# PIR-Based Control for Three-Phase PWM Rectifier with H-Bridge Load

WANG Jian, LI Yongdong, ZHENG Yanwen and YUAN Xibo

Department of Electrical Engineering, Tsinghua University, Beijing, 100084, P.R.China

***Abstract-* In order to eliminate the pulsating instantaneous power of twice of the output voltage frequency in case of single-phase load, a proportional-integral-resonant (PIR) based controller is proposed in this paper for the PWM rectifier in both the dc-link voltage control loop and the ac current control loop. The resonant regulator is very suitable for tracking the ac component, which provides the optimum input power for the output single phase load. Thus, the pulsating power across the dc-link capacitor can be eliminated, reducing the size and cost of the capacitor. The only information required in the control about the load is the H-bridge output frequency, so the control is easy to implement. Simulation and experimental results show good performance of the proposed control strategy.**

## I. INTRODUCTION

Three-phase voltage source pulse-width modulation (PWM) rectifier presents the characteristics of constant dc-link voltage, bidirectional power flow, low input harmonic current contents and controllable input power factor [1-3]. Therefore, it has been widely used in high-performance electric drive systems, especially when the regenerative power should be fed back into the grid in case of frequent acceleration and deceleration.

The load of a three-phase PWM rectifier can be either three-phase or single-phase. For three-phase load, the output power of the inverter is usually constant in the steady state, and the dc-link can have a constant voltage if the power absorbed from the grid is controlled equal to the power delivered to the load by PWM rectifier. In this case, only a small dc-link capacitor is needed.

Nevertheless, when a single-phase full bridge (H-bridge) inverter is used as the load of the three-phase PWM rectifier, e.g. in the power cells of the regenerative cascaded H-bridge multilevel converter [4,5] as shown in Fig. 1 and Fig. 2, the inverter will deliver to the load a pulsating power whose frequency is twice of the output voltage frequency even in the steady state. If the pulsating power cannot be completely exchanged with the power grid by the PWM rectifier, it must be absorbed by the dc-link capacitor, which significantly increases the size and cost of the dc-link capacitor.

Voltage oriented control (VOC) [2] is a common strategy used to control the PWM rectifier with three-phase load. However, it is not well-suited for the PWM rectifiers with single phase H-bridge load and small dc-link capacitance because of the pulsating power. In [5], an extra *pq* equalizer control loop is added to dc-link average voltage control loop, trying to balance the instantaneous power generated by the H-bridge inverter and the power inputted into the PWM rectifier. The instantaneous output power from the H-bridge inverter is obtained by measuring the dc-link voltage, the duty-ratio of the H-bridge legs and the load current, which needs lots of sensors or communication between the load and the rectifier side controllers and complicates the multilevel converter system.

This paper proposes a new control strategy for PWM rectifier with H-bridge load and small dc-link capacitance. By analyzing the power flow between the PWM rectifier and the H-bridge load, it is shown that, in order to balance the power between the load and source, the PWM rectifier needs to absorb active power containing a characteristic frequency component of twice the H-bridge output frequency. Therefore, resonant (R) regulators [6,7] are proposed to be added both in the dc-link voltage control loop and in the current control loop to cooperate with the conventional proportional-integral (PI) regulators, which becomes proportional-integral-resonant (PIR) regulators. The resonant frequency of all the R regulators is tuned to the characteristic frequency in order that the current reference can be the value which provides balanced power with the H-bridge load and that the current can precisely track its reference in the steady state. Furthermore, only the information about the H-bridge inverter output frequency is needed for controlling the PWM rectifier, which simplifies the system configuration. The proposed method is verified by simulation and experimental results.

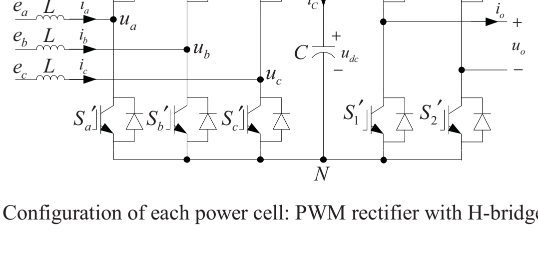


Fig.1 Configuration of each power cell: PWM rectifier with H-bridge load

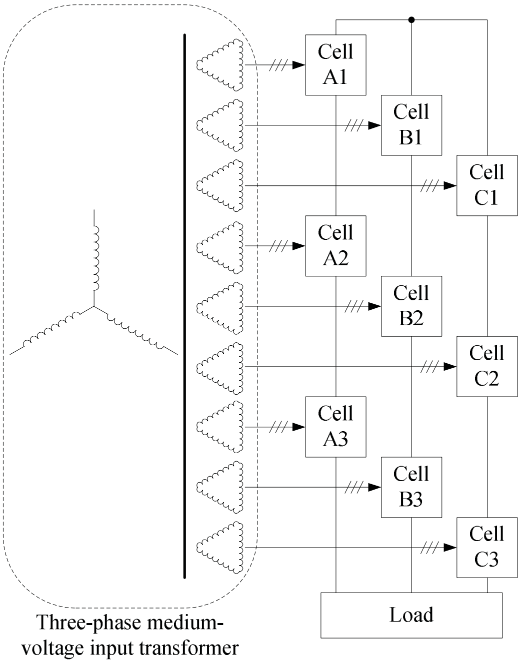


Fig. 2 Topology of regenerative cascaded H-bridge multilevel converter

## II. PWM RECTIFIER WITH H-BRIDGE LOAD

The power cell topology studied in this paper is shown in Fig. 1. The input stage of the cell is a three-phase boost-type PWM rectifier, and the output stage is an H-bridge inverter. This cell can handle bidirectional power flow and is typically used in the regenerative cascaded H-bridge multilevel converters with four quadrants operation capability. The structure of the whole converter is shown in Fig. 2, which is composed of multiple cells. The output stage of the converter is star-connected and each phase leg is made up of several cells connected in series. The number of cells in each leg depends on the required output voltage level. Since all the cells in one converter are identical and the input PWM rectifiers are controlled independently, this paper will focus on one cell only.

In order to derive the control method, the system is first modeled as follows.

On the source side, the PWM rectifier can be modeled by the ac-side voltage equation and dc-side current equation in three-phase natural reference frame as shown in (1)(2)

$$e_k = L\frac{\mathrm{d}i_k}{\mathrm{d}t} + u_k,\ k = a,b,c \tag{1}$$

$$C\frac{\mathrm{d}u_{dc}}{\mathrm{d}t} = i_S - i_L \tag{2}$$

where the meaning of each symbol is indicated in Fig. 1.

If the change of the energy stored in the ac inductance is assumed negligible, the power balance equation of the ac-side and dc-side is given by

$$p_S = e_a i_a + e_b i_b + e_c i_c = u_{dc} i_S \tag{3}$$

where $p_S$ is the instantaneous power on the source side. The VOC is implemented in synchronous reference frame with the $d$-axis aligned to the supply voltage space vector. In this frame, the ac-side voltage equation is formulated by

$$\begin{aligned} L\frac{\mathrm{d}i_d}{\mathrm{d}t} - \omega L i_q &= E - u_d \\ L\frac{\mathrm{d}i_q}{\mathrm{d}t} + \omega L i_d &= -u_q \end{aligned} \tag{4}$$

and the power balance equation is given by

$$p_S = \frac{3}{2}E i_d = u_{dc} i_S \tag{5}$$

where $E$ is the length of the supply voltage vector.

On the load side, the H-bridge inverter can be described by its power balance equation

$$p_L = u_o i_o = u_{dc} i_L \tag{6}$$

where $p_L$ is the load side power and the meaning of other symbols are shown in Fig. 1.

In steady state, the fundamental component of the output voltage and current of the inverter can be described by (7)(8)

$$u_o = U_{om}\cos(\omega_o t) \tag{7}$$

$$i_o = I_{om}\cos(\omega_o t - \varphi) \tag{8}$$

where $\omega_o$ is output voltage frequency, $\varphi$ the motor power factor angle, $U_{om}$ the amplitude of the output voltage and $I_{om}$ the amplitude of the output current. Thus, the inverter output power in steady state is given by

$$p_L = \frac{U_{om} I_{om}}{2}[\cos\varphi + \cos(2\omega_o t - \varphi)] \tag{9}$$

Note that in addition to the dc component, the output power also contains the ac component, which means the output power pulsates even in steady state.

The difference between $p_S$ and $p_L$ is compensated by dc-link capacitor as

$$p_S - p_L = p_C = \frac{\mathrm{d}}{\mathrm{d}t}(\frac{1}{2}C u_{dc}^2) \tag{10}$$

As seen from (10), if the pulsating power $p_L$ is not balanced by $p_S$, it will flow into the dc-link capacitor and cause fluctuation of the dc-link voltage. In steady state, if the dc-link voltage is controlled constant, then $p_L = p_S$. Combining this with (5) and (9) leads to

$$i_d = \frac{U_{om} I_{om}}{3E}[\cos\varphi + \cos(2\omega_o t - \varphi)] \tag{11}$$

Equation (11) represents the required steady state active ($d$-axis) current of the PWM rectifier to balance the H-bridge load power. As observed, the current is composed of a dc component and an ac component of frequency $2\omega_o$ due to the instantaneous power pulsating of the load.

## III. PROPOSED CONTROL STRATEGY

The conventional PI-based VOC strategy is shown in Fig. 3, which is suitable for three-phase load (constant power), but it is not suitable for handling the ac component in (11).

First, the dc-link voltage loop PI regulator cannot give the $2\omega_o$ ac current reference unless the same ac frequency component exists in the voltage error, which means a voltage ripple with frequency of $2\omega_o$ will be present on the dc-link. This is due to the control limitation of the PI regulator with ac component.

Second, the PI-based active current control loop cannot track the current reference in (11) without steady state error. There will be errors in amplitude and phase for the $2\omega_o$ frequency current, and it will get worse when the device's switching frequency gets lower and $\omega_o$ gets higher, which is not uncommon in high-power converters.

To overcome the shortage and limitations of the PI regulator, a PIR-based regulator is proposed in this paper. The R regulator [6-7] has the transfer function in form of (12)

$$G_{res}(s) = \frac{2K_r s}{s^2 + \omega_{res}^2} \tag{12}$$

which has infinite gain at the resonant frequency $\omega_{res}$ and nearly zero gain at other frequencies. This frequency selection property makes it suitable to regulate the ac frequency component in (11). The proposed control strategy is depicted in Fig. 4, where the PI regulator is replaced by the PIR regulator and the transfer function is given by

$$G_{xPIR}(s) = K_{xp} + \frac{K_{xi}}{s} + \frac{2K_{xr} s}{s^2 + (2\omega_o)^2} \tag{13}$$

where $x$=$i$ for current controller; $x$=$u$ for voltage controller, and $\omega_o$ is the H-bridge output voltage frequency.

In the current control loop, the I regulator is used to eliminate the current dc component steady state error, and the R regulator for the $2\omega_o$ component, while the P regulator is responsible for the transient response.

In the dc-link voltage control loop, the PIR regulator is employed as well. The dc-link voltage is related to $p_L$ via a non-linear differential equation, and can be linearized around the reference value as long as the dc-link voltage is well controlled:

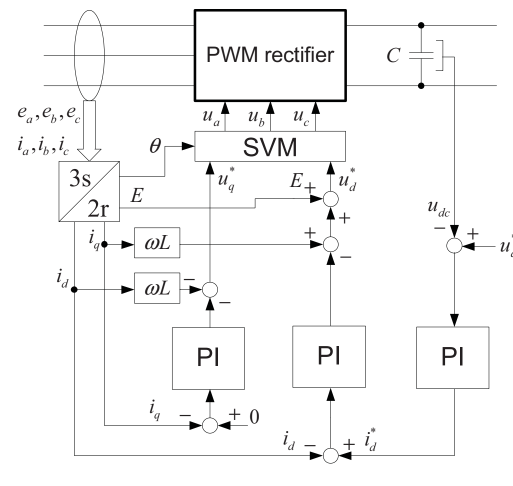


Fig.3 Conventional control strategy for PWM rectifier

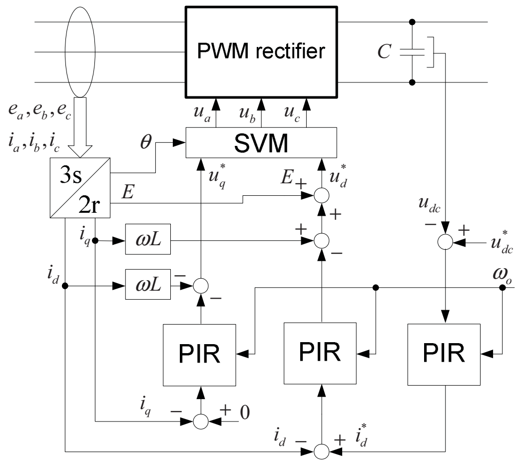


Fig. 4 Proposed control strategy

$$\Delta p_S - \Delta p_L = \Delta p_C = \frac{1}{2} C u_{dc}^* \frac{\mathrm{d}\Delta u_{dc}}{\mathrm{d}t} \tag{14}$$

As seen, around the operating point $u_{dc}^*$, $\Delta u_{dc}$ will have $2\omega_o$ frequency component which is introduced by the disturbance $\Delta p_L$. The R regulator is used specifically to solve this problem. The resonant frequency is tuned to $2\omega_o$ to enable the dc-link voltage track the reference of this frequency. Since the reference is a constant and has no $2\omega_o$ frequency component, the fluctuation of this frequency is effectively eliminated. According to (5), $i_d$ is proportional to $p_S$ with factor $3E/2$. Therefore, the PIR regulator is able to produce the correct active current reference $i_d^*$ in (11).

## IV. Simulation and Experimental Results

Simulation were carried out to verify the previous analysis about the pulsating power of the load and its influence on the dc-link voltage. The performance of the proposed control strategy was also tested with the resonant frequency of the PIR regulator adjusted dynamically according to the change of the output frequency of the H-bridge. After the PWM rectifier started, the H-bridge inverter began to work with open-loop constant volts/hertz control. The output frequency was 10Hz at 0.05s and gradually increased to 45Hz in about 0.2s. The main parameters are listed in Table I. A small dc-link capacitance of 110μF is selected to make the dc-link voltage variation more obvious.

Fig. 5 shows the simulation results of the conventional control strategy in terms of dc-link voltage $u_{dc}$, active ($d$-axis) current reference $i_d^*$, active current $i_d$ and the current error $i_d^* - i_d$. As observed, the power pulsation is getting larger with the increase of the output voltage and frequency, causing the fluctuation of the dc-link voltage. The error between $i_d^*$ and $i_d$ also increases since the amplitude and phase response of the PI-based current control loop is different depending on the frequency. This simulation verifies the previous analysis.

Fig. 6 shows the simulation results of the proposed control strategy, where the dc-link voltage fluctuation is effectively suppressed and $i_d$ tracks its reference $i_d^*$ well. The dynamic changing of resonant frequency does not influence the performance of the R regulator. The proposed control strategy shows a better performance of the dc-link voltage and the *d*-axis current than the conventional one in Fig. 5.

TABLE I

PARAMETERS OF SIMULATION AND EXPERIMENTAL SYSTEM

| PWM rectifier | |
|---|---|
| Grid line voltage | 380V |
| Ac inductor | 5.1mH |
| Switching frequency | 2kHz |
| Dc-link | |
| Capacitor | 110μF |
| Voltage reference | 620V |
| H-bridge inverter load | |
| Inductor | 20mH |
| Resistor | 51Ω |
| Switching frequency | 2kHz |

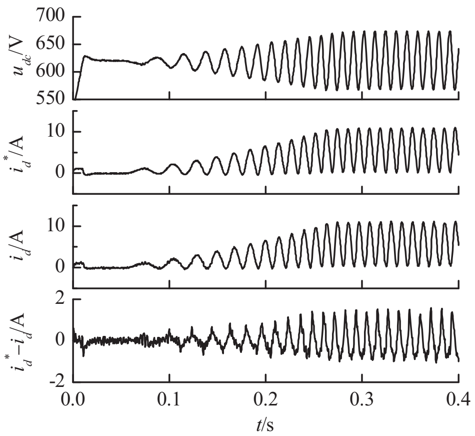

Fig. 5 Simulation results of conventional control strategy

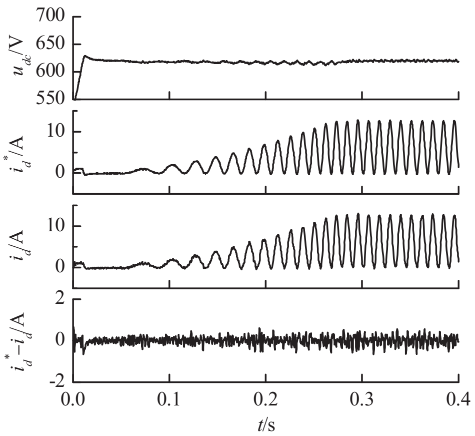

Fig. 6 Simulation results of proposed control strategy

A 7.5kW experimental prototype of one power cell is also built up in the laboratory to verify the proposed control strategy as shown in Fig. 7. Two intelligent power modules (IPMs) PM50RVA120 are used as the power devices for the rectifier and the inverter. The system parameters are the same as in the simulation and the controller is DSP TMS320LF2812.

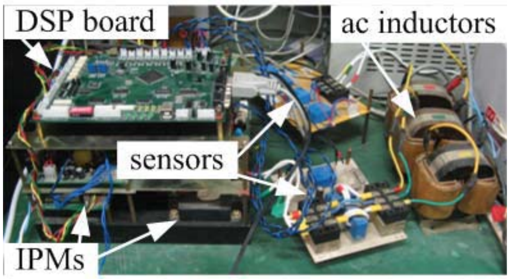


Fig. 7 Experimental prototype of one power cell

The experimental results are depicted in Fig. 8 including dc-link voltage, grid line voltage, input phase current and output current. The output frequency of the H-bridge is 45Hz. The system is first controlled by PI regulator and then the R regulator is enabled at 0.1s. As shown, the peak-peak value of the dc-link voltage ripple is reduced significantly from 128V to 28V due to the R regulator control after a fast transient.

Meanwhile, the input current of the PWM rectifier in each power cell contains lots of harmonics, since the *d*-axis current reference is not a dc value. However, when all the power cells are connected together in the configuration as shown in Fig. 2, the current harmonics will not flow into the power grid, because the harmonic currents in different phases will be cancelled by each other [5].

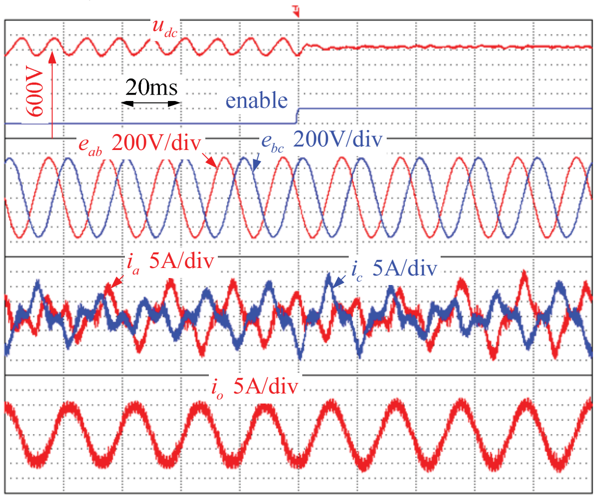


Fig.8 Experimental results

## V. CONCLUSION

In this paper, the three-phase PWM rectifier feeding an H-bridge load which is used as power cells of regenerative cascaded H-bridge multilevel converters has been analyzed and studied. The power pulsation due to H-bridge load may cause large dc-link voltage ripples. A PIR-based controller is proposed for the PWM rectifier in both the dc-link voltage loop and the ac current loop based on the system mathematical model. The dc-link voltage can be maintained constant with the proposed method which makes it possible to reduce the value of the dc-link capacitor significantly. The only information required about the load in the control is the H-bridge output frequency. Therefore the control is easy to implement and the system structure is simplified. Simulation and experimental results validate the performance of the proposed control strategy, and comparison with the conventional method is also included in the paper.

## REFERENCES


[1] M. Malinowski, M. P. Kazmierkowski, S. Hansen, F. Blaabjerg and G. D. Marques, "Virtual-flux-based direct power control of three-phase PWM rectifiers," *IEEE Trans. Ind. Applicat.*, vol.37, pp. 1019-1027, 2001.

[2] V. Blasko and V. Kaura, "A new mathematical model and control of a three-phase AC-DC voltage source converter," *IEEE Trans. Power Electron.,* vol.12, pp. 116-123, 1997.

[3] R. Wu, S. B. Dewan and G. R. Slemon, "A PWM AC-to-DC converter with fixed switching frequency," *IEEE Trans. Ind. Applicat.,* vol.26, pp. 880-885, 1990.

[4] P. W. Hammond, "A new approach to enhance power quality for medium voltage AC drives," *IEEE Trans. Ind. Applicat.* vol.33, pp. 202-208, 1997.

[5] M. A. Perez, J. R. Espinoza, J. R. Rodriguez and P. Lezana, "Regenerative medium-voltage AC drive based on a multicell arrangement with reduced energy storage requirements," *IEEE Trans. Ind. Electron.* vol.52, pp. 171-180, 2005.

[6] D. N. Zmood, D. G. Holmes and G. H. Bode, "Frequency-domain analysis of three-phase linear current regulators," *IEEE Trans. Ind. Applicat.* vol.37, pp. 601-610, 2001.

[7] D. N. Zmood and D. G. Holmes, "Stationary frame current regulation of PWM inverters with zero steady state error," in *Proc. IEEE PESC'99*, pp. 1185-1190.