A flux estimation method for three-phase dual-active-bridge DC/DC converters

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Abstract

This study proposes an estimation method of flux density in high-frequency transformers of three-phase dual-active-bridge transformers. The steady-state analysis shows that the flux in the converter is different in Buck and Boost modes and depending on load and voltage gain condition. From that, a closed-form formula is proposed to estimate the flux swing in the core. Unlike the conventional estimation technique, the proposed one represents the flux as a function of input and output voltages and the phase shift value. Finite element analysis confirms that the proposed formula can predict the flux with less than 4% error. The proposed formula can also help to improve the prediction of temperature profile of the transformers with a high accuracy of less than 1 degree Celsius in the mid- to high-power range as validated in experiments. The proposed flux estimation method can be used in designing the converter or in predicting the converter behaviour under critical conditions such as load changing or starting, and so forth, to help protect the transformers from saturation or DC bias.

1 | INTRODUCTION

Transformers play an essential role in isolated power DC/DC converters. They help to create galvanic isolation and matching the voltage level of two sides. By applying an AC voltage at one side of the transformer (primary), a flux proportional to the core size and number of turns is generated in the core. The flux then induces a voltage at the other sides of the transformer (secondaries), which connects to the output circuitry. The magnetising capability of the core is quantified by a maximum allowable flux density, which is limited and depends on the core material. That allowable limitation is usually provided in the core datasheet or in the material specification sheet from manufacturers such as [1, 2]. Saturation may occur if the flux generated in the core is greater than or equal to that limitation, which, in turn, causes an overheating of the core because of extremely high core loss, current waveform distortion, and excessive current spikes, and so forth, or even converter damage [3]. Therefore, when designing transformers for isolated DC/DC converters, one must pay particular attention to the flux density to avoid core saturation. In order to facilitate the transformer design, the flux is formulated.

The flux density in a converter is usually roughly computed by

\[ \Delta B = \frac{V_1}{N_1 A f s K_f} \]

where \( V_1 \) is the input voltage, \( N_1 \) is the number of turns of the primary winding; \( A \) is the cross-section area of the core; \( f_s \) is the switching frequency; and \( K_f \) is the waveform coefficient, \( K_f = 4.0 \) for square wave and 4.44 for sine wave. Equation (1) has been popularly employed in designing many types of unidirectional isolated converters ranging from small to large power, low to high-frequency [4–8]. The flux estimation according to Equation (1) is independent of the control variable and output voltage level, and thus it supposes to be constant regardless of load changes.

From another aspect, a transformer theoretically changes only the amplitude but repeats the phase and frequency of the input signal. For converter systems where the displacement phase between primary and secondary sides can be manipulated by modulation such as phase-shift-controlled converters, Equation (1) may result in an incorrect flux estimation. From the
design perspective, a mistaken flux estimation leads to an inappropriate number of turns, an undersized or oversized magnetic core, and an erroneous prediction of converter behaviours. This study focuses on such a problem in three-phase dual-active-bridge (DAB) converters, which is one member of the phase-shift-controlled typed converter family.

DAB converters were introduced by De Doncker et al. in [9]. Up to the recent days, they have been applied in many applications such as power system [10], electric vehicle [11], aircraft [12], and so forth. A DAB converter consists of two active bridges and a transformer(s) that allows the power to be transferred bidirectionally via magnetic coupling. By applying an appropriate modulation technique, soft-switching can be achieved, which allows to reduce the electromagnetic interference content, improve the efficiency and reduce the size and volume of the converter [13]. The key point of size reduction is transformer design, and in particular, the integration of inductor into the transformer.

The design of DAB converters can be found in many reports, such as [14–21]. In [14–17], the conventional estimation of flux density expressed by Equation (1) was also employed to design such DAB transformers. In [18, 19], the flux swing in three-phase DAB converters was estimated by Equation (2). Compared to the flux determined by Equation (1), that obtained with Equation (2) is 2.25 times smaller. However, the derivation of Equation (2) was omitted in the articles, and there were no results reported to support the validity of the equation:

\[
\Delta B_{\text{dab}} = \frac{V_1}{9N_1A_s f_s} \tag{2}
\]

In [20, 21], Equation (3) was employed to calculate the flux in single-phase DAB converters:

\[
\Delta B_{\text{dab}} = \frac{V_1}{4N_1A_s f_s} \left(1 + \frac{nV_2}{V_1} (1 - D_p)\right) \tag{3}
\]

where \(V_2\) is the output voltage, \(n\) is the transformer ratio, and \(D_p\) is the phase shift ratio. Equation (3) is very similar to the conventional estimation described in Equation (1) except for the additional factor in the parentheses, which reflects the dependency of flux density on the output voltage and the control variable. However, Equation (3) is dedicated to single-phase DAB converters, and there was no study so far on its application to the three-phase variants. The three-phase topology consists of a three-phase transformer or three single-phase transformers that makes the analysis of magnetisation more complicated due to the presence of the neutral point of transformers.

A design and optimisation of three-phase DAB converters with Y-Y transformer configuration for quick charging applications with a vehicle-to-grid support were reported in [22] by our research team. The capability of the charger is 50 kW, divided into several sub-modules. A sub-module occupied the three-phase DAB topology and was designed and optimised by the genetic algorithm method in which the peak flux density was calculated by Equation (4) where \(\psi\) is the phase shift angle in radians:

\[
B_{pk} = \frac{V_1}{18N_1A_s f_s} \left(1 + \frac{nV_2}{V_1} \left(1 - \frac{3\psi}{2\pi}\right)\right) \tag{4}
\]

Although in [22], we presented the performance indexes of the converter such as temperature, efficiency, and so forth, the derivation of Equation (4) that inspires the motivation of this study was not described.

In this study, a closed-form equation to determine the flux swing in the magnetic core is derived. The equation is then evaluated using finite element analysis to confirm its validity against the other aforementioned flux determination techniques. After that, the temperature profile of the converter is then compared to the theoretical temperature, which is calculated based on the proposed flux model to confirm its validity, as it is difficult to determine the flux in the run time. The proposed closed-form model is derived based on single-phase shift modulation; however, the same approach can be applied to other modulation methods.

2 DERIVATION OF MAGNETIZING VOLTAGE

Figure 1 shows the circuit diagram of a three-phase DAB converter with three single-phase transformer connecting in the Y-Y fashion. In the figures, \(V_1\) and \(V_2\) are terminal DC voltages; \(i_a, i_b\) and \(i_c\) are phase currents; \(n\) is the transformer winding ratio. Assume that power is transferring from terminals 1 to 2. Aiming to simplify the analysis, the following assumptions are made:

1. Three transformers have the same parameters.
2. Equivalent series resistance of transformer windings is negligible.
3. ON-resistance of MOSFET can be ignored.
4. The transition of switches has zero rising and falling time.
5. The transformers operate in the linear zone of the magnetising characteristics.

The primary-referred diagram of the converter is illustrated in Figure 2. The superscript ′ denotes that the corresponding
quantities are referred to the primary side of the transformer:

\[ L'_{k_2} = n^2 L_{k_2} \]  
\[ v'_x = n v_x \]  

where \( x = a, b, c \); \( n \) is the winding ratio, \( n = N_1 / N_2 \); \( N_1 \), \( N_2 \), \( L'_{k_1} \) and \( L_{k_2} \) are the number of turns and leakage inductances of primary and secondary windings, respectively; \( L_M \) is the magnetising inductance of the transformer.

For transformers used in DAB converter, magnetising inductance \( L_M \) is usually much greater than the total leakage inductance \( L_k \) (\( L_k = L'_{k_1} + L_{k_2} \)); therefore, it is usually omitted from the analysis. In this study, in order to investigate the magnetising mechanism of the converter, it is considered an essential part of the circuit.

Now, assuming the potential of the common node \( N \) at the delivery side is zero, let us consider phase A and we have:

\[
\begin{align*}
L'_{k_1} & \frac{di_A}{dt} = v_A - v_{ma} - v_m \\
L'_{k_2} & \frac{di_A}{dt} = v_{ma} - v_a - v_m \\
L_M & \frac{di_{ma}}{dt} = v_{ma}
\end{align*}
\]

(7)

where \( v_{ma} \) and \( v_m \) are the potential of the common points of transformers and the secondary inverter with respect to point \( N \), respectively. Adding the first and second equations in Equation (7) together yields:

\[ L'_{k_1} \frac{di_A}{dt} + L'_{k_2} \frac{di_a}{dt} = v_A - v_a - v_m \]  

(8)

If the three phases are balanced, the summation of currents in each side is zero; therefore, the common-mode voltage \( v_m \) at the output side can be easily derived as

\[ v_m = \frac{1}{3} (v_A + v_B + v_C - v_a - v_b - v_c) \]  

(9)

If \( L'_{k_1} \) and \( L'_{k_2} \) are very small compared to \( L_M \) (i.e. less than one-tenth of \( L_M \)), the magnetising voltage of phase A transformer can be solved from Equations (7) and (9) and represented as Equation (10). Similarly, the magnetising voltage of phase B and C are obtained as shown in Equations (11) and (12):

\[ v_{ma} = \frac{1}{2} (v_A + v_a + v_m) - v_m \]  

(10)

\[ v_{mb} = \frac{1}{2} (v_B + v_b + v_m) - v_m \]  

(11)

\[ v_{mc} = \frac{1}{2} (v_C + v_c + v_m) - v_m \]  

(12)

Since three phases are balanced, solving equations from Equations (9) to (12) for \( v_m \), we have

\[ v_m = \frac{1}{3} (v_A + v_B + v_C) \]  

(13)

Substituting \( v_m \) into Equations (10) to (12) and rearranging for \( v_{ma} \), \( v_{mb} \) and \( v_{mc} \) results in Equation (14) that models the voltage across magnetising inductances with respect to the voltages at switching nodes:

\[
\begin{pmatrix}
    v_{ma} \\
    v_{mb} \\
    v_{mc}
\end{pmatrix} = \frac{1}{6} \begin{pmatrix}
    2 & -1 & -1 \\
    -1 & 2 & -1 \\
    -1 & -1 & 2
\end{pmatrix} \begin{pmatrix}
    v_A + v_a \\
    v_B + v_b \\
    v_C + v_c
\end{pmatrix} \tag{14}
\]

where \( v_a, v_b, v_c \) and \( v_A, v_B, v_C \) are the voltage at the switching nodes with respect to points \( N \) and \( n \), respectively. Values of \( v_x \), \( x \in \{A, B, C, a, b, c\} \), depend on switching states, or in other words, on the modulation strategy. As for three-phase DAB converters, the most popular modulation technique is single-phase-shift (SPS). The next section investigates the magnetisation voltage of transformers in the converter modulated by that scheme. Note here that, the derivation of Equation (14) is independent of modulation strategies; therefore, it can be used to examine the magnetisation voltage of transformers when other modulation techniques are applied.

### 3 | Magnetisation Under SPS Modulation

Conventionally, SPS technique is used to control both the amplitude and direction of the transferred power of DAB-like converters. In each bridge, three phases are interleaving by 120 electrical degrees and the duty cycle of each switch is 50%. Figure 3 demonstrates the phase current and magnetising voltage waveform when power is transferred in the forward direction. In the figure, \( Q_x \) refers to the primary-side switches of the respective phase, and \( S_x \) refers to secondary-side switches of the respective phase, \( x \in \{a, b, c\} \). There are six switching states in each half-cycle. Table 1 lists the magnetising voltage in each state at equilibrium. In the table, \( M = nV_d / V_1 \).
When the converter operates in the Buck mode (i.e. voltage conversion ratio is less than unity, \(M \leq 1\)), the transformer is magnetised from states 1 to 6 because its volt-second product is positive in that interval, and demagnetised for the rest of the present switching cycle because of the negative volt-second product as shown in Figure 3(a). On the other hand, when operating in the Boost mode (i.e. output voltage is greater than input voltage, \(M > 1\)), the transformer is magnetised from states 2 to 7 and demagnetised for the rest of the present switching period and the first state of the next period as can be seen in Figure 3(b). At the steady state, the positive and negative volt-second products must equal each other in amplitude and opposite in polarity, or in other words

\[
\lambda_p = -\lambda_N = \lambda
\]

where \(\lambda\) is the flux linkage amplitude in Webers. Formulating the magnetising voltage in each state and solving for \(\lambda\), the volt-seconds product in each mode is derived as Equation (15) where \(D_p\) is the phase shift ratio in p.u. and \(f_s\) is the switching frequency.

\[
\lambda = \begin{cases} 
\frac{V_1}{9f_s} (1 + M - 3MD_p), & M \leq 1 \\
\frac{V_1}{9f_s} (1 + M - 3D_p), & M > 1 
\end{cases}
\]  

(15)

Let \(A_c\) be the cross-section area of the magnetic core, the flux density swing in the core is determined by

\[
\Delta B = \frac{\lambda}{N_1A_c}
\]

(16)

Flux density determined by Equation (16) is \((1 + M - 3MD_p)\) or \((1 + M - 3D_p)\) times greater than that calculated by Equation (2) depending on the operation mode is Buck or Boost. The extra factors represent the dependency of flux density on phase shift and voltage at both DC terminals. For a given voltage conversion ratio, the maximum flux density is achieved when the phase shift is zero. As the voltage gain increases, the flux swing becomes larger. Therefore, care must be paid when designing transformers for DAB converters operating under a wide voltage range to avoid saturation. Besides, since Equation (16) also reflects the dependence of \(\Delta B\) on the phase shift \(D_p\), it can be used to investigate the flux walking phenomenon when a pulsating load occurs. However, these problems are beyond the scope of this study.

Now, let us compare the differences among all the investigated flux calculation methods. As aforementioned, the flux density in a converter can be computed by Equation (1) where \(K_f\) is the waveform coefficient, \(K_f = 4.0\) for square wave, and 4.44 for sine wave. For three-phase DAB converters with Y-Y transformer configuration, the winding voltage has square form; hence, the \(K_f\) of 4.0 is applied:
The proposed estimation of flux swing can then be expressed as functions of the well-known calculation technique as

\[ \Delta B = \Delta B_{cvk_1} \]

\[ = \Delta B_{dabk_2} \]

\[ = \Delta B_{dabk_3} \]

where \( k_1, k_2, k_3 \) are the relative flux gains determined by

\[ k_1 = \begin{cases} \frac{4}{9} (1 + M - 3MD) & M \leq 1 \\ \frac{4}{9} (1 + M - 3D) & M > 1 \end{cases} \]

\[ k_2 = \begin{cases} 1 + M - 3MD & M \leq 1 \\ 1 + M - 3D & M > 1 \end{cases} \]

\[ k_3 = \begin{cases} \frac{4}{9} \times \frac{1+M-3MD}{1+M-3D} & M \leq 1 \\ \frac{4}{9} \times \frac{1+M-3D}{1+M-3D} & M > 1 \end{cases} \]

The values of \( k_1, k_2 \) and \( k_3 \) represent how the well-known flux estimation techniques differ from the proposed flux formulation. The unity value of \( k_1, k_2 \) or \( k_3 \) means the corresponding flux equation results in the same flux density as the proposed equation.

Figure 4(a) describes the \( k_1 \) characteristics versus voltage conversion ratio \( M \) under several phase-shift ratio conditions. It is showed that the gap between the proposed estimation method and the conventional calculation method is significantly large of about \(-40\% \sim 30\%\) when the voltage gain is in the range of \([0.5, 2]\), which is significantly large. For a given voltage gain, there is an additional 10% variation in the calculation of flux density when using the conventional Equation (1) when the phase shift varies from zero to 30 degrees (0.0833 p.u.) as shown in Figure 4(b).

The same trends can be observed for the \( k_2 \) characteristics as shown in Figure 5 when considering the difference between the flux estimation by Equation (2) and the proposed method. However, in this case, the gap is much greater than that in the previous comparison. For a given phase shift, the deviation in flux when \( M \) varies in the range of \([0.5, 2]\) is from 50% to mostly 300%, which is enormously large. Furthermore, under a given voltage ratio, an additional 25% flux change can be seen. These observations imply that the flux calculated with Equation (2) is much smaller than that obtained by the proposed formula in Equation (16).

On the contrary, the flux estimation by Equation (3), which is dedicated to single-phase DAB converter results in much greater flux density than that obtained with Equation (16) as shown in Figure 6. In the range of voltage ratio being under consideration of \([0.5, 2]\), the relative gain \( k_3 \) has values among 0.405 to 0.44 meaning that the amplitude of flux density is more than 200% greater than that obtained by Equation (16).

Figure 4 Comparison between the proposed estimation (16) and the conventional calculation (1): (a) Relative gain versus voltage ratio, (b) relative gain versus the phase-shift ratio

Now, let us investigate the accuracy of the proposed flux calculation Equation (16) in more detail in the case study below.

4 | CASE STUDY

The proposed flux calculation method is applied to design transformers for a 10 kW three-phase DAB converter. Details of the design and optimisation process were reported in [22]. The converter is designated for the battery quick charger application. Specification of the system is provided in Table 2. The transformers were designed by Ap method described in [7]. Table 3 shows the number of turns of transformer windings when applying the proposed flux calculation in comparison to that obtained with the other approaches. Due to the specification, the transformer winding ratio is chosen to be unity.
FIGURE 5  Comparison between the proposed estimation (16) and the conventional calculation (2): (a) Relative gain versus voltage ratio, (b) relative gain versus the phase-shift ratio

TABLE 2  System specification

| Symbol | Description         | Value    | Unit |
|--------|---------------------|----------|------|
| $V_1$  | Input voltage       | 380      | V    |
| $V_2$  | Output voltage      | 320–420  | V    |
| $P_{\text{max}}$ | Max power         | 10       | kW   |
| $L_k$  | Total leakage inductance | 5     | $\mu$H |
| $L_m$  | Magnetising inductance | 1.3   | mH   |
| $D_p$  | Nominal phase-shift ratio | 0.0426 | p.u. |
| $f_{\text{sw}}$ | Switching frequency | 75       | kHz  |
| $R_{AC}$ | AC resistance      | 36.5     | m$\Omega$ |
| $\Delta B$ | Flux density swing | 270      | mT   |
| $n$    | Winding ratio       | 1:1      |      |

FIGURE 6  Comparison between the proposed estimation (16) and the conventional calculation (3): (a) Relative gain versus voltage ratio, (b) relative gain versus the phase-shift ratio

TABLE 3  Number of turns comparison

| Description | Equation (16) | Equation (17) | Equation (2) | Equation (3) | Unit |
|-------------|---------------|---------------|--------------|--------------|------|
| Pri. turns  | 15            | 18            | 8            | 35           | Turns |
| Sec. turns  | 15            | 18            | 8            | 35           | Turns |

As shown, the conventional calculation using Equation (17) underestimates the number of turns; therefore, it suggests more turns of the winding for the same flux density, compared to the proposed formula. Equation (3), which is originally derived for single-phase DAB converter, even underestimates the number of turns far more as it recommends almost twice more turns for the same flux density. This may lead to a higher utilisation factor of the core, a greater number of layers, higher winding resistance (both DC and AC), and thus higher copper loss. On
| Description               | Equation (16) | Equation (17) | Equation (2) | Equation (3) | Unit |
|---------------------------|---------------|---------------|--------------|--------------|------|
| Flux density $(\Delta B)$ | 264.6 mT      | 317.4 mT      | 141.1 mT     | 621.4 mT     |      |
| Core loss $(\Delta P_{fe})$ | 3.84 W        | 6.22 W        | 0.74 W       | 36.12 W      |      |

the contrary, the number of turns obtained by Equation (2) is overestimated; hence, only half turns are suggested by the equation. Consequently, the core might be saturated in the runtime if the margin is not high enough. Besides, since the actual flux density might be greater, core loss, as well as transformer temperature, might be higher than expected.

Now, assuming a transformer of 15 turns per winding, the flux is calculated according to all the aforementioned approaches. After that, the core loss is computed using the Steinmetz equation and put into comparison as reported in Table 4. As shown, for the same number of primary turns, under the same working condition (i.e. frequency, voltage, control angle), a calculation based on Equations (17) and (3) results in higher flux density and thus higher core loss. In particular, the core loss computed with the flux obtained by Equation (3) is mostly 10 times higher than that achieved with the proposed approach. This might lead to a misapprehension of the temperature rise problem of the core; thus, an extensive cooling effort might be unnecessarily paid or a larger core with a lower utilisation factor might be wastefully chosen. On the other hand, the flux swing obtained with Equation (2) is only half of that with Equation (16) resulting in a very small core loss. As a consequence, an underestimation of the heat management issue of transformers might occur.

5 EXPERIMENTAL EVALUATION

A 10-kW converter system, whose design and optimisation process were reported in [22], is employed to validate the proposed flux estimation formula. The designed transformers have two windings of 15 turns per each and are arranged according to the core-typed style. Suitable insulation is added between the two windings aiming to integrate the inductor into the transformer. The converter is depicted in Figure 7, and its specification is listed in Table 2.

In this study, in order to validate the proposed flux calculation technique, a finite element analysis (FEA) using the finite-element-method-magnetic (FEMM) version 4.2 software [23] is employed. The peak flux density, which is half of the flux swing, is examined at several operation points. In the simulation, the core material is set to linear with a fixed relative permeability. The current flowing through primary and secondary windings are set manually based on offline calculations at each operation point. The relative permeability of the core is set to be 1600 as suggested in [2].

Figure 8 shows the FEA result at 6 kW power transmission. The simulated peak flux density is 112 mT at the centre core. Note that the centre core of ETD54 has a round cross-section area, whereas that seen by FEMM is a rectangle. Taking that into account, one must adjust the value by multiplying the simulated flux density by $\pi/4$, which results in 142 mT. As the calculated value according to Equation (16) is 135.8 mT, which is 6 mT away from the adjusted value, it can be concluded that the calculation gives a good insight into the flux density.

Also, note that the value of 142 mT shown above is the flux density measured at the air gap. The contours seen in Figure 8 indicate that the flux is not evenly distributed throughout the core. At the corner of the window area, the density is much higher than that along the air gap. For instance, for the given case study, the flux density at the inner top right corner is 346 mT that is very close to the saturation flux density of 390 mT (at 100°C) of the core [1]. Therefore, care must be paid when designing the transformer to avoid core saturation. The maximum flux density should be chosen to be no greater than 50% of the saturation level given by the core manufacturer. In this
case study, the selected value is one-third of the saturation flux density, giving the core a safe margin to operate.

At the other operating points, the mismatch between the proposed estimation and FEA simulation results is a little larger but still insignificant as shown in Figure 9. The largest gap of only ±5 mT (about 4% of the simulated value) appears at the minimum and maximum powers. Therefore, it can be concluded that the estimation is consistent with the FEA results.

Since it is difficult to measure the flux in experiments, a transformer temperature evaluation is conducted instead. According to [7], the theoretical temperature rise in °C/W is predicted by Equation (21):

$$\Delta T = 450 \left( \frac{\Delta P_{cu} + \Delta P_{fe}}{A_t} \right)^{0.826}$$  \hspace{1cm} (21)

where $\Delta P_{cu}$ and $\Delta P_{fe}$ are the copper and core losses, respectively; $A_t$ is the total surface area of the transformer in cm$^2$.

Among the two loss components, the core loss $\Delta P_{fe}$ depends on flux density; thus, it is calculated using both the proposed and the conventional method of Equation (1). Meanwhile, in experiments, the transferred power is increased gradually until reaching the rated value. At each point, the saturated temperature of transformers is recorded using a thermal camera.

For instance, the thermal distribution of transformers at 6.6 kW power transmission is illustrated in Figure 10. At this point, the hottest spot is about 60°C, recorded at the transformer winding, whereas the core temperature is about 54°C. The ambient temperature when conducting the experiment was 15°C. In the meanwhile, the estimation using the conventional and the proposed approaches result in 64.7 and 59.1°C, respectively. Obviously, the proposed estimation method gives a better prediction of the transformer temperature. The full power range comparison is indicated in Figure 11.

As shown in Figure 11, the temperature estimation using the proposed formula matches the measured data well, especially in the mid- to high-power range. Although the temperature estimation error is quite remarkable of 6°C at 1.8 kW, the accurate temperature prediction at high power allows to have a suitable cooling effort of the transformers. On the other hand, the conventional flux estimation according to Equation (1) leads to an overestimation of 5 to 11°C compared to the measured data. Estimation using Equation (2) also results in an incorrect heat prediction as it underestimates the temperature by about 13°C. Equation (3) also shows its unsuitability when applying to estimate the flux density of a three-phase DAB converter, as it is derived for a single-phase variant. Therefore, it can be concluded that the temperature estimation can be improved by the proposed flux calculation formula in Equation (16).

Figure 12 shows a phase-to-ground voltage (yellow curve) and a phase current (pink curve) waveforms, measure at the primary side (sending side) of the converter when transferring a 9.3 kW power. The current shape demonstrates a right waveform when the voltage conversion ratio is unity with some ‘quite flat’ intervals. The ‘quite flat’ intervals indicate that the magnetising inductance is not changing during the interval; otherwise, a disrupt change of current would be observed.
Instead, there is a small difference of about 0.5 A between two peaks of the current due to the presence of the magnetising current.

6 | CONCLUSION

This study proposed a flux estimation technique for three-phase DAB DC/DC converters. As analysed, the flux in DAB converters depends on load and voltage gain condition. The conventional method that is based on input voltage, switching frequency, core area, and the number of turns does not reflect the dependence but results in a constant flux value regardless of load or output voltage. That problem is improved with the proposed flux estimation formula. FEA shows that the proposed formula gives a very good prediction of flux swing in the core as the error is less than 4%, compared to the FEA results. The temperature profile of the transformers can also be predicted better thanks to the proposed flux estimation. The formula can be utilised in designing transformer as well as in investigating converter behaviour under critical conditions such as pulsating load, starting processes, and so forth.

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