Journal of Zhejiang University SCIENCE A
ISSN 1673-565X (Print); ISSN 1862-1775 (Online)
www.zju.edu.cn/jzus; www.springerlink.com
E-mail: jzus@zju.edu.cn

Science Letters:

Multi-frequency proportional-resonant (MFPR) current controller for PWM VSC under unbalanced supply conditions*

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Received July 4, 2007; revision accepted Aug. 28, 2007

Abstract: This letter presents a multi-frequency proportional-resonant (MFPR) current controller developed for PWM voltage source converter (VSC) under the unbalanced supply voltage conditions. The delta operator is used in place of the shift operator for the implementation of MFPR by using a low-cost fixed-point DSP. The experimental results with an alternative control strategy validated the feasibility of the proposed MFPR current controller for the PWM VSC during voltage unbalance.

Key words: PWM VSC, Multi-frequency proportional-resonant (MFPR), Voltage unbalance
doi:10.1631/jzus.2007.A1527 **Document code:** A **CLC number:** TM315; TM614

INTRODUCTION

The PWM voltage source converter (VSC) is widely used in the renewable energy conversion systems and electric power systems, i.e., FACTS and HVDC systems. It provides numerous advantages such as sinusoidal input current with controllable power factor, high-quality DC output voltage with a small filter capacitor and capability of independent control of active and reactive powers with bi-directional power flow. These performances cannot be fully achieved under the unbalanced voltage condition, which is common in power systems, particularly for the weak AC systems. It has been reported that as a result of unbalanced supply voltage the negative-sequence components in the AC input currents and even fluctuations in the output DC-link voltage will occur (Stankovic and Lipo, 2001).

It was concluded (Rioual *et al*., 1996; Song and Nam, 1999; Stankovic and Lipo, 2001; Magueed *et al*., 2004; Yazdani and Iravani, 2006) that under unbalanced voltage conditions, the negative-sequence current was independently regulated, which necessitates the use of sequential-component decomposing filters that might reduce the bandwidth and overall stability margin of controller. A proportional-resonant (PR) controller in the stationary frame was proposed and applied in the single-phase and balanced three-phase systems (Zmood and Holmes, 2003). However, so far the application of PR controller to PWM VSC has not been analyzed and implemented during voltage unbalance.

This paper develops a multi-frequency proportional-resonant (MFPR) current controller for the PWM VSC under unbalanced supply voltage. The use of delta operator instead of the shift operator is introduced for the implementation of MFPR by using a low-cost fixed-point DSP. The experiments with an alternative control strategy validated the feasibility of the proposed MFPR current controller for the PWM VSC during voltage unbalance.

ANALYSIS OF PWM VSC

The schematic diagram of a typically three-phase grid-connected PWM VSC is shown in Fig.1.

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* Project (No. 50577056) supported by the National Natural Science Foundation of China

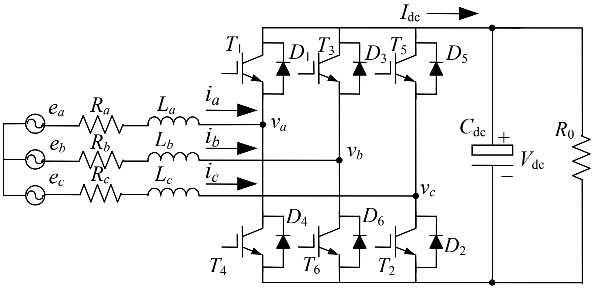

Fig.1 Schematic diagram of a three-phase PWM VSC

Assuming no zero-sequence component existing, the voltage and current of fundamental frequency can be decomposed into positive- and negative-sequence components when the supply voltage and input impedances are unbalanced. The behavior for the AC side of the PWM VSC can be described in terms of positive- and negative-sequence voltage and currents in the dq^+ and dq^- reference frames as (Xu *et al.*, 2005)

$$\begin{cases} \boldsymbol{E}_{dq+}^{+} = L\dfrac{\mathrm{d}\boldsymbol{I}_{dq+}^{+}}{\mathrm{d}t} + R\boldsymbol{I}_{dq+}^{+} + \mathrm{j}\omega L\boldsymbol{I}_{dq+}^{+} + \boldsymbol{V}_{dq+}^{+}, \\ \boldsymbol{E}_{dq-}^{-} = L\dfrac{\mathrm{d}\boldsymbol{I}_{dq-}^{-}}{\mathrm{d}t} + R\boldsymbol{I}_{dq-}^{-} - \mathrm{j}\omega L\boldsymbol{I}_{dq-}^{-} + \boldsymbol{V}_{dq-}^{-}, \end{cases} \tag{1}$$

where $\boldsymbol{E}_{dq+}^{+} = E_{d+}^{+} + \mathrm{j}E_{q+}^{+}$, $\boldsymbol{V}_{dq+}^{+} = V_{d+}^{+} + \mathrm{j}V_{q+}^{+}$ and $\boldsymbol{I}_{dq+}^{+} = I_{d+}^{+} + \mathrm{j}I_{q+}^{+}$ are the space vectors of the positive-sequence component of the AC input supply voltages, the VSC input pole voltages, and the AC input currents in the positive-rotating synchronous frame, respectively. $\boldsymbol{E}_{dq-}^{-} = E_{d-}^{-} + \mathrm{j}E_{q-}^{-}$, $\boldsymbol{V}_{dq-}^{-} = V_{d-}^{-} + \mathrm{j}V_{q-}^{-}$ and $\boldsymbol{I}_{dq-}^{-} = I_{d-}^{-} + \mathrm{j}I_{q-}^{-}$ are the space vectors of the negative-sequence component of the AC input supply voltages, the VSC input pole voltages, and the AC input currents in the negative-rotating synchronous frame, respectively. $L_a=L_b=L_c=L$ and $R_a=R_b=R_c=R$ are the three-phase VSC input inductors and resistors, respectively.

The active and reactive powers at the VSC input pole are expressed as (Xu *et al.*, 2005):

$$\begin{cases} P_{\mathrm{AC}} = P + P_{\cos 2}\cos(2\omega t) + P_{\sin 2}\sin(2\omega t), \\ Q_{\mathrm{AC}} = Q, \end{cases} \tag{2a}$$

where P, $P_{\cos2}$, $P_{\sin2}$ and Q are given by

$$\begin{bmatrix} P \\ Q \\ P_{\cos 2} \\ P_{\sin 2} \end{bmatrix} = \frac{3}{2}\begin{bmatrix} V_{d+}^{+} & V_{q+}^{+} & V_{d-}^{-} & V_{q-}^{-} \\ V_{q+}^{+} & -V_{d+}^{+} & V_{q-}^{-} & -V_{d-}^{-} \\ V_{d-}^{-} & V_{q-}^{-} & V_{d+}^{+} & V_{q+}^{+} \\ V_{q-}^{-} & -V_{d-}^{-} & -V_{q+}^{+} & V_{d+}^{+} \end{bmatrix}\begin{bmatrix} I_{d+}^{+} \\ I_{q+}^{+} \\ I_{d-}^{-} \\ I_{q-}^{-} \end{bmatrix}. \tag{2b}$$

Thus, using the power-balancing equation, the DC side equation under the unbalanced condition can be expressed as

$$P_{\mathrm{ac}} = C_{\mathrm{dc}}\frac{\mathrm{d}V_{\mathrm{dc}}}{\mathrm{d}t} + I_{\mathrm{dc}}V_{\mathrm{dc}}. \tag{3}$$

Consequently, Eqs.(1) and (3) can be used to represent the AC and DC side system model for the three-phase PWM VSC under unbalanced supply conditions.

CONTROL STRATEGIES FOR PWM VSC DURING SUPPLY UNBALANCE

As shown in Eq.(1), under unbalanced supply conditions, both positive- and negative-sequence currents need to be controlled. Apart from controlling the average active and reactive powers, i.e., P and Q shown in Eq.(2b), two more parameters can be controlled. For instance, the system can be designed to operate according to one of the following control targets:

Target I: To eliminate the pulsations of DC-link voltage.

Target II: To obtain balanced three-phase input currents, i.e., no negative-sequence current.

For Target I, the oscillating terms of active power shown in Eq.(2) should be controlled as zero, i.e., $P_{\cos2}=P_{\sin2}=0$.

As the d^+-axis is fixed to the positive input voltage, i.e., $V_{q+}^{+}=0$, the reference values of negative-sequence dq^- currents for the two different targets can be simplified, and compiled in Table 1.

Table 1 Positive- and negative-sequence current references

Target	I_{d+}^{+*}	I_{q+}^{+*}	I_{d-}^{-*}	I_{q-}^{-*}
I	$2V_{d+}^{+}P/(3K_1)$	$2V_{d+}^{+}Q/(3K_2)$	$-k_{dd}I_{d+}^{+*}-k_{qd}I_{q+}^{+*}$	$-k_{qd}I_{d+}^{+*}+k_{dd}I_{q+}^{+*}$
II	$2P/(3V_{d+}^{+})$	$-2Q/(3V_{d+}^{+})$	0	0

$K_1=V_{d+}^{+2}-V_{d-}^{-2}-V_{q-}^{-2}$, $K_2=V_{d+}^{+2}+V_{d-}^{-2}+V_{q-}^{-2}$, $k_{dd}=V_{d-}^{-}/V_{d+}^{+}$, $k_{qd}=V_{q-}^{-}/V_{d+}^{+}$

CURRENT REGULATORS FOR SEQUENTIAL COMPONENTS

Proportional-resonant (PR) regulators

Once the references of positive- and negative-sequence currents are determined, it is critical that the current regulators should control both positive- and negative-sequence components accurately. This paper develops a new current regulator based on the PR (proportional-resonant) principle, which directly regulates the overall current including both the positive and negative components in the stationary $\alpha\beta$ frame. The schematic diagram of the proposed controller is shown in Fig.2. As shown, without involving any sequential-component decomposition, the measured currents fed to the current regulator consist of base-frequency and other odd-harmonic AC components under unbalanced conditions. However, the four current references given in Table 1 are all DC signals, separated in the positive- and negative-rotating synchronous frames. Thus it is necessary to transform the reference values to the stationary $\alpha\beta$ frame.

Unlike the traditional dual current regulators simply having DC signals in the positive- and negative-rotating synchronous frames separately (Rioual *et al*., 1996; Song and Nam, 1999; Stankovic and Lipo, 2001; Magueed *et al*., 2004; Yazdani and Iravani, 2006), the proposed current regulators shown in Fig.3 have AC values with multi-frequency components. In order to eliminate the controlling errors for various current harmonics, such as the 3rd and the 5th, a PR regulator in the stationary $\alpha\beta$ frame is introduced and expanded as a multi-frequency proportional-resonant (MFPR) one, as depicted in Fig.3.

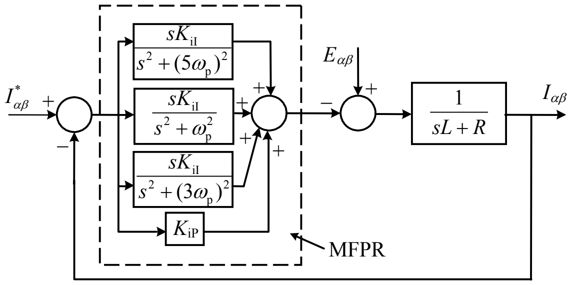

Fig.3 Current control diagram based on the multi-frequency proportional-resonant (MFPR) controller

According to Fig.3, the open-loop transfer function of the current regulator can be derived as

$$C(s)=K_{ip}+\frac{sK_{iI}}{s^2+\omega_p^2}+\frac{sK_{iI}}{s^2+(3\omega_p)^2}+\frac{sK_{iI}}{s^2+(5\omega_p)^2},\quad(4)$$

where K_{ip} and K_{iI} are the proportional and resonant coefficients. $\omega_p=2\pi f$, f=50 Hz is the line frequency. The sensitivity and stability of the PR regulator have been verified theoretically (Zmood and Holmes, 2003; Newman and Holmes, 2003).

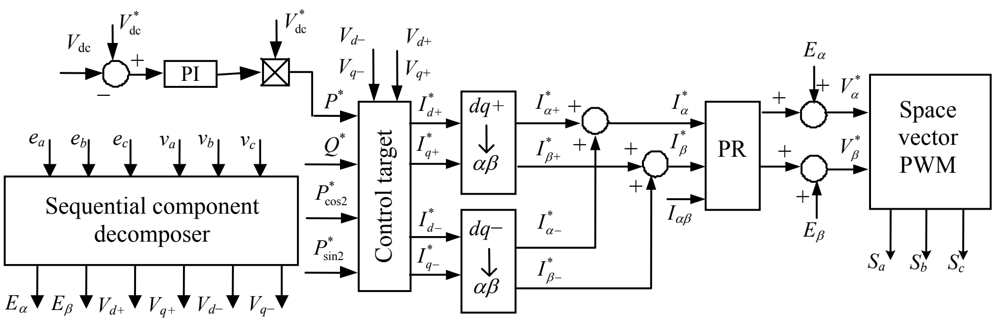

Fig.2 Schematic diagram of the whole control system

Implementation of MFPR regulators

To implement the proposed MFPR regulator digitally by using a low-cost fixed-point Texas Instruments TMS320F240 DSP, the transfer function Eq.(4) must firstly be digitized using an appropriate digitization method. With the conventional shift-operator implementation, the finite-word-length effects resulting from the use of integer variables on a fixed-point DSP lead to some performance degradation especially for the sharply tuned filters such as the resonant components in MFPR regulator (Newman and Holmes, 2003). In order to improve the performance, the use of delta operator δ in place of the shift operator has been investigated. The delta operator has recently gained interest in fast digital control due to its superior finite word length performance (Kauraniemi *et al.*, 1998; Newman and Holmes, 2003). The delta operator δ can be defined in terms of the shift operator z as

$$\delta^{-1} = \Delta z^{-1}/(1-z^{-1}). \tag{5}$$

Essentially, the delta-operator implementation of resonant compensator involves converting a 2nd-order transfer function in z

$$H(z)=\frac{b_0+b_1z^{-1}+b_2z^{-2}}{1+a_1z^{-1}+a_2z^{-2}} \tag{6}$$

into a corresponding 2nd-order transfer function in δ

$$H(\delta)=\frac{\beta_0+\beta_1\delta^{-1}+\beta_2\delta^{-2}}{1+\alpha_1\delta^{-1}+\alpha_2\delta^{-2}} \tag{7}$$

where $\beta_0=b_0$, $\beta_1=(2b_0+b_1)/\Delta$, $\beta_2=(b_0+b_1+b_2)/\Delta^2$, $\alpha_1=(2+a_1)/\Delta$, $\alpha_2=(1+a_1+a_2)/\Delta^2$, and Δ is a positive constant, less than unity, which is carefully chosen to select the appropriate ranges for the α, β coefficients (Kauraniemi *et al.*, 1998).

Eq.(7) is implemented by using digital filter implementation structure of the transposed direct form II (DFIIt) shown in Fig.4. The DFIIt structure is chosen out from the many filter structures available because of its best round-off noise performance for the delta-operator-based filters (Kauraniemi *et al.*, 1998). From Fig.4, the differential equations to be coded for the DSP implementation can be written, in proceeding order, as

$$\begin{cases} w_4(n)=\Delta w_3(n-1)+w_4(n-1), \\ w_2(n)=\Delta w_1(n-1)+w_2(n-1), \\ y(n)=\beta_0x(n)+w_4(n), \\ w_3(n)=\beta_1x(n)-\alpha_1y(n)+w_2(n), \\ w_5(n)=\beta_2x(n)-\alpha_2y(n). \end{cases} \tag{8}$$

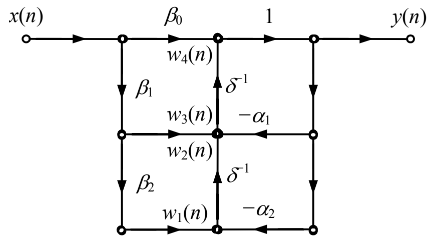

Fig.4 Direct form II transpose (DFIIt) structure for the 2nd-order digital filter

EXPERIMENTAL RESULTS

The experiments with the proposed MFPR controller were conducted on a laboratory prototype PWM VSC with a 10% of supply voltage unbalance. The complete control algorithm was implemented using a fixed-point DSP TI TMS320F240 with both sampling rate and switching frequency of 7.2 kHz.

Figs.5~7 show the performance of PWM VSC with Target I and Q=0. It can be seen clearly that during voltage unbalance the proposed control scheme provides excellent steady-state and transient tracking performance. As shown in Fig.6, during supply voltage unbalance, negative-sequence currents do exist so as to achieve the control objective of eliminating the oscillations of the DC-link voltage, as shown in Fig.7.

CONCLUSION

An MFPR current controller developed for the PWM VSC during supply voltage unbalance was proposed. The delta operator δ, instead of the shift operator, was introduced for the implementation of MFPR by using a low-cost fixed-point DSP. The experimental results with one alternative control target showed that the proposed MFPR current controller is capable of providing pretty good steady-state and transient performances for the PWM VSC under unbalanced supply voltage conditions.

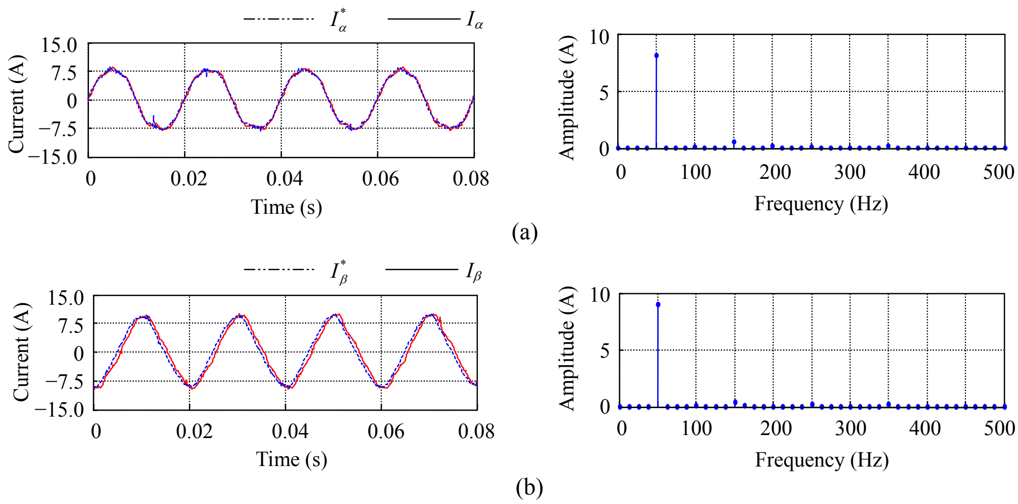

Fig.5 Experimental results of $\alpha\beta$-axis currents and their corresponding spectrum with the proposed MFPR controller under unbalanced conditions R_{dc}=50 Ω. (a) Spectrum of α-axis current; (b) Spectrum of β-axis current

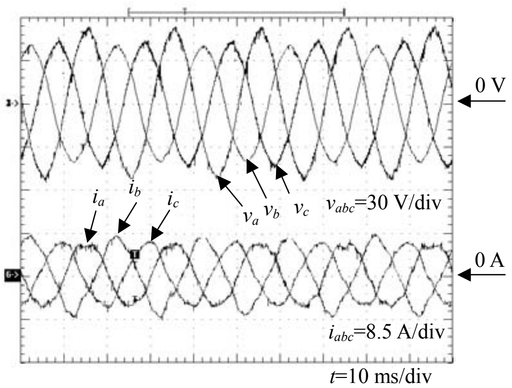

Fig.6 Experimental results of three-phase input pole voltages and currents with the proposed MFPR controller under unbalanced conditions R_{dc}=50 Ω

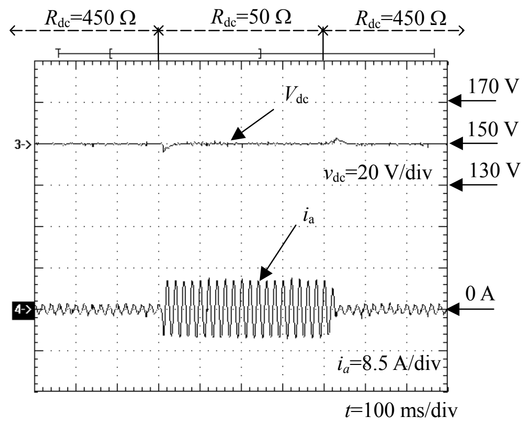

Fig.7 Experimental results of DC-link voltage and A-phase current during load varying (R_{dc}=50 Ω and 450 Ω) with the proposed MFPR controller under unbalanced conditions

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