Method to control flux balancing of high-frequency transformers in dual active bridge dc–dc converters

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Abstract: Due to non-ideal behaviours and switching characteristics, dc bias might be generated in the dual active bridge converters. It could be more serious in the high-frequency transformers because of small loop resistances. It could also restrict the range of operating flux density due to the threat of saturation. This study analyses the steady-state dc bias and divides it into two categories. Then an active flux-balancing control method is proposed to eliminate these dc biases. It uses separate proportional integral controls and adjusts duty ratios of both sides independently to achieve flux balance. A generalised average model is established to verify its dynamic performance and stability. Simulation and experimental verification on a laboratory prototype confirm the performance of this approach.

1 Introduction
With the increasing share of decentralised and renewable power sources, the need for power electronics and especially for efficient high-frequency high-power dc–dc converters is expected to grow [1]. Also, dual-active-bridge (DAB) is a promising structure of dc–dc converters and attracts much attention because of its advantages of high-power density, zero-voltage switching, fast dynamic response, bi-directional power transfer capability, modularity, and symmetric structure [2, 3]. So far, most of the international research on DAB has focused on the following aspects: basic characterisation, control strategy, topology, soft-switching solution, hardware design and optimisation [4]. However, there are few studies on magnetic components of DAB.

Some practical problems about high-frequency transformers are seldom mentioned. In fact, primary attention is supposed to be paid to define an optimal operating flux density in the transformer design, especially transient conditions and a threat of dc bias related to magnetic caused by non-ideal behaviours should be considered [5]. Generally, the loop resistance could make a new balanced working state when dc bias appears. The large resistance could suppress dc component of magnetic current and reduce the risk of transformer saturation. Nevertheless, to decrease the loss of DAB system, the loop resistance is designed to be small, which makes this problem for dc bias in high-frequency transformers more serious. There is a practical example that, as for the shell-type 166 kW/20 kHz efficiency-optimised transformer, primary side equivalent resistance is ~1.7 mΩ and 0.0125% of the relative difference in the duration would suffice to create a dc flux density component of $B_{dc} = 50 \text{mT}$ [6]. Thus, a conservative design practice needs to be used to ensure the safe operation of the transformer even with a dc flux density component. The initial value of the operating flux density selected is usually <50% of the saturation value. Low utilisation of flux density restricts the development of miniaturisation and high efficiency for DAB. Therefore, more research is necessary.

The factors that can cause dc bias in the transformer are as follows: (i) the adjustment of the phase-shift ratio derived from the closed-loop control system; (ii) the inconsistency of the pulse width from the driving control circuit; (iii) the variation in the terminal dc voltage during a period; and (iv) the inconsistency between the characteristic parameters of switches, such as on-state voltage drop, switching speed etc. [2]. These factors can be divided into two categories. The (i) is a dynamic-state factor, the error could be large but exist in a short time. The particularity is that this error caused by the change of phase-shift ratio can be known and controlled, so the step-by-step phase shift method could be used to eliminate this dc bias [1, 2, 7–10]. The (ii), (iii) and (iv) are steady-state factors, the error could be small but exist in a long time. As the mentioned example, this error cannot be neglected because of small resistances. On the one hand, some scholars solved this problem in term of hardware, including the connection of dc-blocking capacitor [11], the application of ‘The Magnetic Ear’ [6, 12], the addition coil-to-control excitation current [13], placing an air gap in the transformer core and placing the flux sensor [14]. These methods, changing topologies of DAB or transformers, need to consider the complexity, the cost, the reliability and the effectiveness. On the other hand, it could be solved by active control of flux bias, extracting the flux bias information from currents [13, 15–17]. In [13, 17], the dc component of primary current or excitation current is employed to eliminate dc bias, but it cannot solve the situation when the errors exist on both primary and secondary side of DAB. In [15, 16], the method using dc components of primary current and excitation current has been proposed to eliminate dc bias. However, it uses the valley-peak value of currents to obtain an average value, which is sensitive to noise and lacks the theoretical analysis for its control method.

According to the previous discussion, this paper presents an active flux-balancing control method and makes a comprehensive theoretical analysis to prove its stability and effectiveness. It extracts the average values of primary current and secondary current based on the fast sampling method and uses separate proportional integral (PI) controls adjusting duty ratios of both sides independently to eliminate steady-state dc bias. The paper also builds the generalised averaging model of DAB and gives an analysis of stability. Moreover, the simulation and experiment verify this method.

2 Generalised averaging model
2.1 State equations of DAB converters
The conventional state-space averaging method is a common method in the model and analysis of switched power converters. However, in the DAB converter, the transformer current has a dominant ac component, which makes it difficult to apply the conventional state-space averaging method [18]. Therefore, the development of a full-order continuous time average model has been presented using the generalised averaging approach [19]. In order to retain the current information in the dc-bias state, the schematic of a DAB dc–dc converter is shown in Fig. 1.
The input capacitance is usually relatively large. Therefore, the dynamics of the input capacitor can be neglected in this model. The voltage source is represented as $V_i$. The primary resistance and secondary resistance, including transistor on-time resistance and transformer winding resistance, are represented by equivalent resistances $R_p$ and $R_o$. Similarly, external inductance and transformer leakage inductance at both windings are lumped as an equivalent inductance $L_m$. The transformer magnetic inductance is represented by $L_m$. The output capacitance is represented as $C_o$. The load is represented as $R_p$. The ac currents on the input side and the output side are expressed as $i_p$ and $i_o$ respectively. The ac voltages on the input side and the output side are expressed as $v_p$ and $v_o$, which is related to the switching states of DAB. Therefore, $v_p$ and $v_o$ can be expressed as follows:

$$
\begin{align*}
  v_p &= s_1(t) V_i \\
  v_o &= s_2(t) v_i
\end{align*}
$$

$(1)$

$s_1(t), s_2(t)$ have the following definition:

$$
\begin{align*}
  s_1(t) &= \begin{cases} 
    1 & Q_i, Q_{on} \\
    -1 & Q_i, Q_{on}
  \end{cases} \\
  s_2(t) &= \begin{cases} 
    1 & Q_o, Q_{on} \\
    -1 & Q_o, Q_{on}
  \end{cases}
\end{align*}
$$

$(2)$

$(3)$

According to phase-shift DAB, the waveforms of $s_1(t), s_2(t)$ are given in Fig. 2. Their duty ratios are represented as $d$, the phase shift between two waveforms is represented as $\phi$.

Selecting the primary current $i_p$, the secondary current $i_o$, and the output capacitor voltage $v_o$ as state variables, the state equations of the DAB converter can be derived as follows:

$$
\begin{align*}
  \frac{di_p}{dt} &= -\frac{R_p}{L_p} i_p - \frac{R_o}{L_o} i_o + \frac{1}{L_p} v_p - \frac{1}{L_o} v_i \\
  \frac{d}{dt} &= -\frac{R_p}{L_p} i_p - \frac{R_o}{L_o} (L_e + L_m) i_o \\
  &= + \frac{1}{L_p} v_p - \frac{1}{L_o} (L_e + L_m) v_i \\
  \frac{dv_o}{dt} &= -\frac{1}{RC_o} v_o + \frac{1}{C_o} i_o \sin(t)
\end{align*}
$$

$(4)$

Applying $(1)$ in $(4)$, then

$$
\begin{align*}
  \frac{dv_p}{dt} &= -\frac{R_p}{L_p} v_p(t) - \frac{R_o}{L_o} v_i(t) \\
  &= + \frac{1}{L_p} V_s(t) - \frac{R_o}{L_o} v_i(t) \sin(t) \\
  \frac{dv_o}{dt} &= -\frac{1}{RC_o} v_o(t) + \frac{1}{C_o} i_o(t) \sin(t)
\end{align*}
$$

$(5)$

It is the complete model of DAB.

2.2 Generalised averaging model

The main idea of the generalised average model is to expand the time-varying state variables in the Fourier series. Then, the circuit could be expressed as a set of differential equations. The new state variables in the equations are the coefficients of the Fourier series. When the generalised averaging technique uses more terms in the Fourier series, the model could be more accurate [19].

Sanders et al. [20] have proposed the degrading process. In general, it is natural to include zeroth and first coefficients in the Fourier series to represent the averages of state variables. Therefore, according to $(5)$, 0th, 1st (the real part of the first coefficient), 11th (the imaginary part of the first coefficient) of each state variable, can be selected as new state variables and the new ninth-order state equations could be constructed as shown in the following equation:

$$
\frac{dX}{dt} = \begin{bmatrix} X_0 \\
X_R \\
X_A \\
X_B \end{bmatrix}
$$

$(6)$

In $(6)$

$$
\begin{align*}
  X_0 &= \begin{bmatrix} x_{i0} \\
x_{i1} \end{bmatrix} \\
x_{i1} &= \begin{bmatrix} x_{i0} \\
x_{i1} \end{bmatrix}
\end{align*}
$$

$(7)$

According to related formulas in [19], the complete ninth-order generalised state model could be obtained, and there are more details in Appendix 1.

3 Small-signal model of steady-state flux bias

In general, the causes of steady-state bias in the DAB system could be classified into two categories, which are reflected in the pulse-width error (duty ratio) and the amplitude error of the variables $s_1(t)$ and $s_2(t)$. With these two types of errors, the model could be
simplified, and the small-signal model of the system could be
deduced.

3.1 Small-signal model with pulse width error
In the actual control of the DAB, due to the imperfection of control system, the square wave voltage excitation of the transformer could be asymmetrical between the positive and negative pulse widths, which influences the duty ratios. In the model, \( d_1 \) and \( d_2 \) represent the duty ratios of \( s_1(t) \) and \( s_2(t) \), respectively, the phase shift between the two waveforms is represented by \( \phi/\pi \). Then, the zeroth and the first coefficient could be expressed as follows:

\[
\begin{align*}
\{s_i\}_0 &= 2d_i - 1 \\
\{s_i\}_R &= -\frac{\sin(2d_i \pi)}{\pi} \\
\{s_i\}_d &= \frac{1}{\pi} [\cos(2d_i \pi) - 1] \\
\{s_i\}_x &= 2d_i - 1 \\
\{s_i\}_R &= -\frac{\sin(2d_i \pi)}{\pi} \cdot \cos(d_i \pi) \\
&\quad + \frac{\cos(2d_i \pi) - 1}{\pi} \cdot \sin(d_i \pi) \\
\{s_i\}_d &= \frac{\cos(2d_i \pi) - 1}{\pi} \cdot \cos(d_i \pi) \\
&\quad + \frac{\sin(2d_i \pi)}{\pi} \cdot \sin(d_i \pi)
\end{align*}
\]

The small-signal model is applied to analyse the state equations. There is a small perturbation in \( d_i \) and \( \Delta x \) and state variables as

\[
\begin{align*}
\dot{d}_i &= D + \Delta d_i \\
\dot{\Delta x} &= X + \Delta x
\end{align*}
\]

where capital variables represent steady states, \( \Delta \) variables mean small signal states, and \( ^\cdot \) variables mean large signal states. With approximated non-linear terms, the following formula could be calculated by using (6)–(11), and it selects \( \Delta x \) as the variables, more details can be found in Appendix 2.

\[
\begin{align*}
\frac{d}{dt} \begin{bmatrix} \Delta x_{\text{R}} \\ \Delta x_{\text{d}} \\ \Delta x_{\text{1}} \\ \Delta x_{\text{2}} \end{bmatrix} &= \begin{bmatrix} A_0 & A_{\text{R}} & A_{\text{d}} & A_{\text{1}} \\ A_{\text{R}} & A_0 & A_{\text{d}} & A_{\text{1}} \\ A_{\text{d}} & A_{\text{1}} & A_0 & A_{\text{R}} \\ A_{\text{1}} & A_{\text{d}} & A_{\text{R}} & A_0 \end{bmatrix} \begin{bmatrix} \Delta x_{\text{R}} \\ \Delta x_{\text{d}} \\ \Delta x_{\text{1}} \\ \Delta x_{\text{2}} \end{bmatrix} \\
&\quad + \begin{bmatrix} C_\text{1} \\ C_\text{1} \end{bmatrix} \begin{bmatrix} \Delta d_1 \\ \Delta d_2 \end{bmatrix}
\end{align*}
\]

When \( D = 0.5 \), \( \{i_{\text{y1}}\}_0 \) and \( \{i_{\text{y2}}\}_0 \) (the average value of ac currents at the input side and the output side) could be decoupled from other state variables. Assuming that there is no voltage fluctuation on the dc voltage of output side \( \{(i_{\text{y1}})\}_\text{R} = 0 \), (11) can be simplified. The small-signal model of \( \{i_{\text{y1}}\}_0 \) and \( \{i_{\text{y2}}\}_0 \) is

\[
\begin{align*}
\frac{d}{dt} \begin{bmatrix} \Delta i_{\text{y1}} \\ \Delta i_{\text{y2}} \end{bmatrix} &= A_i \begin{bmatrix} \Delta i_{\text{y1}} \\ \Delta i_{\text{y2}} \end{bmatrix} + \begin{bmatrix} 2\Delta d_1 \\ 2\Delta d_2 \end{bmatrix}
\end{align*}
\]

where

\[
A_i = \begin{bmatrix} R_{\text{R}} & -\frac{R_{\text{R}}}{nL_1} \\ -\frac{R_{\text{d}}}{nL_1} & R_{\text{R}}(L_1 + L_0) - \frac{R_{\text{R}}}{nL_1} \end{bmatrix}
\]

\[
B_i = \begin{bmatrix} \frac{V_1}{L_1} - \frac{\langle V_{\text{y1}}\rangle_0}{nL_1} \\ \frac{V_1}{nL_1} - \frac{(L_1 + L_0)\langle V_{\text{y2}}\rangle_0}{nL_1} \end{bmatrix}
\]

\( \langle V_{\text{y1}}\rangle_0 \) is the outside-capacitor voltage value of the steady-state operating point.

3.2 Small-signal model with amplitude error
The steady-state bias caused by the amplitude error mainly refers to the asymmetry between the positive and negative amplitudes of the voltage excitation for transformers, due to the differences in switching tube parameters, voltage fluctuations and so on. As a result, the amplitudes of \( v_y \) and \( v_x \) are not same. In the model, small ac perturbations and small dc perturbations express these errors as

\[
\begin{align*}
\dot{s}_1(t) &= s_1(t) + \Delta s_1(t) + \Delta k_{v1} \\
\dot{s}_2(t) &= s_2(t) + \Delta s_2(t) + \Delta k_{v2}
\end{align*}
\]

\[
\begin{align*}
\Delta s_1(t) &= k_1 s_1(t) \\
\Delta s_2(t) &= k_2 s_2(t)
\end{align*}
\]

where, \(-1 < \Delta k_{v1}, \Delta k_{v2} < 1, 0 < k_1, k_2 < 1\).

Similarly, applying the small-signal model, it is deduced near the steady-state operating points. The state equations are as follows, and more details can be found in Appendix 3.

\[
\frac{d}{dt} \begin{bmatrix} \Delta v_0 \\ \Delta v_{\text{R}} \\ \Delta v_{\text{d}} \\ \Delta v_{\text{1}} \\ \Delta v_{\text{2}} \end{bmatrix} = \begin{bmatrix} A_0 & A_{\text{R}} & A_{\text{d}} & A_{\text{1}} & A_{\text{2}} \\ A_{\text{R}} & A_0 & A_{\text{d}} & A_{\text{1}} & A_{\text{2}} \\ A_{\text{d}} & A_{\text{1}} & A_0 & A_{\text{R}} & A_{\text{2}} \\ A_{\text{1}} & A_{\text{d}} & A_{\text{R}} & A_0 & A_{\text{2}} \end{bmatrix} \begin{bmatrix} \Delta v_0 \\ \Delta v_{\text{R}} \\ \Delta v_{\text{d}} \\ \Delta v_{\text{1}} \\ \Delta v_{\text{2}} \end{bmatrix} \\
&\quad + \begin{bmatrix} D_1 \\ D_1 \\ D_2 \\ D_2 \end{bmatrix} \begin{bmatrix} \Delta k_{v1} \\ \Delta k_{v2} \end{bmatrix}
\]

When \( D = 0.5 \), applying the similar simplification, the small-signal model of \( \{i_{\text{y1}}\}_0 \) and \( \{i_{\text{y2}}\}_0 \) with amplitude error is

\[
\frac{d}{dt} \begin{bmatrix} \Delta \{i_{\text{y1}}\}_0 \\ \Delta \{i_{\text{y2}}\}_0 \end{bmatrix} = A_i \begin{bmatrix} \Delta \{i_{\text{y1}}\}_0 \\ \Delta \{i_{\text{y2}}\}_0 \end{bmatrix} + B_i \begin{bmatrix} \Delta k_{v1} \\ \Delta k_{v2} \end{bmatrix}
\]

In (18), \( A_i \) and \( B_i \) are same in (12). Therefore, they have the same expressions.

Considering pulse width and amplitude error, the complete small-signal model is the sum of (11) and (17) with linearisation. When \( D = 0.5 \), the steady-state values of \( \{i_{\text{y1}}\}_0 \) and \( \{i_{\text{y2}}\}_0 \) are zero. Therefore, \( \{i_{\text{y1}}\}_0 = \Delta \{i_{\text{y1}}\}_0 \) and \( \{i_{\text{y2}}\}_0 = \Delta \{i_{\text{y2}}\}_0 \). In summary, the small-signal model is (19), \( A_{p1} \) and \( A_{p2} \) represent the steady-state bias

\[
\begin{align*}
\frac{d}{dt} \begin{bmatrix} \{i_{\text{y1}}\}_0 \\ \{i_{\text{y2}}\}_0 \end{bmatrix} &= A_i \begin{bmatrix} \{i_{\text{y1}}\}_0 \\ \{i_{\text{y2}}\}_0 \end{bmatrix} + B_i \begin{bmatrix} \Delta k_{v1} \\ \Delta k_{v2} \end{bmatrix}
\end{align*}
\]

where

\[
\begin{align*}
\Delta p_1 &= 2\Delta d_1 + \Delta k_{v1} \\
\Delta p_2 &= 2\Delta d_2 + \Delta k_{v2}
\end{align*}
\]

\[
C_i = \begin{bmatrix} 1 \\ 0 \\ 0 \\ 1 \end{bmatrix}
\]
Separate PI controls for steady-state flux bias

4.1 Principle of control method

In DAB system, $i_p^0$ and $i_s^0$ have their actual physical meanings, which represent the dc components of currents on both sides. Their magnitudes also could be used to characterise the flux bias information of high-frequency transformers. When actively suppressing the steady-state bias of the DAB system, the values of $i_p^0$ and $i_s^0$ would be approximately equal to zero. On the basis, a control method is presented, extracting the dc components of input and output currents, and adjusts duty cycles of both sides through feedback PI controls to eliminate the flux bias. The control diagram is shown in Fig. 3a, the detailed diagram of PI control is shown in Fig. 3b.

4.2 Stability analysis of control method

According to the small-signal model obtained in the previous section, the corresponding control block diagram is shown in Fig. 4.

In Fig. 4

$$K_p = \begin{bmatrix} k_{p1} & 0 \\ 0 & k_{p2} \end{bmatrix} \quad (21)$$

$$K_i = \begin{bmatrix} k_{i1} & 0 \\ 0 & k_{i2} \end{bmatrix} \quad (22)$$

It could be known from (19) that $A_1^\dagger B_1$ is a diagonal matrix. Accordingly, in the steady state $\langle i_p^0 \rangle$ is only related to $\Delta p_1$, $\langle i_s^0 \rangle$ is only related to $\Delta p_2$, which means they are decoupled from each other. However, in the dynamic process, $\langle i_p^0 \rangle$ and $\langle i_s^0 \rangle$ have a coupling effect with each other. It is necessary to examine the stability of the system with these two independent PI controls. Therefore, the state variable integration terms are introduced as new state variables, and the system augmentation state matrix is obtained as

$$\frac{d}{dt} \begin{bmatrix} \langle i_p^0 \rangle \\ \langle i_s^0 \rangle \end{bmatrix} = \begin{bmatrix} A_1 - B_1 K_p C_1 & -B_1 K_i C_1 \\ C_1 & 0 \end{bmatrix} \begin{bmatrix} \langle i_p^0 \rangle \\ \langle i_s^0 \rangle \end{bmatrix} \begin{bmatrix} \langle i_p^0 \rangle \\ \langle i_s^0 \rangle \end{bmatrix} + \begin{bmatrix} B_1 \langle \Delta p_1 \rangle \\ 0 \end{bmatrix} \begin{bmatrix} \Delta d_1 \\ \Delta d_2 \end{bmatrix} \quad (23)$$

Applying the Routh criterion, eigenvalues of the matrix can judge the stability of this system, when

$$k_{p1} = -k_{p2} > k_{i1} = -k_{i2} > 0 \quad \left(24\right)$$

the real parts of the eigenvalues are all located on the negative semi-axis, so the system is stable.

4.3 Influence of saturation region

There is a particular situation that the high-frequency transformer operates in the saturation region due to flux bias. According to the fundamental principle of transformers, the following equation is obtained:

$$\langle i_p^0 \rangle < \frac{n^2 V_i}{R_p}$$

Fig. 3  Control diagram
(a) The diagram of current active control, (b) The detailed control block diagram

Fig. 4  Control block diagram

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\[ v_p = R_p i_p + L \frac{di_p}{dt} + \frac{d\psi}{dt} \]
\[
\int_{t_0}^{t_0 + T} v_p dt = R_p \int_{t_0}^{t_0 + T} i_p dt + L \int_{t_0}^{t_0 + T} di_p + \int_{t_0}^{t_0 + T} d\psi
\]

In (25), \( \psi \) is the flux. It can be found that \( i_p \) and \( \psi \) change periodically when the system is in a stable state, so the dc component of \( i_p \) is only dependent on the dc component of \( v_p \) whether saturated or not, the same conclusion for \( i_s \). In fact, when the transformer enters the saturation region, it is reflected in the state equations that \( L_m \) would change with the excitation current. However, from the derivation of the previous section, the stability of the system is independent of the value of \( L_m \). Therefore, regardless of whether the transformer enters the saturated region, the separate PI controls can make the system stable without static differences.

5 Simulation and experiments verification

Simulation and experiments are provided to verify the performance of this method. A DAB converter is built and tested for the steady flux bias. The specifications of the DAB converter are: the external inductor is 3.3 \( \mu \)H, the leakage inductance of the transformer is 0.315 \( \mu \)H, the output capacitance is 200 \( \mu \)F, the switching frequency is 50 kHz, and the turns ratio of the high-frequency transformer is 1:8. The switches of input side are Infineon MOSFET IP1076N15N5, and the switches of output side are CREE SiC MOSFET C3M0065090J. There have same parameters in the simulation model.

5.1 Sampling method

In order to accurately obtain the synchronous and real-time signal for dc component of currents in every switching cycle, sampling takes place ten times in the period as shown in Fig. 5. There is an interruption at the beginning of each cycle (\( n \)), calculating the average value of ten sampling points in the last cycle, adjusting the new duty cycle and loading in the next cycle (\( n + 1 \)). This method is applied in the simulation and experiment.

5.2 Simulation research

The DAB model is built in SIMULINK on the basis. In order to verify the proposed method, according to (19), the flux bias is simulated by supposing \( \Delta p_1 = 0.04 \) and \( \Delta p_2 = 0.02 \). First, according to the deduced state equation with PI controls as (23), the average values of primary and secondary currents can be calculated. Then considering pulse-width error, one condition is simulated with \( d_1 = 0.52 \) and \( d_2 = 0.51 \) in SIMULINK model. Furthermore, the saturation condition is also simulated with same duty cycles. Last, considering amplitude error, another condition is simulated with \( \Delta k_v^2 = 0.04 \), \( \Delta k_s^2 = 0.02 \). The average values of primary and secondary currents are compared in Figs. 6 and 7. These excitation currents simulated in the SIMULINK model are compared in Fig. 8.

It could be found that this proposed method can control the average value of currents and eliminate the flux bias. Figs. 6 and 7 can also verify the formula (19) and (23). It could be proved that this proposed method is still effective if the high-frequency transformer operates in the saturation region in Fig. 8.

5.3 Experimental verification

The DAB control is implemented by using a TMS320F28335 DSP from Texas Instruments. It could be found from the simulation results that pulse-width error and amplitude error have the same dynamic responses. Therefore, the experiments are mainly talking
about the pulse-width error. Fig. 9 shows the experimental waveforms for $V_i = 20\,\text{V}$, $d_1 = 0.52$ and $d_2 = 0.51$, (a) without PI control on both sides, (b) with PI control on both sides, (c) and (d) show the detailed experimental waveforms at the end time. In the waveforms, these curves represent the input voltage ($v_i$), the output voltage ($v_o$), the primary current ($i_p$) and the secondary current ($-i_s$). From Figs. 9a and b, it could be found that the proposed method can eliminate steady flux bias and also can decrease the maximum current caused by the flux bias in primary current and secondary current.

Duty cycles and proportion factors are changed to fully verify the effectiveness and stability of this control method. Fig. 10 shows the experimental waveforms for $V_i = 20\,\text{V}$, $d_1 = 0.52$ and $d_2 = 0.51$, (a) without PI control on both sides, (b) with PI control on both sides, (c) with PI control on the primary side, and proportion factor increased by four times. The details of these waveforms at the end time are shown in Fig. 11. It could be found that the PI control on one side can only influence the average value of current on this side, therefore only if there are PI controls on both sides, the average value of excitation current can be nearly zero and flux bias can be eliminated. In addition, comparing Fig. 10c or Fig. 9b with Fig. 10a, the increased proportion factor can accelerate the control process and it has almost no effect on the stability of the system.

To verify the independence of the PI controls on the primary side and the secondary side, Fig. 12a shows the experimental waveforms for $V_i = 20\,\text{V}$, $d_1 = 0.5$ and $d_2 = 0.51$ with PI control in both sides and Fig. 12b shows the experimental waveforms for $V_i = 20\,\text{V}$, $d_1 = 0.5$ and $d_2 = 0.51$ with PI control in both sides. It could be found that the adjustment of one side cannot affect the other side, and they are decoupled from each other whether dynamic or static processes.

6 Conclusion

The generalised model and small-signal model with steady-state dc bias have been derived. The dc components of the primary and secondary current could decouple from each other. On the basis, it could be proved that this proposed method can be stable in a wide range of PI control parameters. The dynamic response and controlling effect for steady-state dc bias are all verified in simulations and experiments. The dc components whether in primary or secondary currents could be eliminated through two separate PI controls.
Fig. 10  Experimental waveforms for $d_1 = 0.51$ and $d_2 = 0.49$
(a) With PI control on both sides, (b) With PI control on the primary side, (c) With PI control on the secondary side

Fig. 11  Final detailed experimental waveforms for $d_1 = 0.51$ and $d_2 = 0.49$
(a) With PI control on the primary side, (b) With PI control on the secondary side
7 References

[1] Engel, S.P., Soltan, N., Stagge, H., et al.: ‘Improved instantaneous current control for high-power three-phase dual-active bridge DC–DC converters’, IEEE Trans. Power Electron., 2014, 29, (8), pp. 4067–4077.

[2] Zhao, B., Song, Q., Liu, W., et al.: ‘Transient DC bias and current impact effects of high-frequency-isolated bidirectional DC–DC converter in practice’, IEEE Trans. Power Electron., 2016, 31, (4), pp. 3203–3216.

[3] Costinett, D., Maksimovic, D., Zane, R.: ‘Design and control for high efficiency in high step-down dual active bridge converters operating at high switching frequency’, IEEE Trans. Power Electron., 2013, 28, (8), pp. 3931–3940.

[4] Zhao, B., Song, Q., Liu, W., et al.: ‘Overview of dual-active-bridge isolated bidirectional DC–DC converter for high-frequency-link power-conversion system’, IEEE Trans. Power Electron., 2014, 29, (8), pp. 4091–4106.

[5] Vinnikov, D., Laugić, J., Galkin, I.: ‘Middle-frequency isolation transformer design issues for the high-voltage DC/DC converter’. 2008 IEEE Power Electronics Specialists Conf., Rhodes, Greece, 2008, pp. 1930–1936.

[6] Ortiz, G., Fässler, L., Kolar, J.W., et al.: ‘Flux balancing of isolation transformers and application of ‘the magnetic ear’ for closed-loop volt-second compensation’, IEEE Trans. Power Electron., 2014, 29, (8), pp. 4078–4090.

[7] Lin, S.T., Li, X., Sun, C., et al.: ‘Fast transient control for power adjustment in a dual-active-bridge converter’, Electron. Lett., 2017, 53, (16), pp. 1130–1132.

[8] Li, X., Hu, S., Sun, C., et al.: ‘Asymmetric double-sided modulation for fast load transition in a semi-dual-active-bridge converter’, IET Power Electron., 2017, 10, (13), pp. 1698–1704.

[9] Li, X., Li, Y.F.: ‘An optimized phase-shift modulation for fast transient response in a dual-active-bridge converter’, IET Power Electron., 2014, 29, (6), pp. 2661–2665.

[10] Engel, S.P., Soltan, N., Stagge, H., et al.: ‘Dynamic and balanced control of three-phase high-power dual-active bridge DC–DC converters in DC-grid applications’, IEEE Trans. Power Electron., 2013, 28, (4), pp. 1880–1889.

[11] Panov, Y., Jovanović, M.M., Irving, B.T.: ‘Novel transformer-flux-balancing control of dual-active bridge DC–DC converter’, 2014 Int. Conf. Renewable Energy Research Application (ICRERA), Milwaukee, WI, USA., 2014, pp. 490–495.

[12] Baddipadiga, B.P., Ferdowsi, M.: ‘Dual loop control for eliminating DC-bias in a DC–DC dual active bridge converter’, 2014 IEEE Trans. Power Electron., 2015, vol. 30, no. 1, pp. 279–288.

[13] Panov, Y., Jovanović, M.M., Gong, L., et al.: ‘Transformer-flux-balancing control in isolated bidirectional DC–DC converters’. 2014 IEEE Applied Power Electronics Conf. Exposition – APEC 2014, Fort Worth, TX, USA., 2014, pp. 42–49.

[14] Azzimonti, L., Zane, R.: ‘Transformer-flux-balancing control of dual-active bridge DC–DC converters’. 2014 IEEE Applied Power Electronics Conf. Exposition – APEC 2014, Fort Worth, TX, USA., 2014, pp. 42–49.

[15] Zhang, K., Shan, Z., Jantschek, J.: ‘Large- and small-signal average-value modeling of dual-active-bridge DC–DC converter considering power losses’, IEEE Trans. Power Electron., 2017, 32, (3), pp. 1964–1974.

[16] Qin, H., Kimball, J.W.: ‘Generalized average modeling of dual active bridge DC–DC converter’, IEEE Trans. Power Electron., 2012, 27, (4), pp. 2078–2084.

[17] Sanders, S.R., Noworolski, J.M., Liu, X.Z., et al.: ‘Generalized averaging method for power conversion circuits’, IEEE Trans. Power Electron., 1991, 6, (2), pp. 251–259.

8 Appendix

8.1 Appendix 1

\[
\begin{align*}
A_0 &= A_{IR} = A_{10} \\
&= \begin{bmatrix}
-\frac{R_0}{nL_t} & -\frac{R_0}{nL_t} & -\frac{1}{nL_t} (s)_{h_0} \\
0 & -\frac{R_0}{nL_t} - \frac{R_0(L_0 + L_m)}{nL_t} & -\frac{L_0 + L_m}{nL_t} (s)_{h_0} \\
0 & 1 & -\frac{1}{C_0} (s)_{h_0} \\
\end{bmatrix}
\end{align*}
\]

\[
A_{R01} = 2A_{R00} = \begin{bmatrix}
0 & 0 & -\frac{2}{nL_t} (s)_{R01} \\
0 & 0 & -\frac{2(L_0 + L_m)}{nL_t} (s)_{R01} \\
0 & \frac{2}{nL_t} (s)_{R01} & 0 \\
\end{bmatrix}
\]

\[
A_{01} = A_{R0} = \begin{bmatrix}
\omega & 0 & 0 \\
0 & 0 & 0 \\
0 & 0 & \omega \\
\end{bmatrix}
\]

\[
B_{R01} = \begin{bmatrix}
\frac{1}{L_t} (s)_{R01} & 1 & \frac{1}{nL_t} (s)_{R01} & 0 \\
\end{bmatrix}^T
\]

8.2 Appendix 2

\[
\begin{align*}
\langle s \rangle_{01} (X_1 + \Delta X) &= \frac{2(D + \Delta d_1 - 1)}{X_1 + \Delta X_1} \\
& \approx \langle s \rangle_{01} (X_1 + \Delta X_1) + 2X_1 \Delta d_1 \\
\langle s \rangle_{01} (X_1 + \Delta X) & \approx \langle s \rangle_{01} (X_1 + \Delta X_1) + 2\cos(d_{opt})(X_1 \Delta d_1) \\
\langle s \rangle_{11} (X_1 + \Delta X) & \approx \langle s \rangle_{11} (X_1 + \Delta X_1) - 2\sin(d_{opt})(X_1 \Delta d_1)
\end{align*}
\]
\[ (s_i)_1 = (s_i)_0 + 2\Delta d_i \]
\[ (s_i)_{1R} \approx (s_i)_{1R} + 2\Delta d_i \]
\[ (s_i)_{11} = (s_i)_{11} - 2\sin(2D\pi)\Delta d_i \]

\[ C_1 = \begin{bmatrix}
\frac{2V_i}{L_i} & \frac{2V_i}{nL_i} & 0 & \frac{2V_i}{2L_i} & \frac{2V_i}{nL_i} & 0 \\
0 & 2\sin(2D\pi)V_i & 2\sin(2D\pi)V_i & 0 & 0 & 0
\end{bmatrix}^T
\]

\[ C_2 = \begin{bmatrix}
-\frac{1}{nL_i} \cdot K_1 - \frac{L_i + L_m}{n' L_i L_m} \cdot K_1 \cdot \frac{1}{C_o} \cdot K_2 \\
-\frac{1}{nL_i} \cdot K_3 - \frac{L_i + L_m}{n' L_i L_m} \cdot K_3 \cdot \frac{1}{C_o} \cdot K_4 \\
-\frac{1}{nL_i} \cdot K_5 - \frac{L_i + L_m}{n' L_i L_m} \cdot K_5 \cdot \frac{1}{C_o} \cdot K_6
\end{bmatrix}
\]

\[ K_1 = 2(v_0)_0 + 4\cos(d_1\pi)(v_0)_{1R} - 4\sin(d_1\pi)(v_0)_{11} \]
\[ K_2 = 2(i_0)_0 + 4\cos(d_1\pi)(i_0)_{1R} - 4\sin(d_1\pi)(i_0)_{11} \]
\[ K_3 = 2(v_0)_0 + 4\cos(d_1\pi)(v_0)_{1R} - 2\sin(d_1\pi)(v_0)_{11} \]
\[ K_4 = 2(i_0)_0 + 2\cos(d_1\pi)(i_0)_{1R} - 2\sin(d_1\pi)(i_0)_{11} \]

8.3 Appendix 3

\[ D_1 = \begin{bmatrix}
V_i / L_i & \frac{V_i}{nL_i} & 0 & \frac{V_i}{L_i} (s_i)_0 & \frac{V_i}{nL_i} (s_i)_0 & 0 \\
0 & \frac{V_i}{nL_i} & 2V_i / nL_i & 0 & 0 & 0
\end{bmatrix}^T
\]

\[ D_2 = \begin{bmatrix}
V_i / L_i & \frac{V_i}{nL_i} & 0 & 0 & 0 & 0 \\
0 & \frac{V_i}{nL_i} & 0 & 0 & 0 & 0
\end{bmatrix}^T
\]

\[ D_3 = \begin{bmatrix}
-\frac{1}{nL_i} \cdot K_1 - \frac{L_i + L_m}{n' L_i L_m} \cdot K_1 \cdot \frac{1}{C_o} \cdot K_2 \\
-\frac{1}{nL_i} \cdot K_3 - \frac{L_i + L_m}{n' L_i L_m} \cdot K_3 \cdot \frac{1}{C_o} \cdot K_4 \\
-\frac{1}{nL_i} \cdot K_5 - \frac{L_i + L_m}{n' L_i L_m} \cdot K_5 \cdot \frac{1}{C_o} \cdot K_6
\end{bmatrix}
\]

\[ \begin{array}{c}
K_1 = (s_i)_0 (v_0)_0 + 2(s_i)_{1R} (v_0)_{1R} - 2(s_i)_{11} (v_0)_{11} \\
K_2 = (s_i)_0 (i_0)_0 + 2(s_i)_{1R} (i_0)_{1R} - 2(s_i)_{11} (i_0)_{11} \\
K_3 = (s_i)_0 (v_0)_0 + (s_i)_{1R} (v_0)_{1R} - (s_i)_{11} (v_0)_{11} \\
K_4 = (s_i)_0 (i_0)_0 + (s_i)_{1R} (i_0)_{1R} - (s_i)_{11} (i_0)_{11}
\end{array}
\]