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 simultaneously 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. In an electric vehicle battery charger, a dual active bridge (DAB) converter is used as an isolated DC-DC stage. Despite the galvanic isolation it enables very high efficiencies [1] and additionally exhibits a bidirectional power flow capability [2].

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 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. However, 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.

This paper derives analytic expressions which facilitate a direct computation of an optimal set of modulation parameters. The parameters are determined in a way 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 thereby be considered as voltage sources which apply the square-wave voltages $v_{HF1}(t)$ and $v_{HF2}(t)$ 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

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

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 of them should be 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 $\varphi$ (phase shift between the rising edges of $v_{HF1}$ and $v_{HF2}$), $\tau_1$ and $\tau_2$ (pulse widths of the voltages $v_{HF1}$ and $v_{HF2}$, respectively) is used (cf. Fig. 2(d)). 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:

1) Modified Triangular Current Mode (mTCM): Used for low power levels $P \leq P_{\text{max}}^{\text{mTCM}}$. For this modulation scheme, switching modes named $M_5$ and $M_6$ are applied.

2) (Modified) Optimal Transition Mode (OTM): OTM is employed for medium powers, i.e., $P_{\text{max}}^{\text{mTCM}} < P \leq P_{\text{max}}^{\text{otm}}$. This modulation scheme can be realized by switching mode $M_2^+$ for power levels $P > 0$ and $M_2^-$ for $P < 0$.

3) Phase Shift Mode (PSM): Used for high power levels $P_{\text{max}}^{\text{PSM}} < P \leq P_{\text{max}}^{\text{PSM}}$. For this modulation scheme, switching modes $M_1^+$ of $M_1^-$ 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 $\varphi$.

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 $\varphi_{\text{opt}}$, analytical expressions for the currents $i_L(t_i)$ at the respective switching instants $t_i$ and for the transferred power

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

are required. 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 $\varphi$ for all switching modes were proposed in [5] and have been obtained by reformulating (2) with respect to $\varphi$. 

![Fig. 1: (a) DAB converter topology; (b) simplified lossless DAB model.](image1.png)

![Fig. 2: Exemplary waveforms for (a) the modified triangular current mode (mTCM) for $M_5$ and (b) $M_6$, (c) optimal transition mode for buck-mode operation and (d) optimal transition mode for boost-mode operation; $v_{HF1}$ (—), $n v_{HF2}$ (—) and $i_L$ (—).](image2.png)
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 MOSFET 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

\[
\frac{1}{2} L f_{zvs}^2 \geq Q_{oss}(V_{DC}) \cdot V_{DC}. \tag{3}
\]

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

\[
I_1 = -I_{zvs}, \quad I_2 = I_{zvs} \tag{4}
\]

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

B. Closed-form solutions for optimal $\tau_1$, $\tau_2$ and $\varphi$

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

\[
(V_A, V_B, \tau_B, I_A, I_B)^\top = \begin{cases} (V_1, n V_2, \tau_1, \tau_2, I_1, I_2)^\top & \forall \ V_1 \leq n V_2 \\ (n V_2, V_1, \tau_1, \tau_2, I_1, I_2)^\top & \forall \ V_1 > n V_2 \end{cases} \tag{5}
\]

is applied. Note that by this transformation, voltages, currents and pulse widths are interchanged depending on the ratio between $V_1$ and $n V_2$. With this, $V_A \leq V_B$ always applies.

1) Modified Triangular Current Mode: As mentioned before, mTCM is used for low power levels. In contrast to normal TCM, $\tau_A$ and $\tau_B$ 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

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

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

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 $\tau_A$ and $\tau_B$. However, the expression for $\tau_B$ is only valid for $V_A \neq V_B$, i.e. $V_1 \neq n V_2$. An alternative solution can be obtained by applying $V_A = V_B$ to (6) and reformulating with respect to $\tau_B$. Finally, the expressions needed to calculate the respective optimal modulation parameters are

\[
\forall \ |P| \leq P_{\max}^{mtcm}
\]

\[
\begin{align*}
\tau_B &= \left\{ \begin{array}{ll}
\left( \frac{f_s L I_B}{V_A - V_B} \right)^2 - \frac{f_s L |P|}{V_B(V_A - V_B)} + \frac{f_s L I_B}{V_A - V_B} & \forall \ V_1 \neq n V_2 \\
\frac{f_s L I_B}{2 V_B |P|} & \forall \ V_1 = n V_2
\end{array} \right. \\
\tau_A &= \frac{V_B \tau_B - 2 f_s L I_A}{V_A} \\
\varphi &= \text{sgn}(V_1 - n V_2) \frac{\tau_B - \tau_A}{2} + \frac{f_s L P}{2V_A V_B \tau_B} & \forall \ V_1 \neq n V_2 \\
\varphi &= -\frac{\tau_B - \tau_A}{2} + \frac{f_s L P}{2V_A V_B \tau_B} & \forall \ V_1 = n V_2.
\end{align*} \tag{8}
\]

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 $P_{\max}^{mtcm}$. An expression for $P_{\max}^{mTCM}$ can thus be found by inserting $\tau_B$ into $\tau_A$ and solving $\tau_A = 0.5$ for $P$. This yields

\[
P_{\max}^{mtcm} = (V_A + 4 f_s L I_A)^{\top} \left( I_B + \frac{(V_A + 4 f_s L I_A)(V_B - V_A)}{4 f_s L V_B} \right). \tag{9}
\]

In contrast to standard TCM, switching modes $M_5$ and $M_6$ are not only applied for $V_1 < n V_2$ and $V_1 > n V_2$, 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 $\tau_A$, $\tau_B$ and $\varphi$ result and $P_{\max}^{mTCM}$ becomes negative, which is unfeasible. However, appropriate values for $\tau_A$, $\tau_B$, $\varphi$ and $P_{\max}^{mTCM}$ can be obtained by applying (5) to $i_L(t_0)$ and $i_L(t_2)$ of mode $M_5$ and solving...
\( i_L(t_2) = I_A \) and \( i_L(t_0) = I_B \) for \( \tau_A \) and \( \tau_B \), respectively. This results in the modified equations

\[
\forall \ |P| \leq P_{\text{max}}^{\text{mtcm}} \quad \tau_{\text{A}} = \left\{ \begin{aligned}
\frac{-\sqrt{\left( f_s L I_A \right)^2 - f_s L |P| V_A (V_B - V_A)}}{V_A (V_B - V_A)} \\
+ \frac{f_s L I_A}{V_B - V_A}
\end{aligned} \right\} \quad \forall \ V_1 \neq n V_2
\]

\[
\tau_{\text{B}} = \frac{V_A \tau_{\text{A}} - 2 f_s L I_B}{V_B}
\]

\[
\varphi = \text{sgn}(V_1 - n V_2) \frac{\tau_{\text{B}} - \tau_{\text{A}}}{2} + \frac{f_s L P}{2V_A V_B \tau_{\text{A}}}.
\]

which facilitate the calculation of the respective modulation parameters for the operating points mentioned before. For according operating points, the highest possible power for mTCM can be obtained from

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

It should be noted that for the special case \( I_1 = I_2 = 0 \), i.e. standard TCM is applied, the equations of (8), (9), (10) 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, \( \tau_{\text{A}} = 0.5 \) always applies for \( P \geq P_{\text{max}}^{\text{mtcm}} \). A closed-form expression for the calculation of the optimal pulse width \( \tau_{\text{B}} \) regarding minimum conduction losses has been derived in [5] and is given by

\[
e_0 = V_A^2 + V_B^2
\]

\[
e_1 = (4 f_s L P e_0)^2 \left( 8V_A^2 - V_B^2 \right) + 6V_A^6 V_B^2 \left( e_0 + 3V_A^2 \right)
\]

\[
- 3 f_s L |P| V_A^3 V_B^3 \left( 32V_A^4 + 28V_A^2 V_B^2 + 5V_B^4 \right)
\]

\[
e_2 = 6 \sqrt{3} V_A V_B^2 f_s L |P| \left[ V_A^7 V_B^5 - 64 \left( f_s L |P| \right) e_0 \right]^3
\]

\[
+ V_A V_B f_s L |P| \left( 4e_0 - 3V_B^2 \right)
\]

\[
\cdot \left( 8 f_s L |P| e_0^2 - V_A^3 V_B \left( e_0 + \frac{9}{4} V_B^2 \right) \right) \right]^{\frac{1}{2}}
\]

\[
e_3 = f_s L |P| \left( f_s L |P| e_0 - \frac{V_A^3 V_B}{2} \right)
\]

\[
\cdot \left( e_0 - \frac{3}{4} V_B^2 \right) + \left( \frac{V_A^3 V_B}{4} \right)^2
\]

\[
e_4 = 4 \left[ \frac{4 f_s L |P| e_0^2}{V_A V_B} - e_0^2 + V_B^2 \left( e_0 - \frac{3}{4} V_B^2 \right) \right]
\]

An expression for the calculation of the maximum transferable power

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

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
\[ P_{\text{max}} \] 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 \( \text{mOTM} \) is proposed. To guarantee primary-side ZVS for OTM, \( i_L(t_0) \leq -i_{zvs} \) must be fulfilled. Distinguishing whether the DAB operates in buck or in boost mode, \( i_L(t_0) = -i_{zvs} \) must be either solved for \( \tau_1 \) or \( \tau_2 \), respectively. By applying (5) to the solutions, one obtains

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

\[ \tau_{\text{bo}} = \frac{1}{2} - \sqrt{\frac{1}{4} - \frac{2 f_s L \left| P \right|}{V_A V_B} \left( \frac{4 f_s L I_A + V_A}{2 V_B} \right)^2}. \]

It can be seen, that either \( I_B \) or \( I_A \) is exclusively considered for the calculation of \( \tau_B \), which always corresponds to \( I_2 \). For secondary-side ZVS, the expressions for \( \tau_{\text{bu}} \) and \( \tau_{\text{bo}} \) 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 \( \tau_{\text{otm}} = \tau_{\text{bu}} \) or \( \tau_{\text{otm}} = \tau_{\text{bo}} \) 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 \( \tau_B \) 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 \( \tau_B \), \( \tau_{\text{bu}} \) and \( \tau_{\text{bo}} \), respectively. The expressions needed to calculate the respective modulation parameters for OTM are finally given by

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

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

\[
\forall \quad P_{\text{max}} < \left| P \right| \leq P_{\text{max}}
\begin{align*}
\tau_A = & \frac{1}{2} \\
\tau_B = & \frac{1}{2} \\
\varphi = & \text{sgn}(P) \left( 1 - \frac{1}{4} \right) \left( 1 - \frac{8 f_s L \left| P \right|}{V_A V_B} \right). \tag{17}
\end{align*}
\]

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.

IV. RESULTS

The proposed modulation scheme is verified by means of a simulation employing GaN-Systems GS66516B HEMTs as semiconductor switches and considering an additional parasitic layout capacitance of \( C_{\text{par}} = 100 \text{pF} \). The specification of the considered DAB converter is summarized below in Tab. I.
Evaluating (3) for a constant voltage of \( V_1 = 400 \text{ V} \) and an inductance value of \( L = 23 \mu \text{H} \) points out that a minimum current of \( |I_{zvs}| = 2.4 \text{ A} \) 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 \( M_5 \) is not only applied for operating points \( V_1 < nV_2 \) for mTCM as it is the case for standard TCM. Respective operating points are indicated by the shaded area in Fig. 5(a) and require the modified equations (10) and (11) for the calculation of the optimal control variables and the maximum transferable output power.

Fig. 5(b) shows the current values for the primary-side switching instants obtained by the proposed modulation scheme. It was stated that primary-side ZVS can be achieved throughout the entire operating range. The minimum current of \( I_{zvs} = -2.4 \text{ A} \) 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 \( V_1 > nV_2 \) (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 switching modes mentioned above 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 lowered e.g. according to

\[
I_{zvs}(P) = I_{zvs} \min \left( \frac{|P|}{1000}, 1 \right). \tag{18}
\]

V. Conclusion

In this paper, a procedure to analytically calculate the optimal modulation parameters for a dual active bridge converter was 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 eliminating the need of preceding transformations applied to the converter model and/or additional changes in hardware.

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 \mu H |

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