Design Methodology for Three-Phase Four-Wire T-Type Inverter With Neutral Inductor

Li ZHANG, Haoxin YANG, Kui WANG, Yalei YUAN, Yi TANG, and Wai Kuan LOH

Abstract—Three-level (3L) converters have been widely used in industry for decades. Compared to the three-phase-three-wire (3P3W) 3L inverter, the three-phase-four-wire (3P4W) one is able to supply the unbalanced loads but has to afford much larger filter inductors because the neutral wire provides a path for high-switching zero-sequence currents. To save filter inductances, a neutral inductor is proposed to insert in the neutral wire. Meanwhile, a complete design methodology is put forward to design the filter inductors and the neutral inductor. With low-frequency zero-sequence currents flowing through the neutral wire, the three-phase load voltages might become unbalanced and/or distorted. To improve the voltage quality, a resonant controller, with the resonant frequency at fundamental output frequency (f_o), is presented to add into the zero-sequence voltage loop for balancing load voltages; concurrently, the other resonant controller, with the resonant frequency at $3f_o$, is presented to insert in the zero-sequence voltage loop or neutral current loop for mitigating voltage distortion. Finally, all of the proposed works are verified on a 3P4W T-type inverter.

Index Terms—Filter inductor, neutral inductor, neutral wire, three-level inverter, three-phase-four-wire, unbalanced load, voltage distortion.

I. INTRODUCTION

THREE-level (3L) converters have lower voltage stress for switching devices, smaller footprint for filters, and higher power quality for ac loads, with respect to their two-level (2L) counterparts [1]–[4]. Because of higher conversion efficiency, higher power density, and lower voltage/current harmonics, 3L converters have been widely used in uninterrupted power supplies [5], high-speed motor drives [6], more-electric aircrafts [7], more-electric ships [8], renewable energy generation systems [9], and etc, in the past decades.

The three-phase 3L inverter in general has two configurations, namely, three-phase-three-wire- (3P3W) and three-

phase-four-wire- (3P4W) 3L inverter. The only difference is the neutral wire which provides the path for the zero-sequence component in load currents and thus makes the 3P4W 3L inverter capable to supply both balanced and unbalanced ac loads [10]. Nonetheless, the neutral wire also provides the path for high-switching zero-sequence harmonic currents, leading to increased inductor current ripple. Consequently, the filter inductors in the 3P4W 3L inverter are much higher than those in the 3P3W 3L inverter, resulting in degraded power density. To solve this problem, this paper explores the way of reducing the total filter inductance in the 3P4W 3L inverter.

The total filter inductance is not exclusively increased in the 3P4W 3L inverter with respect to the 3P3W counterpart, but also increased in the 3P4W 2L inverter in comparison with the 3P3W one. To cope with this problem, an inductor is inserted in the neutral wire of the 3P4W 2L inverter [11], such that the inductor current ripple can remain the same with lower filter inductors. Unfortunately, however, the neutral inductor in [11] is not specifically designed, let alone has a rigorous design methodology. To fill the gap, a design flow is proposed in [12] for evaluating the minimum required filter inductance in the 3P4W 2L inverter with split dc-link capacitors. In doing so, the filter inductors as well as the neutral inductor can be optimally designed. The neutral inductor has also been reported to insert in other 2L inverter topologies. In [13], neutral inductor is proposed to add in the three-phase four-leg 2L inverter. Nonetheless, this neutral inductor is aimed for reducing the voltage pulsation at the neutral point rather than saving filter inductors. In [14], neutral inductor is introduced in the active split dc-bus four-leg inverter. Different from the one in [11], however, this inductor is used for reducing the current ripple in the neutral module.

Inspired by [11], this paper proposes to insert an inductor in the neutral wire of the 3P4W 3L inverter for reducing the size and volume of the filter inductors. Please note that the minimum required filter inductance is determined by the maximum volt-second of the inductor voltage and the expected inductor current ripple, over a line cycle, and that the volt-second of the inductor voltage is further determined by the inverter switching state. As the switching state of the 3L inverter is far more complex than that of the 2L inverter, the derived formulas in [12] are no longer valid for the 3P4W 3L inverter. For the purpose of providing a guidance to design the filter inductors and the neutral inductor in the 3P4W 3L

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Fig. 1. Topology of (a) 3P3W T-type inverter and (b) 3P4W T-type inverter.

inverter, this paper will also show a complete methodology for inductance derivation and demonstrate how the inserted neutral inductor benefits the reduction for filter inductors.

The load voltage quality is the other significant factor for judging the performance of a 3P4W 3L inverter [15]. When supplying unbalanced loads, the fundamental zero-sequence current will flow through the neutral wire, inducing fundamental voltage ripples in the upper and lower dc bus capacitors, which might further produce fundamental zero-sequence voltage in the three-phase load voltages and therefore make the load voltage become unbalanced. In addition to the high-frequency zero-sequence harmonic currents and the fundamental zero-sequence current, the neutral wire also provides the path for triplen harmonics, which makes it more preferable to adopt the basic level-shifted sinusoidal pulse width modulation (LS-SPWM) to modulate the 3P4W 3L inverter [15]. Different from the 2L inverter, upper and lower dc bus capacitors inherently carry third-order voltage ripples with the LS-SPWM [16]. This might result in third-order harmonic current circulating along the L-C filter and the neutral wire, making the three-phase load voltages become distorted. For having a good voltage quality, this paper will also discuss the zero-sequence voltage control, so as to mitigate the potential voltage unbalance and distortion issue for both balanced and unbalanced loads.

The rest of the paper is organized as follows. Section II briefly illustrates the reason for why 3P4W 3L inverter needs larger filter inductors and proposes to insert an inductor in the

Fig. 2. Key waveforms of (a) 3P3W T-type inverter and (b) 3P4W T-type inverter with level-shifted SPWM.

neutral wire. Thereafter, Section III proposes the methodology for designing the filter inductors and the neutral inductor, and also shows the benefit of the neutral inductor in reducing the minimum required filter inductance. To eliminate the potential unbalancing and distortion in the load voltage, dedicated zero-sequence voltage control is discussed in Section IV. The experimental results are shown in Section V. Finally, Section VI concludes this paper.

II. ROOT OF INCREASED INDUCTOR CURRENT RIPPLE IN THREE-PHASE-FOUR-WIRE T-TYPE INVERTER

Figs. 1(a) and (b) respectively show the topology of the basic 3P3W and 3P4W T-type inverter, where C_{bus1} and C_{bus2} form the split capacitor leg, $Q_{x1}-Q_{x4}$ ($x = a, b, c$) form three phase legs, and L_{fx} and C_{fx} form the three-phase L-C filter. The power stage of these two topologies is quite the same. The only difference is the neutral wire in the 3P4W inverter which connects from the neutral point, i.e., N to the midpoint of the split capacitor leg O . For the ease of illustration, O is hereon seen as a reference point.

Because the neutral wire provides a path for the zero-sequence currents, the basic LS-SPWM is adopted to modulate the 3P4W inverter [15]. To show the differences in the working principle, Figs. 2 (a) and (b) respectively show the key waveforms of the 3P3W and 3P4W T-type inverter with

Fig. 3. Topology of 3P4W T-type inverter with neutral inductor.

Fig. 4. Zero-sequence circuit of the 3P4W T-type inverter with neutral inductor.

LS-SPWM, where v_{tri1} and v_{tri2} are two level-shifted unity-magnitude triangle carrier waveforms; v_{Ma} , v_{Mb} , and v_{Mc} are respectively the modulation waveform for each phase leg; v_{ao} is the phase-leg-a output voltage, v_{No} is the neutral-point voltage, and v_{aN} is the voltage difference of v_{ao} and v_{No} . For the T-type inverter, v_{Mx} ($x = a, b, c$) is compared with v_{tri1} to generate the gate signals for Q_{x1} and Q_{x2} , whereas compared with v_{tri2} to generate the gate signals for Q_{x3} and Q_{x4} [15]. When $v_{Mx} > 0$, Q_{x1} and Q_{x2} conduct in a complementary manner, Q_{x3} is always ON, and Q_{x4} is always OFF. When $v_{Mx} < 0$, Q_{x1} is always OFF, Q_{x2} is always ON, Q_{x3} and Q_{x4} conduct in a complementary manner.

Seen from Fig. 2, v_{No} pulsates at switching frequency in the 3P3W T-type inverter, but remains at 0 in the 3P4W T-type inverter. Since v_{No} in the 3P3W T-type inverter equals to the common-mode voltage harmonics in v_{x0} , the common-mode voltage components in v_{ao} can be fully cancelled by v_{No} , making v_{aN} free of triplen-sideband harmonics. By contrast, the common-mode components in v_{ao} of the 3P4W T-type inverter cannot be cancelled, resulting in high-frequency harmonic currents circulating through the three-phase LC filter and the neutral wire. With the same filter inductances, therefore, the filter inductor current ripple in 3P4W T-type inverter will be higher.

Increasing the impedance of the zero-sequence current path can help reduce the filter inductor current ripple in the 3P4W inverter. To obtain this objective, a neutral inductor, i.e., L_N is proposed to insert into the neutral wire, as shown in Fig. 3. In doing so, the zero-sequence circuit of the power stage can be illustrated as presented in Fig. 4, where $L_{fx} = L_f$, $v_{x0} = (\sum v_{x0})/3$ is the common-mode voltage of v_{x0} , $v_{Cf0} = (\sum v_{Cfx})/3$ is the common-mode voltage of v_{Cfx} , and $i_{x0} = (\sum i_x)/3$ is the common-mode current of load currents. According to Fig. 4, v_{No} is

$$v_{No} = \frac{3L_N}{3L_N + L_f} \cdot \frac{1}{3} \sum_{x=a,b,c} v_{x0} = \frac{3L_N}{3L_N + L_f} v_{x0} \triangleq k v_{x0} \quad (1)$$

Particularly, $k = 1$ if $L_N = \infty$, transforming Fig. 3 into the basic 3P3W T-type inverter, as shown in Fig. 1(a); and $k = 0$ if $L_N = 0$, transforming Fig. 3 into the basic 3P4W T-type inverter, as shown in Fig. 1(b). Obviously, the insertion of L_N makes the common-mode voltage harmonics in v_{x0} partly cancelled by v_{No} .

III. DESIGN METHODOLOGY FOR FILTER INDUCTORS AND NEUTRAL INDUCTOR

The minimum required inductance of L_{fx} , i.e., L_{fmin} , is concurrently determined by the maximum volt-second of v_{Lfx} and the expected inductor current ripple, over a line cycle. Taking the derivation of L_{fmin} for phase a as an example, this section illustrates the methodology of designing L_{fx} and L_N . Because of the symmetrical characterization, it is only necessary to analyze the voltages and currents of phase a in the first quarter of a line cycle, i.e., $\omega t \in [0, \pi/2]$, for deriving the closed-form formulas.

The voltage across L_{fa} , i.e., v_{Lfa} , equals to the difference of v_{aN} and v_{Ca} , of which, v_{Ca} can be assumed constant within a switching cycle because the switching frequency is far higher than the fundamental frequency, whereas v_{aN} pulsates at switching frequency and varies with the inverter switching state which is further decided by the intersection sequence of modulation waveforms and carrier waveforms.

Without the loss of generality, we assume v_{Ma} , v_{Mb} , and v_{Mc} are written as

$$\begin{cases} v_{Ma} = M \sin \omega t \\ v_{Mb} = M \sin(\omega t - 2\pi/3) \\ v_{Mc} = M \sin(\omega t + 2\pi/3) \end{cases} \quad (2)$$

where M is the modulation index and ω is the fundamental angular frequency. Since v_{Ma} equals to v_{Mc} at $\pi/6$, the curve of v_{Ma} will always cross with that of v_{Mc} at $\pi/6$. Similarly, $(1 + v_{Mb})$ equals to $(1 + v_{Mc})$ at $\pi/2$, hence, the curve of $(1 + v_{Mb})$ will always cross with that of $(1 + v_{Mc})$ at $\pi/2$. Let $v_{Ma} = 1 + v_{Mb}$, we have

$$\alpha_1 = \pi/3 - \arccos[1/(\sqrt{3}M)], M \in [\sqrt{3}/3, 1] \quad (3)$$

$$\alpha_2 = \pi/3 + \arccos[1/(\sqrt{3}M)], M \in [\sqrt{3}/3, 1] \quad (4)$$

Obviously, $\alpha_1 < \pi/2$. Nonetheless, $\alpha_2 \leq \pi/2$ when $\sqrt{3}/3 \leq M \leq 2/3$, whereas $\alpha_2 > \pi/2$ when $2/3 < M \leq 1$. Let $v_{Mc} = 1 + v_{Mb}$, we have

$$\alpha_3 = \arccos[1/(\sqrt{3}M)], M \in [\sqrt{3}/3, 1] \quad (5)$$

Obviously, $\alpha_3 < \pi/2$. Let $v_{Ma} = 1 + v_{Mc}$, we have

$$\alpha_4 = 2\pi/3 - \arccos[1/(\sqrt{3}M)], M \in [\sqrt{3}/3, 1] \quad (6)$$

According to (6), $\alpha_4 \leq \pi/2$ when $2/3 \leq M \leq 1$, whereas $\alpha_4 > \pi/2$ when $\sqrt{3}/3 \leq M < 2/3$.

Fig. 5. Plots of v_{Mx} and $1 + v_{Mx}$ in the first quarter of a line cycle. (a) $0 < M < \sqrt{3}/3$. (b) $\sqrt{3}/3 < M < 2/3$. (c) $M = 2/3$. (d) $2/3 < M < 1$.

TABLE I
INTERSECTION SEQUENCE OF MODULATION AND CARRIER WAVEFORMS

M	ωt	Sequence
$[0, \sqrt{3}/3]$	$(0, \pi/6]$	$B \rightarrow C \rightarrow A$
	$(\pi/6, \pi/3]$	$B \rightarrow A \rightarrow C$
	$(\pi/3, \pi/2)$	$C \rightarrow B \rightarrow A$
$(\sqrt{3}/3, 2/3]$	$(0, \alpha_3]$	$C \rightarrow B \rightarrow A$
	$(\alpha_3, \pi/6]$	$B \rightarrow C \rightarrow A$
	$(\pi/6, \alpha_1]$	$B \rightarrow A \rightarrow C$
	$(\alpha_1, \pi/3]$	$A \rightarrow B \rightarrow C$
	$(\pi/3, \alpha_2]$	$C \rightarrow A \rightarrow B$
$2/3$	$(0, \pi/6]$	$C \rightarrow B \rightarrow A$
	$(\pi/6, \pi/3]$	$A \rightarrow B \rightarrow C$
	$(\pi/3, \pi/2)$	$C \rightarrow A \rightarrow B$
$(2/3, 1]$	$(0, \alpha_1]$	$C \rightarrow B \rightarrow A$
	$(\alpha_1, \pi/6]$	$C \rightarrow A \rightarrow B$
	$(\pi/6, \alpha_3]$	$A \rightarrow C \rightarrow B$
	$(\alpha_3, \pi/3]$	$A \rightarrow B \rightarrow C$
	$(\pi/3, \alpha_4]$	$C \rightarrow A \rightarrow B$
	$(\alpha_4, \pi/2)$	$A \rightarrow C \rightarrow B$

Fig. 5 plots the curves of v_{Mx} and $(1 + v_{Mx})$ in the range of 0 to 1, against $\omega t \in [0, \pi/2]$. In accordance with Fig. 5, Table I summarized all the possible cases for intersection sequence of modulation waveforms and carrier waveforms:

- 1) When $0 \leq M < \sqrt{3}/3$, $(1 + v_{Mb})$ has no crossing points with v_{Ma} and v_{Mc} , as shown with the solid line in Fig. 5(a). In this case, v_{Mb} is the first one intersecting with the carrier waveform, v_{Mc} the second, and v_{Ma} the third when $\omega t \in [0, \pi/6]$. For the ease of illustration, intersection sequence is denoted as $B \rightarrow C \rightarrow A$. Likewise, the intersection sequence is $B \rightarrow A \rightarrow C$ when $\omega t \in [\pi/6, \pi/3]$, and $C \rightarrow B \rightarrow A$ when $\omega t \in [\pi/3, \pi/2]$. Specially, when $M = \sqrt{3}/3$, $(1 + v_{Mb})$ is tangent with v_{Ma} and v_{Mc} , as shown with the dashed line in Fig. 5(a), but the intersection sequence remains the same.
- 2) When $\sqrt{3}/3 < M < 2/3$, $(1 + v_{Mb})$ crosses with v_{Ma} at α_1 and α_2 , and also with v_{Mc} at α_3 , as shown in Fig. 5(b). In this case, the intersection sequence is $C \rightarrow B \rightarrow A$ when $\omega t \in [0, \alpha_3]$, $B \rightarrow C \rightarrow A$ when $\omega t \in [\alpha_3, \pi/6]$, $B \rightarrow A \rightarrow C$ when $\omega t \in [\pi/6, \alpha_1]$, $A \rightarrow B \rightarrow C$ when $\omega t \in [\alpha_1, \pi/3]$, $C \rightarrow A \rightarrow B$ when $\omega t \in [\pi/3, \alpha_2]$, and $C \rightarrow B \rightarrow A$ when $\omega t \in [\alpha_2, \pi/2]$.
- 3) When $M = 2/3$, $(1 + v_{Mb})$, v_{Ma} and v_{Mc} come cross at $\pi/6$, while $(1 + v_{Mb})$, v_{Ma} and $(1 + v_{Mc})$ come cross at $\pi/2$, as shown in Fig. 5(c). In this case, the intersection sequence is $C \rightarrow B \rightarrow A$ when $\omega t \in [0, \pi/6]$, $A \rightarrow B \rightarrow C$ when $\omega t \in [\pi/6, \pi/3]$, and $C \rightarrow A \rightarrow B$ when $\omega t \in [\pi/3, \pi/2]$.
- 4) When $2/3 < M \leq 1$, v_{Ma} crosses with $(1 + v_{Mb})$ at α_1 and with $(1 + v_{Mc})$ at α_4 , while v_{Mc} crosses with $(1 + v_{Mb})$ at α_3 , as shown in Fig. 5(b). In this case, the intersection sequence is $C \rightarrow B \rightarrow A$ when $\omega t \in [0, \alpha_1]$, $C \rightarrow A \rightarrow B$ when $\omega t \in [\alpha_1, \pi/6]$, $A \rightarrow C \rightarrow B$ when $\omega t \in [\pi/6, \alpha_3]$, $A \rightarrow B \rightarrow C$ when $\omega t \in [\alpha_3, \pi/3]$, $C \rightarrow A \rightarrow B$ when $\omega t \in [\pi/3, \alpha_4]$, and $A \rightarrow C \rightarrow B$ when $\omega t \in [\alpha_4, \pi/2]$.

As for the three-phase T-type inverter, lower M will result in higher voltage stress and lower conversion efficiency. Considering the voltage drop across the filter inductor, a margin

Fig. 6. Key waveforms of phase *a* for $\omega t \in [0, \alpha_1]$. (a) $0 \leq k \leq 3M \sin \omega t$ (b) $3M \sin \omega t < k \leq 1$.

of 15%–25% is often reserved for M to avoid overmodulation [15]. Hence, M is generally higher than 0.7 in practice. So, we focus on the scenario when $2/3 < M \leq 1$ in this paper. In the following part, the expression of L_{fa} will be derived for each of the subcase as highlighted in Table I. Then, L_{fin} for phase *a* will be determined by the maximum value of L_{fa} over all the subcases.

Case 1: $\omega t \in (0, \alpha_1)$

In this case, v_{Mc} is the first one intersecting with the carrier waveform, v_{Mb} the second, and v_{Ma} the third, as shown in Fig. 6. v_{No} has four voltage levels, i.e., $kV_{dc}/3$, $kV_{dc}/6$, 0, and $-kV_{dc}/6$. Accordingly, v_{aN} also has four voltage levels, i.e., $(1/2 - k/3)V_{dc}$, $-kV_{dc}/6$, 0, and $kV_{dc}/6$.

When $\omega t \in (0, \alpha_1)$, $v_{Cfa} (= V_{dc}v_{Ma}/2) \in (0, V_{dc}/4)$. As $-kV_{dc}/6 \leq 0$ and $V_{dc}/6 \leq (1/2 - k/3)V_{dc} \leq V_{dc}/2$, we have $-kV_{dc}/6 \leq 0 < v_{Cfa} < (1/2 - k/3)V_{dc}$. Hence, i_{La} increases when $t \in [t_0, t_1]$, whereas decreases when $t \in [t_1, t_3] \cup [t_4, t_5]$. So,

- 1) If $0 \leq k < 3M \sin \omega t$, we have $v_{Cfa} > kV_{dc}/6$. Hence, i_{La} also decreases when $t \in [t_3, t_4]$. If $k = 3M \sin \omega t$, then $v_{Cfa} = kV_{dc}/6$,

making i_{La} remain constant. Because i_{La} only increases within $[t_0, t_1]$ when $0 \leq k \leq 3M \sin \omega t$, the current ripple of i_{La} will be determined by the volt-second of $v_{La} (= v_{aN} - v_{Cfa})$ within $[t_0, t_1]$, as highlighted in Fig. 6(a). Defining ΔI_L as the allowed maximum inductor current ripple, the minimum required inductance will be

$$L_{f11} = \frac{1}{\Delta I_L} \left(1 - \frac{2}{3}k - v_{Ma} \right) \frac{V_{dc}}{2} v_{Ma} T_s \quad (7)$$

where T_s is the period of one switching cycle.

- 2) If $k > 3M \sin \omega t$, we have $v_{Cfa} < kV_{dc}/6$. This way, i_{La} not only increases within $[t_0, t_1]$ but also within $[t_3, t_4]$, so that the current ripple of i_{La} within one switching cycle will be determined by the higher volt-second of v_{La} between the one within $[t_0, t_1]$ and the one within $[t_3, t_4]$, as highlighted in Fig. 6(b). Thus, the minimum required inductance will be

$$L_{f12} = \max \left\{ L_{f11}, \frac{1}{\Delta I_L} \left(\frac{1}{3}k - v_{Ma} \right) \frac{V_{dc}}{2} (1 - v_{Mc}) T_s \right\} \quad (8)$$

Case 2: $\omega t \in (\alpha_1, \pi/6)$

In this case, v_{Mc} is the first one intersecting with the carrier waveform, v_{Ma} the second, and v_{Mb} the third, as shown in Fig. 7. v_{No} has four voltage levels, i.e., $kV_{dc}/3$, $kV_{dc}/6$, 0, and $-kV_{dc}/6$. Accordingly, v_{aN} also has four voltage levels, i.e., $(1/2 - k/3)V_{dc}$, $(1/2 - k/6)V_{dc}$, 0, and $kV_{dc}/6$.

For $\omega t \in (\alpha_1, \pi/6)$, $v_{Cfa} (= V_{dc}v_{Ma}/2) \in (0, V_{dc}/4)$. Since $V_{dc}/3 \leq (1/2 - k/6)V_{dc} \leq V_{dc}/2$, we have $(1/2 - k/6)V_{dc} > v_{Cfa} > 0$. When $\omega t < \beta = \arcsin(1/(3M))$, we have $v_{Cfa} < V_{dc}/6$. Further, $v_{Cfa} < kV_{dc}/6$ if $3M \sin \omega t < k \leq 1$, whereas $v_{Cfa} > kV_{dc}/6$ if $0 \leq k < 3M \sin \omega t$. Therefore,

- 1) When $\omega t \leq \beta$ and $0 \leq k < 3M \sin \omega t$, we have $0 < kV_{dc}/6 < v_{Cfa} \leq (1/2 - k/3)V_{dc} \leq (1/2 - k/6)V_{dc}$. This way, i_{La} only increases within $[t_0, t_3]$, as highlighted in Fig. 7(a). Thus, the minimum required inductance will be

$$L_{f21} = \frac{1}{\Delta I_L} \left(1 - \frac{k}{3} - v_{Ma} \right) \frac{V_{dc}}{2} (v_{Ma} - v_{Mb} - 1) T_s + \frac{1}{\Delta I_L} \left(1 - \frac{2}{3}k - v_{Ma} \right) \frac{V_{dc}}{2} (1 + v_{Mb}) T_s \quad (9)$$

- 2) When $\omega t \leq \beta$ and $3M \sin \omega t < k \leq 1$, we have $0 < v_{Cfa} < kV_{dc}/6 \leq (1/2 - k/3)V_{dc} \leq (1/2 - k/6)V_{dc}$. Hence, i_{La} not only increases within $[t_0, t_3]$ but also within $[t_4, t_5]$. In this manner, the minimum required inductance will be

$$L_{f22} = \max \left\{ L_{f21}, \frac{1}{\Delta I_L} \left(\frac{1}{3}k - v_{Ma} \right) \frac{V_{dc}}{2} (1 - v_{Mc}) T_s \right\} \quad (10)$$

When $\omega t > \beta$, we have $v_{Cfa} > V_{dc}/6$. Thus, $v_{Cfa} > kV_{dc}/6$. This manner, i_{La} always decreases from t_3 till t_6 no matter v_{Cfa} is higher than $(1/2 - k/3)V_{dc}$ or not, as highlighted in Fig. 7(c). In this case, the minimum required inductance will be

Fig. 7. Key waveforms of phase a for $\omega t \in [\alpha_1, \pi/6]$. (a) $\omega t \in [\alpha_1, \beta]$ and $0 \leq k \leq 3M \sin \omega t$ (b) $\omega t \in [\alpha_1, \beta]$ and $3M \sin \omega t < k \leq 1$. (c) $\omega t \in [\beta, \pi/6]$.

$$L_{I23} = \frac{1}{\Delta I_L} \left[v_{Ma} \frac{V_{dc}}{2} (v_{Mc} - v_{Ma}) + (v_{Ma} - \frac{1}{3}k) \frac{V_{dc}}{2} (1 - v_{Mc}) \right] T_s \quad (11)$$

Case 3: $\omega t \in (\pi/6, \alpha_3)$

In this case, v_{Ma} is the first one intersecting with the carrier waveform, v_{Mc} the second, and v_{Mb} the third, as shown in Fig.

Fig. 8. Key waveforms of phase a for $\omega t \in [\pi/6, \alpha_3]$.

Fig. 9. Key waveforms of phase a for $\omega t \in [\alpha_3, \pi/3]$.

8. v_{No} has four voltage levels, i.e., $kV_{dc}/3$, $kV_{dc}/6$, 0, and $-kV_{dc}/6$. Accordingly, v_{aN} has four voltage levels, i.e., $(1/2 - k/3)V_{dc}$, $(1/2 - k/6)V_{dc}$, $1/2V_{dc}$, and $kV_{dc}/6$.

For $\omega t \in (\pi/6, \alpha_3)$, $v_{Cfa} (= V_{dc}v_{Ma}/2) > V_{dc}/6 \geq kV_{dc}/6$. Thus, i_{La} always decreases from t_5 till t_6 whichever of v_{Cfa} or $(1/2 - k/3)V_{dc}$ is higher, as highlighted in Fig. 8. Thus, the minimum required inductance will be

$$L_{I3} = \frac{1}{\Delta I_L} \left(v_{Ma} - \frac{1}{3}k \right) \frac{V_{dc}}{2} (1 - v_{Ma}) T_s \quad (12)$$

Case 4: $\omega t \in (\alpha_3, \pi/3)$

In this case, v_{Ma} is the first one intersecting with the carrier waveform, v_{Mb} the second, and v_{Mc} the third, as shown in Fig. 9. v_{No} has four voltage levels, i.e., $kV_{dc}/3$, $kV_{dc}/6$, 0, and $-kV_{dc}/6$. Accordingly, v_{aN} has four voltage levels, i.e., $(1/2 - k/3)V_{dc}$, $(1/2 - k/6)V_{dc}$, $1/2V_{dc}$, and $kV_{dc}/6$.

Similar with Case 3, $v_{Cfa} (= V_{dc}v_{Ma}/2) > V_{dc}/6 \geq kV_{dc}/6$, such that i_{La} always decreases within $[t_5, t_6]$ regardless of the values of v_{Cfa} and $(1/2 - k/3)V_{dc}$, as highlighted in Fig. 9. So, the minimum required inductance for this case will be also

$$L_{I4} = \frac{1}{\Delta I_L} \left(v_{Ma} - \frac{1}{3}k \right) \frac{V_{dc}}{2} (1 - v_{Ma}) T_s \quad (13)$$

Fig. 10. Key waveforms of phase a for $\omega t \in [\pi/3, \alpha_4]$. Fig. 11. Key waveforms of phase a for $\omega t \in [\alpha_4, \pi/2]$.**Case 5: $\omega t \in (\pi/3, \alpha_4)$**

In this case, v_{Mc} is the first one intersecting with the carrier waveform, v_{Ma} the second, and v_{Mb} the third, as shown in Fig. 10. v_{No} has four voltage levels, i.e., $kV_{dc}/6$, 0 , $-kV_{dc}/6$, and $-kV_{dc}/3$. Accordingly, v_{aN} has four voltage levels, i.e., $(1/2 - k/6)V_{dc}$, $1/2V_{dc}$, $kV_{dc}/6$ and $kV_{dc}/3$.

For $\omega t \in (\pi/3, \alpha_4)$, we have $v_{Cfa} (= V_{dc}v_{Ma}/2) \geq \sqrt{3}v_{dc}/3 > kV_{dc}/3 \geq kV_{dc}/6$. Thus, i_{La} always decreases within $[t_3, t_6]$, as highlighted in Fig. 10. Thus, the minimum required inductance will be

$$L_{f5} = \frac{1}{\Delta I_L} \left(v_{Ma} - \frac{1}{3}k \right) \frac{V_{dc}}{2} (v_{Mc} + 1 - v_{Ma}) T_s + \frac{1}{\Delta I_L} \left(v_{Ma} - \frac{2}{3}k \right) \frac{V_{dc}}{2} (-v_{Mc}) T_s \quad (14)$$

Case 6: $\omega t \in (\alpha_4, \pi/2)$

In this case, v_{Ma} is the first one intersecting with the carrier waveform, v_{Mc} the second, and v_{Mb} the third, as shown in Fig. 11. v_{No} has four voltage levels, i.e., $kV_{dc}/6$, 0 , $-kV_{dc}/6$, and $-kV_{dc}/3$. Accordingly, v_{aN} has four voltage levels, i.e., $(1/2 - k/6)V_{dc}$, $1/2V_{dc}$, $1/2V_{dc} + kV_{dc}/6$ and $kV_{dc}/3$.

Like Case 5, i_{La} always decreases within $[t_5, t_6]$. Thus, the

minimum required inductance will be

$$L_{f6} = \frac{1}{\Delta I_L} \left(v_{Ma} - \frac{2}{3}k \right) \frac{V_{dc}}{2} (1 - v_{Ma}) T_s \quad (15)$$

Summarizing from the above-mentioned discussions, L_{fa} can be written as

$$L_{fa} = \begin{cases} L_{f11} & \omega t \in (0, \alpha_1) \cap k \in [0, 3M \sin \omega t) \\ L_{f12} & \omega t \in (0, \alpha_1) \cap k \in [3M \sin \omega t, 1] \\ L_{f21} & \omega t \in [\alpha_1, \beta) \cap k \in [0, 3M \sin \omega t) \\ L_{f22} & \omega t \in [\alpha_1, \beta) \cap k \in [3M \sin \omega t, 1] \\ L_{f23} & \omega t \in [\beta, \pi/6) \\ L_{f3} & \omega t \in [\pi/6, \alpha_3) \\ L_{f4} & \omega t \in [\alpha_3, \pi/3) \\ L_{f5} & \omega t \in [\pi/3, \alpha_4) \\ L_{f6} & \omega t \in [\alpha_4, \pi/2) \end{cases} \quad (16)$$

L_{fmin} for phase a will be determined by the maximum value of L_{fa} over all cases. Thus, L_{fmin} will be

$$L_{fmin} = \max \{ L_{fa} \} \quad (17)$$

Recalling (3) in which k is defined as $3L_N/(L_f + 3L_N)$, we have

$$L_N = \frac{k}{1-k} \frac{1}{3} L_{fmin} \quad (18)$$

So, the total required ac filter inductance will be

$$L_{sum} = L_N + \sum_{x=a,b,c} L_{fx} = \left(\frac{1}{3} \frac{k}{1-k} + 3 \right) L_{fmin} \quad (19)$$

Obviously, L_{fmin} , L_N , and L_{sum} are all functions of k .

Please note that the inductor current and neutral current will induce voltage drop across L_{fa} and L_N which might result in overmodulation if either L_{fa} or L_N is too large. To avoid this problem, L_{fa} and L_N should also satisfy

$$\left| \dot{V}_{Cfa} + j\omega L_{fa} \dot{I}_{La} + j\omega L_N \dot{I}_N \right| < V_{in}/2 \quad (20)$$

In general, the filter capacitor in each phase is designed to keep the capacitor current within 5% of the rated phase current. So, $I_{La} \approx I_a$. Considering the worst case for (20), we have

$$\left| \dot{V}_{Cfa} + j\omega L_{fa} \dot{I}_{La} + j\omega L_N \dot{I}_N \right| < \left| \dot{V}_{Cfa} \right| + \left| j\omega L_{fa} \dot{I}_{La} \right| + \left| j\omega L_N \dot{I}_N \right| = MV_{dc}/2 + \sqrt{2} \omega L_{fa} I_a + \sqrt{2} \omega L_N \lambda I_a < V_{dc}/2 \quad (21)$$

in which $\lambda \in [0, 1]$ is the load unbalanced factor. If the three-phase loads are well balanced, λ is 0. On the contrary, if two phases are with full load whereas the third one is with no load,

TABLE II
SPECIFICATIONS FOR THREE-PHASE T-TYPE INVERTER

Parameter	Symbol	Value
DC Input Voltage	V_{dc}	800 V
AC Phase Voltage	$V_{C_{fa}}, V_{C_{fb}}, V_{C_{fc}}$	220 V (RMS)
Phase Current	I_a, I_b, I_c	15.2 A (RMS)
Output Frequency	f_o	50 Hz
Apparent Power	S	10 kVA

Fig. 12. Plots of L_{fmin} and L_{sum} against k .

λ is 1.

By submitting (17) and (18) into (20), yields

$$L_{fmin} < \frac{1}{\sqrt{2} \omega I_a} \frac{1}{1 + \lambda k / [3(1-k)]} (1-M) \frac{V_{dc}}{2} \quad (22)$$

Using the parameters given in Table II (referring to Section V), Fig. 12 illustrates the plots of (17) and (22). Based on (19), the plot of L_{sum} is also depicted. Within the shaded area (possible area for L_{fmin} and k), the minimum required total filter inductance, $L_{sum, min}$, is 5.25 mH when $k = 0.75$.

In light of (16), Fig. 13 shows the plots of L_{fa} with $k = 0$, $k = 0.75$, and $k = 1$. As seen, the total required filter inductance for the basic 3P4W T-type inverter ($k = 0$) is 7.05 mH (2.35 mH per phase), which is more than twice the total required filter inductance (1.04 mH per phase) for the 3P3W T-type inverter ($k = 1$). Thanks to the insertion of L_N (1.35 mH), the total required filter inductance (5.25 mH) can be saved by 25% compared to the total required filter inductance for the basic 3P4W T-type inverter. Please keep in mind that lower load unbalanced factor results in lower current rating for L_N . This means the size and volume of L_N will be much lower than that of L_{fx} with similar inductance. Therefore, the insertion of L_N will benefit more for filter size and volume reduction in the application where the 3P4W inverter supplies the unbalanced three phase load with low unbalanced factor.

IV. ZERO-SEQUENCE VOLTAGE CONTROL FOR 3P4W T-TYPE INVERTER

Because of the existence of neutral wire, the 3P4W inverter

Fig. 13. Shows the plots of L_{fa} with $k = 0$, $k = 0.75$, and $k = 1$.

Fig. 14. Control block diagram of 3P4W T-type inverter.

can be deemed as three decoupled single-phase inverters with phase voltages interleaved by 120° [17]. As a result, each phase voltage can be independently regulated in the abc-frame, as shown in Fig. 14. In addition to the phase voltage regulation, the upper and lower dc bus voltages, i.e., v_{bus+} and v_{bus-} in Fig. 3, should be balanced [18]–[20]. To attain this objective, we can compare v_{bus+} with v_{bus-} and use the error voltage to regulate the zero-sequence component of the modulation waveforms, as we usually do for the 3P3W T-type inverter.

Different from the 3P3W T-type inverter, however, the neutral wire in the 3P4W T-type inverter provides the path for zero-sequence currents which might make load voltages unbalanced or distorted. To mitigate the potential problems, this section shows the dedicated zero-sequence voltage control for the 3P4W T-type inverter.

A. Zero-Sequence Fundamental Voltage Elimination for Balancing Load Voltages

When supplying unbalanced loads, fundamental zero-sequence current will flow through the neutral wire. Since the

Fig. 15. Insertion of RLC network in the neutral wire.

neutral current flows into the midpoint of the split capacitor leg, there will be fundamental voltage ripples in v_{bus+} and v_{bus-} . Therefore, the error of v_{bus+} and v_{bus-} will have fundamental voltage ripples, leading to fundamental zero-sequence voltage in modulation waveforms. This way, the load voltages will have fundamental zero-sequence component and become unbalanced.

In order to have balanced load voltages, the zero-sequence fundamental voltage should be eliminated from the modulation waveforms. Aiming for this, a resonant controller, written by

$$G_{PR_1st}(s) = \frac{K_{r1st} \omega_i s}{s^2 + \omega_i s + (2\pi f_o)^2} \quad (23)$$

whose resonant frequency is mapped at the fundamental output frequency f_o is inserted into the zero-sequence voltage loop, as highlighted in blue in Fig. 14. This way, the zero-sequence fundamental component will be eliminated from the load voltages because of the particularly high zero-sequence voltage loop gain at f_o , making the load voltages well balanced even if the loads have the highest unbalanced factor ($\lambda = 1$).

B. Third-Order Harmonic Current Elimination for Reducing Load Voltage Distortion

Unlike the 2L voltage-source converter, the midpoint current of the T-type inverter, i.e., i_{mid} inherently pulsates at $3f_o$ with LS-SPWM, inducing third-order harmonic ripples in v_{bus+} and v_{bus-} [16] which will in turn produce third-order harmonic voltage in the output of each phase-leg (v_{x0}).

As illustrated in Fig. 4, the third-order harmonic voltage in v_{x0} will finally forms third-order harmonic current circulating along the L-C filters and the neutral wire. Consequently, the three-phase load voltage will be polluted, leading to voltage distortion and thereby higher THD.

To block the third-order harmonic current, we can insert a paralleled RLC network, with L_0 being parallel-resonant with C_0 at $3f_o$, in the neutral wire, as shown in Fig. 15. The impedance of the RLC network is

$$Z_{RLC}(s) = \frac{1}{1/R_0 + 1/sL_0 + sC_0} = \frac{(L_0/C_0)s}{s^2 + [1/(R_0 C_0)]s + (2\pi 3f_o)^2} \quad (24)$$

whose format is the same as a resonant controller, written by

$$G_{PR_3rd}(s) = \frac{K_{r3rd} \omega_i s}{s^2 + \omega_i s + (2\pi 3f_o)^2} \quad (25)$$

TABLE III
PARAMETERS OF L-C FILTERS IN DIFFERENT CONFIGURATIONS

Configuration	L_f (mH)	L_N (mH)	C_f (μ F)
3P3W	1.04	/	10
3P4W w/o L_N	2.35	/	10
3P4W w/ L_N	1.30	1.35	10

with the resonant frequency mapped at $3f_o$ [22]–[25]. This indicates that the paralleled RLC network can be emulated by inserting a resonant controller in the neutral current control loop, as shown in Fig. 14 with S closed to S1.

As for the three-phase load voltages, the third-order harmonic voltage is also a type of zero-sequence voltage. As an alternative, therefore, we could also eliminate the third-order harmonic voltage by increasing the zero-sequence voltage loop gain at $3f_o$. With this consideration, the other resonant controller, with the resonant frequency at $3f_o$ can be further introduced into the zero-sequence voltage loop, as shown in Fig. 14 with S closed to S2. In this manner, not only the fundamental zero-sequence voltage but also the third-order zero-sequence voltage will be eliminated from the three-phase load voltages, resulting in balanced and sinusoidal three-phase voltage for both balanced and unbalanced loads.

V. EXPERIMENTAL RESULTS

In the lab, a 10-kVA three-phase T-type inverter has been fabricated and tested to validate the proposed work. Table II and III respectively give the prototype specifications and the filter parameters for different configurations.

Figs. 16(a) and (b) show the load voltages and inductor currents for the T-type inverter under balanced full loads in 3P3W and 3P4W configuration, respectively, with $L_f = 1.04$ mH. Despite the same filter inductance, the inductor current ripple in 3P4W configuration is apparently higher than that in 3P3W configuration. By increasing the filter inductor from 1.04 mH to 2.35 mH, the inductor current ripple in 3P4W configuration, as shown in Fig. 16(c), is reduced to the same level (20% I_x) as the inductor current ripple given in Fig. 16(a). This confirms the theoretical analyses in Section III. After inserting the neutral inductor (1.35 mH), the inductor current ripple given in Fig. 16(d) is the same as that shown in Fig. 16(c), but L_f is reduced from 2.35 mH to 1.30 mH. This demonstrates that the inserted neutral inductor can help save the total filter inductance in the 3P4W 3L inverter.

Fig. 17 gives the load voltages and inductor currents for the 3P4W T-type inverter under unbalanced loads (full-load in phase *a* and *b*, whereas 2/3 full-load in phase *c*). Similar to the results for balanced loads, the inductor current ripple with $L_f = 1.04$ mH, as shown in Fig. 17(a), is much higher than that with $L_f = 2.35$ mH, as shown in Fig. 17(b). By inserting the neutral inductor (1.35 mH), the inductor current ripple, as shown in Fig. 17(c), becomes the same as that shown in Fig. 17(b), with L_f reduced to 1.30 mH.

Fig. 18 shows the load voltages and currents for the 3P4W

(a)

(b)

(c)

(d)

Fig. 16. Load voltages and inductor currents under balanced full loads. (a) 3P3W with $L_r = 1.04$ mH. (b) 3P4W with $L_r = 1.04$ mH. (c) 3P4W with $L_r = 2.35$ mH. (d) 3P4W with $L_r = 1.30$ mH and $L_N = 1.35$ mH.

T-type inverter under full balanced loads, where Fig. 18(a) is for the inverter without $G_{PR_1st}(s)$ or $G_{PR_3rd}(s)$; Fig. 18(b) for the inverter with $G_{PR_1st}(s)$ but without $G_{PR_3rd}(s)$; Fig. 18(c) for

(a)

(b)

(c)

Fig. 17. Load voltages and inductor currents under unbalanced loads. (a) 3P4W with $L_r = 1.04$ mH. (b) 3P4W with $L_r = 2.35$ mH. (c) 3P4W with $L_r = 1.30$ mH and $L_N = 1.35$ mH.

the inverter with $G_{PR_1st}(s)$ and $G_{PR_3rd}(s)$ in the zero-sequence voltage loop; and Fig. 18(d) for the inverter with $G_{PR_1st}(s)$ in zero-sequence voltage loop plus $G_{PR_3rd}(s)$ in the neutral current loop. The load-voltage unbalanced factor (UBF) and load-voltage THD are summarized in Table IV. Because of balanced loads, the load voltages in Fig. 18 are all balanced. In contrast with Figs. 18(c) and (d), however, the load voltages in Figs. 18(a) and (b) are distorted because of the third-order harmonic, making the THD increased to 6.0%, indicating that $G_{PR_3rd}(s)$ in the zero-sequence voltage (or current) loop is effective for preventing the load voltages from being distorted.

Keeping the same control parameters, Fig. 19 shows the test results for unbalanced loads (full-load in phase a and b , whereas no-load in phase c). Without adding $G_{PR_1st}(s)$ or $G_{PR_3rd}(s)$, the load voltages are unbalanced and distorted, as

(a)

(b)

(c)

(d)

Fig. 18. Load voltages and inductor currents for 3P4W T-type inverter under balanced full loads. (a) Without $G_{PR_1st}(s)$ or $G_{PR_3rd}(s)$. (b) With $G_{PR_1st}(s)$ but without $G_{PR_3rd}(s)$. (c) With $G_{PR_1st}(s)$ and $G_{PR_3rd}(s)$ in zero-sequence voltage loop. (d) With $G_{PR_1st}(s)$ in zero-sequence voltage loop and $G_{PR_3rd}(s)$ in neutral current loop.

shown in Fig. 19(a). By merely inserting $G_{PR_1st}(s)$ in the zero-sequence voltage loop, the load voltages become balanced, but still have distortion, as shown in Fig. 19(b). With the further

(a)

(b)

(c)

(d)

Fig. 19. Load voltages and inductor currents for 3P4W T-type inverter under unbalanced loads. (a) Without $G_{PR_1st}(s)$ or $G_{PR_3rd}(s)$. (b) With $G_{PR_1st}(s)$ in the zero-sequence voltage loop but without $G_{PR_3rd}(s)$. (c) With $G_{PR_1st}(s)$ and $G_{PR_3rd}(s)$ in the zero-sequence voltage loop. (d) With $G_{PR_1st}(s)$ in the zero-sequence voltage loop and $G_{PR_3rd}(s)$ in the neutral current loop.

addition of $G_{PR_3rd}(s)$ in the zero-sequence voltage (or neutral current) loop, the load voltages are not only balanced but also more sinusoidal, as shown in Figs. 19(c) and (d). These results

TABLE IV
LOAD-VOLTAGE UNBALANCED FACTOR AND LOAO-VOLTAGE THD

Load Type	G_{PR_1st}	G_{PR_3rd}	UBF	THD
Balanced Loads	×	×	0	6.0%/6.0%/6.0%
	√	×	0	6.0%/6.0%/6.0%
	√	S1	0	0.9%/0.9%/0.9%
	√	S2	0	0.9%/0.9%/0.9%
Unbalanced Loads	×	×	31%	5.5%/3.1%/7.4%
	√	×	1%	5.9%/5.5%/6.4%
	√	S2	0.6%	0.9%/0.9%/1.0%
	√	S1	0.6%	0.9%/0.9%/1.0%

validate the role of $G_{PR_1st}(s)$ and $G_{PR_3rd}(s)$ in improving the voltage quality.

VI. CONCLUSION

The 3P4W inverter can supply unbalanced loads but needs to afford larger filter inductors because of the neutral wire. In this paper, a neutral inductor is proposed to insert in the neutral wire for saving the total filter inductances in the 3P4W 3L inverter. Design methodologies are presented for designing the filter inductors and the neutral inductor, indicating that the insertion of the neutral inductor can benefit the size and volume reduction for the filter inductors by at least 25%. For having high quality load voltages, a resonant controller, with the resonant frequency at f_o , is added into the zero-sequence voltage loop for balancing load voltages; concurrently, the other resonant controller, with the resonant frequency at $3f_o$, is inserted into the zero-sequence voltage (or neutral current) loop for mitigating the voltage distortion. Finally, experimental results are shown to verify the feasibility of all proposed work.

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