | 1 | 1 | 0 | 0 | 1 | 0 |


(a)

(b)

Fig. 4. Modulation strategy for a high-frequency bipolar buckboost voltage. (a) High-frequency bipolar buck operation waveform ( $U_{\mathrm{in}}>U_{0} / \mathrm{n}$ ). (b) High-frequency bipolar boost operation waveform ( $U_{\text {in }}<U_{\mathrm{o}} / \mathrm{n}$ ).
relationship between $u_{\mathrm{L} 1}$ and $u_{\mathrm{L} 2}$ is not fixed. Therefore, the operation principle of a conventional buck and boost DC/DC converter can be introduced into DABs when they are operated in the DCM or the BCM . Assuming the input and output voltages of a buck and boost DC/DC converter are also $U_{\text {in }}$ and $U_{0}$, the voltage across the inductor can be expressed as:

$$
\begin{equation*}
u_{\mathrm{L}}=\mathrm{A} U_{\mathrm{in}}-\mathrm{B} U_{\mathrm{o}} \tag{1}
\end{equation*}
$$

Where A and B are the states on the two sides of the inductor in a buck or boost converter, and their values are shown in Table I. There are three modes in a switching cycle. The three modes are the current-increasing stage, the current-decreasing stage and the zero-current stage. If this operation principle is expanded to a high-frequency bipolar topology, the waveforms of $u_{\mathrm{L} 1}, u_{\mathrm{L} 2}$ and $i_{\mathrm{L}}$ are shown in Fig. 4, which form the highfrequency bipolar buck or boost operation principle. Where, $D_{\mathrm{bu}}$ and $D_{\mathrm{bo}}$ are the bipolar buck and boost duty ratios, respectively.

Setting a voltage as threshold value, a proper operation mode can be selected according to the value of $U_{\text {in }}$ and $U_{\mathrm{o}}$, which can guarantee ZVS or ZCS of all the switches. However, this method has a number of obvious shortcomings.

1) When compared with the continuous conduction mode, the peak value of $i_{\mathrm{L}}$ is larger in this method, which results in a larger current stress and a lower efficiency.
2) The converter may be unstable at the time of the transition from the buck mode to the boost mode, and vice versa. Moreover, the converter cannot work when $U_{\text {in }}=U_{\mathrm{o}} / \mathrm{n}$.

In order to retain the characteristic of ZVS and ZCS and to

TABLE II
Unified High-Frequency Bipolar Buck/Boost Control Strategy

| Duty ratio | Positive half cycle |  |  |  | Negative half cycle |  |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
|  | $\mathrm{d}_{1}$ | $\mathrm{d}_{2}$ | $\mathrm{d}_{3}$ | $\mathrm{d}_{4}$ | $\mathrm{d}_{1}$ | $\mathrm{d}_{2}$ | $\mathrm{d}_{3}$ | ${ }_{4}$ |
| A | 1 | 1 | 0 | 0 | -1 | -1 | 0 | 0 |
| B | 0 | 1 | 1 | 0 | 0 | -1 | -1 | 0 |

decrease the current stress of the switches, $i_{\mathrm{L}}$ should be operated in the DCM or the BCM in every half switching cycle. Therefore, a unified high-frequency bipolar buck-boost control strategy can also be explored. Setting the voltage across the inductor is still satisfied with (1). This paper proposed the method shown in Table II.

A switching cycle is divided into a positive and negative half cycle. Each half cycle has four modes and their duty ratios are $d_{1}, d_{2}, d_{3}$ and $d_{4}$, respectively. Their relationships are shown in Fig. 5.

$$
\left\{\begin{array}{l}
d_{1}=\frac{\text { duration time of }(\mathrm{A}=1 \& \mathrm{~B}=0)}{0.5 T_{S}}  \tag{2}\\
d_{2}=\frac{\text { duration time of }(\mathrm{A}=1 \& \mathrm{~B}=1)}{0.5 T_{S}} \\
d_{3}=\frac{\text { duration time of }(\mathrm{A}=0 \& \mathrm{~B}=1)}{0.5 T_{S}} \\
d_{4}=\frac{\text { duration time of }(\mathrm{A}=0 \& \mathrm{~B}=0)}{0.5 T_{S}}
\end{array}\right. \text { positive half cycle }
$$

In a positive half cycle, the operation principle of the former two modes whose duty ratios are $\mathrm{d}_{1}$ and $\mathrm{d}_{2}$ is similar to a boost $\mathrm{DC} / \mathrm{DC}$ converter, and the operation principle of the medium two modes whose duty ratios are $\mathrm{d}_{2}$ and $\mathrm{d}_{3}$ is similar to a buck $\mathrm{DC} / \mathrm{DC}$ converter. That is, the proposed control strategy in Table II is essentially a hybrid control combined with the buck and boost operation principle. The function of increasing or decreasing the voltage can be easily realized if the multi-dutyratios are adjusted. Moreover, the control variables vary continuous. According to the input and output voltage values, not all of the four modes are included in a half switching cycle.

When $\mathrm{n} U_{\text {in }}<U_{\mathrm{o}}$, the converter can be viewed as operating in the boost mode, and Fig. 5 shows the three different conditions. The inductor current $i_{\mathrm{L}}$ is running in the DCM in a half switching cycle and there is no the third mode under a light load, that is $d_{3}=0$, as shown in Fig. 5(a). The converter operating mode is shift from the DCM to the boundary-BCM ( $\mathrm{B}-\mathrm{BCM}$ ) when the output power is equal to the boundary power $\mathrm{P}_{\mathrm{B}}$ boost. In the $\mathrm{B}-\mathrm{BCM}, \mathrm{d}_{3}$ and $\mathrm{d}_{4}$ are always equal to zero. The corresponding waveforms are shown in Fig. 5(b). With the power further increasing, the converter is operated in the $B C M$ and $d_{3}>0$. The waveforms shown in Fig. 5(c) show that part buck operating mode joins in the modulation strategy to guarantee the BCM .

When $\mathrm{n} U_{\text {in }}>U_{\mathrm{o}}$, the converter can be viewed as operating in the buck mode. The corresponding waveforms are shown in


Fig. 5. Modulation strategy for the unified high-frequency bipolar boost principle. (a) DCM. (b) B-BCM. (c) BCM.

(a)

(b)

(c)

Fig. 6. Modulation strategy for the unified high-frequency bipolar buck principle. (a) DCM. (b) B-BCM. (c) BCM.

Fig. 6 and its boundary power is $\mathrm{P}_{\mathrm{B}_{-} \text {buck. The two }}$ boundary powers $\mathrm{P}_{\mathrm{B}_{-} \text {boost }}$ and $\mathrm{P}_{\mathrm{B}_{-} \text {buck }}$ can be obtained from the relationship in Fig. 5(b) and Fig. 6(b).

$$
\left\{\begin{array}{l}
P_{B_{-} \text {boost }}=\frac{\left(U_{o}-n U_{i n}\right) U_{i n}^{2} T_{S}}{4 L U_{o}}  \tag{3}\\
P_{B_{-} \text {buck }}=\frac{\left(n U_{i n}-U_{o}\right) U_{o}^{2} T_{S}}{4 n^{3} L U_{\text {in }}}
\end{array}\right.
$$

The characteristics of the negative power flow are symmetrical with those of the positive power flow. In the following section, only the condition of the positive power flow is discussed.
From Fig. 5 and Fig. 6, it can be seen that all of the modes obey the rule in Table II. However, not all of $d_{1}-d_{4}$ are more than 0 . Thus, a simple and unified algorithm can be used to solve these four duty-ratios. Moreover, there is no circulating current in the converter, and all of the switches can achieve ZVS or ZCS in all occasions. Hence, the object presented in Section II can be satisfied.

## IV. Algorithm to Solve Multiple VARIABLES

In Fig. 5(c) and Fig. 6(c), $y_{1}$ and $y_{2}$ are the current values of $i_{\mathrm{L}}$ at the end of the first and second modes, respectively. Their expressions are independent of the size of $n U_{\mathrm{in}}$ and $U_{0}$. The operation conditions of the DCM and the $\mathrm{B}-\mathrm{BCM}$ are also included in the operation condition of the BCM. For example, if $d_{1}=0$ and $d_{2}+d_{3}<1$, it is operated in the bipolar buck DCM. The other conditions can be summed up in Table III.

It should be noted that $i_{L}$ is operated in the BCM or the DCM, which determines that $i_{\mathrm{L}}$ is equal to 0 at the beginning of the positive or negative half switching cycle. Therefore:

$$
\begin{gather*}
\mathrm{y}_{1}=\frac{U_{i n} d_{1} T_{S}}{2 L}  \tag{4}\\
\mathrm{y}_{2}=\frac{n U_{i n} T_{S} d_{1}+\left(n U_{i n}-U_{o}\right) T_{S} d_{2}}{2 n L} \tag{5}
\end{gather*}
$$

The third duty ratio, $\mathrm{d}_{3}$, can be achieved according to equilibrium of the current increasing and decreasing.

$$
\begin{equation*}
d_{3}=\frac{n U_{i n} d_{1}+\left(n U_{i n}-U_{o}\right) d_{2}}{U_{o}} \tag{6}
\end{equation*}
$$

The part of $i_{\mathrm{L}}$ in modes 2 and 3 can be delivered to the filter capacitor in the HVS, while $i_{\mathrm{L}}$ in mode 1 cannot be delivered to the filter capacitor in the HVS. Hence, the following relationship can be achieved based on the fact that the output current $I_{\mathrm{o}}$ in the primary side is equal to the mean value of $i_{\mathrm{L}}$ in modes 2 and 3 .

$$
\begin{align*}
& d_{2}^{2}\left(n U_{i n}-U_{o}\right)+d_{2}\left(2 n d_{1} U_{i n}+n d_{3} U_{i n}-d_{3} U_{o}\right)  \tag{7}\\
& +n d_{1} d_{3} U_{i n}-4 n^{2} L I_{o} / T_{S}=0
\end{align*}
$$

The variable $\mathrm{d}_{3}$ in (7) can be replaced by (6). Then:

$$
\begin{align*}
& d_{2}^{2}\left(n U_{i n}\left(n U_{i n}-U_{o}\right)\right)+d_{2}\left(2 n^{2} U_{i n}^{2} d_{1}\right)  \tag{8}\\
& +n^{2} U_{i n}^{2} d_{1}^{2}-4 n^{2} L U_{o} I_{o} / T_{S}=0
\end{align*}
$$

Solving the solution of (8), it is possible to achieve:

$$
\begin{gather*}
d_{2}=\frac{x_{2} d_{1}+\sqrt{x_{3} d_{1}^{2}+x_{4}}}{x_{1}}  \tag{9}\\
d_{3}=x_{5} d_{1}+x_{6} d_{2} \tag{10}
\end{gather*}
$$

TABLE III
Relationship Between the Operation Condition and the Duty Ratios

|  | DCM | B-BCM | BCM |
| :---: | :---: | :---: | :---: |
| bipolar | $\mathrm{d}_{3}=0 ; \mathrm{d}_{1}+\mathrm{d}_{2}<1 ;$ | $\mathrm{d}_{1}+\mathrm{d}_{2}=1 ;$ | $\mathrm{d}_{1}+\mathrm{d}_{2}+\mathrm{d}_{3}=1 ;$ |
| boost | $\mathrm{d}_{4}=1-\mathrm{d}_{1}-\mathrm{d}_{2}$ | $\mathrm{~d}_{3}=0 ; \mathrm{d}_{4}=0$ | $\mathrm{~d}_{4}=0$ |
| bipolar | $\mathrm{d}_{1}=0 ; \mathrm{d}_{2}+\mathrm{d}_{3}<1 ;$ | $\mathrm{d}_{2}+\mathrm{d}_{3}=1 ;$ | $\mathrm{d}_{1}+\mathrm{d}_{2}+\mathrm{d}_{3}=1 ;$ |
| buck | $\mathrm{d}_{4}=1-\mathrm{d}_{2}-\mathrm{d}_{3}$ | $\mathrm{~d}_{1}=0 ; \mathrm{d}_{4}=0$ | $\mathrm{~d}_{4}=0$ |

where:

$$
\begin{gathered}
x_{1}=U_{i n}\left(n U_{i n}-U_{o}\right), \quad x_{2}=-n U_{i n}^{2}, \quad x_{3}=n U_{i n}^{3} U_{o}, \\
x_{4}=\frac{4 n U_{i n} U_{o} L I_{o}\left(n U_{i n}-U_{o}\right)}{T_{S}}, \quad x_{5}=\frac{n U_{i n}}{U_{o}}, \quad x_{6}=\frac{\left(n U_{i n}-U_{o}\right)}{U_{o}}
\end{gathered}
$$

According to Fig. 5 and Fig. 6, $i_{\mathrm{L}}$ can be divided into three parts in a half switching cycle, and only the parts in modes 2 and 3 can be delivered to the output filter capacitor. Hence, if $\left(\mathrm{d}_{2}+\mathrm{d}_{3}\right)$ is bigger, i.e. the waveform of $i_{\mathrm{L}}$ is smooth, the current stress of the switches is smaller under the same output power. Setting $y=d_{2}+d_{3}$, the task of the algorithm is to deduce the proper value of $d_{1}, d_{2}$ and $d_{3}$ when $y$ is at its maximum value. Substituting (9) and (10) into the expression of y yields:

$$
\begin{equation*}
y=x_{5} d_{1}+\left(1+x_{6}\right) \frac{x_{2} d_{1}+\sqrt{x_{3} d_{1}^{2}+x_{4}}}{x_{1}} \tag{11}
\end{equation*}
$$

Taking the derivative with respect to $\mathrm{d}_{1}$ yields:
$\frac{d y}{d\left(d_{1}\right)}=x_{5}+\frac{x_{2}\left(1+x_{6}\right)}{x_{1}}+\frac{x_{3}\left(1+x_{6}\right) d_{1}}{x_{1} \sqrt{x_{3} d_{1}^{2}+x_{4}}}=\frac{x_{2}}{x_{1}}+\frac{x_{3}\left(1+x_{6}\right) d_{1}}{x_{1} \sqrt{x_{3} d_{1}^{2}+x_{4}}}$
Setting (12) equal to 0 , the value of $d_{1}$, which is denoted by $\mathrm{d}_{1 \mathrm{y}}$, can be achieved when y is at its maximum value.

$$
\begin{equation*}
d_{1 \mathrm{y}}=\sqrt{\frac{x_{4} x_{2}^{2}}{\left(x_{3}+x_{3} x_{6}\right)^{2}-x_{3} x_{2}^{2}}} \tag{13}
\end{equation*}
$$

Solving the parameters $\mathrm{d}_{1}, \mathrm{~d}_{2}$ and $\mathrm{d}_{3}$ has three possibilities depending on the size of $\mathrm{n} U_{\text {in }}$ and $U_{\mathrm{o}}$ and the sum of $\mathrm{d}_{1}, \mathrm{~d}_{2}$ and $\mathrm{d}_{3}$.

## A. $n U_{\text {in }} \geq U_{o}$

Under this condition, $\mathrm{x}_{1}$ and $\mathrm{x}_{3}-\mathrm{x}_{6}$ are all more than zero, while $\mathrm{x}_{2}<0$. Therefore, the quantity under the square root sign in (9) is automatically more than zero. If $d_{2}$ is guaranteed to be a positive value, it must satisfy (14).

$$
\begin{equation*}
d_{1 \times 1}=0 \leq d_{1} \leq \sqrt{\frac{x_{4}}{\left(x_{2}^{2}-x_{3}\right)}}=d_{1 \mathrm{z} 1} \tag{14}
\end{equation*}
$$

After substituting the values of $\mathrm{d}_{1 \times 1}, \mathrm{~d}_{1 \mathrm{y}}$ and $\mathrm{d}_{1 \mathrm{z} 1}$ into (11), the value of $d_{1}$ can be selected as one of $d_{1 \times 1}, d_{1 y}$ or $d_{1 z 1}$, which guarantees that the value of $y$ is at its maximum.
B. $n U_{\text {in }}<U_{o}$

Under this condition, $\mathrm{x}_{1}, \mathrm{x}_{2}, \mathrm{x}_{4}$ and $\mathrm{x}_{6}$ are all less than zero, while $x_{3}$ and $x_{5}$ are more than zero. The quantity under the square root sign in (9) and the value of $\mathrm{d}_{2}$ must be all more than
zero. Thus, it must satisfy the following expression.

$$
\begin{equation*}
\mathrm{d}_{1 \times 2}=\sqrt{\frac{-x_{4}}{x_{3}}} \leq d_{1} \leq \sqrt{\frac{-x_{4}}{\left(x_{3}-x_{2}^{2}\right)}}=\mathrm{d}_{122} \tag{15}
\end{equation*}
$$

After substituting the values of $\mathrm{d}_{1 \times 2}, \mathrm{~d}_{1 \mathrm{y}}$ and $\mathrm{d}_{1 z 2}$ into (11), the value of $d_{1}$ can be selected as one of $d_{1 \times 2}, d_{1 y}$ or $d_{172}$, which guarantees that y is at its maximum value.

The values of $d_{2}$ and $d_{3}$ can be easily obtained from (9) and (10).

Another possible condition is $\mathrm{d}_{1}+\mathrm{d}_{2}+\mathrm{d}_{3}>1$ which states that a variable-frequency is needed in the modulation. To realize constant -frequency control, the value of $\mathrm{d}_{1}+\mathrm{d}_{2}+\mathrm{d}_{3}$ must be equal to 1 , which is viewed as one of the known conditions.

## C. $d_{1}+d_{2}+d_{3}=1$

First, $\mathrm{d}_{3}$ can be determined according to this known condition.

$$
\begin{equation*}
d_{3}=1-d_{1}-d_{2} \tag{16}
\end{equation*}
$$

Under this condition, (8) is still correct and the $\mathrm{d}_{3}$ in (16) can be substituted by (8). Then:

$$
\begin{equation*}
d_{2}=\frac{U_{o}-\left(n U_{i n}+U_{o}\right) d_{1}}{n U_{i n}} \tag{17}
\end{equation*}
$$

If the $\mathrm{d}_{2}$ in (17) is guaranteed to be a positive value:

$$
\begin{equation*}
d_{1} \leq \frac{U_{o}}{n U_{\text {in }}+U_{o}} \tag{18}
\end{equation*}
$$

Eq. (8) is still correct and the $\mathrm{d}_{2}$ in (17) can be substituted by (8). Then:

$$
\begin{align*}
& \left(n^{2} U_{i n}^{2}+n U_{i n}\left|u_{G}\right|+U_{o}^{2}\right) d_{1}^{2}- \\
& 2 U_{o}^{2} d_{1}+U_{o}^{2}-n U_{i n} U_{o}^{2}+\frac{4 n^{3} U_{i n} L i^{*}}{T_{S}}=0 \tag{19}
\end{align*}
$$

The value of $\mathrm{d}_{1}$ can be determined according to (18) and (19).

$$
\begin{equation*}
d_{1}=\frac{-b-\sqrt{b^{2}-4 a c}}{2 a} \tag{20}
\end{equation*}
$$

where:

$$
\begin{aligned}
& a=n^{2} U_{i n}^{2}+n U_{i n} U_{o}+U_{o}^{2} \\
& b=-2 U_{o}^{2} \quad c=U_{o}^{2}-n U_{i n} U_{o}+\frac{4 n^{3} U_{i n} L I_{o}}{T_{S}}
\end{aligned}
$$

The values of $\mathrm{d}_{2}$ and $\mathrm{d}_{3}$ can be achieved according to (16) and (17). The whole process for the calculation of $d_{1}, d_{2}$ and $d_{3}$ described above is shown in Fig. 7. It is noted that the loadcurrent $I_{0}$ is substituted by the reference current $I_{0}{ }^{*}$. The process in the flowchart shown in Fig. 7 is in accordance with the calculation process (4)-(20), which can be easily realized by a digital signal processor (DSP).

The switch driven signals can be obtained by a proper modulation strategy after achieving the values of $\mathrm{d}_{1}, \mathrm{~d}_{2}$ and $\mathrm{d}_{3}$. The control strategy for a DAB is shown in Fig. 8. It can be seen that the current sensor can be removed and that there is no feedback variable in the current inner-loop when compared


Fig. 7. Flowing chart for solving multi-variables.


Fig. 8. Bipolar buck/boost control strategy for a DAB.
with the conventional method, which can greatly improve the dynamic performance.

Fig. 9 shows that the duty ratio curves vary with the output power and different input voltage when $U_{0}=380 \mathrm{~V}, L=6 \mu \mathrm{H}$ and $\mathrm{n}=380 / 49$. The multi-duty-ratios have the following characteristics.

1. $\mathrm{d}_{1}$ is always more than $\mathrm{d}_{3}$ when $\mathrm{n} U_{\text {in }}<U_{\mathrm{o}}$ because the boost voltage is dominant in this time, and vice versa. In addition, $\mathrm{d}_{1}$ is always less than $\mathrm{d}_{3}$ when $\mathrm{n} U_{\text {in }}>U_{0}$ because the buck voltage is dominant in this time. When $\mathrm{n} U_{\mathrm{in}}=U_{\mathrm{o}}$, $d_{1}$ is always equal to $d_{3}$.
2. The sum of $d_{1}, d_{2}$ and $d_{3}$ becomes gradually larger with the output power increasing, and the DAB is operated in the BCM when $\mathrm{d}_{1}+\mathrm{d}_{2}+\mathrm{d}_{3}=1$.
3. The DAB is operated in the BCM when $U_{\mathrm{in}}=U_{0} / \mathrm{n}$ and the boundary power in this time is zero. The boundary power gradually becomes larger when $U_{\text {in }}$ is far from $U_{\mathrm{o}} / \mathrm{n}$. The lager the difference between $U_{\mathrm{in}}$ and $U_{\mathrm{o}} / \mathrm{n}$ is, the smaller the operation range of the BCM becomes. The boundary power curve is shown in Fig. 10 according to (2) and (3).
4. Not all of the conditions have four modes. For example, the first mode does not exist when $d_{1}=0$ and the fourth mode does not exist when $d_{1}+d_{2}+d_{3}=1$.


Fig. 9. Duty ratio curves varying with the output power. (a) $U_{\text {in }}=42 \mathrm{~V}$. (b) $U_{\text {in }}=45.5 \mathrm{~V}$. (c) $U_{\mathrm{in}}=49 \mathrm{~V}$. (d) $U_{\mathrm{in}}=52.5 \mathrm{~V}$. (e) $U_{\mathrm{in}}=56 \mathrm{~V}$.


Fig. 10. Boundary power curve varying with the output power.

## V. Parameters Design

Two parameters should be deliberatively designed in the main circuit of the DAB shown in Fig. 1. One is the transformer turns ratio n and the other is the inductance $L$. The current stress of the switch is at its minimum when the input voltage $U_{\text {in }}$ is equal to the output voltage in the primary-side of the transformer $\left(U_{0} / \mathrm{n}\right)$. Considering the fluctuation range of $U_{\mathrm{in}}$, n is set to $U_{\mathrm{o}} / U_{\text {in_mid }}$, where $U_{\text {in_mid }}$ is the mid-point value of the fluctuation range of $U_{\text {in }}$.

A typical application of a DAB is the charger and discharger for an accumulator in a DC micro-grid. The typical parameters are: $U_{\text {in }} 42-56 \mathrm{~V}, U_{\mathrm{o}} 380 \mathrm{~V}$, switching frequency 40 kHz , and rated power 500 W .
$U_{\text {in_mid }}$, the mid-point value of the fluctuation range of $U_{\text {in }}$, is equal to 49 V . Thus, the transformer turns ratio n can be determined by $U_{0} / U_{\text {in_mid }}$ and its value is $7.755(380 / 49)$. There are two methods to design the inductance $L$ in [12]. One
method is that the value of $L$ should be as small as possible to decrease the reactive power. The other method aims at expanding the range of ZVS, and the value of $L$ should be as large as possible under the condition of guaranteeing the rated output power. These two methods have their advantages and disadvantages. In this paper, the value of $L$ is designed for the minimum RMS value of $i_{\mathrm{L}}$, which is proportional to the switches RMS current.
The subsection function of $i_{\mathrm{L}}$ in Fig. 5 and Fig. 6 can be achieved according to $\mathrm{y}_{1}$ and $\mathrm{y}_{2}$ in (4) and (5) in a half switching cycle. Thus, the RMS value of $i_{\mathrm{L}}$ can be obtained.

$$
\begin{equation*}
I_{\mathrm{L}}=\sqrt{\frac{\mathrm{y}_{1}^{2}\left(d_{1}+d_{2}\right)+\mathrm{y}_{2}^{2}\left(d_{2}+d_{3}\right)+\mathrm{y}_{1} \mathrm{y}_{2} d_{2}}{3}} \tag{21}
\end{equation*}
$$

The calculation of $I_{L}$ needs to know the values of $d_{1}, d_{2}$ and $d_{3}$. However, they cannot be expressed by a continuous function. A simulation model for calculating $d_{1}, d_{2}$ and $d_{3}$ is built using MATLAB/Simulink. The achieved duty ratios are used to calculate $I_{\mathrm{L}}$. The curves of $I_{\mathrm{L}}$, varying with the input voltage, are shown in Fig. 11 with different output powers and different values of the inductance $L$. It can be seen that $I_{\mathrm{L}}$ is small with a bigger inductance (such as $10 \mu \mathrm{H}$ in the figure) under a light load. However, $I_{\mathrm{L}}$ increases quickly with a larger inductance when compared with a smaller inductance when the output power increases. The value of $I_{\mathrm{L}}$ is big and becomes violent with variations of the input voltage $U_{\text {in }}$ under a smaller inductance (such as $2.5 \mu \mathrm{H}$ in the figure). Therefore, the principle for selecting the inductance is to avoid the two conditions mentioned above. The mediate values, such as $5 \mu \mathrm{H}$


Fig. 11. IL curves varying with $U_{\text {in }}$ using different inductors. (a) $P=100 \mathrm{~W}$. (b) $P=200 \mathrm{~W}$. (c) $P=300 \mathrm{~W}$. (d) $P=400 \mathrm{~W}$. (e) $P=500 \mathrm{~W}$.
and $7.5 \mu \mathrm{H}$, are satisfied with the principle. On the one hand, $I_{\mathrm{L}}$ changes smoothly with a varying of the input voltage. On the other hand, $I_{\mathrm{L}}$ is at its optimized state from a light load to a full load. Considering the overall condition mentioned above, $L=6 \mu \mathrm{H}$ is selected.

## VI. Current Stress Comparison with the Closed form Method

In [24], a closed form solution strategy has been proposed to achieve the optimized RMS value of $i_{\mathrm{L}}$ and the minimum conduction loss. Thus, the RMS value of $i_{\mathrm{L}}$ controlled by the proposed method in this paper is compared with the closed form solution strategy for determining the advantages and disadvantages of the proposed modulation strategy in this paper.

According to the strategy in [24], it is possible to obtain the transformer turns ratio $n_{1}=7.755$ and the optimized inductance $L_{1}=12.86 \mu \mathrm{H}$ with the same input voltage, output voltage and output power. Fig. 12 shows an $I_{\mathrm{L}}$ value comparison between the closed form solution in [24] and the method proposed in this paper. The comparison results are as follows.

1. The $I_{\mathrm{L}}$ value controlled by the proposed method is a little more than that of the strategy in [24] under a light load, which can be seen from 100 W curves.


Fig. 12. $I_{\mathrm{L}}$ value comparison between the closed form solution and the method proposed in his paper.

(b)

Fig. 13. Modulation strategy of TCM in [24] under a light load. (a) $U_{\text {in }}>U_{\mathrm{o}} / \mathrm{n}$. (b) $U_{\text {in }}<U_{\mathrm{o}} / \mathrm{n}$.
2. The $I_{\mathrm{L}}$ value controlled by the proposed method is approximately equal to that of the strategy in [24] when the output power is near 300 W .
3. The $I_{\mathrm{L}}$ value controlled by the proposed method is significantly less than that of the strategy in [24] when the output power is greater than 300 W .
4. The $I_{\mathrm{L}}$ value controlled by the proposed method is less than that of the strategy in [24] in the whole output range when $U_{\mathrm{in}}=U_{\mathrm{o}} / \mathrm{n}$.

It can be seen that the method proposed in this paper has a larger current stress when compared with the strategy in [24] under a light load. However, it can be found that the modulation strategy in the proposed method under a light load is identical to that of the strategy in [24]. The larger current stress under a light load with the proposed strategy results from the small inductance of the buffering inductor. Fig. 13 shows waveforms from [24] using triangular current mode modulation (TCM). There are three parameters in the TCM: the primaryside bridge duty-ratio $D_{1}$, the secondary-side bridge duty-ratio $D_{2}$ and their phase-shift angle $\varphi$. Moreover, $\varphi=D_{1}-D_{2}$ when $U_{\text {in }}<U_{\mathrm{o}} / \mathrm{n}$, and $\varphi=D_{2}-D_{1}$ when $U_{\text {in }}>U_{\mathrm{o}} / \mathrm{n}$. Although the control parameters of the proposed method and the closed-form solution in [24] are different, the essentials of their operation are the same under a light load. The reason for a lager $I_{\mathrm{L}}$ under a light load of the proposed strategy is that it uses a smaller

TABLE IV
Parameters for the DAB Prototype

| Components | Parameters |
| :---: | :---: |
| Input voltage $U_{\text {in }}$ | $42-56 \mathrm{~V}$ |
| Output voltage $U_{\mathrm{o}}$ | 380 V |
| n | 7.755 |
| S1-S4 | IRFB4110 |
| S5-S8 | C2M0080120D |
| $\mathrm{P}_{\max }$ | 500 W |
| L | $6 \mu \mathrm{H}$ |



Fig. 14. Photo of the prototype.
inductance in the proposed method. The larger inductance can make the current waveform smooth when $i_{\mathrm{L}}$ is operated in the DCM, which leads to a smaller RMS value of $i_{\mathrm{L}}$.

Therefore, the requirements in Section II, such as soft switching, no circulating current, low current stress and high efficiency are fulfilled.

## VII. EXPERIMENTAL VERIFICATION

In order to verify the performance of the proposed control strategy, a 500 W prototype is implemented. The circuit components and electric specifications are chosen as Table IV and a photo of the prototype is shown in Fig. 14.

Fig. 15(a)-(c) show waveforms under different input voltages when the output power is 300 W . It can be seen that all three conditions are operated under the $B C M$, i.e. $d_{1}+d_{2}+d_{3}=1$, which guarantees a smaller current stress of the switches. However, the value of $\left(d_{1}+d_{2}+d_{3}\right)$ is less than 1 with a decrease of the output power. Fig. 15(d) shows the condition where the DAB is operated in the DCM when $U_{\mathrm{in}}=42 \mathrm{~V}$ and $P=100 \mathrm{~W}$. At this time, the DAB is at a voltage boost and $d_{3}$ is equal to zero.

Fig. 16 shows waveforms of a leading switch and a lagging switch in the LVS. Fig. 16(a) shows waveforms of the leading switch S1, which include its driven waveform $u_{\mathrm{S} 1}$, the voltage across the drain and source terminals $u_{\mathrm{DS} 1}$, and its through current $i_{\mathrm{DS} 1}$. There is almost no voltage spike in $u_{\mathrm{DS} 1}$. Fig. 16(b) and Fig. 16(c) are the turning on and turning off processes of

(b)

(c)


5 $\mu \mathrm{s} / \mathrm{div}$
(d)

Fig. 15. Waveforms of a DAB with different input voltages and powers. (a) $U_{\mathrm{in}}=42 \mathrm{~V}, P=300 \mathrm{~W}$. (b) $U_{\mathrm{in}}=49 \mathrm{~V}, P=300 \mathrm{~W}$. (c) $U_{\text {in }}=56 \mathrm{~V}, P=300 \mathrm{~W}$. (d) $U_{\text {in }}=42 \mathrm{~V}, P=100 \mathrm{~W}$.


Fig. 16. Soft switching condition of switches in the LVS. (a) Waveforms of the leading switch S1. (b) Turn on process of S1. (c) Turn off process of S1. (d) Waveforms of the lagging switch S2. (e) Turn on process of S2. (f) Turn off process of S2.


Fig. 17. Efficiency of a DAB under the control of the proposed strategy.

S1, respectively. Fig. 16(d) shows waveforms of the lagging switch S2, and Fig. 16(e) and Fig. 16(f) are the turning on and turning off processes of S2, respectively. The soft switching condition of S 2 is same as that of the leading switch S1. It can be seen that the switches in the LVS can also obtain ZVZCS, regardless of the leading switch or the lagging switch.

Efficiency curves of the proposed control strategy for a DAB are shown in Fig. 17. It can be seen that the maximum efficiency is about $95 \%$. The whole efficiency is at its highest when $U_{\text {in }}=49 \mathrm{~V}$ because the current stress is smallest under the same power.

The main loss of a DAB includes the conduction loss of the switches S1-S4 $\mathrm{P}_{\text {con_S14, }}$, the conduction loss of the switches S5-


Fig. 18. Pie diagrams of the loss breakdown of a DAB controlled by the proposed strategy when the input voltage is $U_{\text {in }}=42 \mathrm{~V}$. (a) $P=125 \mathrm{~W}$. (b) $P=250 \mathrm{~W}$. (c) $P=375 \mathrm{~W}$. (d) $P=500 \mathrm{~W}$.

S8 $\mathrm{P}_{\text {con_S58, }}$ the core loss and copper loss of the buffering inductor and transformer $\mathrm{P}_{\mathrm{Fe}}, \mathrm{P}_{\mathrm{Cu}}$. Some other loss $\mathrm{P}_{\text {other, }}$ such as the losses in the driving circuit, auxiliary power supply, DSP and other analog chips, occupies a small percentage of the total
loss. Fig. 18 shows the loss percentage of the different kinds of losses mentioned above. $\mathrm{P}_{\mathrm{Fe}}$ and $\mathrm{P}_{\mathrm{Cu}}$ are the main loss under a light load. Under a heavy load, the conduction loss of the switches are the main loss in a DAB.

## VIII. Conclusion

A control strategy for a DAB based on the high-frequency bipolar buck-boost principle is proposed in this paper. This strategy is derived from the classical DC/DC buck and boost converter operation principle. The proposed strategy guarantees that the inductor current is operated in the DCM or the BCM which leads to ZVS or ZCS of the switches, and high efficiency. An algorithm for solving the multi-variables is proposed and the specific flow is given. The calculation flow is simple and can be easily realized by a DSP. According to the solution of the multi-duty-ratios, the RMS value of $i_{\mathrm{L}}$ can be achieved, which is used to determine the parameters of the turns ratio and inductance. Experimental results verify that the proposed method has high performance.

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