Flux control modulation for the dual active bridge DC/DC converter

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Abstract: This paper introduces a modulation technique for the single-phase dual active bridge (DAB) in which the switching frequency and transformer flux linkage are modulated to achieve improved power transfer characteristics and reduced converter losses. By analysing the DAB, it is noticed that in the traditional single phase shift (SPS) modulation, the transformer flux linkage decreases with increasing power. The proposed flux control modulation (FCM) counteracts this decrease and aims to keep the transformer utilisation constant on the whole load range by a reduction in switching frequency. This has two effects: first, the reduced frequency substantially lowers switching losses and second, the power transfer characteristic of the DAB is linearised, facilitating the control of this converter. The new modulation technique is validated by simulation.

1 Introduction

Power electronics is a key enabling technology in the development of next-generation aircraft systems. The more electric aircraft (MEA) is a concept aiming to replace traditional hydraulic, pneumatic, mechanical, and other systems with more efficient, lightweight, and reliable electrical systems, cutting down fuel consumption and operating cost [1, 2]. However, this trend pushes the aircraft electrical power system (EPS) to the multi-megawatt range. The ASPIRE project deals with the development of an EPS for future aircraft, which can tackle the challenges that the MEA concept introduces. A high power density 3 kW single-phase dual active bridge (DAB) is developed for the implementation of a ‘smart grid’ concept in future EPS.

When analysing the traditional single phase shift (SPS) modulation for the DAB, it becomes clear that the peak flux linkage of the transformer decreases with increasing power transfer. This implies that the transformer has to be designed for no load and that it is not fully utilised during power transfer.

In general, improvement of modulation strategies for the DAB has been subject to research work [3–9]. Qualitatively, variable frequency modulation has been reported in literature as a means of efficiency improvement [10–14]. However, the modulation strategy presented here will provide the mathematical background for optimisation of the frequency adaptation.

Based on thorough analytical modelling of the DAB, this paper introduces an improved modulation technique called flux control modulation (FCM), which is derived from SPS modulation. By variation of the switching frequency, FCM aims to keep the utilisation of the transformer core constant on the whole load range. This not only improves the overall operating efficiency of the DAB but also linearises the power transfer characteristic of the DAB and hence simplifies the control design of the converter.

Section 2 gives a brief overview of the DAB. Section 3 introduces an analytical model of SPS modulation. Based on the outcomes of this analysis, FCM is introduced in Section 4. A detailed comparison of SPS modulation and FCM is presented in Section 5, followed by validation in Section 6.

2 Single-phase dual active bridge

The DAB was first proposed in [15]. The generic single-phase variant of the DAB is depicted in Fig. 1.

The single-phase DAB consists of two active H-Bridges equipped with switching devices such as MOSFETs or IGBTs, and a high-frequency transformer providing galvanic isolation and leakage inductance. $V_P$ denotes the primary-side DC voltage and $V_S$ denotes the secondary-side DC voltage. The original modulation strategy for this topology is called single phase shift (SPS) modulation, which is depicted in Fig. 2 [15]. Both H-Bridges operate at 50% duty cycle, but phase shifted in time by the angle $\phi$. The output voltages of the primary and secondary side H-Bridge are denoted as $v_P(t)$ and $v_S(t)$, respectively. Using the transformer turns ratio $n$, all secondary-side quantities are referred to the primary side, which is indicated by a dash. The waveforms $v_P(t)$ and $v_S(t)$ are shown in Fig. 2a. The voltage difference $v_P(t) - v_S(t)$ across the leakage inductance $L$ of the transformer will result in a quasi-square waveform of high-frequency AC current in the transformer, $i_L(t)$ (see Fig. 2c). As shown in Fig. 1, the input currents of the primary- and secondary-side H-Bridge are denoted by $i_{P1}(t)$ and $i_{S1}(t)$, respectively. They are plotted in Figs. 2d and e, respectively. The current ripple is filtered by the primary- and secondary-side DC capacitors $C_P$ and $C_S$. The average currents provided by the primary- and secondary-side DC buses are denoted by $I_{P1}$ and $I_{S1}$. The transferred power $P$ is given in (1), where $\phi$ denotes the phase-shift angle between $v_P(t)$ and $v_S(t)$ and $f_{sw}$ denotes the switching frequency [15].

$$P = \frac{V_P V_S}{2\pi f_{sw} L} \cdot \phi (\pi - |\phi|)$$  (1)

Maximum power is transferred for a phase-shift angle of $\pi/2$. The turn-off transition in both H-Bridges is generally hard-switched, but for the turn-on transition, zero voltage switching (ZVS) is achieved for most operating points [16].
In this section, an analytical model of the SPS modulation is presented, as FCM is directly derived from SPS modulation. The derived equations will be useful to compare the SPS and the FCM modulation strategies, e.g. in terms of losses. Section 3.1 provides an analysis of the current in the AC link which transfers power between the two DC ports. Section 3.2 gives an analysis of the magnetisation of the transformer.

Due to the waveform symmetries as seen in Fig. 2, it is sufficient to regard half a switching period. The following conventions are made: for positive phase shifts $\phi \geq 0$, the rising edge of the primary voltage $v_p(t)$ is located at $\omega_{sw} t = 0$ and the rising edge of the secondary voltage $v_s(t)$ is delayed to $\omega_{sw} t = |\phi|$. For negative phase shifts $\phi < 0$, the rising edge of the secondary voltage is located at $\omega_{sw} t = 0$ and the rising edge of the primary voltage is delayed to $\omega_{sw} t = |\phi|$. These conventions are visualised in Fig. 3.

### 3.1 Line current waveform

The current waveform $i_L(\omega_{sw} t)$ is obtained by integrating the output voltage difference of the H-Bridges, $v_p(t) - v_s(t)$, which drops across the leakage inductance $L$. For keeping the model simple, all ohmic resistances and the magnetising inductance $L_m$ are neglected in this analysis, as $L_m$ is usually much bigger than the leakage inductance $L$. The resulting current waveform as well as its initial values for $\omega_{sw} t = 0$ and $\omega_{sw} t = |\phi|$ are given as follows:

$$i_L(\omega_{sw} t) = \begin{cases} i_L(0) + \frac{\text{sgn}(\phi)}{\omega_{sw} L} \frac{V_p + V_s}{\omega_{sw} t} & \text{for } 0 \leq \omega_{sw} t < |\phi| \\ i_L(|\phi|) + \frac{V_p - V_s}{\omega_{sw} L} (\omega_{sw} t - |\phi|) & \text{for } |\phi| \leq \omega_{sw} t < \pi \end{cases}$$

The peak value of the AC link current $i_L$ is denoted $\bar{I}_L$:

$$\bar{I}_L = \max \left| i_L(0) \right|, i_L(|\phi|)$$

By squaring (2) and integrating, the RMS value of the AC link current, $I_L$, is also derived:

$$I_L = \frac{1}{\omega_{sw} L} \sqrt{\frac{\pi^2 (V_p - V_s)^2 + (V_p + V_s)^2 - (V_p - V_s)|\phi|^2}{12}}$$

### 3.2 Transformer utilisation in SPS modulation

This section presents an in-depth analysis of the transformer utilisation in SPS modulation. The transformer is modelled by its T-equivalent circuit, which is shown in Fig. 4. The inductances $\lambda_{rP}$ and $\lambda_d$ describe the stray inductances of the transformer on both the primary and secondary windings, but shall include, if present, external series inductors, so that $L = \lambda_{rP} + \lambda_d$. The quantities $d$, $r$, and $\lambda$ as given in (9)–(11) are defined as follows: $r$ is the ratio of the stray inductances $\lambda_{rP}$ over $\lambda_d$ and ranges from $0$ to $\infty$. $d$ is the ratio of the voltages $V_p$ over $V_s$ and describes how well the transformer turns ratio $n$ matches the actual ratio of the DC voltages. $\lambda$ is a new quantity called transformer utilisation factor (see Fig. 5). Its maximum value of 1 is reached for $r = d$, and it
\[
\Delta Q_p = \frac{(2V_p \phi^2 + \pi |2\pi - |\phi| + \pi(V_p - V_S)\phi)}{8 \alpha_{in} L r^2 (V_p - V_S)}
\]
\[
\Delta Q_s = \frac{V_p(2V_p \phi^2 + (\pi - 2|\phi| + \pi(V_p - V_S)\phi)}{4 \alpha_{in} L r^2 (V_p - V_S)}
\]

The flux density \( B \) of the core is

\[
B = \frac{\mu_0 \mu_{r} B_0}{L_m}
\]

for \( V_p > V_S \) and \(|V_{sw,m}| < |I_{L_p}|\)

\[
\bar{i}_{m} = \max \| i_p(0) \| i_p(\phi) \|
\]

(16)

\[
\bar{\psi} = \frac{V_p + rV_S}{2(1 + r)\alpha_{in} \pi} - \frac{(V_p + rV_S) - [V_p - rV_S]}{2(1 + r)\alpha_{in} \pi} \psi
\]

(17)

The flux density \( B \) cannot be calculated as the number of turns and the core area are unknown. However, if a reasonable value is assumed for the maximum flux density, e.g. \( \bar{B}_{max} = 200 \text{ mT} \) for a ferrite core, the per-unit utilisation of the transformer can be derived as given by (19), based on the utilisation factor \( \lambda \) from (11).

The flux linkage in the transformer core equals \( \psi = L_{m}i_{m} \), as shown in Fig. 2b. In analogy to (5), the peak magnetising current \( \bar{i}_{m} \) and the peak flux linkage \( \bar{\psi} \) are calculated:

Fig. 4 Equivalent circuit of the transformer

Fig. 5 Transformer utilisation parameter \( \lambda \) as function of leakage inductance ratio \( r \) and voltage ratio \( d \)
The only unknown quantity is the volume of the transformer core, flux linkage is decreasing with increasing phase shift. Therefore, the transformer is utilised at its maximum for zero phase shift, i.e. zero power transfer. Hence, the transformer has to be designed for no load. This is the main finding of the analysis of the DAB which will be addressed by FCM in Section 4.

The iron losses of the transformer are calculated using the improved generalised Steinmetz equation (iGSE):

\[
P_{Fe} = V_{olc} f_{sw} \int_{0}^{\pi} i_{a} d\theta \Delta B_{pp}^{\phi} \frac{k}{r} \frac{\psi}{\psi_{max}} \frac{\pi f}{V_{\phi}} \frac{\lambda}{\pi} \frac{V_{\phi}}{2(1 + \lambda)}
\]

where

\[
k_{i} = \frac{k}{2\pi^{a} \int_{0}^{\pi} \cos \theta^{a} \cdot 2^{a-\alpha} d\theta}
\]

\[
k, \alpha, \text{ and } \beta \text{ are material specific data, } V_{olc} \text{ is the iron volume and } \Delta B_{pp} \text{ is the peak-to-peak core flux density. The cross-sectional area of the core and the number of turns are not always known. Therefore, the per-unit quantity } B/B_{max} \text{ from (19) has to be used in (20), which now can be solved analytically:}
\]

\[
P_{Fe} = 2^{a} V_{olc} k_{i} f_{sw} \Delta B_{pp}^{\phi} \frac{\psi}{\psi_{max}} \left( 1 - \lambda \frac{\lambda}{\pi} \right) \left( 1 - \lambda_{a} \frac{\lambda}{\pi} \right)
\]

where

\[
\lambda_{a} = 1 - \left( \frac{f - f_{sw}}{f_{sw}} \right)^{a}
\]

The only unknown quantity is the volume of the transformer core, and it is assessed by the stored energy \(E_{m}\). This energy is expressed by the peak flux linkage and the magnetising current on the one hand and by the peak flux density on the other hand:

\[
E_{m} = \frac{1}{2} L_{sat} i_{\mu}^{2} \frac{1}{2} \psi^{2}_{max} i_{\mu} = V_{olc} \frac{B_{max}^{2}}{2k_{i} \mu_{r}}
\]

\[
\Rightarrow V_{olc} = \frac{L_{sat} i_{\mu}^{2}}{B_{max}^{2}}
\]

4 Flux control modulation

Equation (19) shows that the transformer flux linkage decreases with phase shift as long as there is some inductance on both sides of the transformer (\(\lambda \neq 0\)). As already stated, this implies that the transformer is actually designed for no load. The idea of FCM is to reduce the switching frequency linearly as the phase shift increases, to counteract the linear reduction in transformer utilisation as shown in (19). The highest switching frequency, \(f_{sw,\text{max}}\), is used at no load, i.e. \(\phi = 0\). Equations (18) and (19) are thus modified as follows:

\[
\psi_{max} = \frac{V_{p} + r V_{S}}{2(1 + r)\phi_{sw,\text{max}}} \pi
\]

\[
\frac{\dot{B}}{B_{max}} = \frac{\psi}{\psi_{max}} = f_{sw,\text{max}} \left( 1 - \lambda \frac{\lambda}{\pi} \right)
\]

The transformer utilisation is kept constant by setting (25) to unity and solving for the switching frequency. Using this result in the power equation (1) gives a new modulation technique, called FCM, defined by (26) and (27):

\[
f_{sw} = f_{sw,\text{max}} \left( 1 - \lambda \frac{\lambda}{\pi} \right)
\]

\[
P = \frac{V_{p} V_{S}}{2\pi f_{sw,\text{max}} L} \left( \pi - \lambda \frac{\lambda}{\pi} \right)
\]

The power equation (27) is depicted in Fig. 6 for different utilisation parameters \(\lambda\) between 0 and 1. As the utilisation parameter \(\lambda\) approaches unity, the power (27) converges into a linear function, which is convenient for control:

\[
P \approx \frac{V_{p} V_{S}}{2\pi f_{sw,\text{max}} L} \phi \quad \text{for } \lambda \approx 1
\]

Because of the linear decay in switching frequency (26), switching losses are reduced with increasing power. Another big advantage is the linearisation of the power characteristic and its ease of implementation on a control platform. One disadvantage of FCM is increased iron losses, although the iron losses might not play a dominant role in the overall loss balance of the converter. Additionally, FCM covers a broader range of switching frequencies, which may be more demanding for EMI filtering.

The iron losses under FCM are assessed by solving the iGSE (20) and substituting the switching frequency \(f_{sw}\) by the expression in (26):

\[
P_{Fe,\text{FCM}} = 2^{a} V_{olc} k_{i} f_{sw,\text{max}} B_{max}^{\phi} \frac{\psi}{\psi_{max}} \left( 1 - \lambda \frac{\lambda}{\pi} \right)
\]

5 Comparison between SPS and FCM

The SPS modulation is identified as the special case of FCM when \(\lambda = 0\), as in this case (27) degenerates into (1) and (26) implies \(f_{sw} = f_{sw,\text{max}}\). A utilisation factor of \(\lambda = 0\) is particularly the case if there is no inductance on one side of the transformer \((r = 0\) or \(r \rightarrow \infty\)). In any other case, FCM is advantageous to use, especially in the case of \(\lambda = 1\), yielding the maximum reduction in switching losses.

Fig. 6 Power transfer for FCM modulation as a function of phase shift and \(\phi = 0, 0.05, 0.75, 0.9, 0.975, 1\). The linear function corresponds to \(\lambda = 1\). \(V_{p} = 270 \text{ V}, V_{S} = 28 \text{ V}, n = 10, f_{sw,\text{max}} = 100 \text{ kHz}, L = 25 \mu\text{H}\)
The duality between SPS modulation and FCM is presented in the following: in SPS modulation, the frequency is kept at its maximum value and the transformer flux linkage decays with increasing phase shift. However, in FCM, the flux linkage is kept at its maximum value and the switching frequency is reduced. The roles of switching frequency and flux linkage are swapped, as visualised in Fig. 7.

From Fig. 6, it becomes clear that in FCM, a smaller phase shift is needed compared to SPS modulation. The required phase shift angles for SPS modulation and FCM are plotted in Fig. 8a. For SPS modulation, the required phase shift angle \( \phi_{SPS} \) is obtained as a complex equation (30) by solving (1), while the phase shift angle \( \phi_{FCM} \) in FCM is given by a simpler equation (31).

Because of the frequency reduction in FCM, the switching losses \( P_{sw} \) are reduced compared to SPS modulation as given in (32). In contrast, the iron losses \( P_{Fe} \) are increased as shown in (33), as the transformer is fully utilised in FCM.

Some further comparisons between SPS modulation and FCM can be made: If the frequency equation of FCM (26) is substituted into (5), the peak current in the AC link under FCM is obtained. It is plotted, normalised to SPS modulation, in Fig. 8c. Similarly, substituting (26) into (6), the RMS current in the AC link under FCM is plotted in Fig. 8d. Finally, Fig. 8d shows the primary-side ripple charge from (7) in FCM compared to SPS modulation. It is observed that neither peak nor RMS current in the AC link increase as a consequence of frequency reduction, but it is found that this is not the case. The reason for all these effects is that even though the frequency is reduced in FCM, a smaller phase shift than in SPS modulation is needed, as is shown in Fig. 8a.

### 6 Performance validation

The effectiveness of FCM has been validated by a PLECS simulation of the DAB of the ASPIRE project. The system parameters used are given in Table 1.

Fig. 9 shows the efficiency of the DAB, plotted over transferred power, when the same converter is operated with SPS and FCM modulation techniques. The reduction of switching losses in FCM yields an efficiency improvement of 0.46% at full load.

Finally, in Fig. 10, the transformer flux linkage \( \psi(t) \) is plotted during a step load change from 500 W to 3 kW for both modulation strategies. It is clearly visible that in SPS modulation, the peak flux linkage decreases with increasing power. In FCM, however, the magnitude of the flux linkage is kept constant.

### 7 Conclusion

Flux control modulation is proposed in this paper and offers many advantages, such as reduced converter losses, improved transformer utilisation and simplification of control design. It linearises the power transfer characteristic of the DAB and hence making it suitable for analogue control. The advantages of the proposed modulation technique have been validated by simulation.
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Fig. 9 Efficiency of the DAB when operated with SPS modulation and FCM

Fig. 10 Flux linkage in the transformer core during a step load change from 500 W to 3 kW (t = 0) for (a) SPS modulation and (b) FCM

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