Analytic Solutions for Full Operating Range Single-Side ZVS Modulation of Dual Active Bridge Converters

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Abstract—This paper presents analytic solutions for an optimal modulation scheme featuring low switching losses for a bidirectional single-phase dual active bridge (DAB) DC-DC converter used for charging high-voltage batteries of electric vehicles. The proposed modulation scheme facilitates zero-voltage switching (ZVS) for either the primary- or secondary-side full bridge of the DAB converter throughout the whole operating range while maintaining low conduction losses. The expressions are derived based on the current required for an ideal ZVS transition and enable for a direct computation of the respective optimal modulation parameters.

Index Terms—Battery charger, dual active bridge, zero-voltage switching

I. INTRODUCTION

The increasing spread of renewable energy sources, such as photovoltaic systems and fuel cells, as well as electrified vehicles rises the need for advanced power electronic systems. In particular, the importance of their efficiency and power density increases. A dual active bridge (DAB) converter is used as an isolated DC-DC stage in an electric vehicle battery charger. Despite the galvanic isolation it enables very high efficiencies [1] and additionally exhibits a bidirectional power flow capability [2]. The specification of the considered DAB converter is summarized in Tab. I.

Various modulation schemes for DAB converters have been presented in literature, featuring low conduction and/or switching losses. For power MOSFETs, the latter is commonly achieved by applying the principle of zero-voltage switching (ZVS) [3]. Additionally, the proposed modulation schemes enable for an analytic calculation of the respective optimal modulation parameters.

In [4], a modulation scheme is presented which aims at operating the DAB with lowest conduction losses and facilitates an analytical calculation of the respective optimal modulation

TABLE I
SPECIFICATION OF THE INVESTIGATED DC-DC
CONVERTER SYSTEM

Nominal input voltage	V_1	400 V
Output voltage range	V_2	300 V to 450 V
Rated power	P	5.5 kW
Switching frequency	f_s	100 kHz
Transformer turns ratio	n	1
Series inductor	L	23 μ H

parameters. [5] extends this approach by considering the switching frequency as an additional modulation parameter. However, ZVS cannot be achieved throughout substantial regions of the operating range with these modulation schemes.

In [7], a modulation scheme is proposed which ensures full operating range ZVS for all switching devices. Though, this approach relies on hardware modifications as it requires two additional commutation inductors to be inserted into the DAB circuit. These inductors beneficially influence the switching losses but in turn increase the RMS currents and hence the conduction losses.

A modulation scheme which inherently features ZVS for every switching device is presented in [6]. However, ZVS cannot be maintained throughout the whole operating range as it is lost at some transitions between different switching modes.

Finally, in [1], a modulation scheme is used, which enables for primary-side ZVS for all respective operating points. Simultaneously, secondary-side ZVS is ensured by one additional commutation inductor. However, this approach is based on a transformation applied to the DAB model which renders a direct calculation of the optimal modulation parameters infeasible.

Fig. 1: (a) DAB converter topology; (b) simplified lossless DAB model.

This paper derives analytic expressions which facilitate a direct computation of an optimal set of modulation parameters. The parameters are determined such that full operating range ZVS is achieved for the primary-side switches. Analogously to the approach presented in [1], a commutation inductor L_c is added to ensure ZVS for the secondary-side switches. For the sake of clarity, the determination of the optimal modulation parameters is explained solely for positive power transfer and primary-side ZVS, however, the presented equations are also valid for negative power transfer and/or secondary-side ZVS.

II. DAB MODEL AND MODULATION SCHEMES

The topology of the dual active bridge converter is shown in Fig. 1(a). Assuming that the components are lossless, the most simple DAB model shown in Fig. 1(b) can be obtained. The two full bridges can be thereby considered as voltage sources which apply the square-wave voltages $v_{HF1}(t)$ and $n v_{HF2}(t)$ to the high-frequency (HF) AC-link circuit. The latter contains a HF transformer with a turns ratio of $n : 1$, the series inductor L and the commutation inductor L_c , which only introduces reactive power and does not contribute to the power transfer. As mentioned before, the latter is added to achieve ZVS for the secondary-side switches. In discrete time intervals, the current $i_L(t)$ through L can be described by

$$i_L(t_i) = i_L(t_{i-1}) + \frac{v_{HF1}(t_i) - n v_{HF2}(t_i)}{L} (t_i - t_{i-1}). \quad (1)$$

With two full bridges, twelve different switching modes can be realized, each characterized by its individual sequence of rising and falling edges of the voltages v_{HF1} and v_{HF2} . For an efficient DAB operation with low conduction losses, only four different switching modes should be thereof applied [4]. A description of these four modes as well as exemplary waveforms can be found in [5].

To determine the power flow of the DAB converter, the combination of the three modulation parameters φ (phase shift between the rising edges of v_{HF1} and v_{HF2}), τ_1 and τ_2 (pulse widths of the voltages v_{HF1} and v_{HF2} , respectively) is used (cf. Fig. 2(a)). These parameters can be adjusted in the range $-0.5 \leq \varphi \leq 0.5$ and $0 \leq \tau \leq 0.5$.

Three different modulation schemes can be employed with the four switching modes mentioned before, which are:

Fig. 2: Exemplary waveforms for (a) the triangular current mode (TCM), (b) modified triangular current mode (mTCM), (c) optimal transition mode for buck mode operation and (d) optimal transition mode for boost-mode operation; v_{HF1} (—), $n v_{HF2}$ (—) and i_L (—).

- 1) **Modified Triangular Current Mode (mTCM)**: Used for low power levels $P \leq P_{\max}^{\text{mtcm}}$. For this modulation scheme, switching modes M_5 and M_6 are applied.
- 2) **(Modified) Optimal Transition Mode (OTM)**: OTM is employed for medium powers, i.e. $P_{\max}^{\text{mtcm}} < P \leq P_{\max}^{\text{otm}}$. This modulation scheme can be realized by switching mode M_1^+ for power levels $P > 0$ and M_1^- for $P < 0$.
- 3) **Phase Shift Mode (PSM)**: Used for high power levels $P_{\max}^{\text{otm}} < P \leq P_{\max}$. For this modulation scheme, switching modes M_1^+ and M_1^- are used as well. However, $\tau_1 = \tau_2 = 0.5$ always applies for PSM, i.e. the power is solely controlled by the phase-shift φ .

III. ANALYTICAL DETERMINATION OF OPTIMAL MODULATION PARAMETERS FOR PRIMARY-SIDE ZVS

To facilitate an analytic calculation of the optimal control variables $\tau_{1,\text{opt}}$, $\tau_{2,\text{opt}}$ and φ_{opt} , analytical expressions for the

currents $i_L(t_i)$ at the respective switching instants and for the transferred power

$$P = \frac{2}{T_s} \int_{t_0}^{t_0+T_s/2} i_L(t) v_{HF1}(t) dt \quad (2)$$

are needed. The respective switching instants t_i needed for the calculation of the currents and for the evaluation of (2) can be expressed by the modulation parameters introduced before. Expressions for every switching instant as well as solutions for the phase-shift φ for all switching modes were proposed in [5] and have been obtained by reformulating (2) with respect to φ .

A. ZVS Requirements

In order to avoid losses caused by hard-switching transitions, the principle of ZVS is usually applied for MOSFETs. Ideal soft-switching requires an impressed current of an inductive component which charges/discharges the MOSFETs output capacitance C_{oss} within a bridge leg, such as S_1 and S_2 in Fig. 1(a). Two conditions are required for a complete zero-voltage transition [3]:

- 1) If S_1 is turned off and S_2 is turned on, which leads to the charging of $C_{oss,1}$ and the discharging of $C_{oss,2}$, the current must flow out of the bridge leg, i.e. $i_{HF1} > 0$. For a transition from S_2 to S_1 $i_{HF1} < 0$, is required.
- 2) Additionally, the current at the respective switching instant must be large enough, i.e. it must provide enough charge Q_{oss} to fully charge/discharge the MOSFET output capacitance. This can be expressed by the energy balance

$$\frac{1}{2} L I_{zvs}^2 \geq Q_{oss} (V_{DC}) \cdot V_{DC}. \quad (3)$$

If an additional parasitic layout capacitance C_{par} is considered, $\frac{1}{2} C_{par} V_{DC}^2$ has to be added to the right-hand side of (3). Finally, the current required for a complete ZVS transition can be obtained by reformulating (3) with respect to I_{zvs} .

The optimal modulation parameters are determined in such a way, that the energy balance of (3) is fulfilled for every switching instant of v_{HF1} , i.e. primary-side ZVS is achieved throughout the whole operating range. For the calculation of the optimal modulation parameters, the following currents

$$\begin{aligned} I_1 &= -I_{zvs} \\ I_2 &= I_{zvs} \end{aligned} \quad (4)$$

are defined. Note that for secondary-side ZVS the currents must be interchanged.

B. Closed-form solutions for optimal τ_1 , τ_2 and φ

In order to reduce the number of equations needed to fully describe the derivation of the optimal modulation parameters, a transformation according to

$$\begin{aligned} &(V_A, V_B, \tau_A, \tau_B, I_A, I_B)^T \\ &= \begin{cases} (V_1, nV_2, \tau_1, \tau_2, I_1, I_2)^T & \forall V_1 \leq nV_2 \\ (nV_2, V_1, \tau_2, \tau_1, I_2, I_1)^T & \forall V_1 > nV_2 \end{cases} \end{aligned} \quad (5)$$

is applied. Note that by these transformation, voltages, currents and pulse widths are interchanged.

1) *Modified Triangular Current Mode*: As mentioned before, mTCM is used for low power levels. In contrast to normal TCM, τ_A and τ_B are determined such, that the current i_L is offset during the freewheeling intervals, i.e. $|i_L(t)| = I_{zvs} > 0$ for $v_{HF1}(t) = v_{HF2}(t) = 0$ (cf. Fig. 2(a) and Fig. 2(b)). By applying (5) to the currents at the switching instants t_0 and t_2 of mode M_6 the following equations can be obtained

$$i_L(t_0) = I_B = \frac{V_B \tau_B^2 (V_A - V_B) + f_s L P}{2 f_s L V_B \tau_B}, \quad (6)$$

$$i_L(t_2) = I_A = \frac{V_B \tau_B - V_A \tau_A}{2 f_s L}. \quad (7)$$

For mode M_5 the expressions on the right-hand sides of (6) and (7) must be interchanged. To calculate the respective optimal pulse widths, these equations must be solved for τ_A and τ_B . However, the expression for τ_B is only valid for $V_A \neq V_B$, i.e. $V_1 \neq nV_2$. An alternative solution can be obtained by applying $V_A = V_B$ to (6) and reformulating with respect to τ_B . Finally, the expressions needed to calculate the respective optimal modulation parameters are

$$\begin{aligned} \forall |P| &\leq P_{\max}^{\text{mtcm}} \\ \tau_B &= \begin{cases} \sqrt{\left(\frac{f_s L I_B}{V_A - V_B}\right)^2 - \frac{f_s L |P|}{V_B (V_A - V_B)}} \\ + \frac{f_s L I_B}{V_A - V_B} & \forall V_1 \neq nV_2 \\ \frac{|P|}{2 V_B |I_B|} & \forall V_1 = nV_2 \end{cases} \\ \tau_A &= \frac{V_B \tau_B - 2 f_s L I_A}{V_A} \\ \varphi &= \text{sgn}(V_1 - nV_2) \frac{\tau_B - \tau_A}{2} + \frac{f_s L P}{2 V_A V_B \tau_B}. \end{aligned} \quad (8)$$

As mentioned in [5], the duration of the freewheeling interval decreases with increasing power P until it becomes zero at the highest possible output power $P_{\max}^{\text{mtcm}}$. An expression for $P_{\max}^{\text{mtcm}}$ can thus be found by inserting τ_B into τ_A and solving $\tau_A = 0.5$ for P . This yields

$$\begin{aligned} P_{\max}^{\text{mtcm}} &= (V_A + 4 f_s L I_A) \\ &\cdot \left(I_B + \frac{(V_A + 4 f_s L I_A)(V_B - V_A)}{4 f_s L V_B} \right). \end{aligned} \quad (9)$$

In contrast to standard TCM, switching modes M_5 and M_6 are not only applied for $V_1 < nV_2$ and $V_1 > nV_2$, respectively. Corresponding operating points are indicated by the shaded area in Fig. 5(a). If (8) and (9) are evaluated for the operating points mentioned before, incorrect values for τ_A , τ_B and φ result and $P_{\max}^{\text{mtcm}}$ becomes negative, which is unfeasible.

However, appropriate values for τ_A , τ_B , φ and $P_{\max}^{\text{mtcm}}$ can be obtained. This results in the modified equations

$$\forall |P| \leq P_{\max}^{\text{mtcm}} \quad \begin{cases} \tau_A = \begin{cases} -\sqrt{\left(\frac{f_s L I_A}{V_B - V_A}\right)^2 - \frac{f_s L |P|}{V_A(V_B - V_A)}} \\ + \frac{f_s L I_A}{V_B - V_A} \quad \forall V_1 \neq n V_2 \end{cases} \\ \tau_A = \frac{V_A \tau_A - 2 f_s L I_B}{V_B} \\ \varphi = \text{sgn}(V_1 - n V_2) \frac{\tau_B - \tau_A}{2} + \frac{f_s L P}{2 V_A V_B \tau_A}. \end{cases} \quad (10)$$

for the calculation respective modulation parameters. For according operating points the highest possible power for mTCM can be obtained from

$$P_{\max}^{\text{mtcm}} = (V_A + 4 f_s L I_B) \cdot \left(I_A + \frac{(V_B + 4 f_s L I_B)(V_A - V_B)}{4 f_s L V_A} \right). \quad (11)$$

It should be noted that for the special case $I_1 = I_2 = 0$, i.e. standard TCM is applied, the equations of (8), (10), (9) and (11) transform into the ones presented in [5].

2) *Optimal Transition Mode*: OTM is employed for medium power levels and is realized with switching mode M_1^+ . Exemplary waveforms for OTM in buck and boost operation are shown in Fig. 2(c) and Fig. 2(d), respectively. As mentioned before, $\tau_A = 0.5$ always applies for $P \geq P_{\max}^{\text{mtcm}}$. A closed-form expression for the calculation of the optimal pulse width τ_B regarding minimum conduction losses has been derived in [5] and is given by

$$\begin{aligned} e_0 &= V_A^2 + V_B^2 \\ e_1 &= (4 f_s L P e_0)^2 (8 V_A^2 - V_B^2) + 6 V_A^6 V_B^2 (e_0 + 3 V_A^2) \\ &\quad - 3 f_s L |P| V_A^3 V_B (32 V_A^4 + 28 V_A^2 V_B^2 + 5 V_B^4) \\ e_2 &= 6 \sqrt{3} V_A V_B^2 \left[f_s L |P| \left[V_A^7 V_B^5 - 64 (f_s L |P| e_0)^3 \right. \right. \\ &\quad \left. \left. + V_A V_B f_s L |P| (4 e_0 - 3 V_B^2) \right. \right. \\ &\quad \left. \left. \cdot \left(8 f_s L |P| e_0^2 - V_A^3 V_B (e_0 + \frac{9}{4} V_B^2) \right) \right] \right]^{\frac{1}{2}} \\ e_3 &= f_s L |P| \left(f_s L |P| e_0 - \frac{V_A^3 V_B}{2} \right) \\ &\quad \cdot \left(e_0 - \frac{3}{4} V_B^2 \right) + \left(\frac{V_A^3 V_B}{4} \right)^2 \\ e_4 &= 4 \left[\frac{4 f_s L |P| e_0^2}{V_A V_B} - e_0^2 + V_B^2 \left(e_0 - \frac{3}{4} V_B^2 \right) \right] \end{aligned}$$

Fig. 3: Trajectories of the optimal control parameters τ_1 , τ_2 and φ for (a) buck-mode operation, $V_1 = V_B = 400$ V, $n V_2 = V_A = 320$ V, and (b) boost-mode operation, $V_1 = V_A = 400$ V, $n V_2 = V_B = 450$ V. The trajectories of τ_B^{otm} , τ_B^{bu} and τ_B^{bo} , which are required for the determination of τ_B for mOTM intervals are additionally shown.

$$\begin{aligned} e_5 &= \sqrt[3]{f_s L |P| (e_2 - e_1) + 2 V_A^9 V_B^3} \\ e_6 &= \sqrt[3]{2 e_0 \frac{32 e_3 + e_5^2 \sqrt{2}}{V_A V_B e_5}} \\ \tau_B^{\text{otm}} &= \frac{1}{4} \left(\frac{\sqrt{2 e_6 - e_4}}{\sqrt{3} e_0} + \frac{V_A^2}{e_0} + 1 \right) - \left[-\frac{e_6 + e_4}{24 e_0^2} \right. \\ &\quad \left. + \frac{\sqrt{3} V_B (V_A V_B^3 (e_0 + V_A^2) - 8 f_s L |P| e_0^2)}{8 e_0^2 V_A \sqrt{2 e_6 - e_4}} \right]^{\frac{1}{2}}. \end{aligned} \quad (12)$$

An expression for the calculation of the maximum transferable power

$$P_{\max}^{\text{otm}} = \frac{V_B \left(\left(V_A^2 - V_B^2 + V_B \sqrt{V_B^2 - V_A^2} \right) \right)}{4 f_s L V_A} \quad (13)$$

has been presented in [5] as well. However, primary-side ZVS cannot be achieved inherently in OTM for all respective operating points. Especially for power levels slightly above

$P_{\max}^{\text{mtcm}}$ ZVS is lost if standard OTM is used. Thus, a different modulation scheme must be applied for these operating points.

3) *Modified Optimal Transition Mode*: For the operating points mentioned before, a modified version of the optimal transition mode named mOTM is proposed. To guarantee primary-side ZVS for OTM, $i_L(t_0) \leq -I_{\text{ZVS}}$ must be fulfilled. Distinguishing whether the DAB operates in buck or in boost mode, $i_L(t_0) = -I_{\text{ZVS}}$ must be either solved for τ_1 or τ_2 , respectively. By applying (5) to the solutions, one obtains

$$\tau_B^{\text{bu}} = \frac{1}{2 \left((V_A + V_B)^2 + V_A^2 \right)} \left[\left[2V_A^3(V_A + V_B) - 8f_s L V_A^2 \right. \right. \\ \left. \left. \cdot \left(|P| \frac{(V_A + V_B)^2 + V_A^2}{V_A V_B} + I_B (2f_s L I_B + V_B) \right) \right]^{\frac{1}{2}} \right. \\ \left. + V_A (2V_A + V_B) - 4f_s L I_B (V_A + V_B) \right], \quad (14)$$

$$\tau_B^{\text{bo}} = \frac{1}{2} - \sqrt{\frac{1}{4} - \frac{2f_s L |P|}{V_A V_B} - \left(\frac{4f_s L I_A + V_A}{2V_B} \right)^2}. \quad (15)$$

It can be seen, that either I_B or I_A is exclusively considered for the calculation of τ_B , which always corresponds to I_1 . For secondary-side ZVS, the expressions for τ_B^{bu} and τ_B^{bo} must be interchanged, i.e. I_2 is always considered.

Depending on whether the DAB operates in buck or in boost mode, a value for $P_{\text{otm},\min}$ could be obtained by solving $\tau_B^{\text{otm}} = \tau_B^{\text{bu}}$ or $\tau_B^{\text{otm}} = \tau_B^{\text{bo}}$ with respect to P . However, no closed-form expression has been found due to the complexity of the result. Thus, a straightforward approach is presented to determine the respective pulse width τ_B which facilitates primary-side ZVS throughout the whole OTM. This approach is shown in Fig. 3. It is based on a comparison of the trajectories of τ_B^{otm} and τ_B^{bu} and τ_B^{bo} , respectively. The closed-form expressions needed to calculate the respective modulation parameters for OTM are finally given by

$$\forall P_{\max}^{\text{mtcm}} < |P| \leq P_{\max}^{\text{otm}} \begin{cases} \tau_A = \frac{1}{2} \\ \tau_B = \begin{cases} \min \left(\max \left(\tau_B^{\text{otm}}, \tau_B^{\text{bu}} \right), \frac{1}{2} \right) & \forall V_1 > n V_2 \\ \min \left(\min \left(\tau_B^{\text{otm}}, \tau_B^{\text{bo}} \right), \frac{1}{2} \right) & \forall V_1 \leq n V_2 \end{cases} \\ \varphi = \text{sgn}(n V_2 - V_1) \frac{\tau_A - \tau_B}{2} + \text{sgn}(P) \left(\frac{1}{4} - \frac{1}{2} \right. \\ \left. \cdot \sqrt{\frac{1}{4} - \frac{2f_s L |P|}{V_A V_B} - \left(\tau_A - \frac{1}{2} \right)^2 - \left(\tau_B - \frac{1}{2} \right)^2} \right). \end{cases} \quad (16)$$

4) *Phase Shift Mode*: PSM is used at high power levels and inherently features ZVS. The respective modulation parameters can be calculated as follows

$$\forall P_{\max}^{\text{otm}} < |P| \leq P_{\max} \begin{cases} \tau_A = \tau_B = \frac{1}{2} \\ \varphi = \text{sgn}(P) \frac{1}{4} \left(1 - \sqrt{1 - \frac{8f_s L |P|}{V_A V_B}} \right). \end{cases} \quad (17)$$

Finally, the procedure to analytically determine the respective optimal modulation parameters for given n , L , V_1 , V_2 and P is illustrated in Fig. 4 as flow chart.

Fig. 4: Procedure to analytically determine optimal control parameters $\tau_{1,\text{opt}}$, $\tau_{2,\text{opt}}$ and φ_{opt} for given converter parameters $\{n, L\}$ and for a given operating point $\{V_1, V_2, P\}$.

IV. RESULTS

The proposed modulation scheme is verified by means of a simulation employing GaN-Systems GS66516B HEMTs as semiconductor switches and an additional parasitic layout capacitance of $C_{\text{par}} = 100 \text{ pF}$. Evaluating (3) for a constant voltage of $V_1 = 400 \text{ V}$ and an inductance value of $L = 23 \text{ }\mu\text{H}$

Fig. 5: (a) DAB operating modes employed for the proposed modulation scheme; contrary to standard TCM (dotted lines), M_5 is also applied for $V_1 > n V_2$. (b) switching currents obtained with the proposed modulation scheme which enable for full operating range primary-side ZVS.

points out, that a minimum current of $|I_{zvs}| = 2.4$ A is required for a complete ZVS transition. The different modes generated by the proposed modulation scheme are presented in Fig. 5(a). As previously mentioned it becomes clear, that switching mode M_6 is not only applied for operating points $V_1 \geq n V_2$ for mTCM as it the case for standard TCM. Respective operating points are indicated by the shaded area in Fig. 5(a) and require the modified equations for the calculation of the optimal control variables.

Fig. 5(b) shows the current values for the primary-side switching instants obtained by the proposed modulation scheme. It becomes obvious that primary-side ZVS can be achieved throughout the entire operating range. The minimum current of $I_{zvs} = -2.4$ A is maintained for all operating points covered by the modulation schemes mTCM and mOTM, which proves the effectiveness of the proposed modulation strategy. For all OTM and PSM operating points this current at the respective switching instants is exceeded.

It should be mentioned that the trajectories of the modulation parameters become discontinuous at the border between mode M_5 and mOTM for $V_1 > n V_2$ (c.f. Fig. 5(a)). However, this exhibits only a minor drawback for the application at hand as a charger for high-voltage batteries of electric vehicles

usually operates at medium to high power levels. If seamless transitions between the above mentioned switching modes are required and ZVS is not mandatory for low-power operating points, the requested switching currents in the respective regions of the operating range could be reduced e.g. according to

$$I_{zvs}(P) = I_{zvs} \min\left(\frac{|P|}{1000}, 1\right). \quad (18)$$

V. CONCLUSION

In this paper, a procedure to analytically calculate the optimal modulation parameters for a dual active bridge converter has been presented, which inherently facilitates either primary or secondary-side ZVS in the full operating range. The modulation parameters are determined in such a way that the impressed current at the respective switching instants is high enough to guarantee an ideal ZVS transition. ZVS for the opposite bridge can still be achieved by adding an additional commutation inductor. Unlike existing approaches, analytical expressions have been derived which enable for a direct calculation of the respective modulation parameters without the need of preceding transformations applied to the converter model and/or additional changes in hardware.

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