

Chapter 2

Basic Principles for Parallel Power Electronic Filters

Abstract In this chapter, the basic principles for parallel power electronics filters are discussed. In Sect. 2.1, the influences of reactive power, harmonics and unbalance are given, which affect and degrade the practical power network. The basic circuit configuration for injecting and absorbing compensating current is discussed. Furthermore, several parallel power filters are given and compared. In Sect. 2.2, instantaneous power theories which are important for detecting the compensation current components and controlling the fast response of power electronic compensators are introduced and addressed. Finally, different power electronic converters are discussed and compared in Sect. 2.3. The effect of dc voltage variation on the switching functions are also introduced in Sect. 2.3.

2.1 Parallel Compensation for Reactive Power, Current Harmonics and Unbalance

In an ideal case, to reduce the oscillating power components flowing between the source and loads, the voltage and current should be sinusoidal, in-phase and balanced. Thus, the electric system can be operated with higher efficiency and better power quality without reactive power, harmonics and unbalance problems in the power network. However, in practical cases, due to the common loads which are normally inductive and non-linear as well as the unpredictable loadings in different phases in a three-phase system, there will be reactive power oscillating around the system, and distorted unbalance current waveforms would exist. Reactive power, current harmonics and unbalance problems usually occur in practical power networks, in which the influences of reactive power, harmonics and unbalance are given as follows. After that, the basic principle of parallel compensators is introduced in this section.

2.1.1 Influence of Reactive Power

The reactive power occurs when there is a phase angle difference between the voltage and current under a storage passive load. The reactive power in an electric network causes the following issues:

- (1) The reactive power increases the system capacity rating. Increasing the reactive power causes the source current to increase. As a result, the apparent power is increased too. Thus this increases system capacity ratings of generators, transformers, transmission lines, electric applications and measurement systems. Finally, all initial costs are increased proportionally.
- (2) The reactive power is an oscillating power between the source and the load; it increases the unnecessary source current. As a result, the transmission loss is increased by the reactive power flowing inside the system. Under a steady-state condition, the reactive power is expressed as Q . The root mean square (RMS) source current can be decomposed into active current, I_p , and reactive current, I_q . The transmission loss can be represented by (2.1), where R is the transmission line total resistance and V is the RMS source voltage.

$$\Delta P = I^2 R = (I_p^2 + I_q^2) R = \frac{P^2 + Q^2}{V^2} R \quad (2.1)$$

The second term of (2.1), $(Q^2/V^2)R$ causes the additional power loss by reactive power flow. When all the instantaneous reactive power can be compensated, the transmission loss can be reduced.

- (3) The voltage variation of transformers and feeding networks can be increased. Figure 2.1 shows an equivalent circuit for reactive power influence on transformers, where V_s is the source voltage, a transmission line normally is inductive with R (transmission line resistance) and X (transmission line inductance), the feeding terminal voltage V with equivalent loading admittance for transformers such as G (conductance) and B (inductive admittance) and system current I .

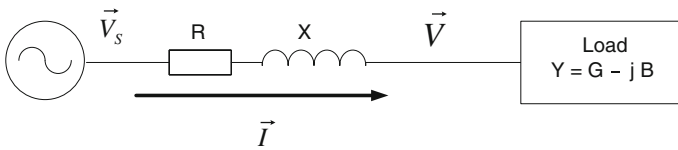
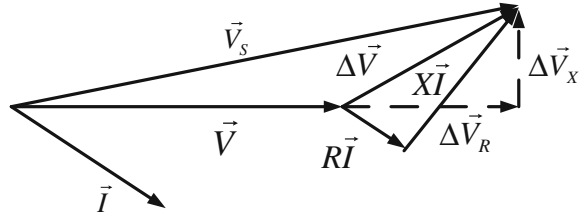


Fig. 2.1 Equivalent circuit for reactive power influence on transformers

Fig. 2.2 Phasor diagram for reactive power influence on transformers



From Fig. 2.2, the phasor diagram shows that the transmission line impedance causes the terminal voltage drop, ΔV . $\Delta \vec{V}$ means a terminal voltage vector that can be decomposed into a real part and an imaginary part respectively.

$$\Delta \vec{V} = \vec{V}_s - \vec{V} = (R + jX) \cdot \vec{I} \quad (2.2)$$

$$\vec{I} = V(G - jB) = \frac{V^2 G - jV^2 B}{V} = \frac{P - jQ}{V} \quad (2.3)$$

$$\Delta \vec{V} = (R + jX) \cdot \left(\frac{P - jQ}{V} \right) = \frac{RP + XQ}{V} + j \frac{XP - RQ}{V} = \Delta V_R + j\Delta V_X \quad (2.4)$$

By checking the phasor diagram, due to the small phase angle difference between $\vec{V}_s$ and $\vec{V}$, it shows that $\Delta V \approx \Delta V_R$. Finally,

$$\Delta V \approx \Delta V_R = \frac{RP + XQ}{V} \quad (2.5)$$

Normally, $X \gg R$. As a result, the active power does not influence too much on the terminal voltage variation. On the other hand, the reactive power causes more terminal voltage variation. Voltage variation can affect the normal operation of other electric applicants and finally the quality of power supplied to the users.

2.1.2 Influence of Harmonics

In the ideal case, the voltage and current should be sinusoidal, which do not have any harmonic component. A harmonic of a wave is a component frequency of a signal that is an integer multiple of the fundamental frequency, where the fundamental frequency is the system operating frequency which is 50 Hz in China and 60 Hz in America. In practical cases, the voltage and current cannot or may not be sinusoidal. Figure 2.3 shows practical measured voltage and current waveforms respectively, it shows that the current is highly distorted and the voltage quality is affected due to the current distortion.

Table 2.1 depicts the harmonic amplitudes of the current waveform of Fig. 2.3, which shows that its Total Harmonic Distortion (THD) is 127.39 %.

Fig. 2.3 Practical voltage and current waveforms

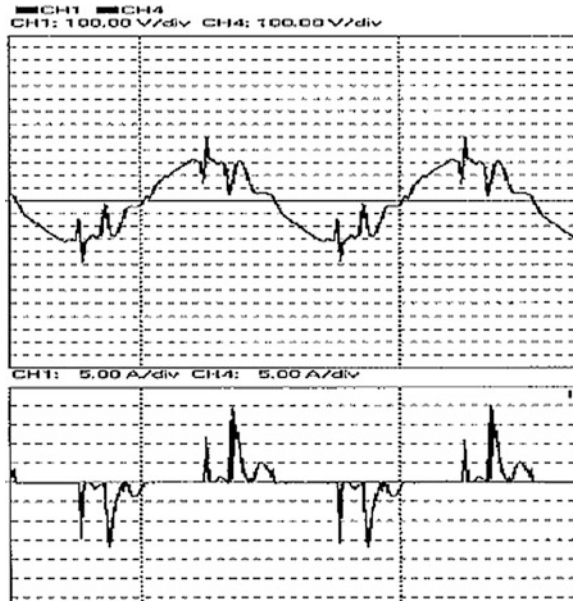


Table 2.1 Harmonic amplitude of current waveforms in Fig. 2.3

CH1: Value harmonic								
Order	(A)		Order	(A)		Order	(A)	
1	100.00	0.48	18	21.98	0.74	35	4.19	0.48
2	5.47	0.49	19	16.13	0.43	36	8.51	0.49
3	52.17	0.40	20	22.16	0.56	37	6.36	0.55
4	5.70	0.43	21	8.54	0.29	38	5.89	0.39
5	24.24	0.30	22	9.87	0.31	39	8.65	0.50
6	15.58	0.41	23	4.93	0.28	40	9.24	0.30
7	44.01	0.33	24	2.28	0.37	41	7.53	0.42
8	24.43	0.41	25	8.27	0.42	42	7.97	0.31
9	59.72	0.35	26	9.49	0.47	43	7.53	0.46
10	23.15	0.35	27	5.39	0.51	44	6.35	0.34
11	40.38	0.22	28	13.71	0.52	45	2.84	0.41
12	14.66	0.33	29	3.62	0.36	46	3.52	0.31
13	17.03	0.30	30	12.95	0.45	47	4.45	0.37
14	11.81	0.56	31	5.79	0.32	48	4.13	0.37
15	14.71	0.44	32	9.54	0.35	49	9.14	0.39
16	16.02	0.68	33	5.64	0.37	50	0.33	0.44
17	19.74	0.49	34	3.28	0.35	THD	127.39	3.00 (%)

The harmonics existed in power networks can cause the following problems:

- (1) Harmonic current can cause additional power loss and reduce the system efficiency of generators, power transmission and electric utilities. When large amounts of triple harmonic current exist in the neutral wire in a three-phase four-wire system, it may cause overheating and even a fire at the neutral wire.
- (2) Harmonics influence normal operations of electric applicants and can reduce the expected life period of electric applicants.
- (3) Harmonics may cause series and parallel resonance in a specific region of the power networks.
- (4) Harmonics may trigger the protection system which causes power breakdown or operates abnormally.
- (5) The power measurement system may include calculation errors.
- (6) The communication system can be influenced by harmonics due to electro-magnetic interference.

2.1.3 Influence of Unbalance

In an ideal balance three-phase system, the three-phase voltage RMS amplitudes should be the same and phase angle difference among them should be 120° apart from each other. However, in practical cases, the voltage and current may not be balanced. Figure 2.4 exhibits the measured three-phase current and the neutral current waveforms in a distributed transformer, which shows that the system is operated under unbalance and with high neutral current amplitude.

The unbalance exists in power networks can cause the following problems:

- (1) In low voltage distribution sites, three-phase unbalance loading may cause user's voltage variation. On one side, the terminal voltage may be higher and it may destroy electric applications, and the electric applicants in low voltage side may not operate normally. Figure 2.5 shows a three-phase four-wire

Fig. 2.4 Three phase current and neutral current waveforms

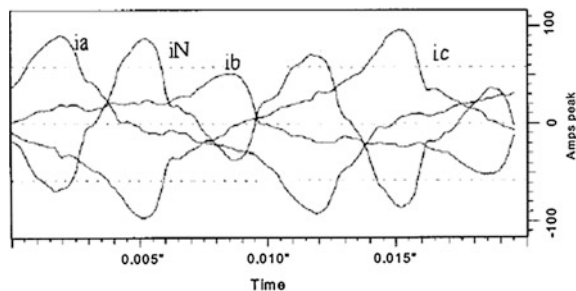


Fig. 2.5 A three-phase four-wire system

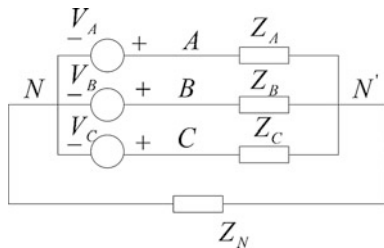
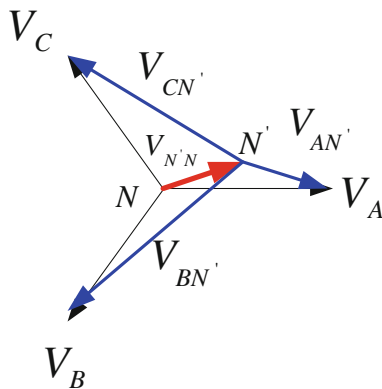


Fig. 2.6 A three-phase four-wire system unbalance phase diagram



system connected with different load impedances, Z_A , Z_B and Z_C . Equation (2.6) shows that the neutral voltage $V_{N'N}$ may not be zero if the loads are unbalance with balanced voltage situation, in which Y_A , Y_B and Y_C are equivalent loading admittance Fig. 2.6 displays the phase diagram for terminal user voltages, $V_{AN'}$, $V_{BN'}$ and $V_{CN'}$, which shows that the terminal user voltages can be varied if the loads are unbalanced and the neutral line impedance, Z_N , is not small enough.

$$V_{N'N} = \frac{V_A Y_A + V_B Y_B + V_C Y_C}{Y_A + Y_B + Y_C + \frac{1}{Z_N}} \quad (2.6)$$

- (2) There is an additional loss in the electric motors and generators due to the negative sequence of the unbalance current, and negative sequence can reduce the resultant torque of a motor and its efficiency, and finally it causes the vibration between the stator and rotor of a motor.
- (3) Abnormal operation of protected systems in a power network degrades the stability and security of the power system.
- (4) Unbalance voltage may cause additional harmonics from the semiconductor converters.

- (5) The system capacity utilization ratio of a generator can be degraded.
- (6) The operation point of a transformer that may shift up or down causes harmonic distortion on the voltage waveforms and additional loss inside the transformer, which increases the temperature of the transformer and its peripheral environment.
- (7) The unbalance operation degrades the electric system efficiency and the quality of telecommunication systems.
- (8) High neutral current causes overheating and failure of the neutral conductor.

2.1.4 Basic Principle of Parallel Compensation

Figure 2.7 shows the basic configuration for parallel current quality compensation. Figure 2.8 shows its corresponding waveforms for compensation. The compensator is connected in parallel with the load. It shows a source supplying a non-linear load that is being compensated by a shunt compensator. As a result, the source current, $\vec{i}_S$ can be balanced and sinusoidal. However, the load current, $\vec{i}_L$ does not change before and after compensation. The load currents of non-linear unbalanced loads are still unbalance, non-sinusoidal and with harmonics. On the other hand, the compensator injected current, $\vec{i}_C$ turns the source current to be balanced and sinusoidal. The ideal compensator is a controlled current source that can inject or draw any set of arbitrarily compensating current reference, $\vec{i}_C^*$.

Basically, there are 4 different kinds of parallel compensators: Capacitor Banks (CB), Passive Power Filters (PPF), Active Power Filters (APF) and Hybrid Active Power Filters (HAPP).

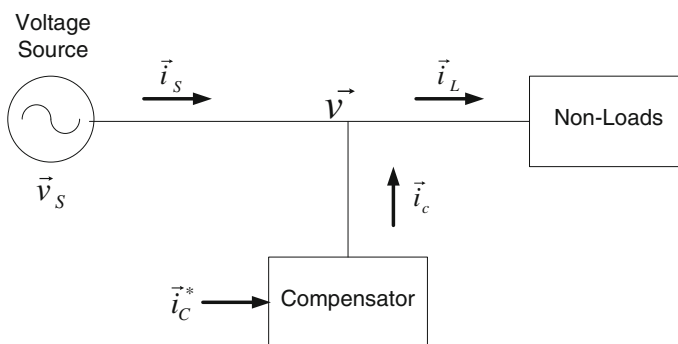


Fig. 2.7 Basic configuration for parallel compensation

Fig. 2.8 Corresponding compensation waveforms

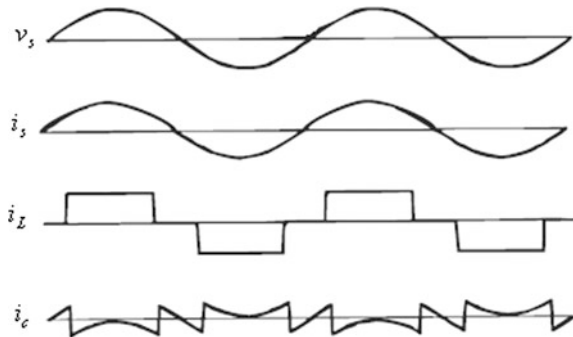
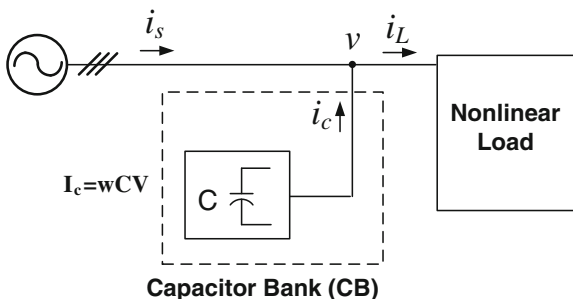


Fig. 2.9 Parallel capacitor banks



2.1.4.1 Capacitor Banks (CB)

Shunt Connected Capacitor Banks (CB) are extensively used in distribution power networks for power factor correction and feeder voltage control due to its relatively low cost and simplicity, which are developed mainly for reactive power compensation as shown in Fig. 2.9. However, since the current passing through the capacitor is proportional to the frequency and rate of change of voltage, the parallel capacitor banks may possibly be damaged and may even cause explosion as shown in Fig. 2.10 if the harmonics level is high. Moreover, CBs have other disadvantages such as static compensation capability (low dynamic performance), slow response time and resonance phenomena.

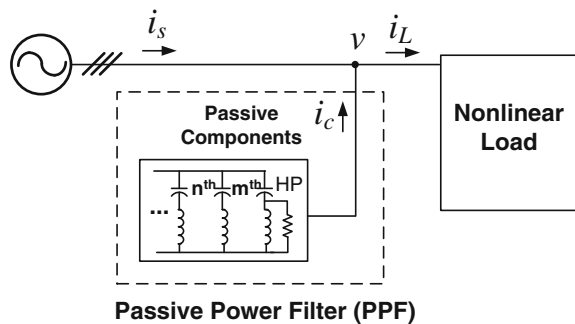
2.1.4.2 Passive Power Filters (PPF)

The first installation of Passive Power Filters (PPF) was in 1940s [1], PPFs have been widely used to suppress current harmonics and to compensate reactive power in distribution power systems due to its low cost, simplicity and high efficiency [2–4]. The system configurations of conventional passive power filters (PPF) are shown in Fig. 2.11. PPFs have been considered a good alternative for current harmonic compensation and displacement power factor correction. In general,

Fig. 2.10 Capacitor bank explosion in a power sub-station



Fig. 2.11 Parallel passive power filters



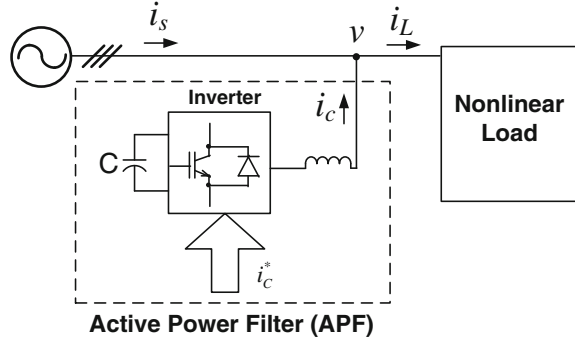
passive tuned filters have been used to eliminate low frequency current harmonics while high-pass passive units have been connected to attenuate the amplitude of high frequency current components [5, 6].

Technical disadvantages of PPFs have been extensively discussed and the inherent drawbacks [2–7] are:

- Source impedance strongly affects filtering characteristics.
- PPF may fall in series resonance when source harmonic voltage produces excessive harmonic current flowing into PPF.
- The shunt PPF may fall in parallel resonance with the source impedance.
- Small design tolerances are acceptable in the values of L and C. Small variations in the values of L or C modify the filter resonant frequency.

In addition, the PPF cannot modify their compensation characteristics according to the dynamic changes of the nonlinear load. The PPF generates fundamental frequency reactive power that changes the system voltage regulation, and if the filter is not designed properly or disconnected during low load operating conditions, over-voltage can be generated at its terminals [8].

Fig. 2.12 Topology of an active power filter (APF)



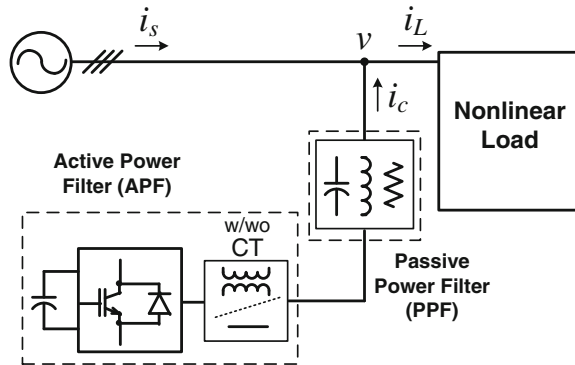
2.1.4.3 Active Power Filters (APF)

Since the concept “Active ac Power Filter” was first developed by Gyugyi in 1976 [9], the research studies of the active power filters (APFs) for current quality compensation prosper. With the remarkable progress in the speed and capacity of semiconductors switching devices, such as GTO thyristors and IGBTs, active power filters consisting of voltage or current source pulse width modulation (PWM) inverters have been studied and put into practical use [9, 10]. The APFs improve the power quality by injecting equal but opposite currents to compensate harmonic distortion, reactive power and unbalance currents in the power network. Figure 2.12 shows the system configuration of an APF. APFs consisting of voltage-fed PWM inverters using IGBTs or GTO thyristors are operating successfully, making significant contributions to improving power quality [11]. The APFs have the ability to overcome the above-mentioned disadvantages inherent in PPFs. However, the initial costs and running costs of the APFs are much higher than those of the PPFs [11, 12] due to the costs of semiconductor switching devices with its drivers, signal conditioning boards and digital controller(s).

2.1.4.4 Hybrid Active Power Filters (HAPF)

The CBs and PPFs cannot achieve dynamic reactive power and current quality compensation but their costs are relatively low. Sometimes, they may cause the operation problems in the power networks and even explosion. The APFs can overcome the above issues, but require high initial and operational costs. There is an alternative that is to combine both PPF and APF to form a Hybrid Active Power Filter (HAPF) [2, 11]. Hybrid active power filters (HAPFs) consist of active and passive filters connected in series or parallel to each other so that the advantages of both filters can be combined. The main purpose of the HAPF is to improve the compensation characteristics of PPF and reduce the costs of the APF. It leads to the effective balance in cost and performance for current quality compensation. Figure 2.13 shows a typical topology of an HAPF.

Fig. 2.13 A typical topology of a hybrid active power filter (HAPF)



In this book, the APF and HAPF are the focus. More details are discussed in Chaps. 3 and 4. In next section, the instantaneous compensation current technique to control the voltage source inverter (VSI) so as to inject or absorb the required compensating current instantaneously will be discussed in detail.

2.2 Instantaneous Power Compensation

Conventionally, the active, reactive and apparent powers are defined under steady-state, which are calculated by root-mean-square (RMS) voltage and current values and/or phase angle difference between the voltage and the current. It means that the compensation or the measurement equipment needs to wait at least one period long, for example 0.02 s in a 50 Hz frequency system, in order to get power components for further operations. As a result, the speed for the response by the compensators is relatively low.

In 1983 [13, 14], the instantaneous active and reactive power theory or the p-q theory was first introduced by Akagi et al. in Japan. The p-q theory [14, 15] pioneered in transforming voltages and currents from abc to $\alpha\beta$ coordinates. This concept has provided an effective method to compensate the instantaneous components of reactive power for the three-phase system without energy storage. However, this instantaneous reactive power theory has a conceptual limitation as pointed out in [16], i.e., the theory is only complete for the three-phase systems without zero-sequence currents and voltages.

To resolve this limitation and associated problems, Willems and Nabae proposed some attractive approaches to define the instantaneous active and reactive currents [16, 17]. However, their approaches are to deal with the decomposition of currents into orthogonal components rather than with the power components. In 1996, a generalized theory of instantaneous reactive power has been proposed [18–21]. The generalized theory is valid for sinusoidal or non-sinusoidal and balanced or unbalanced three-phase systems, with or without zero-sequence currents and/or voltages.

In this part, the definitions of this generalized instantaneous reactive power theory are given both in a - b - c and α - β - 0 coordinates. Finally, the compensating current can be calculated and is discussed in this section.

The instantaneous active power p and reactive power q of a three-phase system are defined as:

$$p = \vec{v} \cdot \vec{i} \quad (2.7)$$

$$\vec{q} = \vec{v} \times \vec{i} \quad (2.8)$$

$$q = \|\vec{q}\| \quad (2.9)$$

where “ $\cdot$ ” denotes the dot (internal) product or scalar product of vectors, “ $\times$ ” denotes the cross (exterior) product of vectors or vector product, v and i are the voltage and current vectors. Since the instantaneous reactive power obtained from Eq. (2.8) is a vector, the magnitude or the length of the instantaneous reactive power is often used, which is defined in Eq. (2.9).

2.2.1 Definitions in a - b - c Frame

For a three-phase system, instantaneous voltages v_a , v_b , v_c and instantaneous currents i_a , i_b , i_c are expressed as instantaneous space vector $\vec{v}$ and $\vec{i}$, i.e.,

$$\vec{v} = v_a \vec{n}_a + v_b \vec{n}_b + v_c \vec{n}_c \quad (2.10)$$

$$\vec{i} = i_a \vec{n}_a + i_b \vec{n}_b + i_c \vec{n}_c \quad (2.11)$$

where $(\vec{n}_a, \vec{n}_b, \vec{n}_c)$ is a basis in a - b - c frame.

By using the definitions of (2.7) and (2.8) the instantaneous active and reactive powers can be expressed as:

$$p = \vec{v} \cdot \vec{i} = v_a i_a + v_b i_b + v_c i_c \quad (2.12)$$

$$\vec{q} = \vec{v} \times \vec{i} = q_a \vec{n}_a + q_b \vec{n}_b + q_c \vec{n}_c \quad (2.13)$$

where

$$q_a = \begin{vmatrix} v_b & i_b \\ v_c & i_c \end{vmatrix}, \quad q_b = \begin{vmatrix} v_c & i_c \\ v_a & i_a \end{vmatrix}, \quad q_c = \begin{vmatrix} v_a & i_a \\ v_b & i_b \end{vmatrix}$$

Finally, instantaneous reactive power is:

$$\vec{q} = (v_b i_c - v_c i_b) \vec{n}_a + (v_c i_a - v_a i_c) \vec{n}_b + (v_a i_b - v_b i_a) \vec{n}_c \quad (2.14)$$

The instantaneous active current vector $\vec{i}_p$ and the instantaneous reactive current vector $\vec{i}_q$ are expressed as:

$$\vec{i}_p = \begin{bmatrix} i_{ap} \\ i_{bp} \\ i_{cp} \end{bmatrix} = \frac{p}{\vec{v} \cdot \vec{v}} \vec{v} = \frac{p}{v^2} \vec{v} \quad (2.15)$$

$$\vec{i}_q = \begin{bmatrix} i_{aq} \\ i_{bq} \\ i_{cq} \end{bmatrix} = \frac{\vec{q} \times \vec{v}}{\vec{v} \cdot \vec{v}} = \frac{\vec{q} \times \vec{v}}{v^2} \quad (2.16)$$

where $v = \|\vec{v}\| = \sqrt{v_a^2 + v_b^2 + v_c^2}$ and $\vec{i} = \vec{i}_p + \vec{i}_q$

2.2.2 Definitions in α - β -0 Frame

The α - β -0 transformation of a three-phase four-wire system gives:

$$\begin{bmatrix} i_0 \\ i_\alpha \\ i_\beta \end{bmatrix} = [P] \cdot \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (2.17)$$

$$\begin{bmatrix} v_0 \\ v_\alpha \\ v_\beta \end{bmatrix} = [P] \cdot \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} \quad (2.18)$$

where

$$\begin{aligned} P &= \sqrt{\frac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ \cos \theta & \cos(\theta - 120^\circ) & \cos(\theta + 120^\circ) \\ -\sin \theta & -\sin(\theta - 120^\circ) & -\sin(\theta + 120^\circ) \end{bmatrix} \\ &= \sqrt{\frac{2}{3}} \begin{bmatrix} 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \\ 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \end{aligned}$$

for $\theta = 0$, then $[P]^T = [P]^{-1}$, the matrix $[P]$ is the Clarke Transformation which transfers voltages and currents from a - b - c to α - β -0 coordinates.

As a result, the instantaneous active power can be expressed as:

$$P_{(\alpha\beta 0)} = \vec{v} \cdot \vec{i} = \vec{v}_{(\alpha\beta 0)} \cdot \vec{i}_{(\alpha\beta 0)} = v_\alpha i_\alpha + v_\beta i_\beta + v_0 i_0 \quad (2.19)$$

where the suffixes “(abc)” and “(αβ0)” denote the corresponding coordinates. Then, its instantaneous active power can be expressed as:

$$\begin{aligned} P_{(abc)} &= \vec{v}_{(abc)} \cdot \vec{i}_{(abc)} = [\vec{v}]^T [\vec{i}] \\ P_{(\alpha\beta 0)} &= \vec{v}_{(\alpha\beta 0)} \cdot \vec{i}_{(\alpha\beta 0)} = ([P][\vec{v}])^T [P][\vec{i}] = [\vec{v}]^T [P]^T [P][\vec{i}] \\ &= [\vec{v}]^T [I][\vec{i}] = [\vec{v}]^T [\vec{i}] \end{aligned}$$

so $P_{(\alpha\beta 0)} = \vec{v} \cdot \vec{i} = \vec{v}_{(\alpha\beta 0)} \cdot \vec{i}_{(\alpha\beta 0)} = \vec{v}_{(abc)} \cdot \vec{i}_{(abc)} = P_{(abc)}$

Furthermore, the instantaneous reactive power can be expressed as:

$$\vec{q}_{(\alpha\beta 0)} = \vec{v}_{(\alpha\beta 0)} \times \vec{i}_{(\alpha\beta 0)} = (v_\beta i_0 - v_0 i_\beta) \vec{n}_\alpha + (v_0 i_\alpha - v_\alpha i_0) \vec{n}_\beta + (v_\alpha i_\beta - v_\beta i_\alpha) \vec{n}_0 \quad (2.20)$$

$$\begin{aligned} \vec{q}_{(\alpha\beta 0)} &= \vec{v}_{(\alpha\beta 0)} \times \vec{i}_{(\alpha\beta 0)} = ([P]\vec{v}_{(abc)}) \times ([P]\vec{i}_{(abc)}) \\ &= [P](\vec{v}_{(abc)} \times \vec{i}_{(abc)}) = [P]\vec{q}_{(abc)} \end{aligned}$$

because $\| [P] \| = 1$ and $\| q_{(abc)} \| = \| q_{(\alpha\beta 0)} \|$ can be deduced.

The previous discussion indicates that the definitions of instantaneous active and reactive power can be expressed in both *abc* and *αβ0* coordinates, and the used coordinates are linear independent to each other.

Similarly, the instantaneous active and reactive current components in *α-β-0* coordinates can be written as:

$$\vec{i}_{p(\alpha\beta 0)} = \frac{P}{\vec{v}_{(\alpha\beta 0)} \cdot \vec{v}_{(\alpha\beta 0)}} \vec{v}_{(\alpha\beta 0)} \quad (2.21)$$

$$\vec{i}_{q(\alpha\beta 0)} = \frac{\vec{q} \times \vec{v}_{(\alpha\beta 0)}}{\vec{v}_{(\alpha\beta 0)} \cdot \vec{v}_{(\alpha\beta 0)}} \quad (2.22)$$

In addition, if the generalized instantaneous reactive power theory is applied to the three-phase three-wire system, i.e., $v_0 = 0$ and $i_0 = 0$, the following results can be derived.

$$P = \vec{v}_{(\alpha\beta 0)} \cdot \vec{i}_{(\alpha\beta 0)} = v_\alpha i_\alpha + v_\beta i_\beta \quad (2.23)$$

$$q = \| \vec{v}_{(\alpha\beta 0)} \times \vec{i}_{(\alpha\beta 0)} \| = v_\alpha i_\beta - v_\beta i_\alpha \quad (2.24)$$

2.2.3 Mixed-Coordinate Instantaneous Compensation

Based on previous parts, three-phase voltages and currents can be described by the 3-dimensional base either in abc or $\alpha\beta 0$ coordinates. In general, the equivalent voltage and current vectors can be expressed by (2.25) and (2.26) in xyz coordinate. The xyz can be abc or $\alpha\beta 0$ coordinates

$$\vec{v}_e = v_x \vec{n}_x + v_y \vec{n}_y + v_z \vec{n}_z \quad (2.25)$$

$$\vec{i}_e = i_x \vec{n}_x + i_y \vec{n}_y + i_z \vec{n}_z \quad (2.26)$$

The instantaneous active and reactive powers can be expressed by (2.27) and (2.28), which can be further decomposed into average and oscillating components as given in (2.29) and (2.30).

$$p = \vec{v}_e \cdot \vec{i}_e = v_x i_x + v_y i_y + v_z i_z \quad (2.27)$$

$$\vec{q} = \vec{v}_e \times \vec{i}_e = (v_x i_y - v_y i_x) \vec{n}_z + (v_y i_z - v_z i_y) \vec{n}_x + (v_z i_x - v_x i_z) \vec{n}_y \quad (2.28)$$

$$p = \bar{p} + \tilde{p} \quad (2.29)$$

$$\vec{q} = \bar{q} + \tilde{q} \quad (2.30)$$

Additionally, the instantaneous active and reactive current can be expressed as follows:

$$\vec{i}_p = \frac{p \cdot \vec{v}_e}{\vec{v}_e \cdot \vec{v}_e} \quad (2.31)$$

$$\vec{i}_q = \frac{q \times \vec{v}_e}{\vec{v}_e \cdot \vec{v}_e} \quad (2.32)$$

Finally, its injecting current for compensation can be expressed by (2.33), which shows that only oscillating component of instantaneous active power and all instantaneous reactive power are compensated.

$$\vec{i}_c^* = \frac{\tilde{p} \cdot \vec{v}_e}{\vec{v}_e \cdot \vec{v}_e} + \frac{\vec{q} \times \vec{v}_e}{\vec{v}_e \cdot \vec{v}_e} \quad (2.33)$$

No matter the voltages or currents are expressed in abc or $\alpha\beta 0$ coordinates, they are describing the same voltage or current vectors in 3-dimensional base. One should be noted that the voltage sources should be assumed to be balanced and sinusoidal. Otherwise, the instantaneous compensated current in (2.33) should be further modified as there are negative and/or zero sequence average power components existed in instantaneous average active power. Without further

modification, the compensated source currents are not sinusoidal and not balanced for unbalanced and/or non-sinusoidal voltages, in which further discussions for unbalanced and non-sinusoidal voltage cases are not included in this book.

The instantaneous compensating currents can be calculated by mixed coordinate signals. As a result, the computation steps and time by a digital controller, such as digital signal processor (DSP) can be reduced. The computation error can also be lowered, in which the detailed algorithm description immediately follows.

The instantaneous compensating active current can be expressed as:

$$\vec{i}_{cp} = \frac{\tilde{p} \cdot \vec{v}_e}{\vec{v}_e \cdot \vec{v}_e} \quad (2.34)$$

where $\tilde{p}$ is the oscillating part of the instantaneous active power which can be calculated by

$$p = v_a i_a + v_b i_b + v_c i_c = p_a + p_b + p_c \quad (2.35)$$

The dot product of the instantaneous voltage is

$$\Delta = \vec{v}_e \cdot \vec{v}_e = v_a^2 + v_b^2 + v_c^2 \quad (2.36)$$

Finally, its instantaneous compensating active current can be expressed as:

$$\vec{i}_{cp} = \frac{\tilde{p}}{\Delta} (v_a \vec{n}_a + v_b \vec{n}_b + v_c \vec{n}_c) = \frac{\tilde{p}}{\Delta} (v_a \vec{n}_a + v_b \vec{n}_b + v_c \vec{n}_c) \quad (2.37)$$

All the instantaneous reactive power components are required for calculating the instantaneous reactive compensating current,

$$\vec{i}_{cq} = \frac{\vec{q} \times \vec{v}_e}{\vec{v}_e \cdot \vec{v}_e} \quad (2.38)$$

The instantaneous reactive compensating power vector components can be projected on $\{\vec{n}_a, \vec{n}_b, \vec{n}_c\}$ base. The instantaneous reactive power can be calculated by (2.39). The cross product of reactive power $\vec{q}$ and voltage vector $\vec{v}_e$ can be derived from (2.40).

$$\vec{q} = (v_b i_c - v_c i_b) \vec{n}_a + (v_c i_a - v_a i_c) \vec{n}_b + (v_a i_b - v_b i_a) \vec{n}_c = A \vec{n}_a + B \vec{n}_b + C \vec{n}_c \quad (2.39)$$

$$\vec{q} \times \vec{v}_e = \begin{vmatrix} \vec{n}_a & \vec{n}_b & \vec{n}_c \\ A & B & C \\ v_a & v_b & v_c \end{vmatrix} = (B v_c - C v_b) \vec{n}_a + (C v_a - A v_c) \vec{n}_b + (A v_b - B v_a) \vec{n}_c \quad (2.40)$$

where

$$A = (v_b i_c - v_c i_b), B = (v_c i_a - v_a i_c) \text{ and } C = (v_a i_b - v_b i_a)$$

Finally, substitute (2.40) into (2.38), its instantaneous compensating reactive current can be described by (2.41).

$$\begin{aligned} \vec{i}_{cq} = \frac{1}{\Delta} [& (v_c^2 i_a + v_b^2 i_a - v_a v_c i_c - v_a v_b i_b) \vec{n}_a \\ & + (v_a^2 i_b + v_c^2 i_b - v_b v_a i_a - v_b v_c i_c) \vec{n}_b \\ & + (v_b^2 i_c + v_a^2 i_c - v_c v_b i_b - v_c v_a i_a) \vec{n}_c] \end{aligned} \quad (2.41)$$

By just taking out the first term of (2.41) and substituting (2.35) into it, after simplification, the instantaneous compensating reactive current in phase a can be obtained as given in (2.42).

$$\begin{aligned} \vec{i}_{cqa} &= \frac{1}{\Delta} (v_c^2 i_a + v_b^2 i_a - v_a v_c i_c - v_a v_b i_b) \\ &= \frac{1}{\Delta} (i_a (v_b^2 + v_c^2) - v_a p_b - v_a p_c) \\ &= \frac{1}{\Delta} (i_a (\Delta - v_a^2) - v_a p_b - v_a p_c) \\ &= \frac{1}{\Delta} (i_a \Delta - v_a p_a - v_a p_b - v_a p_c) \\ &= i_a - \frac{p}{\Delta} v_a \end{aligned} \quad (2.42)$$

By considering phase b and phase c as well, the instantaneous compensating reactive current in 3-dimensional case can be given in (2.43).

$$\vec{i}_{cq} = \left(i_a - \frac{p}{\Delta} v_a\right) \vec{n}_a + \left(i_b - \frac{p}{\Delta} v_b\right) \vec{n}_b + \left(i_c - \frac{p}{\Delta} v_c\right) \vec{n}_c \quad (2.43)$$

The final instantaneous compensating current is composed of instantaneous compensating active and reactive currents as shown in (2.44). After simplification, the final instantaneous compensating current for abc coordinates is shown in (2.45).

$$\vec{i}_c^* = \vec{i}_{cp} + \vec{i}_{cq} \quad (2.44)$$

$$\vec{i}_c^* = \left(i_a - \frac{\bar{p}}{\Delta} v_a\right) \vec{n}_a + \left(i_b - \frac{\bar{p}}{\Delta} v_b\right) \vec{n}_b + \left(i_c - \frac{\bar{p}}{\Delta} v_c\right) \vec{n}_c \quad (2.45)$$

Moreover, the voltage and current vectors can also be expressed in $\alpha\beta 0$ coordinates; its compensating current can be described by (2.46).

$$\vec{i}_c^* = \left(i_\alpha - \frac{\bar{p}}{\Delta} v_\alpha\right) \vec{n}_\alpha + \left(i_\beta - \frac{\bar{p}}{\Delta} v_\beta\right) \vec{n}_\beta + \left(i_0 - \frac{\bar{p}}{\Delta} v_0\right) \vec{n}_0 \quad (2.46)$$

where $\Delta = v_a^2 + v_b^2 + v_c^2$, $p = v_a i_a + v_b i_b + v_c i_c$ and $\bar{p}$ is the average value of p . Figure 2.14 shows the mixed coordinate instantaneous compensation algorithm.

Simulation verifications are performed by Matlab/Simulink. Figure 2.15 shows the three-phase source voltages while Fig. 2.16 shows the three-phase currents, the currents are obviously unbalanced and with harmonics. Figure 2.17 shows the source currents after transformation in α , β and 0 coordinates.

Figure 2.18 shows the compensated source currents with the compensation consideration of instantaneous reactive current only. It clearly shows that the source currents are not perfectly sinusoidal and balanced after compensation.

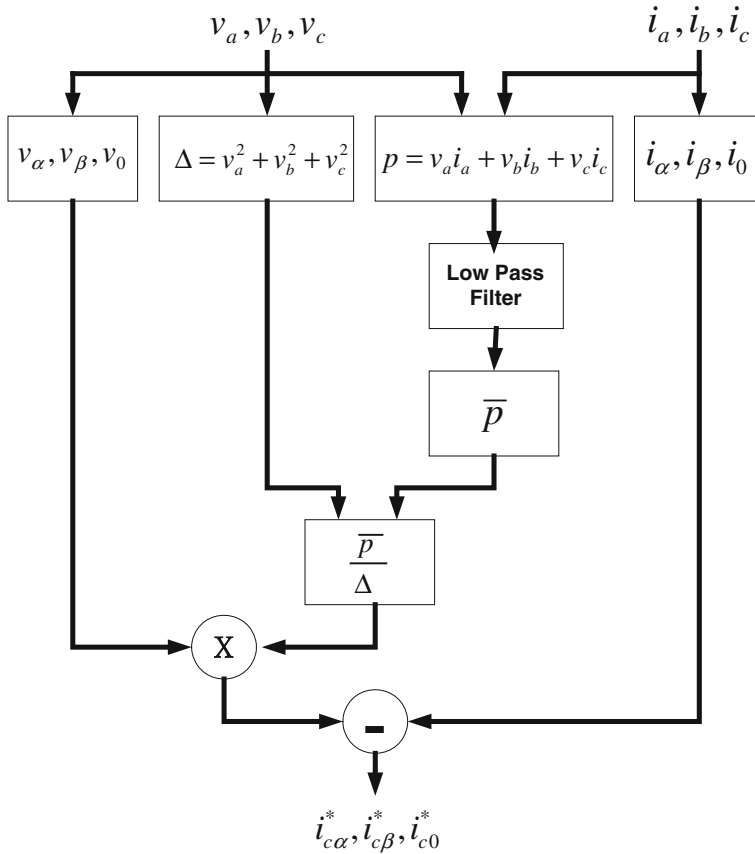


Fig. 2.14 Mixed coordinate instantaneous compensation signal flow diagram

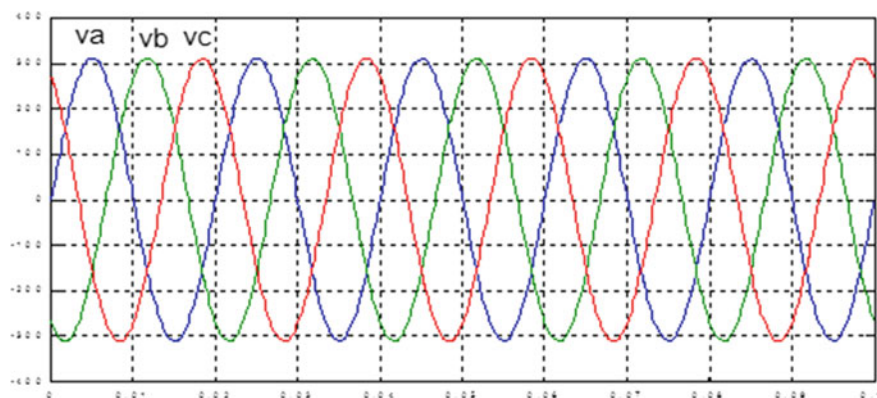


Fig. 2.15 Three-phase source voltage

Fig. 2.16 Three-phase currents

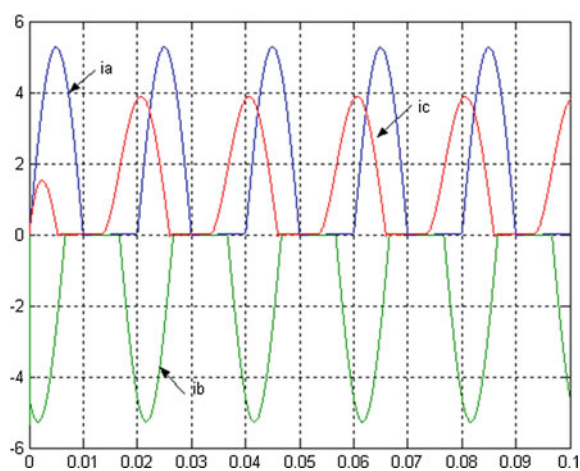
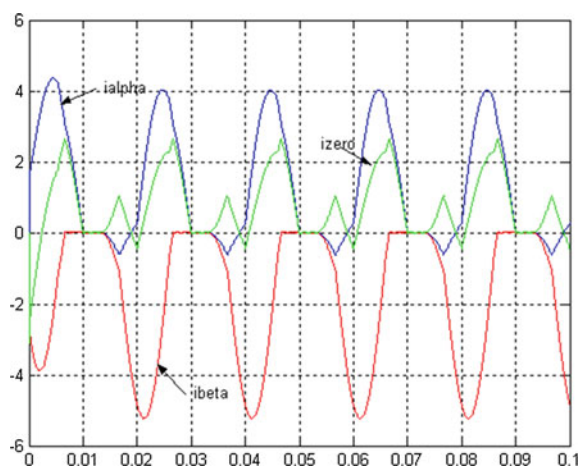


Fig. 2.17 Three-phase source currents in $\alpha\beta 0$ coordinates



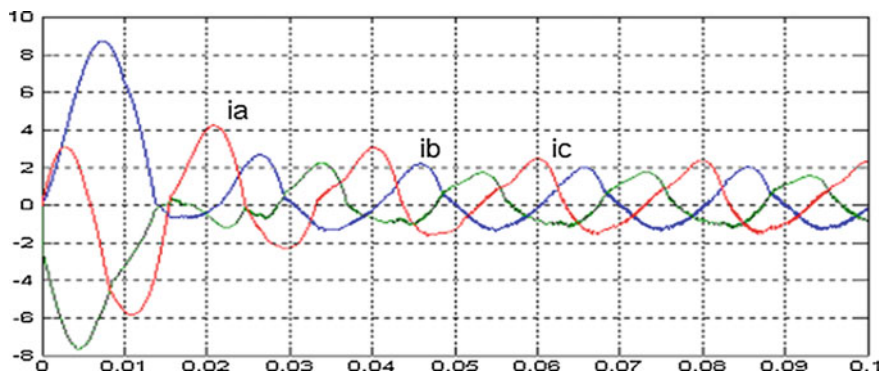


Fig. 2.18 Compensated source currents with compensation consideration of instantaneous reactive current only

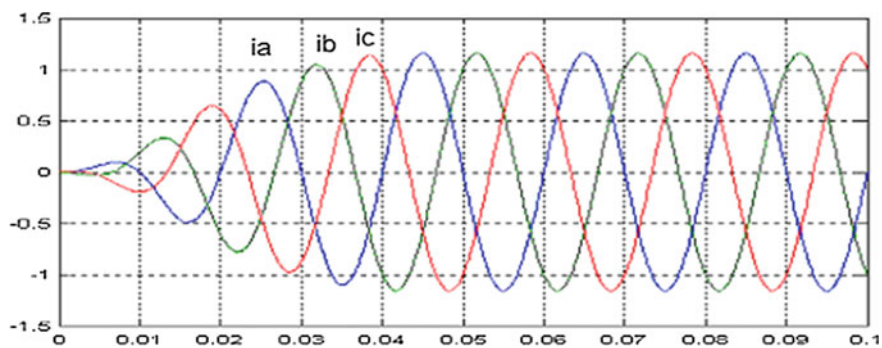
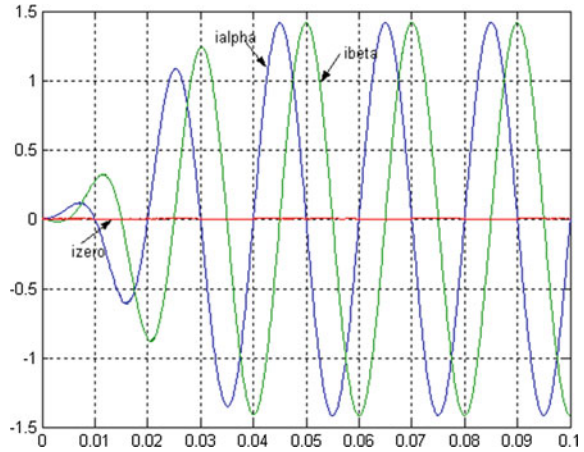


Fig. 2.19 Compensated source currents by mixed coordinate instantaneous compensation algorithm

Figure 2.19 shows the simulated result by Mixed Coordinate Instantaneous Compensation Algorithm. It clearly shows that three-phase source currents are sinusoidal and balanced after compensation. Figure 2.20 shows the α , β and 0 coordinate source currents after compensation by Mixed Coordinate Instantaneous Compensation Algorithm.

From the above simulation results, there is a transient part between $t = 0$ s and $t = 0.03$ s. Under the transient period, there is a large influence due to the transient response of the low-pass filter to obtain the instantaneous average power. The response of the low-pass filter affects this transient response. The faster the response of the low-pass filter, the shorter the transient period will be. However, the low-pass filter design is out of the scope of this section. The above mixed coordinate instantaneous power can simplify the calculation steps comparing to the original instantaneous reactive power theories.

Fig. 2.20 α , β and 0 coordinate source currents after compensation



2.3 Three-Phase Converters and Their Discussions

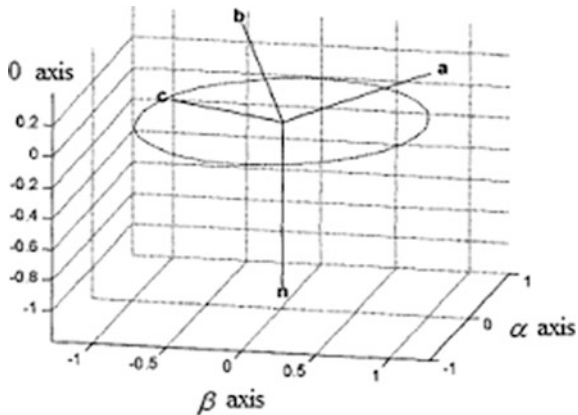
In this section, three basic three-phase converter topologies are analyzed, compared and discussed in two-level and multi-level topologies such as three-leg, three-leg center-split and four-leg inverters in three-phase three-wire and three-phase four-wire systems. Since Year 2000, the 3-Dimensional Pulse Width Modulation (3D PWM) technique has become one of the popular research topics in three-phase four-wire applications, such as distributed power generators and active power filters [22–25]. The challenges to investigate the three-phase four-wire converters are:

- (1) To choose a suitable circuit topology,
- (2) To have a decoupling transformation mathematical model for analysis,
- (3) To develop a control strategy with high performance, and
- (4) To apply the system in high power applications.

The three-phase voltage source inverters (VSI) normally have two ways of providing a neutral connection for the three-phase four-wire systems: Three-Leg Center-Split and Four-Leg inverters [22]. However, according to the past studies [23, 25–27], most of the three-phase four-wire converter investigations mainly focus on the four-leg topology due to two reasons. First, the maximum achievable peak value of line-to-neutral voltage of a three-leg center-split inverter is lower than a four-leg inverter, which means that the utilization of dc linked voltage, V_{dc} , of a three-leg center-split inverter is lower. Second, large dc linked capacitors are needed in a three-leg center-split inverter to maintain a small ripple dc voltage. Based on the above points, the development of the control of the three-leg center-split topology is slower than that of the four-leg.

On the other hand, the control strategies of three-phase four-wire inverters are studied from abc Hysteresis controls to $\alpha\beta 0$ space vector modulations [23–30] by abc- $\alpha\beta 0$ transformations. Recently, there was 4×4 Quad-Transformation matrix

Fig. 2.21 Four-leg inverter primary voltage vectors [27, 30]



[27] which describes the direct transformation between the 4 Degree-of-Freedom Leg-Output Voltage Space of the Four-Leg Inverter and its 3 Degree-of-Freedom output voltage space. The Quad-Transform [27, 30] was proposed to analyze the four-leg systems in $\alpha\beta 0$ coordinates. The leg-voltages of a four-leg inverter could be modeled as a projection into the 3 dimensional output-voltage space along 4 vectors through a Quad-transform. These vectors were denoted as the primary voltage-vectors of the inverter and were depicted in Fig. 2.21. The a, b, c, and n vectors in Fig. 2.21 formed the corners of a symmetrical tetrahedron [27, 30].

A unique 4×4 decoupling transformation matrix was presented for the four-leg circuit topology, which was orthonormal, and hence invertible. The row and column vectors formed a basis for the 4 Degree-Of-Freedom leg-voltage space. However, the concept of “Primary Voltage Vectors” that was defined as “An N-leg inverter has an N-dimensional leg-voltage space, but only controls an $(N - 1)$ dimensional output voltage space. The leg voltages of an N-Leg inverter project into the output-voltage space along the N vectors of equal magnitude equally arranged in an $(N - 1)$ Degree-Of-Freedom space [27, 30]”. In this chapter, a 3 dimensional basis is used as follows. Furthermore, when four-leg and three-leg center-split multilevel converters are taken into account in three-phase four-wire high power applications, the advantages of the three-leg center-split multilevel topology will be explained in the following sections.

2.3.1 Realization of 3 Dimensional Coordinates for Three-Phase Systems

One should note that the voltage space vector development hereafter is based on the assumption that instantaneous power is constant after and before transformation from one into other coordinates in this book. In a three-phase three-wire or three-phase four-wire system, an instantaneous voltage vector, $\vec{v}$, can be represented

by a linear combination of the vectors, $\vec{n}_a$, $\vec{n}_b$ and $\vec{n}_c$ which can form a basis, $B = \{\vec{n}_a, \vec{n}_b, \vec{n}_c\}$, for a vector to span with. The vectors, $\vec{n}_a$, $\vec{n}_b$ and $\vec{n}_c$ are linear independent to each other in a physical sense. Basically, the dimension of this instantaneous voltage vector is 3.

$$\vec{v} = K_a \vec{n}_a + K_b \vec{n}_b + K_c \vec{n}_c \quad (2.47)$$

where K_a , K_b and K_c are scalars.

Let $M = \{\vec{n}_1, \vec{n}_2, \dots, \vec{n}_n\}$ and $M' = \{\vec{w}_1, \vec{w}_2, \dots, \vec{w}_m\}$ be the two bases for a finite-dimensional vector space $\mathcal{Q}$. Since M is a basis and M' is a linearly independent set, which implies that $m = n$. As a result, any two bases for a finite-dimensional vector space have the same number of vectors. An instantaneous voltage vector can be represented by $B = \{\vec{n}_a, \vec{n}_b, \vec{n}_c\}$ and $B' = \{\vec{n}_\alpha, \vec{n}_\beta, \vec{n}_0\}$, and it can span in a 3 dimension space. The 3-phase instantaneous voltages can be transferred from abc into $\alpha\beta 0$ coordinates by the coordinate transformation matrix [19], $P_{\alpha\beta 0}$, (2.48) under a linear power transformation consideration.

$$\begin{bmatrix} v_\alpha \\ v_\beta \\ v_0 \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \\ 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \end{bmatrix} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} = P_{\alpha\beta 0} \begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix} \quad (2.48)$$

$$\text{where } P_{\alpha\beta 0} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos 0^\circ & \cos(120^\circ) & \cos(240^\circ) \\ \sin 0^\circ & \sin(120^\circ) & \sin(240^\circ) \\ \sin(45^\circ) & \sin(45^\circ) & \sin(45^\circ) \end{bmatrix}.$$

Under the linear power transformation consideration [19], there are two conditions that need to be fulfilled, (2.49) and (2.50).

$$I = [P_{\alpha\beta 0}]^T [P_{\alpha\beta 0}] \quad (2.49)$$

$$\|P_{\alpha\beta 0}\| = 1 \quad (2.50)$$

From (2.49), it implies that $[P_{\alpha\beta 0}]^{-1} = [P_{\alpha\beta 0}]^T$, which has the property that the matrix $P_{\alpha\beta 0}$ is said to be an orthogonal matrix and I is an identity matrix. As a result, vector coordinates can be described as shown in Fig. 2.22. Another important characteristic is noticed as follows. The determinant of $P_{\alpha\beta 0}$ is equal to 1 so that the volumes bounded after and before the transformation are the same [31].

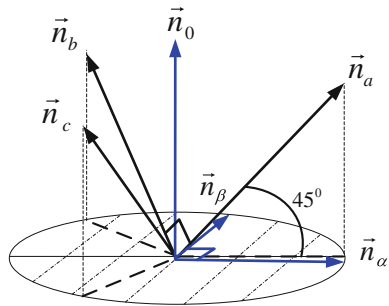
$$|P_{\alpha\beta 0}| = 1 \quad (2.51)$$

It means that an instantaneous voltage vector can be expressed as (2.52).

$$\vec{v} = K_\alpha \vec{n}_\alpha + K_\beta \vec{n}_\beta + K_0 \vec{n}_0 \quad (2.52)$$

where K_α , K_β and K_0 are scalars.

Fig. 2.22 Transformation basis allocation



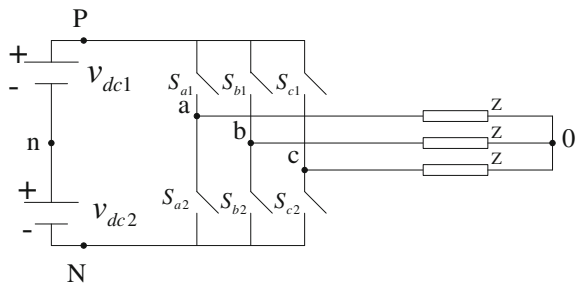
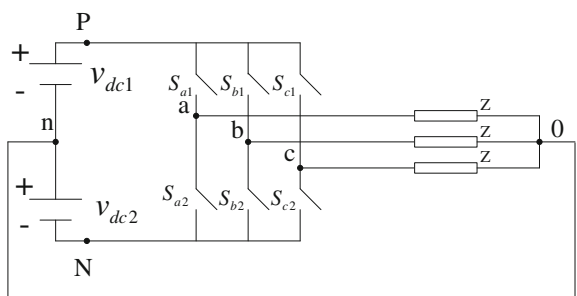
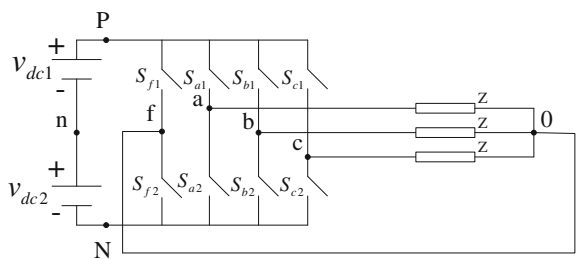
The zero sequence is the projection of vectors in a, b and c coordinates with 45° on it. When the vector coordinates are considered from top view of Fig. 2.22, the abc coordinates are apart from each other with 120° , but it should be noticed that they are orthogonal to each other as shown in Fig. 2.22. All the transformations taken by abc and $\alpha\beta 0$ coordinates are in 3 dimensions, not 3 to 2 dimensions in the three-phase systems. It can also be explained why phase angles of abc coordinates are 120° apart with orthogonal basis to each other. The above concept can be applied in multi-phase systems such as the four-phase systems described by a 4D transformation for analysis. However, it is not within the discussion of this book.

2.3.2 Basic Three-Phase Converters, 3D Coordinates and Their Comparisons

The literatures on controlling three-phase three-wire converters are rich; recently more and more attention is paid to the three-phase four-wire converters [23–27, 30]. In this section, in contrast to the earlier approaches, the comparisons among the basic three-phase converters of their voltage vectors in 3-dimensional coordinates are performed, rather than using the switching tables only. These basic converters are of three-leg, three-leg center-split and four-leg. All the corresponded converters are illustrated in 2-level equivalent circuits, as shown in Figs. 2.23, 2.24 and 2.25. In this section, $v_{dc} = v_{dc1} = v_{dc2}$ is assumed for comparisons and none of the switching losses are being considered.

2.3.2.1 Two-Level Three-Leg Converters

From Sect. 2.3.1, it is found that all the voltage vectors of three-phase converters can be analyzed by a 3-D transformation basis. According to Fig. 2.23, the switching functions can be given as $S_j = \{+1, -1\}$, where $j = \{a, b, c\}$. For example, when $S_a = +1$, S_{a1} is turned on and S_{a2} is turned off, and vice versa.

Fig. 2.23 Two-level three-leg converters**Fig. 2.24** Two-level three-leg center-split converters**Fig. 2.25** Two-level four-leg converters

The converter output voltages in a-b-c coordinates can be expressed in (2.53). It is required to transform the voltages from a-b-c into α - β -0 coordinates for comparisons through (2.54).

$$\begin{bmatrix} v_{a0} \\ v_{b0} \\ v_{c0} \end{bmatrix} = \frac{1}{6}(2v_{dc}) \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \quad (2.53)$$

where $S_j = \{+1, -1\}$ and $j = \{a, b, c\}$

$$\begin{bmatrix} v_{\alpha 0} \\ v_{\beta 0} \\ v_{00} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} v_{a0} \\ v_{b0} \\ v_{c0} \end{bmatrix} \quad (2.54)$$

Furthermore, (2.53) can be inserted into (2.54), which gets (2.55).

$$\begin{aligned} \begin{bmatrix} v_{\alpha 0} \\ v_{\beta 0} \\ v_{00} \end{bmatrix} &= \sqrt{\frac{2}{3}} \frac{(2v_{dc})}{6} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \\ &= v_{dc} \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \\ 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \end{aligned} \quad (2.55)$$

As a result, an instantaneous converter output voltage vector of a three-leg converter can be expressed as (2.56) so that the scalars, K_α , K_β and K_0 , are $\sqrt{2/3}v_{dc}S_\alpha$, $1/\sqrt{2}v_{dc}S_\beta$ and 0 correspondingly. It should be noticed that v_{dc} is the half value of v_{PN} as shown in Fig. 2.23. There is no zero-component coincidentally, which changes 3-D a - b - c coordinates on 2-D α - β plane.

$$\vec{v}_{three-leg} = v_{dc} \left[\sqrt{\frac{2}{3}} S_\alpha \cdot \vec{n}_\alpha + \frac{1}{\sqrt{2}} S_\beta \cdot \vec{n}_\beta \right] \quad (2.56)$$

where $S_\alpha = S_a - \frac{1}{2}S_b - \frac{1}{2}S_c$ and $S_\beta = S_b - S_c$.

2.3.2.2 Two-Level Three-Leg Center-Split Converters

Comparing Fig. 2.24 with Fig. 2.23, node “O” is connected with node “n”. According to the defined switching functions, the voltages are defined as (2.57) in a - b - c coordinates. It forms a two-level three-leg center-split converter shown in Fig. 2.24.

$$\begin{bmatrix} v_{a0} \\ v_{b0} \\ v_{c0} \end{bmatrix} = v_{dc} \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \quad (2.57)$$

where $S_j = \{+1, -1\}$ and $j = \{a, b, c\}$

Furthermore, the converter output voltages in α - β -0 coordinates are expressed in (2.58). An instantaneous converter output voltage vector of a three-leg center-split converter can be represented by (2.59).

$$\begin{aligned}
 \begin{bmatrix} v_{\alpha 0} \\ v_{\beta 0} \\ v_{00} \end{bmatrix} &= v_{dc} \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \\
 &= v_{dc} \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1/2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \\ 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \quad (2.58)
 \end{aligned}$$

In this case, scalars, K_α , K_β and K_0 , are $\sqrt{2/3}v_{dc}S_\alpha$, $1/\sqrt{2}v_{dc}S_\beta$ and $1/\sqrt{3}v_{dc}S_0$ correspondingly.

$$\vec{v}_{center-split} = v_{dc} \left[\sqrt{\frac{2}{3}} S_\alpha \cdot \vec{n}_\alpha + \frac{1}{\sqrt{2}} S_\beta \cdot \vec{n}_\beta + \frac{1}{\sqrt{3}} S_0 \cdot \vec{n}_0 \right] \quad (2.59)$$

where $S_\alpha = S_a - \frac{1}{2}S_b - \frac{1}{2}S_c$, $S_\beta = S_b - S_c$ and $S_0 = S_a + S_b + S_c$.

2.3.2.3 Two-Level Four-Leg Converters

There is a growing interest in the four-leg converters for three-phase four-wire applications. Figure 2.25 shows the equivalent circuits of a four-leg converter. The mid-point “F” of an addition leg is connected with node “O”. When switching functions are only considered with +1 and -1, there are 2 operation modes for the four-leg converter, as shown in Figs. 2.26 and 2.27.

Fig. 2.26 Mode 1 when switching function of S_f is +1

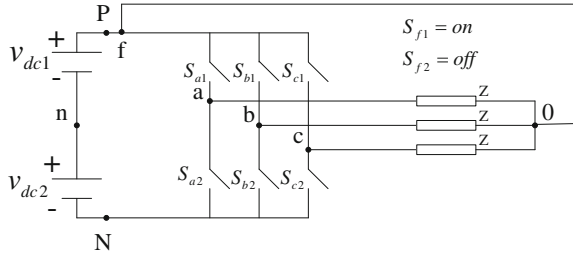


Fig. 2.27 Mode 2 when switching function of S_f is -1

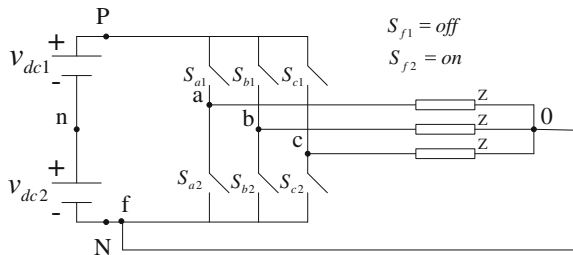
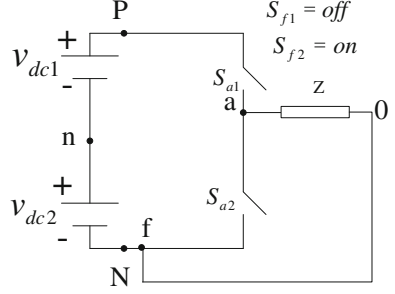


Fig. 2.28 One-leg converter in Mode 2



When S_f is +1, the node “O” is connected with node “P”. In another mode, node “O” is connected with node “N” if S_f is -1.

For simplification only one leg of them is considered in mode 2 as shown in Fig. 2.28. It has $v_{a0} = v_{dc1} + v_{dc2} = 2v_{dc}$. When $S_{a1} = 1$, $v_{a0} = 2v_{dc}$. In another way, when $S_a = -1$, $v_{a0} = 0$. As a result, the instantaneous voltage can be deduced as (2.60).

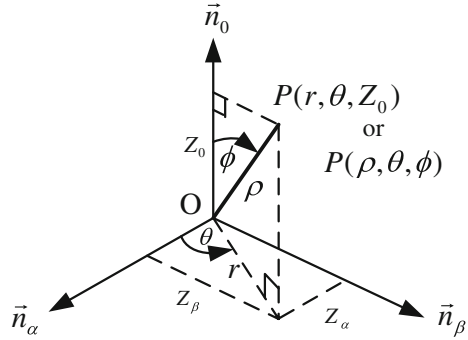
$$v_{a0} = (S_a - S_f) \cdot \frac{2v_{dc}}{2} \quad (2.60)$$

As a result, the instantaneous converter output voltages can be defined in a - b - c coordinates as shown in (2.61). The four-leg inverter can produce three independent output voltages, regardless of load. Furthermore, the instantaneous converter output voltages in α - β -0 coordinates from a - b - c coordinates are expressed in (2.62). An instantaneous converter output voltage vector of a four-leg converter can be represented by (2.63). Scalars, K_α , K_β and K_0 , are $\sqrt{2/3}v_{dc}S_\alpha$, $1/\sqrt{2}v_{dc}S_\beta$ and $1/\sqrt{3}v_{dc}S_0$ correspondingly.

$$\begin{bmatrix} v_{a0} \\ v_{b0} \\ v_{c0} \end{bmatrix} = v_{dc} \begin{bmatrix} 1 & 0 & 0 & -1 \\ 0 & 1 & 0 & -1 \\ 0 & 0 & 1 & -1 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \\ S_f \end{bmatrix} \quad (2.61)$$

$$\begin{aligned} \begin{bmatrix} v_{\alpha 0} \\ v_{\beta 0} \\ v_{00} \end{bmatrix} &= \sqrt{\frac{2}{3}}v_{dc} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix} \begin{bmatrix} 1 & 0 & 0 & -1 \\ 0 & 1 & 0 & -1 \\ 0 & 0 & 1 & -1 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \\ S_f \end{bmatrix} \\ &= \sqrt{\frac{2}{3}}v_{dc} \begin{bmatrix} 1 & -1/2 & -1/2 & 0 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 & 0 \\ 1/\sqrt{2} & 1/\sqrt{2} & 1/\sqrt{2} & -3/\sqrt{2} \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \\ S_f \end{bmatrix} \end{aligned} \quad (2.62)$$

Fig. 2.29 Defined parameters in coordinates



$$\vec{v}_{four-leg} = v_{dc} \left[\sqrt{\frac{2}{3}} S_\alpha \cdot \vec{n}_\alpha + \frac{1}{\sqrt{2}} S_\beta \cdot \vec{n}_\beta + \frac{1}{\sqrt{3}} S_0 \cdot \vec{n}_0 \right] \quad (2.63)$$

where $S_\alpha = S_a - \frac{1}{2}S_b - \frac{1}{2}S_c$, $S_\beta = S_b - S_c$ and $S_0 = S_a + S_b + S_c - 3S_f$.

2.3.2.4 Three-Phase Converters, Voltage Space Vectors and Their Comparisons

In 3-D aspect, a vector or a switching state can be represented by Rectangular, Cylindrical or Spherical Coordinates. According to parameters defined in Fig. 2.29, the instantaneous converter output voltage vector can be further considered as follows.

$$\vec{v} = v_{dc} \cdot (Z_\alpha \vec{n}_\alpha + Z_\beta \vec{n}_\beta + Z_0 \vec{n}_0) \quad (2.64)$$

where $Z_\alpha = \sqrt{\frac{2}{3}} S_\alpha$, $Z_\beta = \frac{1}{\sqrt{2}} S_\beta$ and $Z_0 = \frac{1}{\sqrt{3}} S_0$.

The parameters defined in Fig. 2.29 can be expressed as:

$$\rho = \sqrt{Z_\alpha^2 + Z_\beta^2 + Z_0^2} \quad (2.65)$$

$$r = \sqrt{Z_\alpha^2 + Z_\beta^2} \quad (2.66)$$

$$\theta = \tan^{-1} \left(\frac{Z_\beta}{Z_\alpha} \right) \quad (2.67)$$

$$\phi = \cos^{-1} \left(\frac{Z_0}{\rho} \right) = \sin^{-1} \left(\frac{r}{\rho} \right) \quad (2.68)$$

Table 2.2 Two-level three-leg converter voltage vector parameters

	S_a	S_b	S_c	S_x	S_β	S_0	r	θ	Z_0	ρ	ϕ
$\vec{V}_1$	1	-1	-1	2	0	0	1.633	0°	0	1.633	90°
$\vec{V}_2$	1	1	-1	1	2	0	1.633	60°	0	1.633	90°
$\vec{V}_3$	-1	1	-1	-1	2	0	1.633	120°	0	1.633	90°
$\vec{V}_4$	-1	1	1	-2	0	0	1.633	180°	0	1.633	90°
$\vec{V}_5$	-1	-1	1	-1	-2	0	1.633	240°	0	1.633	90°
$\vec{V}_6$	1	-1	1	1	-2	0	1.633	300°	0	1.633	90°
$\vec{V}_{0p}$	1	1	1	0	0	0	0	*	0	0	90°
$\vec{V}_{0n}$	-1	-1	-1	0	0	0	0	*	0	0	90°

However, $v_{dc} = 1$ is assumed for simplifying the comparison among three-phase converters as follows.

A. A Two-Level Three-Leg Converter and Its Voltage Space Vector

In a two-level three-leg converter, there are $2^3 = 8$ available voltage vectors but with only 7 achievable switching states. Coincidentally, there is no zero sequence component of voltage space vectors in α - β -0 coordinates. The available vectors are projected from 3D a - b - c coordinates on a 2D α - β plane. Table 2.2 summarizes parameters of the 2-level voltage vectors for a two-level three-leg converter. The mark, *, means “undefined” in following Tables. Figure 2.30 shows the allocation of available output voltage vectors of a two-level three-leg converter.

In conventional 2-D Pulse Width Modulation techniques, voltage space vector allocation can be illustrated on a α - β plane, which is shown in Fig. 2.30. There are totally 8 available voltage vectors with 6 directional vectors and 2 zero vectors. These 2 zero vectors are utilized in optimizing the switching sequence in order to reduce the switching loss.

Fig. 2.30 Allocation of available output voltage vectors of a two-level three-leg converter

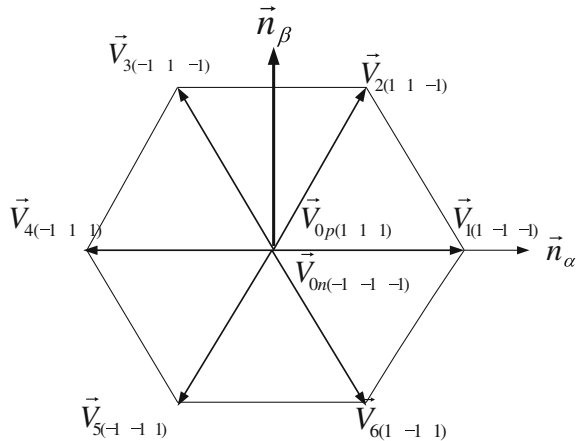


Table 2.3 Two-level three-leg center-split converter voltage vector parameters

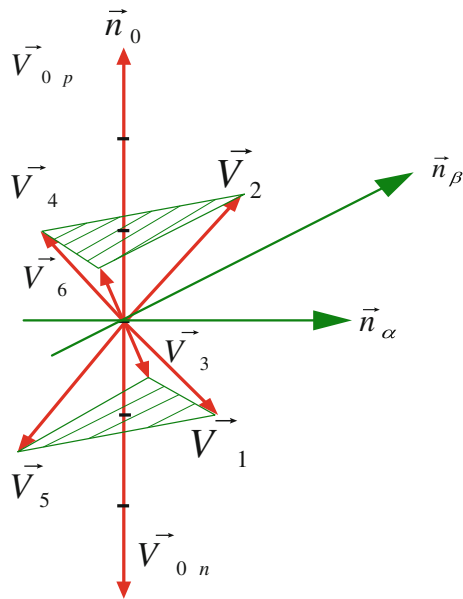
	S_a	S_b	S_c	S_α	S_β	S_0	r	θ	Z_0	ρ	ϕ
$\vec{V}_1$	1	-1	-1	2	0	-1	1.633	0°	-0.577	$\sqrt{3}$	109.46°
$\vec{V}_2$	1	1	-1	1	2	1	1.633	60°	0.577	$\sqrt{3}$	70.54°
$\vec{V}_3$	-1	1	-1	-1	2	-1	1.633	120°	-0.577	$\sqrt{3}$	109.46°
$\vec{V}_4$	-1	1	1	-2	0	1	1.633	180°	0.577	$\sqrt{3}$	70.54°
$\vec{V}_5$	-1	-1	1	-1	-2	-1	1.633	240°	-0.577	$\sqrt{3}$	109.46°
$\vec{V}_6$	1	-1	1	1	-2	1	1.633	300°	0.577	$\sqrt{3}$	70.54°
$\vec{V}_{0p}$	1	1	1	0	0	3	0	*	$\sqrt{3}$	$\sqrt{3}$	0°
$\vec{V}_{0n}$	-1	-1	-1	0	0	-3	0	*	$-\sqrt{3}$	$\sqrt{3}$	180°

B. A Two-Level Three-Leg Center-Split Converter and Its Voltage Space Vector

In a two-level three-leg center-split converter, there are $2^3 = 8$ available voltage vectors with 8 achievable switching states. The available vectors are projected from 3D a - b - c coordinates into 3D α - β - 0 coordinates. Table 2.3 summarizes the voltage vector parameters for a two-level three-leg center-split converter. Figure 2.31 shows the allocation of available output voltage vectors of a two-level three-leg center-split converter.

Comparing Figs. 2.30 and 2.31, it shows that the number of available voltage vectors is equal to the number of available states in a two-level three-leg center-split converter, which is not the case in a two-level three-leg converter. Those two zero

Fig. 2.31 Allocation of available output voltage vectors of a two-level three-leg center-split converter



vectors can be employed for a neutral line current compensation as well as optimizing the switching sequence to reduce switching loss. All the vectors of the three-leg center-split converter have zero components. It is noticed that the amplitude, r , of $\{\vec{V}_1, \vec{V}_2, \vec{V}_3, \vec{V}_4, \vec{V}_5, \vec{V}_6\}$ vectors of the three-leg center-split converter is the same as that of the original three-leg converter, which is the projection of $\{\vec{V}_1, \vec{V}_2, \vec{V}_3, \vec{V}_4, \vec{V}_5, \vec{V}_6\}$ vectors on the α - β plane.

C. A Two-Level Four-Leg Converter and Its Voltage Space Vector

In a two-level four-leg converter, there are $2^4 = 16$ available vectors with 15 achievable switching states. The available voltage vectors are projected from 3D a - b - c coordinates into 3D α - β - 0 coordinates. Table 2.4 summarizes the voltage vector parameters for a two-level four-leg converter. Figure 2.32 shows the output voltage allocation of available vectors of a two-level four-leg converter. The amplitude, r , of $\{\vec{V}_{1p}, \vec{V}_{2p}, \vec{V}_{3p}, \vec{V}_{4p}, \vec{V}_{5p}, \vec{V}_{6p}, \vec{V}_{1n}, \vec{V}_{2n}, \vec{V}_{3n}, \vec{V}_{4n}, \vec{V}_{5n}, \vec{V}_{6n}\}$ vectors of the four-leg converter is the same as that of the original three-leg and the three-leg center-split converters, which are the projections of those vectors on the α - β plane.

D. Comparisons and Discussions

Table 2.5 summarizes the above analysis results. In previous studies, under the center-split capacitor approach, the three-phase converter essentially becomes three single-phase half-bridge converters. Thus, it suffers a lower utilization of dc link

Table 2.4 Two-level four-leg converter voltage vector parameters

	S_f	S_a	S_b	S_c	S_x	S_β	S_0	r	θ	Z_0	ρ	ϕ
$\vec{V}_{1p}$	-1	1	-1	-1	2	0	2	1.633	0°	1.155	2	54.73°
$\vec{V}_{2p}$	-1	1	1	-1	1	2	4	1.633	60°	2.309	2.828	35.27°
$\vec{V}_{3p}$	-1	-1	1	-1	-1	2	2	1.633	120°	1.155	2	54.73°
$\vec{V}_{4p}$	-1	-1	1	1	-2	0	4	1.633	180°	2.309	2.828	35.27°
$\vec{V}_{5p}$	-1	-1	-1	1	-1	-2	2	1.633	240°	1.155	2	54.73°
$\vec{V}_{6p}$	-1	1	-1	1	1	-2	4	1.633	300°	2.309	2.828	35.27°
$\vec{V}_{0p}$	-1	1	1	1	0	0	6	0	*	3.464	3.464	0°
$\vec{V}_{00n}$	-1	-1	-1	-1	0	0	0	0	*	0	0	90°
$\vec{V}_{1n}$	1	1	-1	-1	2	0	-4	1.633	0°	-2.309	2.828	144.73°
$\vec{V}_{2n}$	1	1	1	-1	1	2	-2	1.633	60°	-1.155	2	125.27°
$\vec{V}_{3n}$	1	-1	1	-1	-1	2	-4	1.633	120°	-2.309	2.828	144.73°
$\vec{V}_{4n}$	1	-1	1	1	-2	0	-2	1.633	180°	-1.155	2	125.27°
$\vec{V}_{5n}$	1	-1	-1	1	-1	-2	-4	1.633	240°	-2.309	2.828	144.73°
$\vec{V}_{6n}$	1	1	-1	1	1	-2	-2	1.633	300°	-1.155	2	125.27°
$\vec{V}_{0n}$	1	-1	-1	-1	0	0	-6	0	*	-3.464	-3.464	180°
$\vec{V}_{00p}$	1	1	1	1	0	0	0	0	*	0	0	90°

Fig. 2.32 Allocation of available output voltage vectors of a two-level four-leg converter

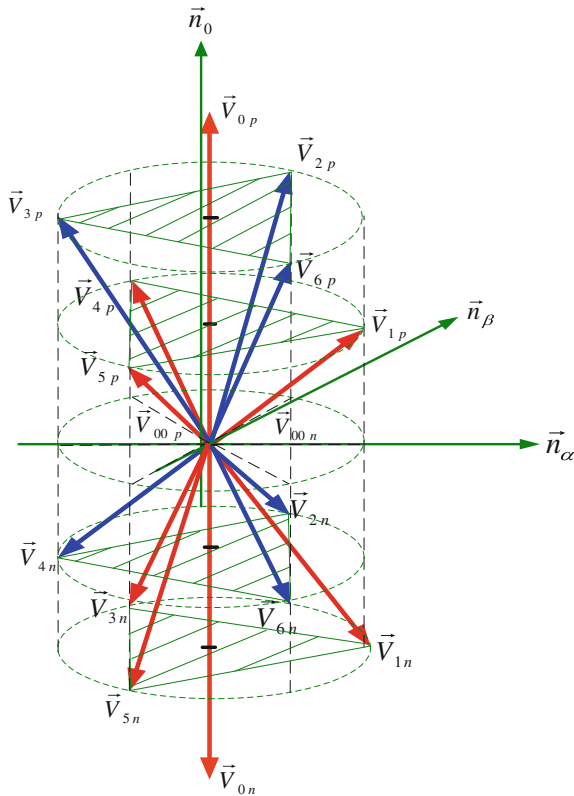


Table 2.5 Comparisons among three basic three-phase converters

	Maximum phase voltage	Voltage capacity on α - β plane	Voltage capacity on zero axis
Three-leg converters	$\frac{4}{3}v_{dc}$	Same	0
Three-leg center-split converters	v_{dc}	Same	Middle
Four-leg converters	$2v_{dc}$	Same	Highest

voltage as compared to the four-leg. Based on the maximum phase voltage that could be performed by the three-phase converters, reaching the above conclusion was straightforward. However, according to Tables 2.2, 2.3 and 2.4, all of the three-phase converters have the same amplitude, r , of the projected non-zero vectors on the α - β plane. It means that when the converters have the same dc linked voltage, they have the same α - β voltage capacity for converter's operation, and the four-leg converter has higher zero-current injection ability as it has a higher Z_0 than that of the three-leg center-split. As a result, a three-leg center split and four-leg inverters have the same voltage capacity on the α - β plane when they have the same dc linked voltage.

Table 2.6 Simulation results

$v_{dc} = v_{dc1} = v_{dc2} = 100 \text{ V}$ $v_{PN} = 200 \text{ V}$	Phase voltage (V)	$r \cdot v_{dc} \text{ (V)}$	$Z_0 \cdot v_{dc} \text{ (V)}$
Three-leg converters [S _a S _b S _c] = [1]	133.3	163.3	1.641×10^{-14}
Three-leg center-split converters [S _a S _b S _c] = [1]	99.99	163.3	-57.73
Four-leg converters [S _a S _b S _c S _f] = [1]	200	163.3	115.4

A simulation test of the three-phase three-leg, the three-leg center-split and the four-leg inverters is performed when $v_{dc} = v_{dc1} = v_{dc2} = 100 \text{ V}$ and $v_{PN} = v_{dc1} + v_{dc2} = 200 \text{ V}$ with balanced loads are connected. The switching functions are given by Table 2.6. It is noticed that the total dc linked voltage is 200 V. Table 2.6 summarizes the simulation results, where r and Z_0 are defined in Tables 2.2, 2.3 and 2.4 respectively. All the simulated values followed the above theoretical results. It shows the validity of the above theoretical conclusions. First, phase voltage is the highest in the four-leg inverter and the three-leg center-split converter has the smallest phase voltage. Second, all the inverters have the same voltage capacity on the α - β plane when they have the same dc linked voltage. Third, the four-leg inverter has the highest zero injection capacity. Fourth, only one transformation can be employed to analyze all three-phase converters.

In left columns of Tables 2.2, 2.3 and 2.4, the switching functions in a - b - c and α - β -0 coordinates are identical to those in the other researches [23–30]. Experimental results were performed in [25, 32] to show the validity of switching function tables. However, without using a 3D voltage model, such as (2.56), (2.59) and (2.63), to describe the output vectors of three-phase converters, it was hard to compare the relationship among the converters. Based on the mathematical and theoretical deduction with only one 3D transformation basis, the converters can be comparable directly.

2.3.3 DC Voltage Unbalance, Variation, Switching Functions and 3D Coordinates

In the three-phase converters, the unbalance in dc-link neutral point voltages, for example, in a three-phase three-leg center-split converter, may affect PWM performance and it may cause a significant distortion in generating a current waveform. In this section, $v_{dc1} \neq v_{dc2}$ is assumed in order to consider the influence of the unbalance dc upper and lower-leg voltage variations on 3D voltage space vectors of the three-phase converters. Switching functions are considered due to the unbalance between dc upper and lower capacitor voltages. V_{dc} is defined as (2.69). The upper-leg and lower-leg switching functions in a, b or c phases can be defined as

(2.70) accordingly, where S_j^N is defined as this new proposed switching function due to dc voltage unbalance and $j = a, b$ or c .

$$V_{dc} = \frac{v_{dc1} + v_{dc2}}{2} \quad (2.69)$$

$$S_j^N = \begin{cases} \frac{v_{dc1}}{V_{dc}}, & \text{upper leg} \\ -\frac{v_{dc2}}{V_{dc}}, & \text{lower leg} \end{cases} \quad (2.70)$$

Conventionally, the switching function can be 1 or -1 , but this proposed switching function can either be smaller or larger than 1 due to the dc voltage deviation. When $v_{dc1} = v_{dc2} = V_{dc}$, $S_j^N = S_j$. However, when $v_{dc1} \neq v_{dc2} \neq V_{dc}$, $S_j^N \neq \pm 1$. Furthermore, S_j^N can be expressed as (2.71). The proposed switching function is found to be affected by the dc voltage unbalance.

$$S_j^N = \begin{cases} \frac{v_{dc1}}{V_{dc}} = \frac{2v_{dc1}}{v_{dc1} + v_{dc2}} = \frac{(v_{dc1} + v_{dc2}) + (v_{dc1} - v_{dc2})}{v_{dc1} + v_{dc2}} = 1 + \Delta S, & \text{upper leg} \\ -\frac{v_{dc2}}{V_{dc}} = \frac{-2v_{dc2}}{v_{dc1} + v_{dc2}} = \frac{-(v_{dc1} + v_{dc2}) + (v_{dc1} - v_{dc2})}{v_{dc1} + v_{dc2}} = -1 + \Delta S, & \text{lower leg} \end{cases} \quad (2.71)$$

where $\Delta S = \frac{v_{dc1} - v_{dc2}}{2V_{dc}}$.

The relationship between the new switching function and the original one can be expressed as:

$$S_j^N = S_j + \Delta S \quad (2.72)$$

Therefore, the output voltage vector of a three-leg inverter can be recalculated by S_a^N , S_b^N and S_c^N due to the dc voltage unbalance. And, V_{dc} can either be larger or smaller due to the dc voltage variation.

2.3.3.1 A Three-Leg Converter

The new instantaneous voltage vector $\vec{v}^N$ can be expressed as (2.73) due to the dc upper-leg and lower-leg voltage unbalance.

$$\vec{v}_{three-leg}^N = V_{dc} \left[\sqrt{\frac{2}{3}} S_\alpha^N \cdot \vec{n}_\alpha + \frac{1}{\sqrt{2}} S_\beta^N \cdot \vec{n}_\beta \right] \quad (2.73)$$

where $S_\alpha^N = S_a^N - \frac{1}{2} S_b^N - \frac{1}{2} S_c^N$

$$S_\beta^N = S_b^N - S_c^N$$

Substituting (2.72) into (2.73), no influence is found on the space vector voltage of a three-leg converter, that is $\vec{v}_{three-leg} = \vec{v}_{three-leg}^N$. However, the instantaneous voltage bounded area can either be larger or smaller due to the dc voltage variation, V_{dc} .

2.3.3.2 A Three-Leg Center-Split Converter

The instantaneous voltage vector $\vec{v}^N$ of a three-leg center-split converter in 3D can be expressed as (2.74).

$$\vec{v}_{center-split}^N = V_{dc} \left[\sqrt{\frac{2}{3}} S_{\alpha}^N \cdot \vec{n}_{\alpha} + \frac{1}{\sqrt{2}} S_{\beta}^N \cdot \vec{n}_{\beta} + \frac{1}{\sqrt{3}} S_0^N \cdot \vec{n}_0 \right] \quad (2.74)$$

where $S_{\alpha}^N = S_a^N - \frac{1}{2}S_b^N - \frac{1}{2}S_c^N$

$$S_{\beta}^N = S_b^N - S_c^N$$

$$S_0^N = S_a^N + S_b^N + S_c^N$$

Substituting (2.72) into (2.74), only the zero switching function is found to be affected by the dc voltage unbalance.

$$S_{\alpha}^N = S_{\alpha} \quad (2.75)$$

$$S_{\beta}^N = S_{\beta} \quad (2.76)$$

$$S_0^N = S_0 + 3\Delta S \quad (2.77)$$

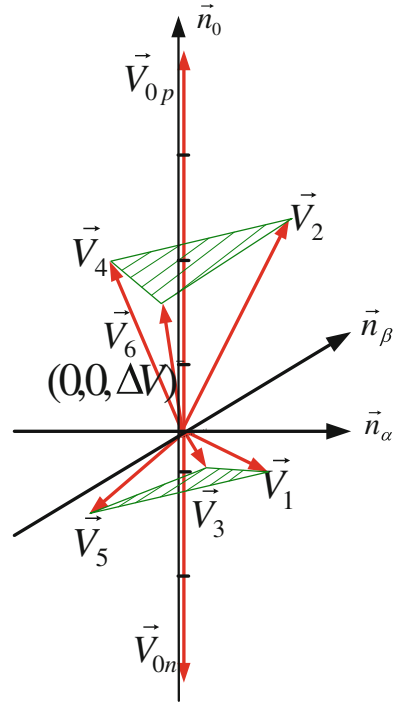
The instantaneous voltage vector $\vec{v}_{center-split}^N$ can be expressed as (2.68).

$$\vec{v}_{center-split}^N = \vec{v}_{center-split} + \vec{v}_0^N \quad (2.78)$$

where $\vec{v}_0^N = \sqrt{3}V_{dc}\Delta S \cdot \vec{n}_0$.

According to (2.78), the voltage vectors may vary upward or downward along $\vec{n}_0$ axis due to the unbalance of dc voltage variation as shown in Fig. 2.33. However, the instantaneous voltage bounded volume can either be larger or smaller due to the dc voltage variation, V_{dc} .

Fig. 2.33 2-Level 3D vector allocation due to dc voltage unbalance



2.3.3.3 A Four-Leg Converter

The instantaneous voltage vector $\vec{v}^N$ of a four-leg converter in 3D can be expressed as (2.79).

$$\vec{v}_{four-leg}^N = V_{dc} \left[\sqrt{\frac{2}{3}} S_{\alpha}^N \cdot \vec{n}_{\alpha} + \frac{1}{\sqrt{2}} S_{\beta}^N \cdot \vec{n}_{\beta} + \frac{1}{\sqrt{3}} S_0^N \cdot \vec{n}_0 \right] \quad (2.79)$$

where $S_{\alpha}^N = S_a^N - \frac{1}{2} S_b^N - \frac{1}{2} S_c^N$

$$S_{\beta}^N = S_b^N - S_c^N$$

$$S_0^N = S_a^N + S_b^N + S_c^N - 3S_f^N$$

Substituting (2.72) into (2.79), it is found that there is no influence on the space vector voltages of a four-leg converter due to dc voltage unbalance, which is $\vec{v}_{four-leg} = \vec{v}_{four-leg}^N$. However, the instantaneous voltage bounded volume can either be larger or smaller due to the dc voltage variation, V_{dc} .

From the above calculations, only the space voltage vectors of the three-leg center-split converter are affected due to the dc voltage unbalance. Reversely, the dc unbalance voltages can be controlled by shifting upward and downward of the reference voltage vector accordingly. Details and experimental results for a three-leg center-split converter can be found in [32]. However, the instantaneous voltage bounded volume can either be larger or smaller due to the dc voltage variation, V_{dc} .

2.4 Summary

Three fundamental three-phase converter circuit topologies are investigated and compared under only one coordinate transformation and a 3D model proposed in this chapter. Several conclusions are drawn:

- (1) Phase voltage is the highest in a four-leg inverter and a three-leg center-split converter has the smallest phase voltage.
- (2) All the inverters have the same voltage capacity on α - β plane when they have the same dc-linked voltage.
- (3) A four-leg inverter has the highest zero injection capacity.
- (4) Switching functions can be defined either smaller or larger than 1 due to the dc voltage unbalance.
- (5) Space voltage vectors of a three-leg center-split converter are affected due to the dc voltage unbalance by shifting upward and downward of the reference voltage vector along zero-axis accordingly.
- (6) Space voltage vectors bounded volume can either be larger or smaller due to dc voltage variation.

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