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Novel control scheme for 3-phase PWM current-source converters under unbalanced source voltage conditions*

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Abstract: Under unbalanced source voltage supply, considerable output second harmonics and input low-order harmonics in 3-phase PWM current-source converters (PWM-CSC) are generated. This paper proposes a new deadbeat controller based on compensation for unbalanced source voltage and current. With the proposed scheme, the second harmonics of the output current are eliminated and low-order harmonics of the source current are reduced effectively. Simulation and experimental results confirmed the feasibility of the proposed method.

Key words: PWM current-source converters (PWM-CSC), Unbalanced source voltage conditions, Deadbeat control
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INTRODUCTION

With the development of super conductor, the PWM-CSC is being increasingly studied in recent years owing to its advanced features including sinusoidal source current at unity power factor and high quality dc output current (Zhang and Ooi, 1993; Kazerani and Ye, 2002; Vincenti and Jin, 1994; Wang, 2000). However, under unbalanced source voltage, considerable output second harmonics and input low-order harmonics are produced due to negative sequence voltage (Vincenti and Jin, 1994; Wang, 2000), resulting in unbalanced 3-phase source current, large losses, bad work performance, even failure. Therefore, some compensation methods for such undesirable characteristics are needed.

Several schemes that counter or eliminate the adverse effects of input and output harmonics had been proposed (Stankovic and Lipo, 2000; Song *et al*., 2003; Song and Nam, 1999; Enjeti and Ziogas, 1990; Vincenti and Jin, 1994). In (Stankovic and Lipo, 2000; Song *et al*., 2003; Song and Nam, 1999), control schemes, for eliminating harmonics of the PWM voltage-source converters under unbalanced source voltage conditions, were derived. However, these methods do not work well to PWM-CSC.

In (Enjeti and Ziogas, 1990), the output second harmonics of the PWM-CSC were shown to be the cross product terms between the positive sequence components of the converter transfer function and the negative sequence components of the source voltage. From this analysis, an algorithm to cancel the second harmonic component was presented in (Vincenti and Jin, 1994), which was done by suitably regulating the converter switch gating signals. Unfortunately, with this method, the source current of the converter cannot maintain good sinusoidal waveform performance. Additionally, other methods including conventional PI controller, had not been considered in the PWM-CSC under unbalanced voltage.

In this paper, deadbeat controller with unbalanced source voltage and current compensation is employed in PWM-CSC. Accordingly, the current is exactly equal to its reference value at the next sampling instant and the dynamic response of the system

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is improved greatly. Compared to the proposed compensation deadbeat controller without compensation deadbeat controller, consequently, the proposed scheme yields better control performance.

ANALYSIS OF UNBALANCED SOURCE VOLTAGE

Under unbalanced source voltage conditions, it is necessary to calculate the positive and negative source voltages. Assuming unbalanced source voltage in *abc* coordinates (e_a, e_b, e_c) can be expressed as:

$$\begin{cases} e_a = k_a E_\mathrm{m}\cos(\omega t+\theta_a), \\ e_b = k_b E_\mathrm{m}\cos(\omega t+\theta_b-120°), \\ e_c = k_c E_\mathrm{m}\cos(\omega t+\theta_c+120°), \end{cases} \quad (1)$$

where k_j,θ_j are coefficients ($0\le k_j\le 1$, $-\pi\le\theta_j\le\pi$); $j=a,b,c$; E_m is the amplitude of the source voltage; ω is angular frequency.

Without considering the effect of zero sequence voltage, the positive and negative *dq*-components of the source voltage in synchronous frames (e_d^p,e_d^n,e_q^p,e_q^n) are obtained from the measured *abc* coordinate source voltage (Zhang and Zhang, 2003)

$$\begin{cases} e_d^\mathrm{p} = \dfrac{E_\mathrm{m}}{3}[k_a\cos\theta_a + k_b\cos\theta_b + k_c\cos\theta_c], \\ e_d^\mathrm{n} = \dfrac{E_\mathrm{m}}{3}[k_a\cos\theta_a + k_b\cos(\theta_b+120°) + k_c\cos(\theta_c-120°)], \\ e_q^\mathrm{p} = \dfrac{E_\mathrm{m}}{3}[k_a\sin\theta_a + k_b\sin\theta_b + k_c\sin\theta_c], \\ e_q^\mathrm{n} = \dfrac{-E_\mathrm{m}}{3}[k_a\sin\theta_a + k_b\sin(\theta_b-120°) + k_c\sin(\theta_c+120°)]. \end{cases} \quad (2)$$

The positive and negative sequence components of Eq.(2) are converted into the variables in the original synchronous frame model. Consequently, the new original synchronous frame voltages can be regulated pertinently according to input voltage as follows:

$$\begin{bmatrix} e_d \\ e_q \end{bmatrix} = \begin{bmatrix} e_d^\mathrm{p} \\ e_q^\mathrm{p} \end{bmatrix} + \boldsymbol{R}(-2\omega t)\begin{bmatrix} e_d^\mathrm{n} \\ e_q^\mathrm{n} \end{bmatrix}, \quad (3)$$

where $\boldsymbol{R}(-2\omega t) = \begin{bmatrix} \cos 2\omega t & \sin 2\omega t \\ -\sin 2\omega t & \cos 2\omega t \end{bmatrix}$.

PROPOSED CONTROL METHOD IN UNBALANCED SOURCE VOLTAGE

The 3-phase PWM-CSC scheme is shown in Fig.1. In case of balanced source voltage, its dynamic equation in the synchronous frame can be described as (Wang, 2000):

$$\begin{cases} L\dfrac{\mathrm{d}}{\mathrm{d}t}i_d + R_\mathrm{s}\cdot i_d - \omega L\cdot i_q + v_{cd} = e_d \\ L\dfrac{\mathrm{d}}{\mathrm{d}t}i_q + R_\mathrm{s}\cdot i_q + \omega L\cdot i_d + v_{cq} = e_q \end{cases}, \quad (4)$$

$$\begin{cases} i_d = C\dfrac{\mathrm{d}}{\mathrm{d}t}v_{cd} - \omega C\cdot v_{cq} + i_{pd} \\ i_q = C\dfrac{\mathrm{d}}{\mathrm{d}t}v_{cq} + \omega C\cdot v_{cd} + i_{pq} \end{cases}, \quad (5)$$

where e_d=0, $e_q = -\sqrt{2/3}E_\mathrm{m}$, E_m is the amplitude of the source voltage.

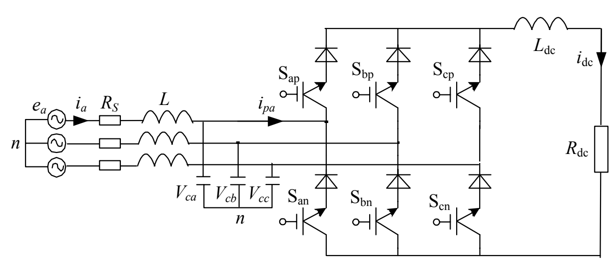


Fig.1 Scheme of the 3-phase PWM-CSC

In the PWM-CSC, the source currents are controlled by input currents of the converter. From Eqs.(4) and (5), the input current can be expressed as:

$$\begin{cases} i_{pd} = LC\dfrac{\mathrm{d}^2}{\mathrm{d}t^2}i_d - 2\omega LC\dfrac{\mathrm{d}}{\mathrm{d}t}i_q + RC\dfrac{\mathrm{d}}{\mathrm{d}t}i_d \\ \qquad -\omega RCi_q + (1-\omega^2 LC)\cdot i_d + \omega Ce_q, \\ i_{pq} = LC\dfrac{\mathrm{d}^2}{\mathrm{d}t^2}i_q + 2\omega LC\dfrac{\mathrm{d}}{\mathrm{d}t}i_d + RC\dfrac{\mathrm{d}}{\mathrm{d}t}i_q \\ \qquad +\omega RCi_d + (1-\omega^2 LC)\cdot i_q - \omega Ce_d. \end{cases} \quad (6)$$

Based on the assumption that the switching period T_s is set to be a small value, Eq.(6) can be written approximately as:

$$\begin{cases} i_{pd}(k)=\dfrac{LC}{T_s}\left(\dfrac{i_d(k)-i_d(k-1)}{T_s}-\dfrac{i_d(k-1)-i_d(k-2)}{T_s}\right) \\ \quad -2\omega LC\cdot\dfrac{i_q(k)-i_q(k-1)}{T_s}+RC\cdot\dfrac{i_d(k)-i_d(k-1)}{T_s} \\ \quad -\omega RC\cdot i_q(k)+(1-\omega^2 LC)\cdot i_d(k)+\omega Ce_q(k), \\ i_{pq}(k)=\dfrac{LC}{T_s}\left(\dfrac{i_q(k)-i_q(k-1)}{T_s}-\dfrac{i_q(k-1)-i_q(k-2)}{T_s}\right) \\ \quad +2\omega LC\cdot\dfrac{i_d(k)-i_d(k-1)}{T_s}+RC\cdot\dfrac{i_q(k)-i_q(k-1)}{T_s} \\ \quad +\omega RC\cdot i_d(k)+(1-\omega^2 LC)\cdot i_q(k)-\omega Ce_d(k). \end{cases} \tag{7}$$

Eq.(7) is the deadbeat controller that forces the *k*th instant input current of PWM-CSC to be exactly equal to the reference value at the (*k*+1)th sample instant.

Generally, in Eq.(7), the values of synchronous frame voltage (e_d, e_q) and current (i_d, i_q) are obtained by park transformation (park transformation denotes transformation from *abc* coordinate to *dq* coordinate). If this method is adapted directly to unbalanced source voltage PWM-CSC, the result undesirable input and output performance. In this paper, unbalanced source voltage and current are compensated simultaneously, which can be described in detail as follows.

Considering only the first harmonic component of the source voltage and current, the input complex power of the converter is given in Eq.(8) (Song and Nam, 1999)

$$\tilde{S}=p+\mathrm{j}q=(\mathrm{e}^{\mathrm{j}\omega t}\boldsymbol{E}_{dq}^{\mathrm{p}}+\mathrm{e}^{-\mathrm{j}\omega t}\boldsymbol{E}_{dq}^{\mathrm{n}})\left(\overline{\mathrm{e}^{\mathrm{j}\omega t}\boldsymbol{I}_{dq}^{\mathrm{p}}+\mathrm{e}^{-\mathrm{j}\omega t}\boldsymbol{I}_{dq}^{\mathrm{n}}}\right), \tag{8}$$

where, p, q is active power and reactive power respectively, $\left(\overline{\mathrm{e}^{\mathrm{j}\omega t}\boldsymbol{I}_{dq}^{\mathrm{p}}+\mathrm{e}^{-\mathrm{j}\omega t}\boldsymbol{I}_{dq}^{\mathrm{n}}}\right)$ is conjugate complex of $\left(\mathrm{e}^{\mathrm{j}\omega t}\boldsymbol{I}_{dq}^{\mathrm{p}}+\mathrm{e}^{-\mathrm{j}\omega t}\boldsymbol{I}_{dq}^{\mathrm{n}}\right)$.

In Eq.(8), p and q can be written as:

$$\begin{cases} p=p_0+p_{\mathrm{c}2}\cos(2\omega t)+p_{\mathrm{s}2}\sin(2\omega t) \\ q=q_0+q_{\mathrm{c}2}\cos(2\omega t)+q_{\mathrm{s}2}\sin(2\omega t) \end{cases}.$$

Thus, six real and imaginary terms of the complex power are obtained in Eq.(9):

$$\begin{cases} p_0=\frac{3}{2}\left(e_d^{\mathrm{p}}i_d^{\mathrm{p}}+e_q^{\mathrm{p}}i_q^{\mathrm{p}}+e_d^{\mathrm{n}}i_d^{\mathrm{n}}+e_q^{\mathrm{n}}i_q^{\mathrm{n}}\right), \\ p_{\mathrm{c}2}=\frac{3}{2}\left(e_d^{\mathrm{p}}i_d^{\mathrm{n}}+e_q^{\mathrm{p}}i_q^{\mathrm{n}}+e_d^{\mathrm{n}}i_d^{\mathrm{p}}+e_q^{\mathrm{n}}i_q^{\mathrm{p}}\right), \\ p_{\mathrm{s}2}=\frac{3}{2}\left(e_q^{\mathrm{n}}i_d^{\mathrm{p}}-e_d^{\mathrm{n}}i_q^{\mathrm{p}}-e_q^{\mathrm{p}}i_d^{\mathrm{n}}+e_d^{\mathrm{p}}i_q^{\mathrm{n}}\right), \\ q_0=\frac{3}{2}\left(e_q^{\mathrm{p}}i_d^{\mathrm{p}}-e_d^{\mathrm{p}}i_q^{\mathrm{p}}+e_q^{\mathrm{n}}i_d^{\mathrm{n}}-e_d^{\mathrm{n}}i_q^{\mathrm{n}}\right), \\ q_{\mathrm{c}2}=\frac{3}{2}\left(e_q^{\mathrm{p}}i_d^{\mathrm{n}}-e_d^{\mathrm{p}}i_q^{\mathrm{n}}+e_q^{\mathrm{n}}i_d^{\mathrm{p}}-e_d^{\mathrm{n}}i_q^{\mathrm{p}}\right), \\ q_{\mathrm{s}2}=\frac{3}{2}\left(e_d^{\mathrm{p}}i_d^{\mathrm{n}}+e_q^{\mathrm{p}}i_q^{\mathrm{n}}-e_d^{\mathrm{n}}i_d^{\mathrm{p}}-e_q^{\mathrm{n}}i_q^{\mathrm{p}}\right), \end{cases} \tag{9}$$

where, p_0 and q_0 are average active and reactive power, $p_{\mathrm{c}2}$ and $p_{\mathrm{s}2}$ are the second harmonics active power cosine and sinusoidal amplitude, $q_{\mathrm{c}2}$ and $q_{\mathrm{s}2}$ are the second harmonic reactive power cosine and sinusoidal amplitude respectively.

In order to cancel second harmonics of the output dc current, $p_{\mathrm{c}2}$ and $p_{\mathrm{s}2}$ are set to zero. And the average reactive power q_0 is regulated to zero, which will lead to unity power factor. These conditions are incorporated into Eq.(9), the positive and negative *dq* components of the currents in the synchronous frame are regulated to the reference values as Eq.(10).

$$\begin{cases} \left(i_d^{\mathrm{p}}\right)^*=\dfrac{2e_d^{\mathrm{p}}p_0^*}{3\left[\left(e_d^{\mathrm{p}}\right)^2+\left(e_q^{\mathrm{p}}\right)^2\right]-3\left[\left(e_d^{\mathrm{n}}\right)^2+\left(e_q^{\mathrm{n}}\right)^2\right]}, \\ \left(i_q^{\mathrm{p}}\right)^*=\dfrac{2e_q^{\mathrm{p}}p_0^*}{3\left[\left(e_d^{\mathrm{p}}\right)^2+\left(e_q^{\mathrm{p}}\right)^2\right]-3\left[\left(e_d^{\mathrm{n}}\right)^2+\left(e_q^{\mathrm{n}}\right)^2\right]}, \\ \left(i_d^{\mathrm{n}}\right)^*=\dfrac{-2e_d^{\mathrm{n}}p_0^*}{3\left[\left(e_d^{\mathrm{p}}\right)^2+\left(e_q^{\mathrm{p}}\right)^2\right]-3\left[\left(e_d^{\mathrm{n}}\right)^2+\left(e_q^{\mathrm{n}}\right)^2\right]}, \\ \left(i_q^{\mathrm{n}}\right)^*=\dfrac{-2e_q^{\mathrm{n}}p_0^*}{3\left[\left(e_d^{\mathrm{p}}\right)^2+\left(e_q^{\mathrm{p}}\right)^2\right]-3\left[\left(e_d^{\mathrm{n}}\right)^2+\left(e_q^{\mathrm{n}}\right)^2\right]}. \end{cases} \tag{10}$$

However, when the current control is performed with its references of Eq.(10), it is required to analyze the feedback current with the positive and negative sequence component. To simplify, the source current references are determined in the same format as that of Eq.(4) and are the corresponding values of source current in the original synchronous frame.

In this paper, the PI current controller is used in the dc-link. The reference of average active power p_0^* results from the product of dc voltage and current.

The diagram of the proposed control block converter system is shown in Fig.2.

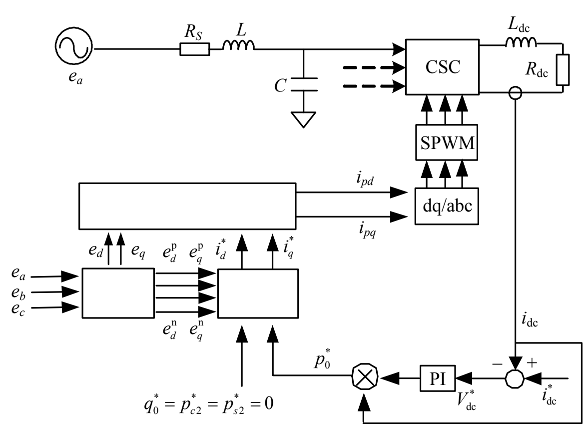


Fig.2 The proposed control block diagram

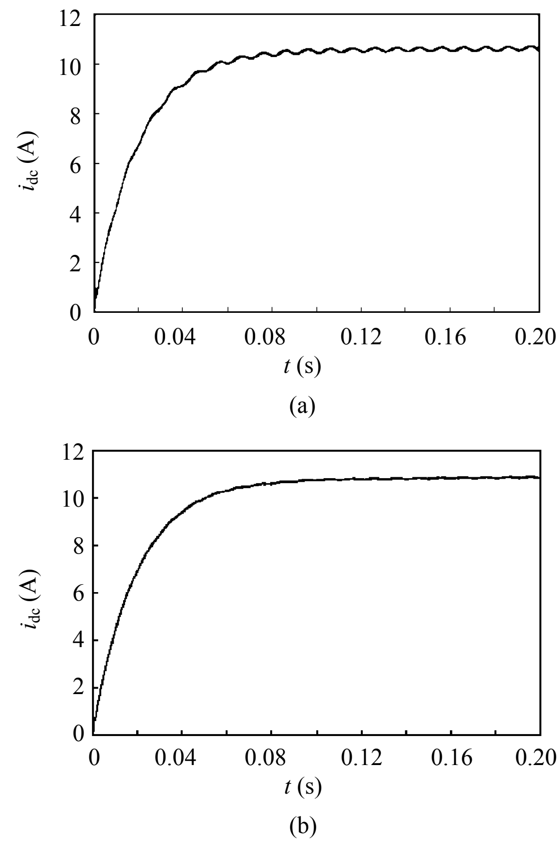


Fig.3 Output dc current (a) without compensation; (b) with compensation

SIMULATION AND EXPERIMENTAL RESULTS

The parameters for simulation and experiment. are as follows: 3-phase unbalanced source voltage are e_a=110sinωt V, e_b=100sin(ωt−120°) V, e_c=90sin(ωt+120°) V, ω=100π rad/s, R_s=0.5 Ω, L=0.8 mH, C=56 μF, L_{dc}=200 mH, equivalent load R_{dc}=15 Ω.

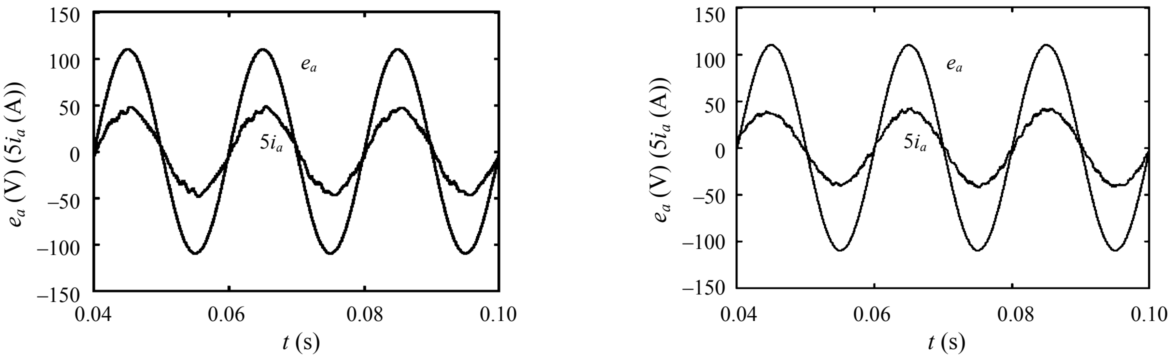
The proposed control scheme was simulated using MATLAB. The complete control was implemented in a DSP processor TMS320F2407 with a sampling rate and a switching frequency of 5 kHz.

Fig.3 and Fig.4 illustrate simulation results for the without unbalance compensation deadbeat-controller and the proposed compensation controller. Obviously, there is reduction of the output dc current ripple in Fig.3b compared with that of Fig.3a. Fig.4 shows that source current waveforms have little distortion and better power factor with the proposed scheme.

The experimental results for the deadbeat controller with and without compensation are shown in Fig.5 and Fig.6. The experimental results are accord with the simulation results. The detailed frequency spectra are shown in Fig.7 and Fig.8 (where, Amplitude ratio=harmonic amplitude/dc component amplitude), which verify that the proposed compensation controller leads to satisfactory elimination of second harmonics of the dc output current and considerable reduction in the harmonic amplitude of the source current.

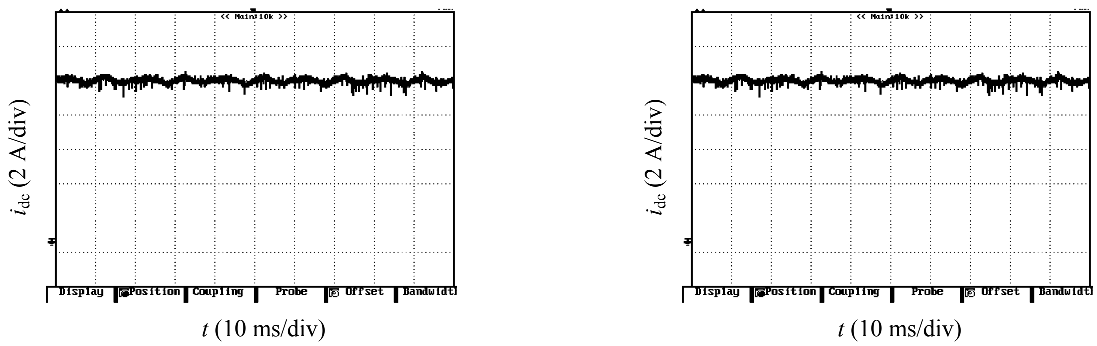
CONCLUSION

In this paper, an unbalanced source voltage and current compensation deadbeat controller is presented for 3-phase PWM-CSC. The proposed method can reduce harmonics and reactive power components of the system, which result in sinusoidal and unity power factor source current and regulate dc output current as



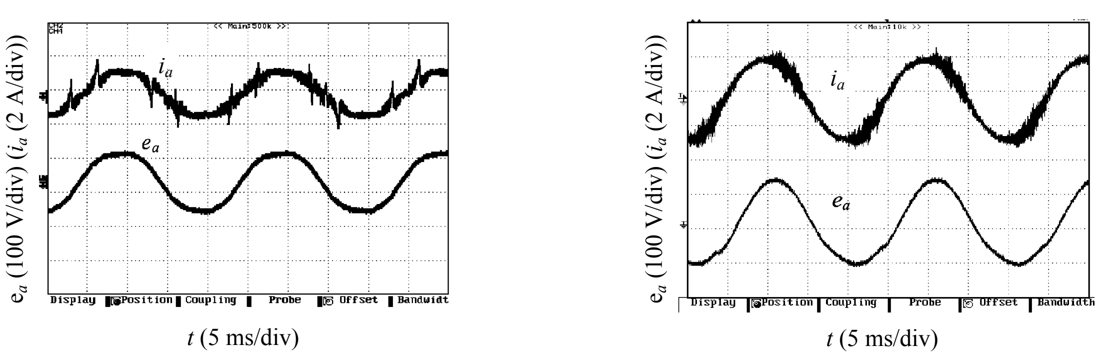
(a) (b)

Fig.4 Steady A-phase source voltage and current (a) without compensation; (b) with compensation



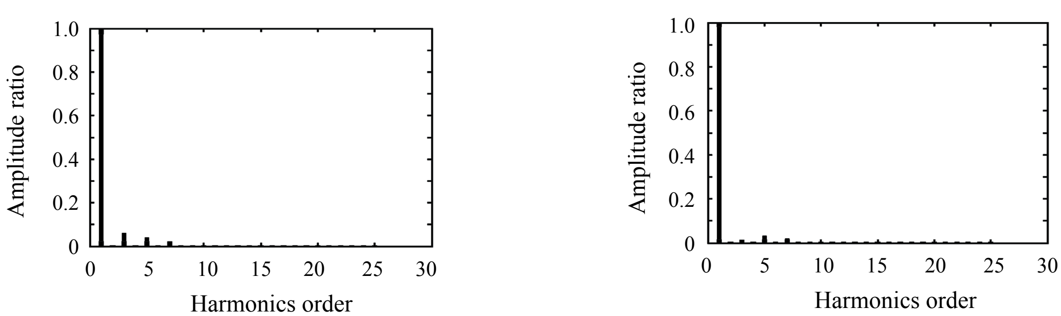
(a) (b)

Fig.5 Steady output dc current (a) without compensation; (b) with compensation



(a) (b)

Fig.6 Steady A-phase source voltage and current (a) without compensation; (b) with compensation



(a) (b)

Fig.7 Dc current frequency spectrum (a) without compensation; (b) with compensation

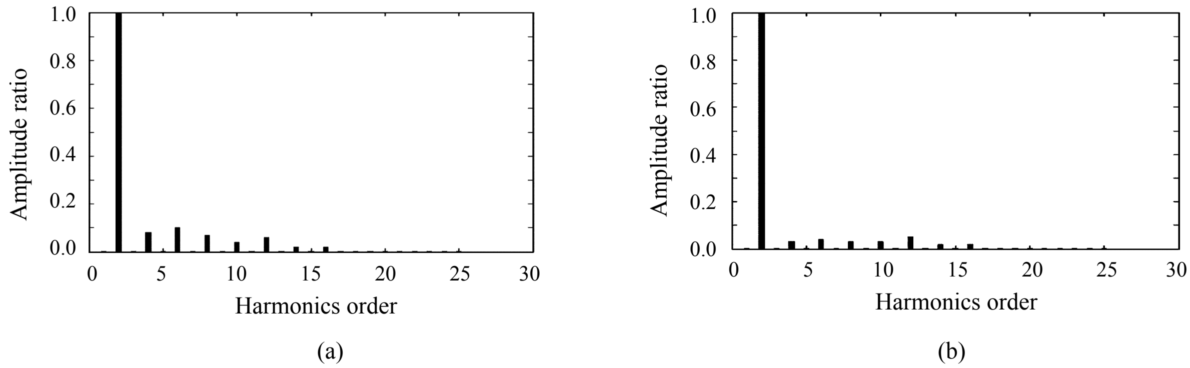


Fig.8 A-phase source current frequency spectra (a) without compensation; (b) with compensation

a constant under unbalanced source voltage conditions. This method does not require current sensors and is easy to employ with digital control technique. Moreover, the proposed scheme can be extended for other converter applications.

References

Enjeti, P.N., Ziogas, P.D., 1990. Analysis of a static power converter under unbalanced: a novel approach. *IEEE Trans. on Industrial Electronics*, **37**(1):91-93. [doi:10.1109/41.45849]

Vincenti, D., Jin, H., 1994. A three-phase regulation PWM rectifier with on-line feedforward input unbalanced correction. *IEEE Trans. on Industrial Electronics*, **41**(5):526-532. [doi:10.1109/41.315271]

Kazerani, M., Ye, Y., 2002. Comparative evaluation of three-phase PWM voltage- and current-source converter topologies in FACTS applications. *IEEE Power Engineering Society Summer Meeting PESS-02*, **1**:473-479. [doi:10.1109/ PESS.2002.1043279]

Song, H.S., Nam, K., 1999. Dual current control scheme for PWM converter under unbalanced source voltage conditions. *IEEE Trans. on Industrial Electronics*, **46**(5):953-959. [doi:10.1109/41.793344]

Song, H.S., Joo, I.W., Nam, K., 2003. Source voltage sensorless estimation scheme for PWM rectifiers under unbalanced conditions. *IEEE Trans. on Industrial Electronics*, **50**(6):1238-1245. [doi:10.1109/TIE.2003.819685]

Stankovic, A.V., Lipo, T.A., 2000. A novel control method for input output harmonic elimination of the PWM boost type rectifier under unbalanced operating conditions. *Conf. Rec. APEC*, **1**:413-419.

Wang, C.Y., 2000. Multi-modular Converter with Phase-shifted SPWM Technique and Its Application to Active Power Filter. Ph.D Thesis, Zhejiang University, Hangzhou (in Chinese).

Zhang, Z.C., Ooi, B.T., 1993. Multimodular current-source SPWM converters for superconducting a magnetic energy storage system. *IEEE Trans. on Power Electronics*, **8**(3): 250-255. [doi:10.1109/63.233280]

Zhang, C.W., Zhang, X., 2003. PWM Rectifiers and Control. China Machine Press, Beijing (in Chinese).