Generalized modeling and optimization of a bidirectional dual active bridge DC-DC converter including frequency variation

Felix Jauch and Jürgen Biela, Member, IEEE,

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Abstract—The paper presents a novel modeling approach of the power flow in a bidirectional dual active bridge DC-DC converter. By using basic superposition principles, the mathematical distinction of cases is avoided in the modeling process of high-frequency transformer currents for different types of modulation. The generalized model is used in an optimization of converter losses of a 3.3 kW electric vehicle battery charger with an input voltage of 400 V and a battery voltage range of 280 V to 420 V. Besides the commonly used control variables such as phase-shift and clamping intervals, also the variation of switching frequency is considered in the optimization process. The optimal modulation including frequency variation leads to an increase of converter efficiency up to 8.6 % using IGBTs and 17.8 % using MOSFETs in the most critical point compared to phase-shift modulation at fixed switching frequency.

Index Terms—DC-DC Converter, Dual Active Bridge, Frequency Variation, Modeling, Optimization

I. INTRODUCTION

During the last few decades the environmental impact of petroleum-based transportation infrastructure gained more and more significance. Fossil fuel-powered vehicles lead to large emissions of CO2 and other pollutions. In order to reduce those impacts, electric vehicles will play an important role in our future transportation infrastructure as the use of renewable energy sources is constantly increasing.

For charging the batteries of electric vehicles or storage systems in general, suitable power electronic systems are necessary. Usually, a basic two-stage approach comprising a boost Power Factor Correction (PFC) rectifier and a subsequent high-frequency isolated DC-DC converter is used for a charging system connected to the low-voltage AC grid. Additionally, for implementing Vehicle-2-Grid (V2G) concepts, the converter systems feature bidirectional power flow capability. Suitable DC-DC converters comprise the Dual Half-Bridge (DHB) [1], Dual Active (Full-)Bridge (DAB) [2] or resonant DC-DC converters [3].

For the DAB, several modulation methods like phase-shift modulation [2], triangular and trapezoidal current mode modulation [4] have been investigated. Further adjustments of these methods have been presented in [5]. Usually, for each of these modulation methods, piecewise linear equations are used to describe the currents where several mathematical cases depending on the control variables have to be distinguished. A general optimization of converter losses becomes relatively complex and demands high computational power.

In this paper, a novel generalized modeling approach of the power flow is presented to include all possible modulation methods at once. The general power flow equation is then used in an optimization procedure to find the optimal modulation, which leads to highest converter efficiency over the whole operating range. Besides the commonly used control variables like phase-shifts and clamping intervals, also the switching frequency is considered to control a DAB [6].

First, in section II the DAB converter topology with its modulation methods is introduced. The mathematical derivation of a novel generalized power flow equation depending on the control variables is presented in section III. Then, section IV shows the optimization of converter losses of a DAB prototype system. Finally, the optimal modulation including frequency variation is compared to conventional modulation methods with respect to converter efficiency.

II. TOPOLOGY AND MODULATION

In the following, first the topology of the DAB DC-DC converter with its operating principle is shortly explained. Afterwards, commonly used modulation methods with their limits are summarized. These are the phase-shift modulation as well as the trapezoidal and the triangular current mode modulation.

A. Dual Active Bridge DC-DC Converter

Fig. 1 shows the converter topology of a DAB DC-DC converter. The converter consists of a primary and a secondary full-bridge with unidirectional switches. The two full-bridges are connected to the windings of a two-winding transformer and generate high-frequency (HF) square-wave voltages with amplitudes of the DC port voltages $V_1, V_2$. The converter is operated by phase-shift control where the control variables are
B. Phase-Shift Modulation

In phase-shift modulation, square-wave voltages without clamping intervals are applied to the transformer windings as shown in Fig. 2. The power transferred from primary to secondary side is controlled by the phase-shift $\phi_2$ between the two voltages $v_1, n v_2$ and given by

$$P_{12} = -\frac{V_1 n V_2 \phi_2 (\pi - |\phi_2|)}{\pi \omega_s L_\sigma}$$  \hspace{1cm} (1)$$

with $\phi_2 \in [-\pi, \pi]$. $V_1, V_2$ are the input and output voltage, $\omega_s = 2\pi / T_s$ the angular switching frequency, $n = N_1 / N_2$ the turns ratio and $L_\sigma$ the leakage inductance of the transformer.

The maximum transferable power is

$$P_{12,\text{max}} = \pm \frac{\pi V_1 n V_2}{4 \omega_s L_\sigma}. \hspace{1cm} (2)$$

The soft-switching range of the modulation where Zero-Voltage-Switching (ZVS) can be achieved is strongly dependent on the voltage ratio $V_1 / n V_2$ as well as the power level $P_{12}$ [2]. Especially for voltage ratios $V_1 / n V_2 \ll 1$ and $V_1 / n V_2 \gg 1$ at low loads, ZVS cannot be maintained. Disadvantages like limited soft-switching range and high RMS transformer currents can be overcome by using the trapezoidal current mode modulation explained in the next section.

C. Trapezoidal Current Mode Modulation

In trapezoidal current mode modulation, square-wave voltages with clamping intervals are applied to the transformer windings as shown in Fig. 3 for the general modulation mode.
By setting $\tau_0 = \tau_1$ and $\tau_4 = \tau_5$ in Fig. 3 the trapezoidal current mode modulation given in Fig. 4 is obtained. The transferred power is then described by

$$P_{12} = -\text{sign}(\phi_2) V_1 n V_2 (\pi |\phi_2| - 2\delta_1^2 + 2\delta_1 \delta_2)$$

(3)

with $\delta_1 = f(\phi_2) [0, \pi /2]$, $\delta_2 = f(\phi_2) [0, \pi /2]$ and $\phi_2 \in [-\pi, \pi]$. The maximum transferable power is

$$P_{12,\text{max}} = \pm \frac{\pi V_1^2 n^2 V_2^2}{2 \omega_s L_s (V_1^2 + n^2 V_2^2 + V_1 n V_2)}.$$  

(4)

The current $i_{L_s}$ reaches zero at switching instants $\tau_0, \tau_3, \tau_6$ where Zero-Current-Switching (ZCS) is possible [4]. At $\tau_1, \tau_2, \tau_4, \tau_5$ ZVS is possible as far as the minimum commutation current needed for the resonant transition is reached. Nevertheless, the modulation method cannot be applied for low output power. This leads to the triangular current mode modulation with a seamless transition between the modulation methods.

**D. Triangular Current Mode Modulation**

In triangular current mode modulation, also square-wave voltages with clamping intervals are applied to the transformer windings as can be seen from Fig. 5 for the general modulation mode. Considering $\tau_3 = \tau_4$ and $\tau_7 = \tau_8$ in Fig. 5a as well as $\tau_2 = \tau_3$ and $\tau_6 = \tau_7$ in Fig. 5b the triangular current mode modulation shown in Fig. 6 is obtained. The transferred power can then be written as

$$P_{12} = -\frac{V_1 n V_2 \phi_2 (\pi - 2\delta_1)}{\pi \omega_s L_s}, \quad V_1 > n V_2$$

(5)

$$P_{12} = -\frac{V_1 n V_2 \phi_2 (\pi - 2\delta_2)}{\pi \omega_s L_s}, \quad V_1 < n V_2$$

(6)

with $\delta_1 = f(\phi_2) [0, \pi /2]$, $\delta_2 = f(\phi_2) [0, \pi /2]$ and $\phi_2 \in [-\pi, \pi]$. The maximum transferable power is

$$P_{12,\text{max}} = \pm \frac{\pi V_1^2 n^2 V_2^2 (V_1 - n V_2)}{2 \omega_s L_s V_1}, \quad V_1 > n V_2$$

(7)

$$P_{12,\text{max}} = \pm \frac{\pi V_1^2 n^2 V_2^2 (V_1 - n V_2)}{2 \omega_s L_s n V_2}, \quad V_1 < n V_2.$$  

(8)

The current $i_{L_s}$ reaches zero at switching instants $\tau_0, \tau_2, \tau_3, \tau_5, \tau_6$ where ZCS is possible [4]. At instants $\tau_1, \tau_4$ ZVS is possible as far as the minimum commutation current needed for the resonant transition is reached. Due to the mathematical complexity, especially for high port numbers in multi-port converters [7], the following analysis uses basic superposition principles to find the general analytical formula for the power flow. With this approach, there is no need for mathematical distinction of cases.

The mathematical analysis of the power flow is based on the primary side referred equivalent circuit of the converter topology shown in Fig. 7a. The full-bridges are modeled by HF square-wave voltage sources $v_1, v_2$ with clamping intervals as shown in Fig. 7a.

The power flow over one switching cycle $T_s = 2\pi /\omega_s$ between two ports (from a first port 1 to a second port 2) applying square-wave voltages with clamping intervals as shown in Fig. 7a is based on the well-known power flow equation [8] (power from primary port 1 to secondary port 2)

$$P_{ps} = \frac{V_p n V_s (\phi_p - \phi_s)}{\omega_s L_s} \left(1 - \frac{|\phi_p - \phi_s|}{\pi}\right).$$  

(9)

There, two square-wave voltages with 50% duty cycles, amplitudes $V_p, V_s$ and phases $\phi_p, \phi_s \in [-\pi, \pi]$ are applied across the windings of a two-winding transformer with primary referred leakage inductance $L_s$, negligible large magnetizing inductance and turns ratio $n = N_p /N_s$. The phase angles are measured against a given reference, a positive angle defines a leading signal and a negative angle a lagging signal with respect to the reference.

The two-port circuit with clamping intervals given in Fig. 7a can be modeled by the equivalent four-port circuit shown in Fig. 7b where only square-wave voltages without clamping
intervals and duty cycles of 50% occur. This is done by splitting up voltage \( v_1 \) with clamping interval into a sum \( v_{1(1)} + v_{1(2)} \) of two voltages with 50% duty cycle, no clamping interval and a phase-shift of \( 2\delta_1 \) against each other as depicted in Fig. 8. Analogously, this is done for the voltage \( v_2 \). The power transferred from port 1 to port 2 is then given by

\[
P_{12} = \frac{1}{T_s} \int_{0}^{T_s} v_{1(1)} i_{L\sigma} \, dt + \frac{1}{T_s} \int_{0}^{T_s} v_{1(2)} i_{L\sigma} \, dt
\]

with the two power shares of voltage sources \( v_{1(1)}, v_{1(2)} \) (see Fig. 7b). The leakage inductance current \( i_{L\sigma} \) is split up into three parts \( i_{L\sigma(1)} + i_{L\sigma(II)} + i_{L\sigma(III)} \) which are obtained by applying the superposition principle as shown in Fig. 9 by selectively short-circuiting voltage sources. In this way, the power exchange of source \( v_1 \) with sources \( v_{1(2)}, v_{2(1)}, v_{2(2)} \) is described. The power share \( P_{12(1)} \) in (10) can then be written as

\[
P_{12(1)} = \frac{1}{T_s} \int_{0}^{T_s} v_{1(1)} i_{L\sigma(1)} \, dt + \frac{1}{T_s} \int_{0}^{T_s} v_{1(1)} i_{L\sigma(II)} \, dt
\]

Analogously, the second power share \( P_{12(2)} \) is described. From Fig. 9 and (11) it is concluded, that the power shares \( P_{12(1)(I)}, P_{12(1)(II)}, P_{12(1)(III)} \) are given by (9). This is also the case for the power shares \( P_{12(2)(I)}, P_{12(2)(II)}, P_{12(2)(III)} \). By summing up all the power shares, the resulting power transferred per switching cycle from port 1 to port 2 applying square-wave voltages \( v_1, v_2 \) with clamping intervals \( 2\delta_1, 2\delta_2 \) and phases \( \phi_1, \phi_2 \) as shown in Fig. 8 is thus given as

\[
P_{12} = \frac{1}{4\omega_L L_\sigma} \left[ (\phi_1 - \delta_1)(\phi_2 - \delta_2) \left( 1 - \frac{|\phi_1 - \delta_1 - \phi_2 + \delta_2|}{\pi} \right) + (\phi_1 - \delta_1)(\phi_2 + \delta_2) \left( 1 - \frac{|\phi_1 - \delta_1 - \phi_2 - \delta_2|}{\pi} \right) + (\phi_1 + \delta_1)(\phi_2 - \delta_2) \left( 1 - \frac{|\phi_1 + \delta_1 - \phi_2 - \delta_2|}{\pi} \right) + (\phi_1 + \delta_1)(\phi_2 + \delta_2) \left( 1 - \frac{|\phi_1 + \delta_1 - \phi_2 + \delta_2|}{\pi} \right) \right]
\]

In general, the proposed superposition method for deriving analytical power flow equations can be applied to any number of ports which are connected in series in multi-port converters as for instance shown in [9] for an isolated three-phase bidirectional AC-DC converter.

IV. OPTIMIZATION OF CONVERTER LOSSES

The general power flow equation (12) is used in an optimization of converter losses of a DAB prototype system designed for a nominal output power of 3.3 kW. The loss model includes the main loss shares as conduction and switching losses of the semiconductor devices, skin and proximity effect losses of the transformer windings as well as the core losses of the transformer.

A. Prototype System

As a prototype system to find optimal control variables, a 3.3 kW electric vehicle battery charger to connect to a fixed 400 V DC-link with an output voltage range of 280 V to 420 V of a lithium-ion battery is considered. The primary and secondary switching devices are chosen to be 650 V IGBTs of type IKW50N65F5 for a first converter solution and 650 V MOSFETs of type IPW65R019C7 for a second converter solution, both from Infineon [10]. The system parameters are listed in detail in Table I.

| Input voltage | \( V_1 \) | 400 V |
| Battery voltage | \( V_2 \) | 230 V ... 420 V |
| Output power | \( P_2 \) | 3.3 kW |
| Switching frequency | \( f_s \) | 20 kHz ... 50 kHz |
| Transformer turns ratio | \( n \) | 8/7 |
| Transformer leakage inductance | \( L_\sigma \) | 181 \( \mu \)H |
| Phase-shift modulation | \( L_{\sigma(1)} \) | 158 \( \mu \)H |
| Triangular/Trapezoidal current modulation | \( L_{\sigma(2)} \) | 158 \( \mu \)H |
| Optimized control variables modulation | \( L_{\sigma(3)} \) | 158 \( \mu \)H |
| Transformer magnetizing inductance | \( L_m \) | Neglected |

B. Loss Models

For optimizing the DAB converter system, its main losses occurring in the semiconductor devices as well as in the transformer have to be modeled. For the semiconductor devices, both IGBT and MOSFET loss models based on datasheet parameters are used and given in the following.

1) IGBT Losses: For calculating the conduction losses, the typical output characteristic \( i_C = f(v_{CE}) \) of the IGBT and the typical diode forward current as a function of forward voltage \( i_D = f(i_D) \) of the anti parallel diode at a junction temperature of \( T_{j,max} - 25^\circ C = 150^\circ C \) from the datasheet are considered. Given the current \( i_C \) flowing through the IGBT and the current \( i_D \) through the diode over the interval \([0, T_s]\) of a switching period, the conduction losses of an IGBT co-pack device can be calculated by

\[
P_{c,s} = \int_{0}^{T_s} i_C(\tau) \cdot v_{CE}(i_C(\tau)) \, d\tau,
\]
When optimizing and designing a converter system, the details of the gate drives and the parasitics of the commutation path are usually unknown, so that only approximations of switching losses can be performed. For estimating the switching losses of an IGBT co-pack device, the following assumptions are made: For a device which is turned on
- diode losses at zero-voltage turn-on are neglected,
- IGBT losses $E_{on} = f(iC)$ at turn-on including diode reverse recovery losses are taken from datasheet and are linearly scaled with voltage according to datasheet.

For a device which is turned off
- diode losses at turn-off are neglected,
- IGBT losses $E_{off} = f(iC)$ at turn-off are taken from datasheet and are linearly scaled with voltage according to datasheet.

For soft-switching in terms of ZVS at the switching instant $\tau_s$, the switching losses of an IGBT co-pack device are approximated by

$$P_s = f_s \cdot E_{off}(iC(\tau_s)) \frac{V_{CE}(\tau_s)}{V_{CE}},$$  

whereas for hard-switching in terms of forced diode commutation, the switching losses are estimated according to

$$P_s = f_s \cdot E_{on}(iC(\tau_s)) \frac{V_{CE}(\tau_s)}{V_{CE}},$$

with $V_{CE}$ being the collector-emitter voltage where switching losses were measured according to the datasheet. For soft-switching in terms of ZCS for small $i_C(\tau_s)$, losses according to (16) and (17) become negligible small.

2) MOSFET Losses: Also for MOSFETs, the typical output characteristic $i_D = f(v_{DS})$ at a junction temperature of $T_{j,max} = -25^\circ C = 125^\circ C$ from the datasheet can be used to calculate the conduction losses. Given the current $i_D$ flowing through the MOSFET over the interval $[0, T_s]$ of a switching period, the conduction losses are given by

$$P_c = \frac{1}{T_s} \int_0^{T_s} i_D(\tau) \cdot v_D(i_D(\tau)) \, d\tau, \quad (14)$$

$$P_c = P_{c,S} + P_{c,D}. \quad (15)$$

The approximation of switching losses of a MOSFET device is based on the following assumptions: For ZVS conditions when stored energy in the output capacitance is transferred from one MOSFET to another, difference between released and absorbed energies are negligible small. In other words, losses caused during the commutation of the inductive current are neglected, $P_s = 0$. For hard-switching, two loss effects are modeled, these are
- dissipation of energy $E_{oss} = f(v_{DS})$ stored in the output capacitance at turn-on,
- body diode reverse recovery losses occurring in turn-on device squarely scaled with voltage and linearly scaled with current.

Switching losses at time instant $\tau_s$ are then estimated using

$$P_{oss} = f_s \cdot E_{oss}(v_{DS}(\tau_s)), \quad (19)$$

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- dissipation of energy $E_{oss} = f(v_{DS})$ stored in the output capacitance at turn-on,
- body diode reverse recovery losses occurring in turn-on device squarely scaled with voltage and linearly scaled with current.
In the loss model, the core losses per volume are calculated by applying the improved Generalized Steinmetz Equation (iGSE) [14]. The skin and proximity effect losses per unit length in litz wires for each current harmonic are determined according to [15]. The external magnetic field strength for evaluating proximity effect losses is derived by a 1D approximation using the Dowell method [16].

4) Auxiliary Losses: Besides the load dependent loss shares shown in the previous sections, a constant loss share for gate drives, control, sensing and fans of 8 W is considered.

C. Optimization Procedure

The optimal control variables in terms of clamping intervals $\delta_1$, $\delta_2$, phase-shift $\phi_2$ and switching frequency $f_s$ are numerically determined by minimizing the total converter losses (semiconductor losses $P_{sw}$, transformer losses $P_{tr}$ and auxiliary losses $P_{aux}$) subject to power flow constraint. The optimization procedure is shown in Fig. 11. For given output voltage $V_2 \in [280 V, 420 V]$ and reference output power $P_{2r} \in [0.33 kW, 3.3 kW]$, the optimization routine calculates the optimal control variables. The optimization problem is stated as

$$\min_{x} [P_{sw} + P_{tr} + P_{aux}] \text{ with respect to } x = \begin{bmatrix} \delta_1 \\ \delta_2 \\ \phi_2 \\ f_s \end{bmatrix}$$ (22)

with

$$x_{lb} = \begin{bmatrix} 0 \\ 0 \\ -\pi \\ 20 kHz \end{bmatrix}, \quad x_{ub} = \begin{bmatrix} \pi/2 \\ \pi/2 \\ \pi \\ 50 kHz \end{bmatrix}$$ (23)

where $x$ denotes the vector of control variables which is restricted to lower and upper bounds $x_{lb}, x_{ub}$ respectively. The equality constraint is given by setting the power transfer $P_{t2} = P_{12}$ using (12).

D. Optimization Results

Calculated relative converter efficiencies applying phase-shift modulation, combined triangular/trapezoidal current mode modulation and modulation with optimized control variables including frequency variation are shown in Fig. 13 and Table III for an IGBT solution and in Fig. 14 and Table IV for a MOSFET solution.

For phase-shift modulation, it can be seen that for the MOSFET solution in the soft-switching area higher efficiencies are achieved than for the IGBT solution (compare Fig. 14a to Fig. 13a). This is mainly due to the fact, that IGBT turn-off losses cannot be substantially reduced by using ZVS. In the hard-switching region, ZVS is lost and forced diode commutations occur. The switching losses in this region are strongly dependent on the characteristics of the anti parallel diode of the IGBT and the body diode of the MOSFET respectively.

To improve efficiencies, especially in the hard-switching region, the combined triangular/trapezoidal current mode modulation can be used. There, also for low output power in the areas of low and high output voltages, soft-switching (ZCS combined with ZVS) can be achieved.

Considering also frequency variation, efficiencies can be slightly increased compared to triangular/trapezoidal current mode modulation. Soft-switching is achieved in the whole operating range: ZCS and ZVS for the IGBT solution and only ZVS for the MOSFET solution. The modulation modes, which are found by the optimization procedure, are given in Fig. 13c and Fig. 14c. Fig. 12 shows the resulting switching frequencies found by the optimization, both for the IGBT and the MOSFET solution. Especially for the MOSFET solution, the frequency is varied over a wide area of the operating range. With decreasing power, frequency can be increased to lower the power transfer and achieve high efficiencies at the same time. Nevertheless, at low input powers, it is more attractive to decrease the switching frequency and change the modulation mode when necessary (see Fig. 14c).
For a 14.2 kWh battery pack with 11.5 Ah lithium iron phosphate (LiFePO4) cells [17], efficiencies of a typical charging process from 10 % to 90 % state-of-charge (battery voltage from 385 V to 418 V) with constant input power \( P \) = 3.3 kW are calculated and given in Table V for different modulation methods applying IGBTs and MOSFETs. It can be seen, that the charging efficiencies at maximum input power do not differ much for the same semiconductor technology applying different modulation methods.

### V. CONCLUSION

A novel modeling approach of the power flow in a bidirectional DAB DC-DC converter is presented. By using basic superposition principles, the mathematical distinction of cases is avoided in the modeling process of HF transformer currents for different types of modulation. The generalized model is used in an optimization of converter losses for an IGBT and a MOSFET solution considering all types of modulations, also including variation of frequency for converter control. For a 3.3 kW electric vehicle battery charger, the efficiency increases up to 8.6 % using IGBTs and 17.8 % using MOSFETs in the most critical point compared to phase-shift modulation at fixed switching frequency.

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### REFERENCES


Fig. 13. Relative converter efficiencies applying IGBTs for an input power range of 10% to 100% of maximum input power and an output voltage range of 280 V to 420 V for phase-shift modulation (a) and triangular/trapezoidal current mode modulation (b) both with fixed switching frequency at 20 kHz and for modulation with optimized control variables with variable switching frequency from 20 kHz to 50 kHz (c). For phase-shift modulation in (a), soft- and hard-switching areas are given, whereas with modulations in (b) and (c) soft-switching is always achieved. In (c), also the modulation mode found by the optimization is given.

Fig. 14. Relative converter efficiencies applying MOSFETs for an input power range of 10% to 100% of maximum input power and an output voltage range of 280 V to 420 V for phase-shift modulation (a) and triangular/trapezoidal current mode modulation (b) both with fixed switching frequency at 20 kHz and for modulation with optimized control variables with variable switching frequency from 20 kHz to 50 kHz (c). For phase-shift modulation in (a), soft- and hard-switching areas are given, whereas with modulations in (b) and (c) soft-switching is always achieved. In (c), also the modulation mode found by the optimization is given.