Effects of Voltage Unbalance and Harmonics on Three-Phase PWM Ac-to-Dc Converters with Instantaneous Power Feedback

Non-member Wen-Inne Tsai (National Cheng Kung University)
 Non-member York-Yih Sun (National Cheng Kung University)
 Non-member Ming-Tsung Tsai (Nan-Tai Junior College of Industrial and Commercial Management)

A PWM three-phase ac-to-dc converter with unity input power factor can be approached by setting the required active power and setting reactive power zero in the instantaneous reactive power theory. The instantaneous reactive power theory is proposed by Akagi and is usually used to design reactive power and harmonic compensators. In the power distribution system, voltage unbalance and harmonics always exist. The effects of the harmonics and voltage unbalance are presented and discussed in this paper. Moreover, a low pass filter is suggested to add into the coordinate transformation in the instantaneous reactive power theory to reduce the effects of harmonics and unbalance in the power source. Finally, the theoretical results are verified both by simulated and experimental results in a laboratory prototype system.

Keywords : Instantaneous reactive power, power factor, voltage unbalance

1. Introduction

Traditionally, conversion of ac line voltages from utilities has been done by a combination of a phase-controlled rectifier and a bulk capacitor. Such a conversion approach leads to pulsed currents being drawn from the ac distribution networks. These currents contain an amount of harmonics, resulting in serious problems in the electric power system⁽¹⁾.

As a method of overcoming these problems, pulse-width modulated ac-to-dc voltage source converters have been proposed. They have the merits of a nearly sinusoidal input current, a good power factor, and a regeneration ability. In the past, various control strategies have been proposed^{(2)~(6)}, and some of them have been put into practice. Fig. 1 shows the main circuit of the three-phase PWM ac-to-dc converter, which consists of a three-phase voltage source converter, three input inductors and a smoothing capacitor in the dc side.

A control strategy which is based on the instantaneous reactive power theory and is suitable for three-phase PWM ac-to-dc converter is studied in this paper. The instantaneous reactive power theory, proposed by Akagi, is usually used to design the reactive power and harmonics compensators⁽⁷⁾⁽⁸⁾. The basic idea of this

control strategy is that if the power source only supplies the active power to the converter, then this converter will have a unity input power factor. Therefore, the current references of the three-phase ac-to-dc converter can be calculated, according to the instantaneous reactive power theory and the power demand of the converter. These current references are used to control the input inductor currents by three independent current regulators.

In the three-phase ac distribution network, voltage unbalance and harmonics typically exist. The effects of voltage unbalance and harmonics on the three-phase PWM ac-to-dc converter based-on the instantaneous reactive power theory are investigated. Moreover, a modified current reference calculator (MCRC) is proposed to reduce the effects of the non-ideal power source. To confirm the effectiveness of the proposed modified current reference calculator and the analytical results shown in this paper, some simulation and experimental results are presented. A test converter is made for the experiment.

2. System configuration

Fig. 2 shows the simplified functional block diagram of the proposed system. It mainly consists of a three-

Fig. 1. Main circuit of the three-phase PWM ac-to-dc converter.

Fig. 2. Simplified functional block diagram of the three-phase PWM ac-to-dc converter with instantaneous power feedback.

phase voltage source converter (VSC3) whose main circuit is shown in Fig. 1, a notch filter and a noise filter, a carrier generator, a pulse width modulator (PWM), a voltage regulator (AVR), a current reference calculator (CRC) and three independent current regulators (ACR). The function of each control circuit shown in Fig. 2 will be described in more detail later on.

2.1 Instantaneous Reactive Power Theory

The instantaneous reactive power theory is briefly reviewed as follows. According to the instantaneous reactive power theory, the instantaneous real (active) power P_L and the instantaneous reactive power q_L on the load side can be defined as follows. First, transformation of the instantaneous phase voltages (e_R , e_S , and e_T) of the power source and the line currents (i_R , i_S , and i_T) into the d - q orthogonal coordinates (e_d and e_q , i_d and i_q) gives the following expressions:

$$\begin{bmatrix} e_d \\ e_q \end{bmatrix} = \sqrt{\frac{2}{3}} C \begin{bmatrix} e_R \\ e_S \\ e_T \end{bmatrix} \quad \dots\dots\dots (1)$$

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \sqrt{\frac{2}{3}} C \begin{bmatrix} i_R \\ i_S \\ i_T \end{bmatrix} \quad \dots\dots\dots (2)$$

where C is the coordinate transformation matrix and is given by

$$C = \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix}.$$

Then, the instantaneous real power p_L and the instantaneous reactive power q_L are defined by

$$\begin{bmatrix} p_L \\ q_L \end{bmatrix} = \begin{bmatrix} e_d & e_q \\ -e_q & e_d \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad \dots\dots\dots (3)$$

If the three-phase PWM ac-to-dc converter only draws the real power from the power source, the converter will have a unity input power factor. The current reference calculator is used to compute the desired input currents so as to achieve the operation condition.

2.2 Current Reference Calculator

In the current reference calculator (CRC), the three input phase voltages which are sensed by three star connecting potential transformers are converted into the d - q coordinates (denoted by v_d , v_q) by the coordinate transformation matrix:

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = C \begin{bmatrix} k_{pt} e_R \\ k_{pt} e_S \\ k_{pt} e_T \end{bmatrix} \quad \dots\dots\dots (4)$$

where k_{pt} is the turn ratio of the potential transformer. Note that the scaling factor $\sqrt{2/3}$ is not used for ease of implementation because this will not affect the calculation in the CRC. Therefore, the current references can be given by the following expression from the instantaneous reactive power theory.

$$\begin{bmatrix} i_{Rref} \\ i_{Sref} \\ i_{Tref} \end{bmatrix} = \frac{1}{k_m} C^T \begin{bmatrix} v_d & v_q \\ -v_q & v_d \end{bmatrix}^{-1} \begin{bmatrix} p \\ 0 \end{bmatrix} \\ = C^T \frac{p}{V_{dq}} \begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad \dots\dots\dots (5)$$

where C^T denotes the transpose matrix of C , V_{dq} denotes $k_m(v_d^2 + v_q^2)$ and p denotes the required power of the three-phase ac-to-dc converter which is the sum of the load power in the dc side and the power loss of the converter. The operations of multiplication and division in the CRC is performed by four analog multipliers and one analog divider, in which the gain of the multiplier is k_m ($=0.1$) and the gain of the divider is $1/k_m$. The control circuit picks up the load power (p_{load}) by multiplying the load current (i_{load}) by a scaling factor k_s and the power loss of the converter is represented by the output of the voltage regulator (p_{avr}) as shown in Fig. 2. The input inductor currents can nearly track the current references by properly designing the current regulators (ACR). Therefore, the three-phase ac-to-dc converter has a unity input power factor.

2.3 Current Regulator

From the main circuit of the three-phase PWM ac-to-dc converter shown in Fig. 1, the dynamic equations of the input inductors can be expressed as

$$e_R = Ri_R + L \frac{di_R}{dt} + d_1^* \frac{v_{dc}}{2} + v_{no} \quad (6)$$

$$e_S = Ri_S + L \frac{di_S}{dt} + d_2^* \frac{v_{dc}}{2} + v_{no} \quad (7)$$

$$e_T = Ri_T + L \frac{di_T}{dt} + d_3^* \frac{v_{dc}}{2} + v_{no} \quad (8)$$

where

$$d_1^* = \begin{cases} 1 & \text{if } Q_1 \text{ or } D_1 \text{ ON, denoted SW 1 ON} \\ -1 & \text{if } Q_4 \text{ or } D_4 \text{ ON, denoted SW 1 OFF} \end{cases}$$

$$d_2^* = \begin{cases} 1 & \text{if } Q_2 \text{ or } D_2 \text{ ON, denoted SW 2 ON} \\ -1 & \text{if } Q_5 \text{ or } D_5 \text{ ON, denoted SW 2 OFF} \end{cases}$$

$$d_3^* = \begin{cases} 1 & \text{if } Q_3 \text{ or } D_3 \text{ ON, denoted SW 3 ON} \\ -1 & \text{if } Q_6 \text{ or } D_6 \text{ ON, denoted SW 3 OFF} \end{cases}$$

For a three-phase system without a neutral line, then

$$i_R + i_S + i_T = 0 \quad (9)$$

If the ac supply is a balanced source,

$$e_R + e_S + e_T = 0 \quad (10)$$

Thus, the voltage v_{no} can be obtained by adding Eqs. (6)~(8) together:

$$v_{no} = -\frac{1}{6} v_{dc} \sum_{k=1}^3 d_k^* \quad (11)$$

The ON or OFF states of switches in Fig. 1 are determined by the pulse width modulator (PWM). If the modulating signal (i_{rm} , i_{sm} , and i_{tm}) is smaller than the carrier, then SW_k is ON, otherwise OFF.

From equation (11), v_{no} is a coupled item and it will be considered as a disturbance in the current regulator design, since three independent current regulators are used. Therefore, the disturbance rejection of the current loop must be good enough to reduce the effects of v_{no} .

2.4 Voltage Regulator

Assume that the input voltages are

$$\begin{bmatrix} e_R \\ e_S \\ e_T \end{bmatrix} = V \begin{bmatrix} \sin(\omega t) \\ \sin(\omega t - \frac{2}{3}\pi) \\ \sin(\omega t + \frac{2}{3}\pi) \end{bmatrix} \quad (12)$$

where V is the peak value of the input voltage. Then, v_d and v_q in the CRC can be expressed as

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{3}{2} k_{pt} V \begin{bmatrix} \sin(\omega t) \\ \cos(\omega t) \end{bmatrix} \quad (13)$$

Substituting Eq. (13) into Eq. (5) gives the current references

$$\begin{bmatrix} i_{Rref} \\ i_{Sref} \\ i_{Tref} \end{bmatrix} = \frac{2}{3} \frac{p}{k_{pt} k_m} V \begin{bmatrix} \sin(\omega t) \\ \sin(\omega t - \frac{2}{3}\pi) \\ \sin(\omega t + \frac{2}{3}\pi) \end{bmatrix} \quad (14)$$

Assume that the converter is lossless, that is, the switching and conducting loss of the switching elements and the loss of the reactive components are neglected. From the viewpoint of the power balance, the power of

the AC-side must be equal to the power in the DC-side.

$$e_R i_R + e_S i_S + e_T i_T = v_{dc} i_o \quad (15)$$

The inductor currents are controlled by three independent current regulators to track the current references from the CRC. The relationship of the inductor currents and the current references are

$$\begin{bmatrix} i_R \\ i_S \\ i_T \end{bmatrix} = k_{acr} \begin{bmatrix} i_{Rref} \\ i_{Sref} \\ i_{Tref} \end{bmatrix} \quad (16)$$

where k_{acr} is the closed-loop gain of the current loop in the line frequency ω .

Substituting Eqs. (12) and (16) into Eq. (15) and considering the dynamic behavior of the smoothing capacitor, we have the system dynamic equations as follows.

$$\frac{k_{acr}}{k_{pt} k_m} p = v_{dc} i_o \quad (17)$$

$$i_o = C \frac{dv_{dc}}{dt} + i_{load} \quad (18)$$

$$p = p_{avr} + k_s i_{load} \quad (19)$$

where i_o is the output current of the converter, i_{load} is the load current, and p_{avr} is the output of the voltage regulator.

From Eqs. (17)~(19), it is clear that the output dc-bus voltage (v_{dc}) is independent of the input source voltage. That means the output dc-bus voltage is not affected by the variations of the input source voltage, in other words, the converter has zero audio susceptibility.

For the derivation of the small signal model, let

$$v_{dc} = V_{dc} + \hat{v}_{dc} \quad (20)$$

$$i_o = I_o + \hat{i}_o \quad (21)$$

$$p = P + \hat{p} \quad (22)$$

$$p_{avr} = P_{avr} + \hat{p}_{avr} \quad (23)$$

$$i_{load} = I_{load} + \hat{i}_{load} \quad (24)$$

where capitalized quantities refer to steady state values and carets indicate small perturbations. Substituting the above perturbed quantities into Eqs. (17)~(19) and neglecting higher order terms, the small signal control block diagram can be get as shown in Fig. 3 and the corresponding parameters are given by

Fig. 3. Small signal model of the voltage loop.

$$Z_L(s) = \frac{1}{C_s}$$

$$G_z = \frac{I_o}{V_{dc}}$$

$$G_k = \frac{k_{acr}}{k_{pt}k_m V_{dc}}$$

In Fig. 3, $G_{avr}(s)$ is the voltage regulator and $H_{fb}(s)$ denotes the noise filter and notch filter. It is easy to show that if k_s is chosen as

$$k_s = G_k \quad (25)$$

the perturbed output dc voltage will also not be affected by the load changes.

The above derivations are based on the averaged small signal analysis, it shows the converter has a nearly perfect performance. In practice, the load change may be not in a small signal variation. This problem is well known to all approximate linearization techniques. Therefore, some experiments and simulation are needed to check the true behavior of this converter in large signal variations. However, small signal analysis will give useful information about this converter.

Based on this model, a simple PI compensator is designed in this work as a voltage regulator (AVR) to keep the output dc-bus voltage in the specified level and to have a fast transient response.

3. Unbalanced power source

Fig. 4(a) shows a typical three-phase ac supply system. Voltage unbalance and harmonic of the power source of the PWM ac-to-dc converter is caused by unbalanced loads and nonlinear loads in the same feeder.

(a) Single wired diagram of a general ac distribution network.

(b) Operating range in the terminal of the PWM ac-to-dc converter.

Fig. 4. Voltage and phase unbalance in three-phase voltage source.

Therefore, the range of the phase voltage at the common coupling point (CCP) corresponding to the rated current of the distribution transformer can be shown in Fig. 4(b).

The voltage unbalance ($VUB\%$) is defined as⁽⁹⁾

$$(VUB\%) = 100 \frac{V_{\max} - V_{\min}}{V_{\text{avg}}} \quad (26)$$

where V_R , V_S and V_T are the peak values of the phase voltage, and

$$V_{\text{avg}} = \frac{V_R + V_S + V_T}{3}$$

$$V_{\max} = \max(V_R, V_S, V_T)$$

$$V_{\min} = \min(V_R, V_S, V_T)$$

And the phase unbalance (PUB) is defined as⁽⁹⁾

$$PUB = \max(|\phi_R - \phi_S - 120^\circ|, |\phi_S - \phi_T - 120^\circ|, |\phi_T - \phi_R - 120^\circ|) \quad (27)$$

where ϕ_R , ϕ_S , and ϕ_T are the phase angles of the phase voltage. From Fig. 4(b), one can find that the maximum voltage unbalance is

$$(VUB\%)_{\max} = 100 \frac{2x}{1 - \frac{x}{3}} \quad (28)$$

and the maximum phase unbalance is

$$(PUB)_{\max} = 2\theta \quad (29)$$

where x denotes $\Delta V / V_n$, V_n is the voltage at no load, ΔV is the maximum voltage drop in the line impedance at rated current, and $\theta = \sin^{-1}(x)$. For example, if $x = 10\%$, then the maximum voltage unbalance is 20% and the maximum phase unbalance is 11.5°.

Therefore, the instantaneous voltages of one of the maximum voltage unbalance and the maximum phase unbalance can be expressed in the following, respectively.

- One case of maximum voltage unbalance :

$$\begin{bmatrix} e_R \\ e_S \\ e_T \end{bmatrix} = \begin{bmatrix} (V_n - \Delta V) \sin(\omega t) \\ (V_n - \Delta V) \sin(\omega t - 120^\circ) \\ (V_n + \Delta V) \sin(\omega t + 120^\circ) \end{bmatrix} \quad (30)$$

- One case of maximum phase unbalance :

$$\begin{bmatrix} e_R \\ e_S \\ e_T \end{bmatrix} = \begin{bmatrix} V_p \sin(\omega t - \theta) \\ V_p \sin(\omega t - 120^\circ + \theta) \\ V_p \sin(\omega t + 120^\circ + \theta) \end{bmatrix} \quad (31)$$

where $V_p = V_n \sqrt{1 - x^2}$

4. Effects of harmonics and voltage unbalance

The heart of the PWM converter is the CRC. The effects of the non-ideal power source containing the harmonics and unbalanced voltages on the CRC are investigated in this section. First, the CRC in the ideal power source is discussed for comparing with other cases. It is noted that the scaling factor in the practical

hardware are omitted and the desired output power is constant in the following discussions.

4.1 Ideal Power Source

The sensed voltages in case of the ideal power source can be expressed as

$$\begin{bmatrix} v_{rn} \\ v_{sn} \\ v_{tn} \end{bmatrix} = V \begin{bmatrix} \sin(\omega t) \\ \sin(\omega t - 120^\circ) \\ \sin(\omega t + 120^\circ) \end{bmatrix} \dots\dots\dots (32)$$

where V is the amplitude of the phase voltage. From (4), the parameters in the CRC can be given by

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{3}{2} V \begin{bmatrix} \sin(\omega t) \\ -\cos(\omega t) \end{bmatrix} \dots\dots\dots (33)$$

and

$$\begin{aligned} V_{dq} &= v_d^2 + v_q^2 \\ &= \frac{9}{4} V^2 \dots\dots\dots (34) \end{aligned}$$

If the desired output power is P , the reference currents become

$$\begin{bmatrix} i_{ref} \\ i_{sref} \\ i_{tref} \end{bmatrix} = \frac{3}{2} P V^{-1} \begin{bmatrix} \sin(\omega t) \\ \sin(\omega t - 120^\circ) \\ \sin(\omega t + 120^\circ) \end{bmatrix} \dots\dots\dots (35)$$

Therefore, the three-phase PWM ac-to-dc converter has a unity power factor if the input currents exactly tracks the reference currents. From the above derivations, it is clear that V_{dq} is a dc term if the power source is ideal.

4.2 Harmonics

The input voltages containing one balanced high frequency harmonic are given as

$$\begin{bmatrix} v_{rn} \\ v_{sn} \\ v_{tn} \end{bmatrix} = V_1 \begin{bmatrix} \sin(\omega t) \\ \sin(\omega t - 120^\circ) \\ \sin(\omega t + 120^\circ) \end{bmatrix} + V_2 \begin{bmatrix} \sin(k\omega t) \\ \sin(k\omega t - 120^\circ) \\ \sin(k\omega t + 120^\circ) \end{bmatrix} \dots\dots\dots (36)$$

According to k , three cases are discussed as follows.

1. Zero-phase sequence, i. e. $k=3, 6, 9, \dots$

The parameters in the CRC are given by

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{3}{2} V_1 \begin{bmatrix} \sin(\omega t) \\ -\cos(\omega t) \end{bmatrix} \dots\dots\dots (37)$$

and

$$V_{dq} = \frac{9}{4} V_1^2 \dots\dots\dots (38)$$

2. Positive-phase sequence, i. e. $k=4, 7, 10, \dots$

The parameters in the CRC are given by

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{3}{2} V_1 \begin{bmatrix} \sin(\omega t) \\ -\cos(\omega t) \end{bmatrix} + \frac{3}{2} V_2 \begin{bmatrix} \sin(k\omega t) \\ -\cos(k\omega t) \end{bmatrix} \dots\dots\dots (39)$$

and

$$V_{dq} = \frac{9}{4} (V_1^2 + V_2^2) + \frac{9}{2} V_1 V_2 \cos((k-1)\omega t) \dots\dots (40)$$

3. Negative-phase sequence, i. e. $k=2, 5, 8, \dots$

The parameters in the CRC are given by

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{3}{2} V_1 \begin{bmatrix} \sin(\omega t) \\ -\cos(\omega t) \end{bmatrix} + \frac{3}{2} V_2 \begin{bmatrix} \sin(k\omega t) \\ \cos(k\omega t) \end{bmatrix} \dots\dots\dots (41)$$

and

$$V_{dq} = \frac{9}{4} (V_1^2 + V_2^2) + \frac{9}{2} V_1 V_2 \cos((k+1)\omega t) \dots\dots (42)$$

4.3 Maximum voltage amplitude unbalance

The voltages in the case of the maximum voltage amplitude unbalance can be expressed as

$$\begin{bmatrix} v_{rn} \\ v_{sn} \\ v_{tn} \end{bmatrix} = \begin{bmatrix} (V - \Delta V) \sin(\omega t) \\ (V - \Delta V) \sin(\omega t - 120^\circ) \\ (V + \Delta V) \sin(\omega t + 120^\circ) \end{bmatrix} \dots\dots\dots (43)$$

Then, the parameters in the CRC can be obtained as

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{3}{2} V \begin{bmatrix} \sin(\omega t) \\ -\cos(\omega t) \end{bmatrix} + \Delta V \begin{bmatrix} -\sin(\omega t) - \frac{\sqrt{3}}{2} \cos(\omega t) \\ \frac{\sqrt{3}}{2} \sin(\omega t) \end{bmatrix} \dots\dots (44)$$

and

$$\begin{aligned} V_{dq} &= \frac{9}{4} V^2 + \frac{5}{4} (\Delta V)^2 - \frac{3}{2} \Delta V V \\ &\quad - \frac{1}{2} ((\Delta V)^2 - \frac{3}{2} \Delta V V) \cos(2\omega t) \\ &\quad + \frac{\sqrt{3}}{8} (4(\Delta V)^2 - 9\Delta V V) \sin(2\omega t). \dots\dots (45) \end{aligned}$$

4.4 Maximum phase unbalance

The voltages in the case of the maximum phase unbalance may be expressed as

$$\begin{bmatrix} v_{rn} \\ v_{sn} \\ v_{tn} \end{bmatrix} = \begin{bmatrix} V \sin(\omega t - \theta) \\ V \sin(\omega t - 120^\circ + \theta) \\ V \sin(\omega t + 120^\circ + \theta) \end{bmatrix} \dots\dots\dots (46)$$

Then, the parameters in the CRC can be obtained as

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \frac{3}{2} V \cos(\theta) \begin{bmatrix} \sin(\omega t) \\ -\cos(\omega t) \end{bmatrix} + \frac{3}{2} V \sin(\theta) \begin{bmatrix} -\frac{1}{3} \cos(\omega t) \\ \sin(\omega t) \end{bmatrix} \dots\dots\dots (47)$$

and

$$\begin{aligned} V_{dq} &= V^2 \left(\frac{9}{4} \cos^2(\theta) + \frac{5}{4} \sin^2(\theta) \right. \\ &\quad \left. - \cos(2\omega t) - \frac{3}{2} \sin(2\theta) \sin(2\omega t) \right) \dots\dots\dots (48) \end{aligned}$$

4.5 Modified CRC

From the above derivations, if the power source is non-ideal then V_{dq} contains undesired components. These will contaminate the reference current generated by the CRC. For the maximum voltage unbalance and the maximum phase unbalance, from Eqs. (43)~(48), these equations can be expressed in a general form as

$$V_{dq} = V_{dc} (1 - k_1 \sin(2\omega t + \alpha)) \dots\dots\dots (49)$$

$$v_d = k_2 \sin(\omega t + \beta) \dots\dots\dots (50)$$

$$v_q = k_3 \sin(\omega t + \gamma) \dots\dots\dots (51)$$

where

$$k_1 < 1$$

Then,

Fig. 5. Modified coordinate transformation.

$$\begin{aligned} V_{dq}^{-1} &= V_{dc}^{-1} (1 + k_1 \sin(2\omega t) + k_2^2 \sin^2(2\omega t) + \dots) \\ &\cong V_{dc}^{-1} (1 + k_1 \sin(2\omega t)) \end{aligned} \quad (52)$$

Therefore,

$$\begin{aligned} \begin{bmatrix} i_{Rref} \\ i_{Sref} \\ i_{Tref} \end{bmatrix} &= C^T \frac{p}{V_{dc}} \begin{bmatrix} k_2 \sin(\omega t + \beta) + \frac{k_1 k_2}{2} \sin(\omega t + \alpha - \beta) \\ k_3 \sin(\omega t + \gamma) + \frac{k_1 k_3}{2} \sin(\omega t + \alpha - \gamma) \\ -\frac{k_1 k_2}{2} \sin(3\omega t + \alpha + \beta) \\ -\frac{k_1 k_3}{2} \sin(3\omega t + \alpha + \gamma) \end{bmatrix} \end{aligned} \quad (53)$$

From the above derivation, the current references contain third harmonics and their amplitude is proportional to the component with frequency of 2ω in V_{dq} .

A modified CRC, shown in Fig. 5, is proposed in this paper to reduce the effects of voltage unbalance and harmonics. This is done by adding a low pass filter into the CRC to filter out the components with frequency greater than 2ω in V_{dq} . The circuit of the MCRC is shown in Fig. 5(b).

5. Simulation and experimental results

For verification, a laboratory prototype three-phase PWM ac-to-dc converter system is setup. An input three-phase transformer is used to match the utility voltage and the designed input voltage, and a three-phase input filter is used to eliminate the high frequency components in the inductor currents. Three 50A 600V IGBT modules are used as switching elements. Five analog multipliers are used to perform the operations in the CRC.

5.1 System parameters

The parameters used in this three-phase PWM ac-to-dc converter are

1. the source phase voltage is 80 V ;
2. the dc output voltage is 240 V ;
3. the load resistor is 10.8 Ω ;
4. the inductance of the input inductor is 1 mH ;
5. the switching frequency is 11 kHz ;
6. the capacitance of the smoothing capacitor is 7200 μ F.

Fig. 6 shows the measured waveforms in the balanced

(a) Transient of output voltage and load current.

(b) Input phase voltage and input current.

Fig. 6. Experimental waveforms in the balanced power source.

power source. Fig. 6(a) shows the transient response of the output voltage at no-load to full-load and Fig. 6(b) shows the corresponding input current and phase voltage. From Fig. 6, it shows the converter has a unity power factor and a nearly perfect voltage dynamic regulation.

5.2 Simulation results

The proposed three-phase PWM ac-to-dc converter system including a power electronic stage and a control stage is simulated by the PC based SIMNON program⁽¹⁰⁾. Fig. 7 and Fig. 8 show the simulation results of the maximum voltage amplitude unbalance and maximum phase unbalance with $x=10\%$, respectively. V_{dq} in the unbalanced power source contains second order harmonics which lead to the distorted input current.

For demonstrating the design of the LPF in the MCRC, three Butterworth low pass filters (BLPF) are illustrated.

1. Second-order BLPF with cutoff frequency=34 Hz: -20 dB for 120 Hz;

(a) Waveforms of input phase voltages.

(b) Waveform of V_{dq} . (c) Waveform of V_{dqf} .

(d) Waveforms of input currents, in the CRC.

(e) Waveforms of input currents, in the MCRC.

Fig. 7. SIMNON simulation results in case of the maximum voltage amplitude unbalance.

2. Second-order BLPF with cutoff frequency=12 Hz: -40 dB for 120 Hz;
3. Fourth-order BLPF with cutoff frequency=34 Hz: -40 dB for 120 Hz.

The input current total harmonic distortion (THD) and the ripple factor of the output voltage are chosen to evaluate the effects of the BLPF, and the results are shown in Fig. 9(a) for the ripple factor and Fig. 9(b) for the current THD respectively. From Fig. 9, it shows the case (2) and case (3) have nearly same results and the case (2) and case (3) are slightly better than the case (1) in the current THD of the input currents and the ripple factor of the output voltage. That means the design of LPF in the MCRC does not strongly affect the three-phase current waveforms i_R, i_S and i_T .

The PLL method, which is possible to obtain balanced and purely sinusoidal three-phase current reference signals by introducing a phase-lock-loop circuit even if three-phase voltages are unbalanced, and the CRC method are also evaluated and are shown in Fig. 9. From Fig. 9, we can find three facts:

1. The ripple factor of the CRC method is the smallest.

Fig. 8. SIMNON simulation results in case of the maximum phase unbalance.

2. The current THD of the CRC method is the largest.
3. The voltage ripple and the current THD of the PLL method are smaller than the ones of the MCRC method.

These facts can be explained in the following. Since the current references in the CRC are based on the instantaneous reactive power theory, the ripple in the output voltage is zero theoretically even if the voltage source is unbalanced. However, the input currents will be distorted. The input current THD of the MCRC is greatly improved but the ripple factor becomes higher because the phase differences exist between the current references and the input voltages. This can be seen in Eq. (53). The PLL method is better than the MCRC method in the voltage ripple, but it losses the feedforward compensation of the input line voltages. Considering the feedforward compensation of the input line voltages, the MCRC having a second-order BLPF with cutoff frequency = 34 Hz is chosen in experiments.

5.3 Experimental results

To demonstrate the effects of the modified CRC, a single-phase auto-transformer (about 3 kVA) is used to

(a) Ripple factor of the output dc voltage.

(b) Total harmonic distortion of the input current.

Fig. 9. Comparison of the CRC, MCRC and PLL methods.

Fig. 10. Scheme of the experimental equipment.

adjust the amplitude of the phase-T voltage to the desired voltage unbalance and the setup of this experiment is shown in Fig. 10. Here, the phase-T voltage is increased to 110%. The input phase-voltages in this experiment are slightly contaminated by harmonics. The spectrums of the input phase-voltages are measured and the averaged THD is 2.6%. Fig. 11(a) shows the waveform of V_{dq} and its spectrum. It shows that V_{dq} contains an amount of second order harmonic, this is agreed with the analytic results. Fig. 11(b) and Fig. 11(c) show the experimental waveforms of the input inductor currents with the CRC and with the MCRC, respectively. The measured averaged THD of the input

(a) Waveforms of V_{dq} and its spectrum.

(b) Waveforms of input currents, in the CRC.

(c) Waveforms of input currents, in the MCRC.

Fig. 11. Experimental results in the unbalanced power source.

currents is 3.96% when the modified CRC is used and is 5.05% for the CRC.

6. Conclusions

In this paper, effects of non-ideal power source on a three-phase PWM ac-to-dc converter based on the instantaneous reactive power theory has been presented.

Moreover, a modified current reference calculator was proposed to reduce the effects.

A brief conclusions of this paper are given in the following.

1. The three-phase PWM ac-to-dc converter with instantaneous power feedback has a nearly perfect dynamic voltage regulation.
2. Harmonics will appear in the current references if the power source of the converter is unbalanced or contains harmonics.
3. The effects of voltage unbalance and harmonics can be compensated by a low pass filter which completely attenuates 2ω components or higher ones.
4. Both simulation and experimental results are presented and these are agreed with the analytic ones.

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Wen-Inne Tsai (Non-member) He was born in Tainan, Taiwan, 1963. He graduated from the Department of Electrical Engineering of National Institute of Technology (NKIT) in 1984 and received the M. S. degree in electrical engineering from National Cheng Kung University (NCKU) in 1989. Since 1989 he has been working towards the Ph. D. degree in electrical engineering at NCKU. Currently, his research interests include modelling and control of power electronic systems, soft switched power converters, active power filters and the advanced motion control.

Wing-Tsung Tsai (Non-member) He was born in Tainan, Taiwan, 1964. He graduated from the Department of Electrical Engineering of National Institute of Technology (NKIT) in 1984 and received the M. S. degree in electrical engineering from National Cheng Kung University (NCKU) in 1991. He is presently a lecturer in the Electrical Engineering Department, Nantai Junior College of Industrial and Commercial Management. Currently, his research interests include active power filter, battery energy storage system, and residential photovoltaic power conditional system.

York-Yih Sun (Non-member) He was born in Amoy, Fukien Province, China, 1933. He received the B. S. E. E. degree from the National Cheng Kung University, Taiwan, Republic of China, in 1957, and the M. S. E. E. degree from Purdue University, Lafayette, IN, in 1968. He is presently a Professor in the Electrical Engineering Department, National Cheng Kung University where he has been teaching for 32 years. His current interests are in power electronics, digital control and adaptive signal processing.