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Published in:
Proceedings of 10th International Conference on Power Electronics

Publication date:
2020

Document Version
Peer reviewed version

[Link back to DTU Orbit](#)

Citation (APA):
Huang, J., Zhang, Z., Xiao, Y., Sun, B., & Andersen, M. A. E. (2020). An Integrated Three-phase Transformer for Partial Parallel Dual Active Bridge Converter. In *Proceedings of 10th International Conference on Power Electronics* [8797262] IEEE.

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An Integrated Three-phase Transformer for Partial Parallel Dual Active Bridge Converter

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Abstract-- In this paper, an integrated three-phase transformer is utilized in the partial parallel dual active bridge (PPDAB) converter. Due to the symmetrical structure, the flux generated by the primary windings is equally distributed to the three outer legs. Given the same number of turns of the windings in each parallel modules, the ac currents can inherently balance in the parallel modules. In order to enlarge the leakage inductor to serve as the interfacing ac inductor in the PPDAB converter, the coupling between the primary windings and secondary windings is weakened by placing the primary and secondary windings on the center and outer legs, respectively. Compared to the discrete three-phase transformers, the winding loss will also decrease with the proposed structure. A 1kW 400V/50V 500 kHz prototype with an efficiency of 96.72% is built and tested to verify the theoretical analysis.

Index Terms-- DAB, dc-dc, integrated magnetics, three-phase transformer

I. INTRODUCTION

Dual active bridge (DAB) converter is widely used in energy storage systems, automotive applications, smart transformers, etc., due to its attractive characteristics such as bidirectional power flow capability, symmetrical configuration and zero voltage switching (ZVS) [1]-[8].

In [6], an overview of the DAB converter for high-frequency-link power conversion systems was given.

In [7], a novel neutral point clamped (NPC) DAB converter with a blocking capacitor is proposed for energy storage systems. By inserting a blocking capacitor in the primary loop of the NPC DAB converter, the proposed topology adapts to wide output voltage range applications.

In [8], one DAB-based architecture, quadruple active bridge (QAB) converter is proposed to be used as a basic module of a modular three-stage solid-state transformer (SST) or smart transformer. A 20kW prototype of the smart transformer was built to demonstrate the high potential of the QAB converter on efficiency and cost.

Achieving high power density is also emerging as a goal in power electronics applications. Passive components such as inductor, capacitor, and transformer often occupy the vast majority of the volume of the converter. Besides, increasing the switching frequency to reduce the size of passive components, integrated magnetics are also widely used in power electronic converters [9]-[15].

Moreover, integrated magnetic techniques are widely applied to high power applications.

In [14], a matrix transformer structure was proposed for a 1 MHz, 380V/12V, 800W LLC converter. The concept

Fig. 1. Topology of the partial parallel dual active bridge converter

of flux cancelation is used to reduce core loss and core size.

In [15], a 6-layer PCB winding three-phase transformer is reported. The leakage inductor can be controlled to the desired value by adjusting the reluctance of the additional center post to serve as the resonant inductor. However, the additional center post will also increase the core loss.

In this paper, a new integrated three-phase transformer is proposed for the partial parallel dual active bridge (PPDAB) converter, shown in Fig. 1. The PPDAB converter was first proposed and researched in [4]. Based on the dual active bridge (DAB) topology, the transformer windings on the high-voltage (HV) side are connected in series; on the other hand, the full-bridges units on the low-voltage (LV) side are connected in parallel, making the voltage gain and the transmission power significantly increased.

For the PPDAB converter, interfacing inductors are needed in the ac link in order to enable regulation of power flow between the HV and LV sides. Therefore, the proposed three-phase transformer integrates not only three transformers but also the ac inductor into one magnetic structure so that further reduces the size and volume of magnetic components. Moreover, the winding loss can be reduced accordingly with the proposed magnetic integration method. The symmetrical structure leads to the flux generated by the primary windings equally distributed to the three outer legs. Given the same number of turns of the windings in each parallel modules, the high-frequency ac currents are balanced on the LV side.

Fig. 2. Typical waveform of SPS control

This paper is organized as follows. After the introduction, Section II analyzes the operation of the PPDAB converter. Section III gives the detailed analysis on the magnetic integration of three-phase transformers and ac inductors, and compares it with the discrete three-phase transformer, which is comprised of three sets of UI core. Section IV presents the experimental results on the 1kW three-phase PPDAB converter to verify the converter's analysis and operation with the newly proposed integrated three-phase transformer. Finally, conclusion is given in Section V.

II. PARTIAL PARALLEL DUAL ACTIVE BRIDGE DC/DC CONVERTER

The typical waveforms of the PPDAB converter with single-phase-shift (SPS) control are shown in Fig. 2, where φ represents the phase-shift angle as a percentage of the switching period T_s , f_s is the switching frequency, and L_{ac} is the total leakage inductance referred to the HV side. Moreover, the theoretical analysis below is based on the following assumptions: 1) All the switches are ideal. 2) The input voltage V_{in} and output voltage V_{out} are constant.

[t_0, t_1]

In this subinterval, the voltage across the ac inductor is $V_{in} + 3nV_{out}$. The ac inductor current increases linearly.

$$i_{Lac}(t_1) = i_{Lac}(t_0) + \frac{V_{in} + 3nV_{out}}{L_{ac}} \cdot \varphi T_s \quad (1)$$

[t_1, t_2]

In this subinterval, the voltage across the ac inductor is $V_{in} - 3nV_{out}$.

$$i_{Lac}(t_2) = i_{Lac}(t_1) + \frac{V_{in} - 3nV_{out}}{L_{ac}} \cdot \left(\frac{1}{2} - \varphi\right) T_s \quad (2)$$

Similarly, $i_{Lac}(t_3)$ and $i_{Lac}(t_4)$ can be derived as expressed in (3) and (4),

$$i_{Lac}(t_3) = i_{Lac}(t_2) - \frac{V_{in} + 3nV_{out}}{L_{ac}} \cdot \varphi T_s \quad (3)$$

$$i_{Lac}(t_4) = i_{Lac}(t_3) + \frac{-V_{in} + 3nV_{out}}{L_{ac}} \cdot \left(\frac{1}{2} - \varphi\right) T_s \quad (4)$$

Due to the symmetry of the ac inductor current, $i_{Lac}(t_0) = -i_{Lac}(t_2)$. So $i_{Lac}(t_0)$ can be derived,

$$i_{Lac}(t_0) = -\frac{V_{in} + 3nV_{out}}{4L_{ac}} \cdot (4\varphi - 1) T_s \quad (5)$$

The average current of ac inductor L_{ac} in a half period is expressed in (6).

$$\begin{aligned} \bar{I} &= [i_{Lac}(t_0) + i_{Lac}(t_1)] \cdot \varphi + [i_{Lac}(t_1) + i_{Lac}(t_2)] \cdot \left(\frac{1}{2} - \varphi\right) \\ &= \frac{3nV_{out}}{L_{ac}} \varphi (1 - 2\varphi) T_s \end{aligned} \quad (6)$$

Hence, the PPDAB converter's steady-state power equation can be calculated,

$$P = V_{in} \cdot \bar{I} = \frac{3nV_{in}V_{out}}{f_s L_{ac}} \varphi (1 - 2\varphi) \quad (7)$$

Assume the load Z_{load} is fixed, the voltage gain can be derived as (8).

$$\frac{V_{out}}{V_{in}} = \frac{3nZ_{load}}{f_s L_{ac}} \varphi (1 - 2\varphi) \quad (8)$$

Compared to the DAB converter, the PPDAB converter has the following advantages:

- The voltage gain is proportional to the number of paralleled modules.
- The ac current balancing between the parallel modules is inherently ensured by the series connection of the windings on the HV side.
- In addition to single-phase-shift (SPS), the PPDAB converter can be regulated by adding phase shifts among the parallel modules.

III. INTEGRATED THREE-PHASE HIGH-FREQUENCY TRANSFORMER

A. Integrated three-phase transformer

As shown in Fig. 1, the three-phase transformer is a key component to achieve galvanic isolation and voltage converting. In order to improve the performance of the PPDAB converter, a newly integrated three-phase transformer is proposed. The structure is shown in Fig. 3(a) and the equivalent circuit is depicted in Fig. 3(b). The dash lines in Fig. 3(b) means the coupling between the windings. The coils of the HV side, winding-P are wound around the center leg. In addition, the coils of the three parallel modules, winding-A, winding-B, winding-C are wound around three outer legs, respectively. The cross-sectional area of the center leg is equal to the total area of the three outer legs to ensure the same flux density along the magnetic paths.

Fig. 3. Three-phase integrated transformer and ac inductors. (a) configuration of the integrated three-phase transformer, and (b) the equivalent circuit of the integrated three-phase transformer

B. Analysis on ac current balancing

The reluctance model of the proposed integrated three-phase transformer is shown in Fig. 4. Where $F=NI$ is the magneto motive force (MMF), N is the number of turns in the coil and I is the current through the circuit. $\mathcal{R}=l/(\mu A)$, where l is the length of flux path, μ is the permeability of the material and A is the cross-sectional area. Due to the symmetrical structure, the following assumptions are made:

- The reluctances of three outer legs are the same. Define: $\mathcal{R}_A = \mathcal{R}_B = \mathcal{R}_C = \mathcal{R}_S$
- The reluctances between two arbitrary legs are the same. These reluctances are relative to the thickness of the air gap. Define: $\mathcal{R}_{AB} = \mathcal{R}_{AC} = \mathcal{R}_{BC} = \mathcal{R}_{SS}$

As shown in Fig. 4, the reluctance model of the center leg is split into three parts. Combine Ampere's Circuital Law, (9) can be obtained.

$$\begin{cases} \phi_A = \frac{N_A I_A}{3R_p + R_s} - \frac{N_p I_p}{3R_p + R_s} + \frac{(2N_A I_A - N_B I_B - N_C I_C) \cdot 9R_p^2}{(3R_p + R_s)(9R_p R_s + 6R_p R_{ss} + 2R_s R_{ss})} \\ \phi_B = \frac{N_B I_B}{3R_p + R_s} - \frac{N_p I_p}{3R_p + R_s} + \frac{(2N_B I_B - N_A I_A - N_C I_C) \cdot 9R_p^2}{(3R_p + R_s)(9R_p R_s + 6R_p R_{ss} + 2R_s R_{ss})} \\ \phi_C = \frac{N_C I_C}{3R_p + R_s} - \frac{N_p I_p}{3R_p + R_s} + \frac{(2N_C I_C - N_A I_A - N_B I_B) \cdot 9R_p^2}{(3R_p + R_s)(9R_p R_s + 6R_p R_{ss} + 2R_s R_{ss})} \\ \phi_p = \frac{N_A I_A + N_B I_B + N_C I_C}{3R_p + R_s} - \frac{3N_p I_p}{3R_p + R_s} \end{cases} \quad (9)$$

Fig. 4. Reluctance model

where ϕ_p is the total flux in the center leg, and ϕ_A, ϕ_B, ϕ_C are the flux in the three outer legs, respectively.

Due to the symmetrical structure and the same switching patterns of the three parallel modules, the following equation can be derived:

$$\phi_A = \phi_B = \phi_C = \frac{1}{3} \phi_p \quad (10)$$

Combining (9) and (10), (11) can be obtained.

$$N_A I_A = N_B I_B = N_C I_C \quad (11)$$

Therefore, it can be seen that when the numbers of turns of winding-A, winding-B, winding-C are kept the same, the ac currents in each parallel modules will be balanced.

C. Leakage inductance

By placing the primary and secondary windings on the center leg and outer legs, respectively, the leakage inductance of the transformer can be increased. In order to ensure the leakage inductor is large enough to serve as the ac inductor in the PPDAB converter, the finite element analysis (FEA) is used to design the windings structure of the transformer.

D. Comparison between the integrated three-phase transformer and discrete three-phase transformer

Winding loss is frequency-dependent due to the skin effect and proximity effect. Both effects cause the current unevenly distributed in the cross section of the conductor and cause a high winding resistance at high frequency.

The windings of the integrated three-phase transformer and the discrete three-phase transformer are depicted in Fig. 5.

With the FEA method, the winding loss can be simulated. In order to make a fair comparison, the parameters of the two three-phase transformers are listed in Table I and the simulation results are listed in Table II. From this comparison, we can conclude that the proposed integrated three-phase transformer has a lower winding loss.

Fig. 5. Comparison of two three-phase transformer structures. (a) discrete and (b) integrated.

TABLE I
PARAMETERS OF THE TWO THREE-PHASE TRANSFORMERS

Parameter	Value	Description
	U-37.5-8.525-18.5	Core dimensions(integrated)
	U-32.6-8.525-18.5	Core dimensions(discrete)
f / kHz	500	Switching frequency
w_{HV} / mm	6	Winding width on the HV side
w_{LV} / mm	4.25	Winding width on the LV side
h	2oz (0.07mm)	Winding thickness
δ / mm	0.5	Distance between any two arbitrary turns
σ / mm	0.254	Space between the core and turns
$Tr-P,A,B,C$	8:3:3:3	Number of turns
m_{HV}	2	Number of layers of windings in HV side (in series)
m_{LV}	2	Number of layers of windings in LV side (in parallel)
I_{HV} / A	3	Current amplitude of HV side
I_{LV} / A	8	Current amplitude of LV side

TABLE II
SIMULATION RESULT OF THE WINDING LOSS

	Integrated three-phase transformer	Discrete three-phase transformer
Winding loss / W	1.9125	2.5065

Fig. 6. Picture of the PPDAB converter hardware prototype

TABLE III
SPECIFICATIONS OF THE PROTOTYPE

Parameter	Value	Description
V_{in_nom}	400V	Nominal DC input voltage
V_{out_nom}	50V	Nominal DC output voltage
P_r	1kW	Rated power
f_s	500kHz	Switching frequency
$Tr-P,A,B,C$	8:3:3:3	Number of turns
Transformer	ML91S	Material
Transformer	U-37.5-12.2-18.5	Core dimensions
L_{ac}	22.0213μH	Total leakage inductance referred to the HV side
ϕ	10%	Phase shift between the HV side and LV side
Digital Controller	TMS320F28379D	Digital signal processing
Power semiconductor	GS66504B	Switches on the HV side
Power semiconductor	EPC2021	Switches on the LV side

IV. EXPERIMENTAL RESULTS

In order to verify the proposed converter and its associated magnetic integration, a 1kW three-phase PPDAB converter with the newly integrated three-phase transformers was constructed and tested. The photo of the built prototype is shown in Fig.6 and the prototype specifications are summarized in Table III.

Two winding techniques i.e. wire-wound and print circuit board (PCB) winding are tested and compared for the proposed magnetic structure. Fig.7 shows the wire-wound transformer and its windings arrangement, and Fig.8 shows the solution with PCB windings and its windings arrangement. The transformer in Fig. 8 has a lower leakage inductance due to the windings interleaving.

Fig. 7. Wire-wound transformer. (a) wire-wound transformer and (b) winding arrangement

Fig. 8. Planar transformer with PCB windings. (a) planar transformer and (b) winding arrangement

Fig. 9. Equivalent circuit on the HV side

TABLE IV
SIMULATION AND TEST DATA OF TRANSFORMERS

	Transformer with PCB winding		Transformer with Litz wire	
	Simulation	Measured	Simulation	Measured
$L_m / \mu H$	322.22	306.239	323.82	300.079
$L_k / \mu H$	7.578492	9.75077	19.78898	22.0213
C_p / pF	—	652.255	—	12.6

However, the large surface area of the PCB windings increases the winding parasitic capacitance. The simulation and measured results of the transformers with these two different windings are listed in Table IV, where L_m is the magnetizing inductance of the windings on the HV side and L_k is the total leakage inductance referred to the HV side and C_p is the parasitic capacitance of the windings on the HV side. The equivalent circuit of the HV side is shown in Fig. 9. L_w is the inductance of a wire which is placed in series in the circuit path for ease on current probing.

The waveform of the three-phase PPDAB converter with the planar transformer with PCB windings working at 200 kHz is shown in Fig. 10. The waveform is labeled according to Fig. 1.

It shows that after S1 and S3 turn off, the L_w firstly resonates with output capacitance C_{oss} of the semiconductor devices on the HV side. The resonant frequency f_{res1} can be calculated,

$$f_{res1} = \frac{1}{2\pi\sqrt{L_w \cdot C_{oss}}} \quad (16)$$

After S2 and S4 turn on, the LCL form by L_w , C_p , and L_k results in a high amplitude current resonance and causes higher conduction loss of the primary devices.

The resonant frequency f_{res2} can be expressed,

$$f_{res2} = \frac{1}{2\pi\sqrt{\frac{L_k L_w}{L_k + L_w} \cdot C_p}} \quad (17)$$

Finally, the calculation results of the resonant frequency are listed in Table V.

TABLE V
CALCULATION RESULT OF THE RESONANT FREQUENCY

Parameter	Value
$L_k / \mu H$	9.75077
C_p / pF	652.255
L_w / nH	95.1
C_{oss} / pF	80
f_{res1} / MHz	57.7
f_{res2} / MHz	20.306

(a)

(b)

(c)

Fig. 10. (a)The working waveform of the PPDAB converter with a planar transformer with PCB windings. (b)zoom in the resonant rings and measure the LC resonant frequency (c)zoom in the resonant rings and measure the LCL resonant frequency

It can be seen that the measured f_{res2} matches the calculated result, which verifies the resonance is because of LCL resonance of C_p , L_k , and L_w . However, there are still errors on the calculated f_{res1} and it will be studied further in the future work.

Therefore, in the final test, a wire-wound transformer shown in Fig. 7(a) is employed in the prototype. Fig. 11 provides the thermal picture of the proposed three-phase PPDAB converter working at 1kW, 400V/50V. The temperature of the devices in the HV side reaches 72°C, which is still within the safe temperature range. Fig. 12 gives the experimental waveforms of the proposed three-phase PPDAB converter operating at 1kW, 400V/50V. The waveforms are labeled according to Fig. 1. The proposed three-phase PPDAB converter has an efficiency of 96.72% at the rated power.

Fig. 11. Thermal image

(a)

(b)

Fig. 12. Measured waveforms at the rated power

V. CONCLUSIONS

In this paper, a new integrated three-phase transformer is proposed for the PPDAB converter. Compared to the discrete three-phase transformer, the proposed integrated one has a lower winding loss. In order to ensure the leakage inductor is large enough to serve as the ac inductor in the PPDAB converter, the FEA method is used to design the windings structure of the proposed transformer. From the experimental results, we are able to verify the theoretical analysis and can see the three high-frequency ac currents in the paralleled branches are well balanced. Finally, the efficiency is 96.72% at the rated power. Modeling the leakage inductance of the proposed integrated three-phase transformer and optimizing the PCB windings to reduce the parasitic capacitance will be carried out in the future work.

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