Current-Sensorless Robust Control for Bidirectional DC-DC Dual Active Bridge Converter with Current Stress Optimization

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***Abstract*—This paper proposes a current-sensorless robust control method for the output voltage of the dual active bridge converter under dual phase-shift modulation while minimizing the current stress. The deadbeat control-based Lagrange multiplier method is employed to optimize current stress. Then, the extended state observer is used to observe the load current and inject it directly into the predicted phase-shift duty ratios after each switching cycle. The proposed observer is very simple to construct and delivers excellent performance without the need for any additional compensation controller.**

Keywords—Dual active bridge, current sensorless, dual phase-shift, extended state observer

I. Introduction

The dual active bridge (DAB) converter is widely used because of its advantages such as high efficiency, galvanic isolation, and bidirectional transmission power. Due to capacity limitations from the semiconductor manufacturer, converters must be connected in series or parallel to increase the transmission power, distribution reliability, and reduce the passive filter size. As a result, cost reduction, simplicity, and small-size hardware are regarded as important issues. In addition, compared to the other control methods, deadbeat control is regarded as a potential strategy for improving dynamic performance. If accurate parameters and current stress optimization (CSO) methods are used in deadbeat control, this can yield optimal phase-shift duty ratios immediately for the next switching period. However, the current sensor is rather expensive and bulky. Besides, the extended state observer (ESO) [1] is one of the most well-known and effective methods to observe state variables. It requires the least amount of model information to suppress disturbance and provide an accurate observed value, resulting in the number of sensors being reduced.

This paper proposes a current-sensorless control to ensure robustness and excellent performance of the output voltage of DAB. In the proposed method, the ESO is used to observe the load current, while the additional proportional-integral (PI) controller is removed. The phase-shift duty ratios are directly determined as deadbeat properties under CSO by the Lagrange multiplier method (LMM). Furthermore, the effect of output capacitor mismatch is investigated to further demonstrate the effectiveness of the proposed method.

The paper is divided into four sections. Section II proposes current-sensorless control, which includes the ESO to observe the load current, and the CSO is adopted also. Section III displays the experimental results, which demonstrate the effectiveness of the proposed method. Finally, section IV concludes this paper.

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II. Proposed Current-Sensorless Control

Fig. 1 and Fig. 2 depict the DAB converter and its waveforms under dual phase-shift (DPS) modulation, respectively. The transferred power when $0 \le D_1 \le D_2 \le 1$ (Mode A) and $0 \le D_2 < D_1 \le 1$ (Mode B) are derived as (1) and (2), respectively, where n is the transformer turn, f is the switching frequency, and phase-shift duty ratios are denoted by D_1 and D_2.

$$P = \frac{nv_1v_2\left(D_2\left(1-D_2\right)-0.5D_1^2\right)}{2fL} \qquad \text{Mode A,} \tag{1}$$

$$P = \frac{nv_1v_2D_2\left(1-D_1-0.5D_2\right)}{2fL} \qquad \text{Mode B,} \tag{2}$$

From (1) and (2), the unified transferred power p_u and the unified current i_u are derived [2], [3]:

$$p_u = 4D_2\left(1-D_2\right)-2D_1^2 \qquad \text{Mode A,} \tag{3}$$

$$p_u = 4D_2\left(1-D_1\right)-2D_2^2 \qquad \text{Mode B,} \tag{4}$$

$$i_u = 2\left(M\left(1-D_1\right)+\left(D_1+2D_2-1\right)\right), \tag{5}$$

where $M = v_1/(nv_2)$ is the voltage conversion ratio.

In order to minimize the current stress, the Lagrange function is adopted as

$$F\left(\boldsymbol{d},\psi\right) = i_u\left(\boldsymbol{d}\right)+\varphi\left[p_u\left(\boldsymbol{d}\right)-p_{uref}\right], \tag{6}$$

where p_{uref} is the reference value of p_u, $\boldsymbol{d} = [D_1, D_2]^T$, and φ is the Lagrange multiplier.

By differentiating the Lagrange function F with respect to D_1, D_2, and φ to zero, D_1 is derived in case of $\tau < p_u \le 1$ as shown in (7), where $\tau = ((M+1)^2-4)/(2M^2)$.

$$D_1 = \left(\frac{\left(1-p_u\right)\left(M-1\right)^2}{2\left(M^2-2M+3\right)}\right)^{1/2}. \tag{7}$$

In the case of $0 \le p_u \le \tau$, D_1 is similarly given, and the unified transferred power p_u is calculated:

$$p_u = \frac{8fLi_2}{nv_1}. \tag{8}$$

By replacing the secondary current $i_s = P/v_2$ into the dynamic equation of C_2 when v_2 is controlled as $v_2 = v_{2ref}$, D_2 and i_s in Mode A can be derived:

$$D_2 = \frac{1}{2}-\left(\frac{1}{4}-\frac{1}{2}D_1^2-\frac{2LC_2}{T^2nv_1}\left(\frac{T}{C_2}i_2+v_{2ref}-v_2\right)\right)^{1/2}, \tag{9}$$

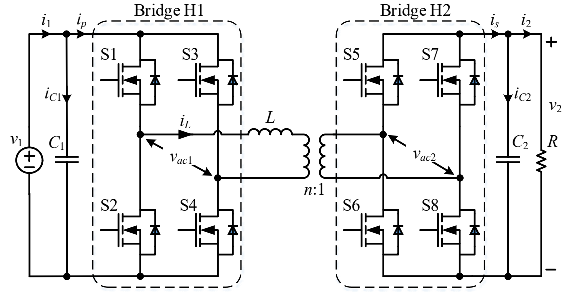

Fig. 1. DAB converter.

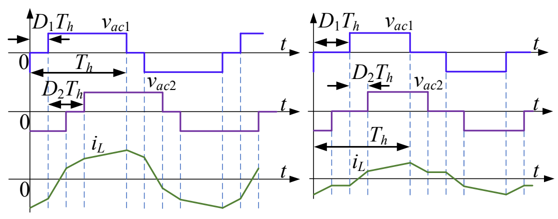
(a) Mode A (b) Mode B

Fig. 2. Waveforms of DAB converter under DPS.

$$i_s = \frac{C_2}{T}\left(v_{2ref} - v_2\right) + i_2, \tag{10}$$

where v_{2ref} is the reference value of v_2. Likewise, D_2 in Mode B is easily derived.

In order to provide a simplified load current observer, the dummy variable is defined as follows

$$f_i = -i_2, \tag{11}$$

In (11), f_i is supposed to be the unknown component that must be observed. As a result, the ESO can be constructed:

$$\frac{d\hat{v}_2}{dt} = \frac{1}{C_2} i_s + \frac{1}{C_2}\hat{f}_i + \delta_1\left(v_2 - \hat{v}_2\right), \tag{12}$$

$$\frac{d\hat{f}_i}{dt} = C_2\delta_2\left(v_2 - \hat{v}_2\right), \tag{13}$$

where $\hat{v}_2$ and $\hat{f}_i$ are the observed values of v_2 and f_i, respectively; δ_1 and δ_2 are the observer gains, that have a significant impact on the observer's performance and robustness. In this paper, δ_1 and δ_2 are chosen as $\delta_1 = 6 \times 10^3$ (rad/s) and $\delta_2 = 9 \times 10^6$ (rad/s).

By discretizing (12) and (13) using forward Euler approximation, (14) and (15) are derived:

$$\hat{v}_2[k] = \hat{v}_2[k-1] + \frac{T}{C_2} i_s[k-1] + \frac{T}{C_2}\hat{f}_i[k-1] + T\delta_1\left(v_2[k-1] - \hat{v}_2[k-1]\right), \tag{14}$$

$$\hat{f}_i[k] = \hat{f}_i[k-1] + TC_2\delta_2\left(v_2[k-1] - \hat{v}_2[k-1]\right). \tag{15}$$

The observed load current from (11), (14), and (15) is immediately sent into the CSO computation. Obviously, the proposed method is simple to implement because only two voltage sensors are required, this is due to i_2 can be obtained from (11). Besides, there is no further compensation controller is necessary.

III. Experimental Results

Fig. 3 shows experimental results when the load current steps up from 1.6 A to 3.2 A. In transient response, the proposed method has a peak value around 1.25 V and an observer settling time of 1.6 ms. Fig. 4 depicts the dynamic performance when C_2 varies by 20%. Experimental results demonstrate the adaptability and robustness of the proposed method when the steady-state error of v_2 is eliminated. The output voltage v_2 is very stiff when the observed current i_{2obs} is perfectly consistent with the values measured by the sensor i_{2sen}. It can be concluded that the proposed method has excellent steady-state and dynamic response performance even though it does not have any PI controller and the current sensor.

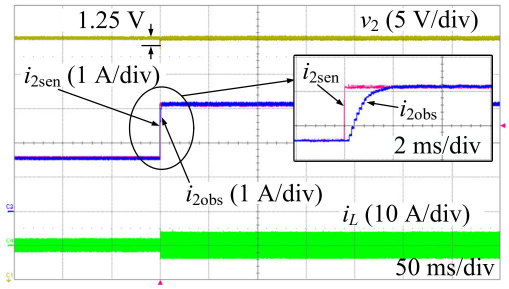

Fig. 3. Experimental results when the load current steps up from 1.6 A to 3.2 A

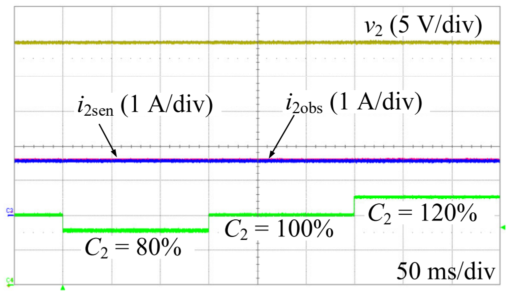

Fig. 4. Experimental results when C2 varies by 20%.

IV. Conclusions

This paper proposed a current-sensorless robust control for DPS modulated DAB converter while the current stress is minimized by LMM. The observed load current is injected directly into the predicted phase-shift duty ratios after each switching cycle. The current sensor is eliminated, and no additional PI controller is required, resulting in cost savings, system simplicity, and increased reliability. The proposed observer is robust to the output capacitor mismatches, resulting in good observer performance. Furthermore, the proposed method provides an excellent steady-state and dynamic performance.

References

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