Capacitor Voltage Ripple Minimization of a Modular Three-Phase AC/DC Power Electronics Transformer With Four-Winding Power Channel

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ABSTRACT Power electronic transformer (PET) with modular cascaded structure has become a flexible interface between AC and DC grids. Nonetheless, the notable 2\textsuperscript{nd}-order ripple power introduced from the AC source requires considerable energy storage in the module capacitors, which is likely to degrade the overall power density and increase system cost. In this paper, a power channel based four-winding PET (PC-PET) is proposed to address such an issue. Owing to the phase-crossing power channels, the low-frequency ripples can cancel each other through magnetic flux counteracting. To improve the accuracy of power decoupling, the effect compromised by inconsistent windings is considered. Accordingly, a proportional-resonant (PR) control based power decoupling strategy for PC-PET incorporated with power balance is proposed to decouple the three-phase low-frequency ripple powers, and minimize the voltage ripple in module capacitors. Additionally, the improvement of power density is discussed based on analyses on power losses, capacitor volume and transformer volume. Simulation and experimental results verified the validity of proposed control strategies.

INDEX TERMS PET, power channel, power decoupling, QAB.

I. INTRODUCTION

Power electronic transformer (PET) contributes to intelligent power flow management while providing reduced weight/volume, improved power quality, enhanced reliability and so forth [1]–[3]. It is playing an important role in modern power system as energy routers to replace line-frequency transformer (LFT) [4], [5]. In three-phase AC/DC application, the PET can reduce the required volume and weight to nearly one third, halve the power loss while the material cost can still compete with LFT [6].

The configuration of AC/DC PET can be categorized into two general types: AC/HFAC (high-frequency AC)-HFAC/DC, and AC/DC-DC/HFAC-HFAC/DC. Tough control issues of matrix converters are hiding the former configuration from high voltage/high power applications. On the contrary, the AC/DC-DC/HFAC-HFAC/DC one is attracting more and more attentions owing to its mature control schemes [1], [7]. Cascaded H-bridge (CHB) and modular multilevel converter (MMC) based PET are both with AC/DC-DC/HFAC-HFAC/DC configuration. Featuring multiport capability and modular structure, these two types of PETs are commonly studied [8]. The MMC-PET [9]–[11] are usually applied when high-voltage DC bus is required. Generally, the topology of a CHB based PET inter-connecting three-phase medium-voltage (MV) AC and low-voltage (LV) DC grid is shown in Fig. 1 with N PET modules in each phase. Each module consists of a single-phase PWM rectifier as active front end and a dual-active-bridge (DAB) or multi-active bridge converter [5], [7], [12]. The 2\textsuperscript{nd}-order ripple...
powers are introduced from each AC-input-port and stored in DC capacitors of the modules. In order to reduce the 2nd-order ripple in capacitor voltage, large DC capacitance is usually required. This will increase the volume of the system and therefore reduce the power density.

![Diagram](image1.png)

**FIGURE 1.** The topology of normal three-phase AC/DC CHB-PET.

Power decoupling in single-phase PWM rectifier normally requires large passive capacitors to trap low frequency ripple powers [13]–[17]. As for three-phase systems, interaction between phases makes the ripple powers cancelling each other possible. Quad-active-bridge (QAB) equipped with three primary ports and one secondary port is an ideal way to function as power interaction channels between phases [18], [19]. The power channel based PET (PC-PET) can serve as high-power-density power electronic interface between three-phase three-wire AC and DC grid [20]. However, the three-phase power imbalance issue cannot be ignored.

In this paper a proportional-resonant (PR) control based power decoupling strategy for three-phase AC/DC PC-PET is proposed. Realized in stationary reference frame, the strategy provides suppression to multiple harmonics by paralleled controllers. The low-frequency voltage ripples in module capacitors can be minimized even in the situation of asymmetrical three-phase. The capacitance and volume of module capacitors can be reduced, thereby the power density can be improved.

In Section II, the PC-PET configuration is firstly analyzed. Then the 2nd-order ripple voltage in module capacitors and operating principle of PC-PET are discussed. In Section III, the equivalent circuit model of the QAB and its power dispatching characteristics are investigated. In Section IV, the overall control scheme including the proposed ripple voltage suppression strategy is illustrated. In Section V, simulation and experimental results are presented to demonstrate the proposed topology and control schemes. In Section VI, the power losses as well as capacitor volume are calculated based on analytical models. Finally, Section VII concludes this paper.

**II. SECOND-ORDER RIPPLE VOLTAGE ANALYSIS AND OPERATING PRINCIPLE OF PC-PET**

The circuit of PC-PET is shown in Fig. 2, with $N$ modules in each phase leg ($a_{1\sim a_N}$, $b_{1\sim b_N}$, $c_{1\sim c_N}$). The PC-PET is composed of three power conversion stages, i.e. level 1, level 2 and level 3 sub-converters. For modular design, the sub-converters from a same stage are identical with each other. The level 1 converter is selected as a H-bridge (HB) that functions as a PWM rectifier (RECM$_x$,$j$, $x = a, b$ and $c$, $j = 1 \sim N$). The level 2 converter (three HBs) and level 3 converter (one HB) act as a QAB. Each QAB is equipped with a four-winding high-frequency transformer with three primary ports and one secondary port. The primary ports are connected to the HFAC-outputs of the three power modules QABM$_x$,$j$, corresponding to the $j$th ($j = 1 \sim N$) module in phase $x$ ($x = a, b$ and $c$). The secondary port is tied with the HFAC-input of the level 3 converter. The total $N$ DC ports are paralleled to form LVDC-bus.

**A. SECOND-ORDER RIPPLE VOLTAGE ANALYSIS**

The 2nd-order ripple power imported from the AC port and the induced 2nd-order ripple voltage in DC capacitors are analyzed as follows.

Assume the input voltage and current of phase $a$ are:

$$u_{sa} = U_s \sin(\omega t),$$

$$i_{sa} = I_s \sin(\omega t - \theta),$$

where $U_s$ and $I_s$ are the amplitudes, $\omega$ and $\theta$ are fundamental angular frequency and power factor angle respectively. Taking RECM$_a$,$j$ for example, the power flowing to the AC port is expressed as:

$$p_{s,aj} = \frac{u_{sa} \cdot i_{sa}}{N} = \frac{U_s I_s}{2N} \cos \theta - \frac{U_s I_s}{2N} \cos(2\omega t - \theta).$$

It is supposed that losses of the all the active and passive components can be ignored. The DC component of $p_{s,aj}$ is

![Diagram](image2.png)

**FIGURE 2.** The topology of modular PC-PET interfacing three-phase MVAC and LVDC power grids.
directly determined by the DC load.

\[ \frac{U_s I_s}{2N} \cos \theta = \frac{p_o}{3N} \quad (4) \]

where \( p_o \) is the power consumed by the load.

If the 2nd-order ripple powers of \( p_{s,aj} \) (\( p_{s,aj} \)) are totally absorbed by module capacitors, the instantaneous power of capacitor in RECM \( _{a,j} \) is

\[ p_{C,aj} = u_{C,aj} \cdot i_{C,aj} = \frac{u_{sa} \cdot i_{sa} - i_{sa} \cdot L_s \frac{di_{sa}}{dt}}{N} - p_{dc} \quad (5) \]

where \( u_{C,aj} \) and \( i_{C,aj} \) are the capacitor voltage and capacitor current of RECM \( _{a,j} \), \( L_s \) is the line-frequency inductance. \( p_{dc} \) is the power transferred to the next conversion stage, which only contains DC component.

\[ p_{dc} = \frac{p_o}{3N} \quad (6) \]

Hence \( u_{C,aj} \) mainly contains DC and 2nd-order components, as expressed by:

\[ u_{C,aj} = U_{Cdc,aj} + u_{C2,aj} \quad (7) \]

Then \( i_{C,aj} \) can be rewritten to:

\[ i_{C,aj} = C \cdot \frac{d(u_{C,aj})}{dt} = C \cdot \frac{d(U_{Cdc,aj} + u_{C2,aj})}{dt} \]

\[ = C \cdot \frac{du_{C2,aj}}{dt} \quad (8) \]

where \( C \) is the capacitance of RECM.

Substituting (2) - (4) and (6) - (8) to (5), devices (9).

\[ (U_{Cdc,aj} + u_{C2,aj}) \cdot C = \frac{du_{C2,aj}}{dt} \]

\[ = \frac{-U_s I_s}{2N} \cos(2\omega t - \theta) - \frac{\omega L_s I^2_s}{2N} \sin(2\omega t - 2\theta) \quad (9) \]

Considering zero initial voltage for \( u_{C2,aj} \) and integrating both sides of (9) by time, (10) can be derived:

\[ \frac{1}{2} C u^2_{C2,aj} + C U_{Cdc,aj} u_{C2,aj} \]

\[ = \frac{-U_s I_s}{40N} \sin(2\omega t - \theta) + \frac{L_s I^2_s}{4N} \cos(2\omega t - 2\theta) \quad (10) \]

Then \( u_{C2,aj} \) can be solved from (10), as expressed by (11).

\[ u_{C2,aj} = -U_{Cdc,aj} + \sqrt{ \frac{U^2_{Cdc,aj}}{A} - A \cdot \sin(2\omega t - 2\theta + \psi) } \quad (11) \]

where

\[ A = \sqrt{ \frac{p_o^2}{9(\omega CN)^2} + \frac{p_o}{3\omega CN} \sin \theta - \frac{4L_s I^2_s}{18U^2_s CN \cos^2 \theta} } \]

\[ \psi = \arctan \left( \frac{\tan \theta - \frac{2\omega L_s p_o}{3U^2_s C \cos^2 \theta}}{} \right) \]

The relationship between the amplitude of \( u_{C2,aj} \) (\( U_{C2,aj} \)). \( C \) and \( p_o \) is shown in Fig. 3. \( U_{C2,aj} \) will increase with the increasing of \( p_o \) or the decreasing of module capacitance.

Similarly, 2nd-order ripple voltage of module capacitors of phase b and phase c are expressed as:

\[ u_{C2,bj} = -U_{Cdc,bj} + \sqrt{ \frac{U^2_{Cdc,bj} - A \cdot \sin(2\omega t - 2\theta + 2\pi/3 + \psi)}{A} } \quad (12) \]

\[ u_{C2,cj} = -U_{Cdc,cj} + \sqrt{ \frac{U^2_{Cdc,cj} - A \cdot \sin(2\omega t - 2\theta + \pi/3 + \psi)}{A} } \quad (13) \]

### B. OPERATING PRINCIPLE OF PC-PET

If the powers flowing through three phases are balanced, the input powers of RECM \( _{b,j} \) and RECM \( _{c,j} \) are:

\[ p_{s,bj} = \frac{U_s I_s}{2N} \cos \theta - \frac{U_s I_s}{2N} \cos(2\omega t - \theta + 2\pi/3) \quad (14) \]

\[ p_{s,cj} = \frac{U_s I_s}{2N} \cos \theta - \frac{U_s I_s}{2N} \cos(2\omega t - \theta - 2\pi/3) \quad (15) \]

The 2nd-order ripple powers of three phase legs are symmetrical with negative sequence. If the ripple powers are forced to flow through the three primary ports of the four-winding transformer, they will be counteracted through flux coupling. That is the reason why the QAB can be regarded as a "power channel". Only DC component will be reflected on the LVDC bus.

Consequently, there is no low-frequency voltage ripple in all the capacitors of PC-PET. Module capacitors are just required for the purpose of filtering high-frequency switching harmonics. It can sharply reduce the capacitance, the volume as well. In addition, owing to the highly integrated multi-winding structure, the transformer volume is smaller than conventional two-winding ones. All of these make the PC-PET very competitive in terms of power density.

### III. EQUIVALENT CIRCUIT MODELING OF THE QAB IN PC-PET

Considering switching frequency components, the controlled voltage source based equivalent circuit of QAB is shown in Fig. 4.

Windings \( N_1 \sim N_3 \) are the three primary windings of QAB, corresponding to phase \( a \sim c \). \( N_4 \) is the secondary winding. The inductor \( L_k \) \((k = 1 \sim 3)\) represents the self-leakage inductance of \( N_1 \sim N_3 \).

Assuming the phase-shift control (PSC) is employed, the averaged power transferred from port \( k \) to port 4 in a
The switching period can be expressed as [20]:

\[ p_{k4} = \frac{1}{T_s} \int_0^{T_s} v_k i d\tau = \frac{v_k v_4}{2\pi^2 f_k L_{k4}} \varphi_{k4}(\pi - \varphi_{k4}). \]  

(16)

where \( v_k \) and \( v_4 \) are the voltages of port \( k \) and port 4, \( f_k \) is the switching frequency, \( T_s \) is the switching period, \( L_{k4} \) is the equivalent inductance between port \( k \) and port 4, and \( \varphi_{k4} \) is the phase-shift angle between port \( k \) and port 4.

\[ v_{TH4_{k4}} \] is the Thevenin equivalent voltage, that is calculated by

\[
\begin{align*}
    v_{TH4_{k4}} &= \frac{L_2 L_3}{L_1 L_3 + L_1 L_2 + L_2 L_3} \cdot v_4 \\
    v_{TH4_{k4}} &= \frac{L_1 L_2 + L_2 L_3 + L_1 L_3}{L_1 L_4} \cdot v_4 \\
    v_{TH4_{k4}} &= \frac{L_1 L_3 + L_2 L_3 + L_1 L_2}{L_1 L_4} \cdot v_4.
\end{align*}
\]

(17)

Due to the imperfect manufacturing of multi-winding transformer, the leakage inductance of each winding will not be completely consistent. Supposing that the difference value between \( L_1 \) and \( L_2 \) is \( \Delta L_{12} \), and \( \Delta L_{13} \) is the difference value between \( L_1 \) and \( L_3 \). The power deviation caused by difference between leakage inductances is addressed. The \( L_{TH} \) can then be expressed as:

\[
\begin{align*}
    L_{TH1} &= \left(1 + \frac{1}{L_{m1} + \Delta L_{12} + \frac{1}{L_1}} + \frac{1}{L_4}\right)^{-1} \\
    L_{TH2} &= \left(1 + \frac{1}{L_{m1} + \frac{1}{L_1}} + \frac{1}{L_4} + \frac{1}{L_3 + \Delta L_{12}}\right)^{-1} \\
    L_{TH3} &= \left(1 + \frac{1}{L_{m1} + \frac{1}{L_1}} + \frac{1}{L_4} + \frac{1}{L_3 + \Delta L_{12}}\right)^{-1}.
\end{align*}
\]

(19)

Power transferred from port \( k \) to port 4 can be expressed as:

\[ p_{k4} = \frac{v_k v_{TH4_{k4}}}{2\pi^2 f_k (L_m + L_{TH})} \varphi_{k4}(\pi - \varphi_{k4}). \]  

(20)

According to (16), (17), (19), (20), \( L_{k4} \) can be figured out:

\[
\begin{align*}
    L_{14} &= (L_1 + L_{TH1}) \\
            &\times \left[ L_4 \left(1 + \frac{1}{L_{m1} + \frac{1}{L_1}} + \frac{1}{L_3 + \Delta L_{12}}\right) + 1\right] \\
    L_{24} &= (L_2 + L_{TH2}) \\
            &\times \left[ L_4 \left(1 + \frac{1}{L_{m1} + \frac{1}{L_1}} + \frac{1}{L_3 + \Delta L_{12}}\right) + 1\right] \\
    L_{34} &= (L_3 + L_{TH3}) \\
            &\times \left[ L_4 \left(1 + \frac{1}{L_{m1} + \frac{1}{L_1}} + \frac{1}{L_3 + \Delta L_{12}}\right) + 1\right].
\end{align*}
\]

(21)

Substituting (21) to (16), taking \( p_{14} \) as the base value, the power difference value in per unit is:

\[
\begin{align*}
    \Delta p_{12} \text{(p.u.)} &= \frac{p_{14} - p_{24}}{p_{14}} = 1 - \frac{L_{14}}{L_{24}} \\
    \Delta p_{13} \text{(p.u.)} &= \frac{p_{14} - p_{34}}{p_{14}} = 1 - \frac{L_{14}}{L_{34}}.
\end{align*}
\]

(22)
of three-phase system, the voltage regulation of phase $c$ is acquired by the voltage loop of phase $a$ and phase $b$.

The control block diagram of QAB is shown in Fig. 8(b). The overall control of QAB contains PSC, ripple power decoupling control and power balance control. In the start-up process, the primary HB of QAB works in PWM mode. Duty cycle gradually increases from 0% to 50%, and then switches to PSC.

### IV. Proposed Capacitor Voltage Ripple Suppression Method

The overall control scheme of PC-PET is shown in Fig. 8. The dual-loop control of PWM rectifier is adopted. And the control block diagram for RECM$_{x,j}$ ($x = a, b$ and $c$, $j = 1\sim N$) is shown in Fig. 8(a). As the physical constraints
B. POWER BALANCE CONTROL

Power imbalance among phases commonly occurs due to inevitable inconsistent circuit parameters, such as different leakage inductance of transformer windings. Consequently, the ripple powers are hard to be completely counteracted, which degrades the effect of ripple power decoupling. Therefore, a power balance control strategy is introduced. Its control block diagram is shown in Fig. 10.

FIGURE 8. The control scheme of PC-PET. (a) Control strategy of the PWM rectifiers. (b) Control strategy of QAB involving the whole operating process.

FIGURE 9. The ripple power decoupling control strategy for PC-PET.

FIGURE 10. The power balance control strategy of QAB in PC-PET.
TABLE 1. Simulation parameters of proposed PC-PET.

| Simulation parameters | Value   |
|-----------------------|---------|
| AC line voltage: $V_{a}$ (rms) | 380 V |
| Rated power: $P$       | 4 kW   |
| No. of module per phase | 2      |
| DC-bus voltage: $V_{dc}$ | 400 V |
| Rated voltage of module: $V_{c}$ | 400 V |
| Switching frequency: $f_s$    | 10 kHz |
| Capacitance in module: $C$ | 100 μF |
| Capacitance at DC bus: $C_{dc}$ | 820 μF |
| Line frequency          | 50 Hz  |

TABLE 2. Leakage inductances.

| Leakage inductance | Value   |
|--------------------|---------|
| Winding 1, phase $a$: $L_{a}$ | $254e^3$ H |
| Winding 2, phase $b$: $L_{b}$ | $264e^3$ H |
| Winding 3, phase $c$: $L_{c}$ | $277.9e^3$ H |

The power balance control strategy is composed of power balance of three phases and power balance in phase $x$ ($x = a$, $b$ and $c$). According to the power balance control strategy, the compensation on phase-shift angles of QAB ($\Delta \psi_{aj}$, $\Delta \psi_{bj}$ and $\Delta \psi_{cj}$) are finally obtained.

V. SIMULATION AND EXPERIMENTAL VERIFICATION

A. SIMULATION RESULTS AND ANALYSIS

To demonstrate the proposed techniques, a simulation model of PC-PET with two QAB modules per phase is set up in MATLAB/Simulink. Parameters of the power circuit are summarized in Table 1.

To illustrate the influence of inconsistently coiled windings, and to verify the effect of proposed power balancing control, the leakage inductances of the four-winding transformer in QAB are deliberately set as the specific values shown in Table 2. The largest parameter difference is 9.4%.

Fig. 11 and Fig. 12 depict the steady-state as well as dynamic performance of proposed power balancing control for PC-PET. In Fig. 11(a), imbalance percentage of three-phase grid currents caused by inconsistent windings can be as large as 28.3%. While in Fig. 11(b), three-phase grid currents are well balanced. Fig. 12 shows that balancing adjustment is achieved within an interval of 11 periods.

The voltage of module capacitor and its FFT analysis results are shown for comparison in Fig. 13. Without ripple power decoupling, the peak-to-peak of capacitor voltage can be as large as 10%, and the 2nd-order ripple component reaches 5.11% (shown in Fig. 13(a)). With ripple power decoupling applied, the peak-to-peak ratio of the capacitor voltage is decreased to 0.5%, and the 2nd-order ripple component remains only 0.125% (shown in Fig. 13(b)). The results reveal that the proposed PC-PET control strategy can effectively suppress the voltage ripple in module capacitors.

Fig. 14 shows the primary-side current of four-winding transformer and phase-shift angle of QAB with ripple power decoupling. The 2nd-order component of phase-shift angle contributes to the transferring of 2nd-order ripple power. And the transformer current is high-frequency pulsating with double-line-frequency ripple.

The steady-state of PC-PET with the overall control scheme is presented in Fig. 15. As shown, the unit power factor at grid side is always guaranteed. The capacitor voltages and the output voltage of QAB are all well balanced. Through FFT analysis, the THD of grid current is 3.18%, the 2nd-order harmonic component is 0.02%, and other harmonic components are all below 2%.

The effects of power imbalance on ripple power decoupling are shown in Fig. 16. Without power balance control, the three-phase grid currents become asymmetrical as shown in Fig. 16(a). The imbalanced three-phase ripple powers cannot be completely counteracted in power decoupling, leaving
considerable voltage ripple in module capacitor even though power decoupling is applied (shown in Fig. 16(b)). The residual ripple powers are transferred to the secondary side of QAB and lead to low-frequency voltage ripples at the DC bus, as illustrated in Fig. 16(c).

In the same PC-PET without proposed power decoupling, 3000 µF capacitance in power modules is required to achieve the effect that the ratio of peak to peak ripple voltage equals to 0.5%. The waveforms of capacitor voltages in both scenarios are shown in Fig. 17. The histogram in Fig. 18 indicates that module capacitance can be sharply reduced by using proposed ripple power decoupling.

Fig. 19 shows the transient processes of PC-PET with 50% load change. It can be seen that with step load changing, the power balance control and ripple power decoupling control can function as expected. During the load change, the system remains operating stably and adjusts to new steady state within 3 fundamental periods.

B. EXPERIMENTAL RESULTS AND ANALYSIS

In order to further verify the effectiveness of the proposed control strategy, a 4:1 downscaled PC-PET prototype is built up. Parameters of the prototype is listed in Table 3.

Fig. 20 shows the main control board and the power unit used. The overall setup is shown in Fig. 21. On the main control board, the FPGA (EP1C12Q240I7) receives the modulating signals from the DSP (TMS320F28335) and performs the pulse-width modulation. Based on modular design, the main circuit of PC-PET is constituted by 7 identical power modules.
The power devices and driving circuit are integrated on one power unit. Each unit has a H-bridge composed of 4 IXFH40N85X MOSFETs (850V/40A), an EP4CE10E22C8 FPGA and other auxiliary circuits. The MOSFETs gating signals are generated by the main control board and transferred to power units by optical fibers.

Fig. 22 shows the module capacitor voltages and DC output voltage of PC-PET during the whole operating interval. Firstly, three single-phase PWM rectifiers are working in diode rectifying mode. During the rectifying process, the module capacitor voltages rise from 311V to 400V. Afterwards, the QAB is softly started with a certain ramp rate.
of the pulse width. Once the output voltage reaches 398V, QAB enters PSC operating mode. Finally, the ripple power decoupling control is imposed. The output voltage of QAB is well regulated and stabilized at the expected value, 400V.

![Power balance control implementation](image)

**FIGURE 23.** Three-phase grid-side currents of PC-PET during the starting of power balance control.

Since the three primary leakage inductances of the four-winding transformer are different, the power balance control should be carried out after entering to the PSC mode. Fig. 23 illustrates the corresponding transient process. When the power balance control is not imposed, imbalance percentage of grid currents can be as large as 11.7%. After when the power balance control effects, the three-phase currents gradually tend to be symmetrical.

![Steady-state waveforms of grid-side currents](image)

**FIGURE 24.** Steady-state waveforms of grid-side currents. (a) Without power balance control. (b) With power balance control.

Fig. 24 shows the steady state waveforms of grid side currents with/without the power balance control. Originally, the three grid currents show considerable amplitude differences due to asymmetrical circuit parameters. Such differences are finally resolved by using the power balance control.

Fig. 25 shows the transient process for ripple power decoupling. As ripple power decoupling control started, the low-frequency ripples in module capacitor voltages are suppressed effectively.

![Transient process for ripple power decoupling](image)

**FIGURE 25.** Transient process for ripple power decoupling.

Fig. 26 demonstrates the effect of voltage ripple suppression. Without ripple power decoupling, the peak to peak ratio of ripple voltage is 6.75%. With proposed power decoupling, the peak to peak ratio of ripple voltage is decreased to 1.25%.

The three primary currents of the four-winding transformer with ripple power decoupling are shown in Fig. 27. The waveforms match with simulation results well, and the three primary currents are symmetrical with each other.
FIGURE 25. Module capacitor voltages during the starting process of proposed power decoupling control.

FIGURE 26. Comparison of module capacitors voltages in PC-PET. (a) Without ripple power decoupling. (b) With proposed ripple power decoupling control.

Fig. 28 shows grid side currents and grid side voltage of phase a with ripple power decoupling. Taking phase a for example, the current is in phase with the voltage at grid side. Thus, it is proved that ripple power decoupling will not affect the power factor control of PWM rectifier in PC-PET.

VI. POWER LOSSES ANALYSIS

The power dissipation comes from different parts, i.e., the semiconductors, the high-frequency transformers, the reactive components. So the power loss calculation are performed from part to part, based on the simulation results.

A. POWER LOSSES IN POWER DEVICES

IGBT power losses can be divided into two groups: conduction loss and switching loss. The conduction loss of IGBT \( (p_{CT}) \) and the anti-parallel diode \( (p_{CD}) \) can be expressed as [23]:

\[
\begin{align*}
p_{CT} &= \frac{1}{T_s} \int_0^{T_s} \left( u_{CE0} \cdot i_C(t) + r_C \cdot i_C^2(t) \right) dt \\
p_{CD} &= \frac{1}{T_s} \int_0^{T_s} \left( u_{DD} \cdot i_D(t) + r_D \cdot i_D^2(t) \right) dt
\end{align*}
\]
where $i_{C}(t)$ is the current flowing through the collector of IGBT, $i_{D}(t)$ is the current flowing through the anti-parallel diode. $u_{CEO}$ and $r_{C}$ represent the on-state voltage the on-state resistance of IGBT respectively, $u_{D0}$, $r_{D}$ represents the on-state voltage the on-state resistance of the anti-parallel diode respectively. $u_{CEO}$, $r_{C}$, $u_{D0}$ and $r_{D}$ can be calculated referring to datasheet [24]. Switching loss ($p_{sw}$) can be expressed as:

$$p_{sw} = (E_{onT} + E_{offT} + E_{onD}) \cdot f_{s} \ (28)$$

where $E_{onT}$ and $E_{offT}$ are turn-on and turn-off energy losses in IGBT. $E_{onD}$ is turn-on energy loss in anti-parallel diode. Parameters $E_{onT}$, $E_{offT}$ and $E_{onD}$ can be found in [24]. The total power losses of an IGBT ($p_{I}$) can be calculated with:

$$p_{I} = p_{CD} + p_{CT} + p_{sw} \ (29)$$

B. POWER LOSSES IN FOUR-WINDING TRANSFORMER

Power losses of the four-winding transformer is composed by two parts: core loss and winding loss. Based on the improved Steinmetz equation, the core losses can be calculated as [25],

$$p_{C} = F K f^{a} B^{\beta} V \ (30)$$

For square excitation, the flux waveform factor $F$ equals to $\pi/4$. The values of $K$, $a$ and $b$ are $4.8810^{-5}$, 1.63 and 6.22 respectively, corresponding to the ferrite core. $B$ is the flux density, and $V$ is the volume of core.

Neglecting proximity effect and skin effect, the winding loss $p_{w}$ is calculated by equivalent resistance of the coils. The total power losses ($p_{T}$) of four-winding transformer can be calculated by:

$$p_{T} = p_{C} + p_{w} \ (31)$$

C. POWER LOSS IN STORAGE CAPACITOR

Power loss of capacitor $p_{C}$ is mainly consumed by the equivalent series resistance (ESR) that can be calculated as [26]:

$$R_{C} = \frac{\tan(\delta)}{2\pi f_{s} C} \ (32)$$

where $\tan(\delta)$ is the dissipation factor [27]. Then $p_{C}$ can be calculated as:

$$p_{C} = I_{C}^{2} \cdot R_{C} \ (33)$$

D. POWER LOSSES AND VOLUME ESTIMATION

Power losses and capacitor volume are calculated based on the selected components listed in Table 4. Similarly, both of the conditions with and without ripple power decoupling are accounted. The comparison results are illustrated in Fig. 30. In Fig. 30(a), $P_{I_{rec}}$, $P_{Ipri}$ and $P_{I_{sec}}$ represent power loss of IGBTs in RECM, primary and secondary side of QAB, $P_{T}$ is the total loss of four-winding transformers and $P_{C}$ is total loss of all capacitors in PC-PET. Without ripple power decoupling, the total power loss of PC-PET is 345.7674W.
While with ripple power decoupling, the total power loss is 359.39W. The efficiency is 95.47% and 95.13% respectively. As we shall see, the efficiencies of the conditions with and without power decoupling would be very similar. Since the focus of this paper is on minimizing of the capacitor voltage ripples, a comparison is done as following, in order to demonstrate the effect of capacitance reduction.

In Fig. 30(b), the volume of 100\(\mu F\) module capacitor \((V_{100}\mu F)\) is 38.013cm\(^3\). The volume of 3000\(\mu F\) module capacitor \((V_{3000}\mu F)\) is 692.72cm\(^3\). By using the power decoupling strategy, the volume of module capacitor can be remarkably reduced while the efficiency of the system almost remains the same. Thereby power density of the system is promoted.

For one power module in a conventional three-phase PET, three two-winding transformers are needed. Supposing that the rated power is 2kW, the ferrite core of transformer is EE55/28/25, and the volume is 78.7cm\(^3\). Thus the total volume of transformers in a conventional PET is about 236.1cm\(^3\). For one power module in the proposed PC-PET, only one four-winding transformer is needed. The ferrite core is EE85/45/31, and the total transformer volume in PC-PET is 162cm\(^3\), which is much less than the total one of three two-winding transformers. The comparison of transformer volume between the two types of PET is illustrated in Fig. 31.

VII. CONCLUSION

The low-frequency ripple power in module capacitors prevents PET from promoting its power density. In this paper, a power channel based four-winding PET (PC-PET) is proposed to solve this issue. The PC-PET principle is firstly introduced, which is followed by the mathematical analysis of the ripple power and its induced ripple voltages. Then the negative effect of inconsistent windings on power decoupling is analyzed. Furthermore, the proposed PR control based power decoupling strategy is presented in details. The simulation and experiment results verify that by employing the proposed techniques, the three-phase ripple powers are completely eliminated through power channels without flowing to loads. And the low-frequency ripples voltage in module capacitors are effectively suppressed even with asymmetric circuit parameters, which reduces the required module capacitance sharply. Intentionally, the power losses, the capacitor volume and the transformer volume are estimated based on the simulation data. The results show that PC-PET with the power decoupling control has the similar efficiency with before, but benefits a much higher power density.

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