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A Space Vector Modulation Based Three-level PWM Rectifier under Simple Sliding Mode Control Strategy

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ABSTRACT

In this paper, a voltage oriented control strategy for three-level PWM rectifier based on Sliding Mode Control (SMC) is introduced in order to obtain fast and accurate response of dc-bus voltage. To verify the validity of the analysis and the feasibility of the proposed control method a set of simulation tests have been conducted using Matlab/Simulink. The simulation results show that compared to the conventional PI controller, the SMC can reduce drastically the three-level rectifier's voltage fluctuation and improve the dynamic response of dc-bus significantly.

Keywords: Three-level; PWM Rectifier; Voltage Oriented Control; Sliding Mode Control; Unbalanced Input Voltage

1. Introduction

Recent development of high-power and high switching frequency power electronic devices and their large-scale application have led to the study of converter systems performing near unity power factor and digital implementation. This new wave of research has paved the way to eliminate the power grid pollution and provide green power requirements. Thus, research interest in three-phase pulse with modulation (PWM) rectifiers has grown rapidly due to their numerous advantages such as bidirectional power flow, low harmonic distortion of source current, near unity power factor, and adjustable dc-bus voltage [1-14]. Moreover, the three-level neutral point clamped (NPC) converter presents more advantages over the conventional two-level converter in high power applications, such as lower voltage stress of semiconductors, smoother waveform, less distortion and less switching frequency stresses [15,16]. The PWM rectifier based on three-level NPC technique is an attractive method suitable for high power applications since it provides the merits of both PWM rectifier and three-level converter. The most prevalent control scheme for PWM rectifier is the voltage oriented control (VOC) [17], which is implemented by PI controllers for inner current control and outer voltage control loops. The outer voltage loop is traditionally implemented by fixed-gain proportional-integral (PI) or proportional-integral-derivative (PID) controller. However, the design of such a controller depends on the precise system mathematical model used which is difficult to develop.

Recently, much attention has been given to a sliding mode controller (SMC) in order to overcome the above drawbacks. (SMC) is a discontinuous system, and the control character can force the system to move in tiny extent and in high frequency according to the specified state track under certain conditions. Because of the merits of high speed response, insensitivity to the variable parameters, and ease of implementation, the SMC has been widely used in the non-linear system.

In this paper, a simple control strategy for three-level PWM rectifier with voltage oriented control to improve the system's robustness and dynamic response of the dc-bus voltage is proposed. The sliding mode control is used in the outer voltage loop. In order to improve the dynamic performances of the source current loop, the anti-windup IP controller of inner current controller is used instead of the conventional PI controller [18-21]. Simulation results show that compared to the conventional PI controller, the SMC can reduce the three-level rectifier's voltage fluctuation and improve the dynamic response of the dc-bus significantly.

2. Topology and Mathematical Model of Three Level PWM Rectifier

The input of the rectifier is connected to the power network, and the output is in the dc side. The main objective of the control strategy of the rectifier is to make the input current follow a sine wave and the output voltage to be a controllable dc voltage.

The topology of the three-level PWM rectifier is shown in **Figure 1**, [22-24]. L_s and R_s are the equivalent inductance and resistance of the three phase reactor inserted between the grid source and the rectifier, C_{dc1}

and C_{dc2} are the dc-bus capacitances, V_{dc1} and V_{dc2} are voltages of the two capacitors, V_{dc} is the sum of V_{dc1} and V_{dc2}. e_i, i_i and v_i $i = a,b,c$ are the three-level grid voltage, grid current and ac-side voltage of the rectifier, respectively. Assuming that s_{ip}, s_{id}, s_{in} ($i = a,b,c$) are the switching variables of the three level PWM rectifier when the three phases of power source voltages (e_a, e_b, e_c) are sinusoidal and symmetrical. Then, they can be defined according to different switch states of the four switches in each phase as:

$s_{ip} = 1$, $s_{io} = 0$, $s_{in} = 0$, when S_{1i}, S_{2i} on and S_{3i}, S_{4i} off.

$s_{ip} = 0$, $s_{i0} = 1$, $s_{in} = 0$, when S_{2i}, S_{3i} on and S_{1i}, S_{4i} off.

$s_{ip} = 0$, $s_{i0} = 0$, $s_{in} = 1$, when S_{3i}, S_{4i} on and S_{1i}, S_{2i} off.

Assuming that the three phase source voltages are balanced, sinusoidal and symmetrical, the phase angle of voltage e_a is θ, E denotes the RMS value of the source phase voltage, thus

$$\begin{cases} e_a = \sqrt{2}E\cos\theta \\ e_b = \sqrt{2}E\cos(\theta - 2\pi/3) \\ e_c = \sqrt{2}E\cos(\theta + 2\pi/3) \end{cases} \tag{1}$$

The transformation equation from abc coordinates to static $\alpha-\beta$ coordinates and then to synchronous rotating $d-q$ coordinates are

$$\begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1/\sqrt{2} & -1/\sqrt{2} \\ 0 & -\sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \tag{2}$$

$$\begin{bmatrix} x_d \\ x_q \end{bmatrix} = \begin{bmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{bmatrix} \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} \tag{3}$$

According to "equation (1)",

$$e_a + e_b + e_c = 0 \tag{4}$$

In the three- phase inverter- wire system

$$i_a + i_b + i_c = 0 \tag{5}$$

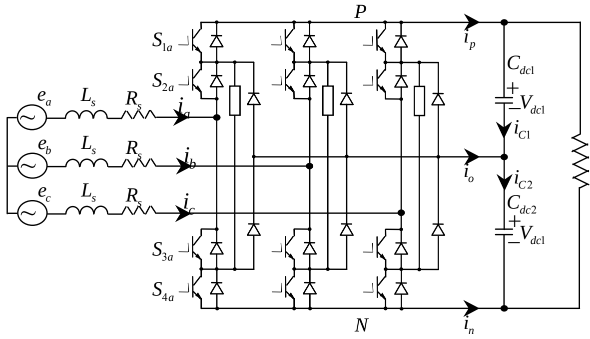


Figure 1. Topology of three-level PWM rectifier.

Therefore, "equation 2" is simplified to "equation 6" in order to reduce the number of current sensors and improve the quality of voltage.

$$\begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} = \sqrt{\frac{3}{2}} \begin{bmatrix} 1 & 0 \\ 1/\sqrt{3} & 2/3 \end{bmatrix} \begin{bmatrix} x_a \\ x_b \end{bmatrix} \tag{6}$$

After some tedious mathematical processes on the above equations, the mathematical model of the system in static abc coordinates is as follows:

$$Z\dot{x} = Ax + Be \tag{7}$$

where

$$Z = diag\begin{bmatrix} L_s & L_s & L_s & C_{dc1} & C_{dc2} \end{bmatrix}$$

$$x = \begin{bmatrix} i_a & i_b & i_c & v_{dc1} & v_{dc2} \end{bmatrix}^T$$

$$B = diag\begin{bmatrix} 1 & 1 & 1 & -1 & -1 \end{bmatrix}$$

$$e = \begin{bmatrix} e_a & e_b & e_c & i_2 & i_2 \end{bmatrix}^T$$

$$A = \begin{bmatrix} -R_s & 0 & 0 & -(s_{ap} - s_p^{'}) & (s_{an} - s_n^{'}) \\ 0 & -R_s & 0 & -(s_{bp} - s_p^{'}) & (s_{bn} - s_n^{'}) \\ 0 & 0 & -R_s & -(s_{cp} - s_p^{'}) & (s_{cn} - s_n^{'}) \\ s_{ap} & s_{bp} & s_{cp} & 0 & 0 \\ -s_{an} & -s_{bn} & -s_{cn} & 0 & 0 \end{bmatrix}$$

and

$$s_p^{'} = \frac{s_{ap} + s_{bp} + s_{cp}}{3}, \quad s_n^{'} = \frac{s_{an} + s_{bn} + s_{cn}}{3}$$

The physical meaning of the mathematical model in abc coordinates is pellucid, but variable parameters of ac reactors are unstable which is not suitable for the design of control system, so the mathematical model in the rotating $d-q$ coordinates is:

$$Z^{'}\dot{x} = A^{'}x + B^{'}e \tag{8}$$

where

$$Z^{'} = diag\begin{bmatrix} L_s & L_s & C_{dc1} & C_{dc2} \end{bmatrix}$$

$$x = \begin{bmatrix} i_d & i_q & v_{dc1} & v_{dc2} \end{bmatrix}^T$$

$$B = diag\begin{bmatrix} 1 & 1 & -1 & -1 \end{bmatrix}$$

$$e = \begin{bmatrix} e_d & e_q & i_L & i_L \end{bmatrix}^T$$

$$A^{'} = \begin{bmatrix} -R_s & \omega L_s & -s_{dp} & s_{dn} \\ -\omega L_s & -R_s & -s_{qp} & s_{qn} \\ s_{dp} & s_{qp} & 0 & 0 \\ -s_{dn} & -s_{qn} & 0 & 0 \end{bmatrix}$$

If we suppose that v_d and v_q are the voltages of

d-axis in the $d-q$ coordinates, it can be shown that:

$$\begin{cases} v_d = e_d + \omega L_s i_q - \left(L_s s + R_s\right) i_d \\ v_q = e_q - \omega L_s i_d - \left(L_s s + R_s\right) i_q \end{cases} \tag{9}$$

where, s is the arithmetic operator of differential coefficient.

Considering $C_{dc1} = C_{dc2} = C_d$, then

$$C_d \frac{dv_{dc}}{dt} = \left(s_{dp} - s_{dn}\right) i_d + \left(s_{qp} - s_{qn}\right) i_q - 2\frac{v_{dc}}{R_L} \tag{10}$$

From the aforementioned model, the equivalent circuit of the three-level PWM rectifier in the $d-q$ coordinates can be obtained as shown in **Figure 2**.

3. Control Strategy for the Three-level PWM Rectifier Based on SVPWM

Similarly to the two-level PWM rectifier [26-34], the control target of the three-level PWM rectifier is to make the dc output voltage v_{dc} follow its reference value v_{dc}^*, while keeping the input currents (i_a, i_b, i_c) approximately sinusoidal and in phase with the corresponding grid voltages (e_a, e_b, e_c).

Furthermore, to ensure the reliability of the system [35] the two special requirements of three-level PWM rectifier: balance of neutral-point voltage and avoidance of excessive voltage jump in phase and line-to-line voltages must be satisfied.

Voltage oriented control (VOC) which is the classical and most popular control strategy for the three-level PWM rectifiers [32] provides excellent steady-state performance, acceptable dynamic performance and constant switching frequency for the rectifier compared with other strategies. The block diagram scheme of VOC strategy based on sliding mode control for the three level PWM rectifier is illustrated in **Figure 3**.

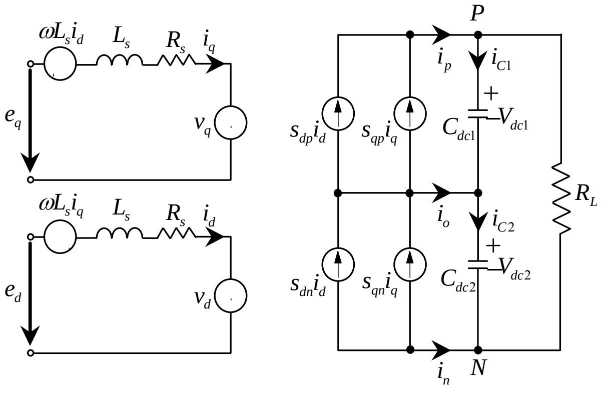
There are three control loops in the VOC strategy. The error between the reference dc-bus voltage v_{dc}^* and the sampled dc-bus voltage v_{dc} is processed by SMC, which produces the reference active current i_d^*. As in the inner loops, d-axis currents loop and q-axis current loop use anti-windup IP controllers to make the actual currents (i_d and i_q) track their reference values (i_d^* and i_q^*). Generally, and in order to achieve near unity power factor condition, the controlled value of the q-axis current is set to zero.

Then, the errors are processed in two conventional anti-windup IP controllers to produce the output signals of v_d^* and v_q^*, after coordinates transformation, v_α^* and v_β^* which can be obtained and used to produce switching signals S_a, S_b and S_c by three-level space vector pulse with modulation (SVPWM).

As shown in **Figure 3**, in the three-level PWM grid phase voltages (e_a, e_b), two grid phase currents (i_α, i_β) and two dc-link voltages (v_{dc1}, v_{dc2}) are sampled.

3.1. Sliding Mode Control Design of the Output Voltage Loop

The main goal of the voltage control of the rectifier is keep the output voltage constant, ripple of the voltage small, and overshoot small and the regulation course short during transient conditions.



(a) (b)

Figure 2. Equivalent circuit of the three-level PWM rectifier in *d* - *q* coordinates.

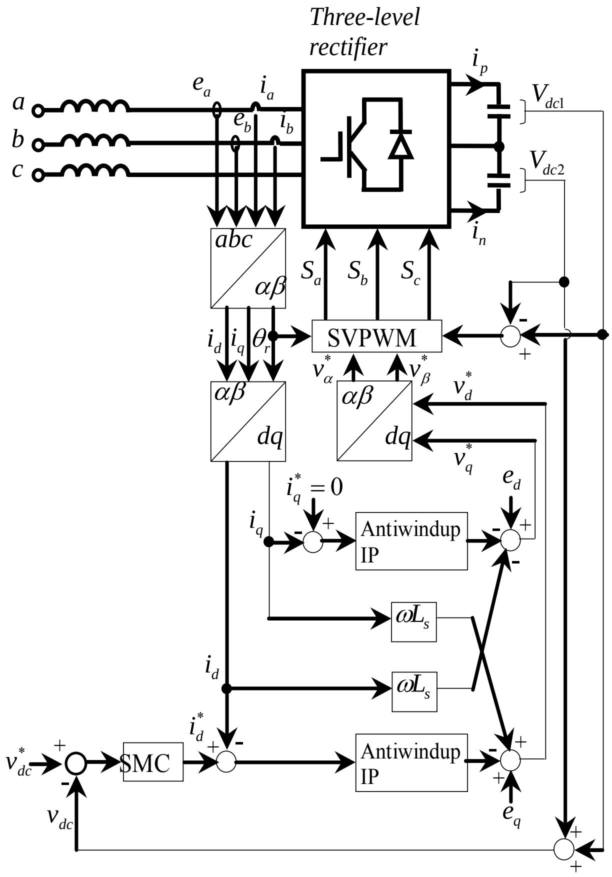


Figure 3. Three-level rectifier control system.

There are two external variables (v_{dc} and i_q for the three-level PWM rectifier, where v_{dc} is determined by s_d, and i_q is controlled by s_q. Considering v_{dc} and i_q as contestable output variables, standard state space can be obtained as

$$\frac{d}{dt}\begin{pmatrix} i_q \\ v_{dc} \end{pmatrix} = \begin{pmatrix} -\omega i_d - \frac{R_s}{L_s} i_q - \frac{s_q}{2L_s} v_{dc} + \frac{1}{L_s} e_q \\ \frac{s_d i_d + s_q i_q + 2i_L}{C_d} \end{pmatrix} \tag{11}$$

Substituting the error between reference and fact variable into "equation (10)", then

$$\frac{d}{dt}\begin{pmatrix} e_{iq} \\ e_{vdc} \end{pmatrix} = \begin{pmatrix} y_1(E) - x_1(t) + z_1(E) s_q v_{dc} \\ e_\varphi \end{pmatrix} \tag{12}$$

where,

$$e_{iq} = i_{iq_{ref}} - i_q,\ e_{v_{dc}} = v_{dc_{ref}} - v_{dc},$$

and

$$e_\varphi = \varphi_{ref} - \varphi$$

So, it can be concluded that selecting the two following sliding surfaces, the stability and robustness of the system can be achieved:

$$s_1 = k_{e_{iq}} e_{iq} = k_{e_{iq}} \left(i_{q_{ref}} - i_q \right) = 0 \tag{13}$$

$$s_2 = k_1 e_{v_{dc}} + k_2 \frac{de_{v_{dc}}}{dt} = e_{v_{dc}} + \beta \frac{de_{v_{dc}}}{dt} = 0 \tag{14}$$

By combining the above equations, then s_2 can be rewritten as

$$\begin{aligned} s_2 &= \left(v_{dcref} - v_{dc} \right) + \beta \frac{dv_{dc}}{dt} - \beta \left(\frac{s_d i_d + s_q i_q}{C_d} - \frac{2i_L}{C_d} \right) \\ &= \left[\left(v_{dcref} - v_{dc} \right) + \beta \frac{dv_{dc}}{dt} - \frac{\beta}{C_d} \left(2i_L - s_q i_q \right) \right] \times \frac{C_d}{\beta s_d} - i_d \end{aligned} \tag{15}$$

In $d-q$ coordinates, $e_d = \sqrt{3} u_{RMS}$, $e_q = 0$, in the ideal sliding mode state, s_q can be calculated and simplified as

$$s_q \approx -\frac{2L_s \omega i_d}{v_{dc}} \tag{16}$$

Similarly, the output voltage v_{dc} will follow the reference v_{dc}^* accurately, and based on the principle of power balance, s_d is obtained as:

$$s_d \approx \frac{e_d - R_s i_d}{v_{dc}} \approx \frac{\sqrt{3} u_{RMS} - R_s i_d}{v_{dc}} \tag{17}$$

By substituting "equation (16)" and "equation (17)" into "equation (15)", "equation (18)" can be deduced as

$$\begin{aligned} s_2 = &\left[\left(v_{dcref} - v_{dc} \right) + + \beta \frac{dv_{dc}}{dt} - \frac{2\beta}{C_d} \left(i_L + \frac{L_s \omega i_d i_q}{v_{dc}} \right) \right] \\ &\times \frac{C_d v_{dc}}{\beta \left(\sqrt{3} u_{RMS} \right) - R_s i_d} - i_d = 0 \end{aligned} \tag{18}$$

Therefore, s_d and s_q will not be relevant to the choice of sliding mode surface, and the sliding mode surface can be obtained as

$$s_1 = k_{e_{iq}} e_{iq} = k_{e_{iq}} \left(i_{q_{ref}} - i_q \right) = 0 \tag{19}$$

$$s_2 = i_{d_{ref}} - i_d = 0 \tag{20}$$

From "equation (18)" and "equation (20)", the control rule for the outer voltage loop can be described as

$$\begin{aligned} i_{d\,ref} = &\left[\left(v_{dcref} - v_{dc} \right) + + \beta \frac{dv_{dc}}{dt} - \frac{2\beta}{C_d} i_L \right] \\ &\times \frac{C_d v_{dc}}{\beta \left(\sqrt{3} u_{RMS} \right) - R_s i_d} \end{aligned} \tag{21}$$

The block scheme of the VOC strategy based on SMC for the three-level PWM rectifier is shown in **Figure 3**. The error between reference dc-bus voltage v_{dc}^* and the sampled dc-bus voltage v_{dc} is processed by SMC, which produces the reference active current i_d^*. i_d^* and i_q^*, (for unity power factor control, $i_q^* = 0$) are compared with the measured grid current, i_d and i_q, respectively. Then, the errors are processed in two anti-windup IP controllers to produce the output signals of v_q^* and v_d^*, after coordinates transformation, v_α^* and v_β^* that can be obtained and used to produce switching signals S_a, S_b and S_c by the three- level space vector pulse width modulation (SVPWM).

3.2. Three-level Space Voltage Vector Modulation Algorithm

There are 27 output voltage vectors in the three-level VSI as shown in **Figure 4**. In **Figure 5**, suppose the desired reference voltage vector lies in the triangle B which is in the first 60° sector (sector I).

Then, the function time of each output voltage vector should be calculated first as well as the corresponding time for the power devices to turn on or turn off.

The desired output voltage vector consists of V_1, V_3 and V_4 by the adjacent three vector compounding principle. Based on the volt-second balancing principle [36]:

$$\begin{aligned} T_a &= 2T_s \left[1 - k \sin\left(\pi/3 + \theta \right) \right] \\ T_b &= T_s \left[2k \sin\left(\pi/3 - \theta \right) - 1 \right] \\ T_c &= 2kT_s \sin\theta \end{aligned} \tag{23}$$

where, $k = V_{ref} \big/ \sqrt{3}$ is the modulation depth, T_s is the system sampling control cycle, V_{ref} and θ is the amplitude and angle of the reference voltage vector V_{ref}.

In the same way, the function time of the adjacent three-vectors could be fixed when it lies in the triangle A, C and D. The vector function time of the other five vectors could be deduced in a symmetrical manner.

According to the function time of each vector and the centro-symmetric vector sending sequence, the three phase output vectors sequential chart could be fixed when the reference vector V_{ref} lies in the triangle A, B, C and D in sector I, which also gets the space voltage vector modulation mode.

There are some similar SVPWM modes when the reference vector lies in other vectors. According to the SVPWM mode and the function time of each vector corresponding to each sector, the power devices driven signal of the three phase arms could be obtained to control the three-level inverter in SVPWM mode.

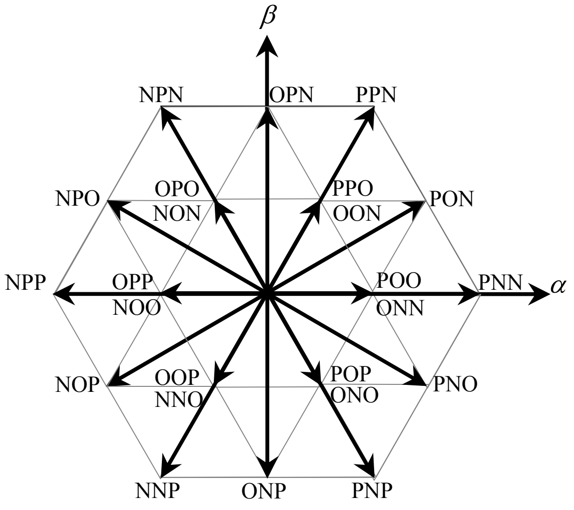


Figure 4. Space voltage vectors in three-level rectifier.

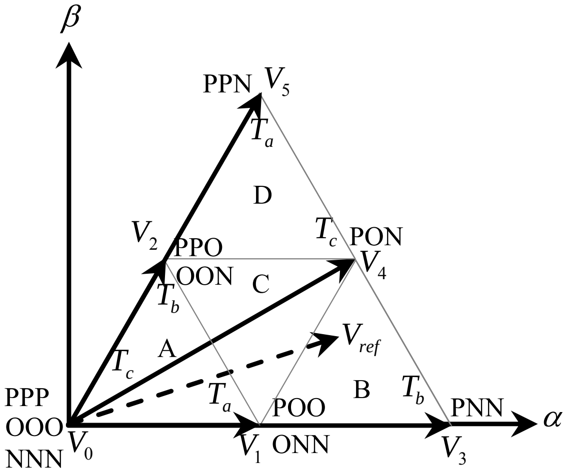


Figure 5. Synthesized reference vector in the first 60° sector.

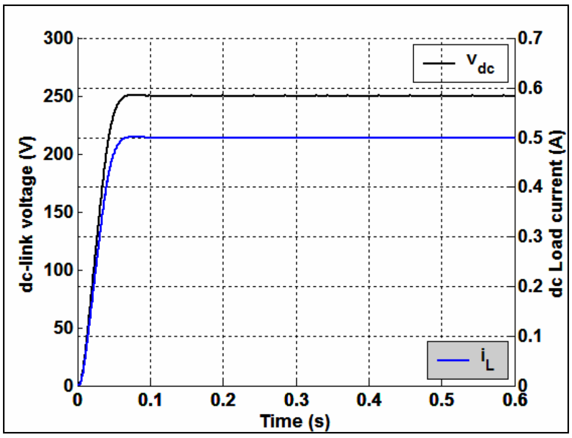
4. Simulation Results

To validate the proposed control scheme proposed in this paper, a series of simulation tests have been conducted under Matlab/Simulink environment. The main parameters of the simulation system are given in **Table 1**.

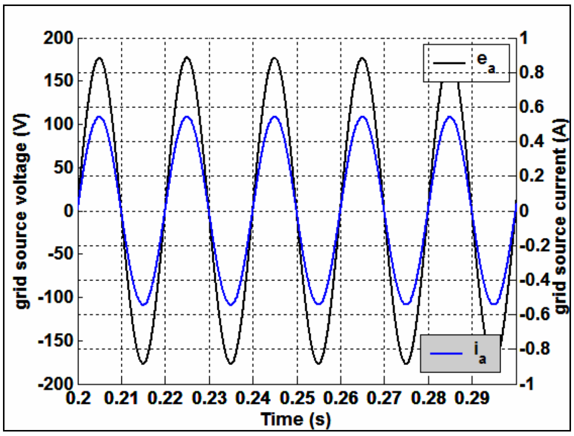
Figure 6(a) shows the DC voltage and current waveforms where the DC output voltage reaches the given stable value (250 V) of the voltage in a short time. **Figure 6(b)** shows the grid phase voltage (e_a) and current (i_a) waveforms. It can be seen that the grid current is in phase with the grid voltage, and the power factor is higher than 0.997.

Table 1. Rectifier parameter.

The input phase voltage	125 V
The Power source frequency	50 Hz
The input inductance	37 mH
The input resistance	0,3 Ω
DC-bus capacitor	1100μF
DC-bus voltage	250 V



(a) Output voltage and current



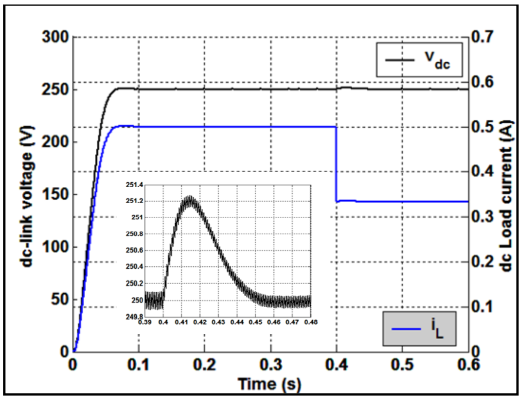
(b) Grid source side voltage and current

Figure 6. Simulation results of system.

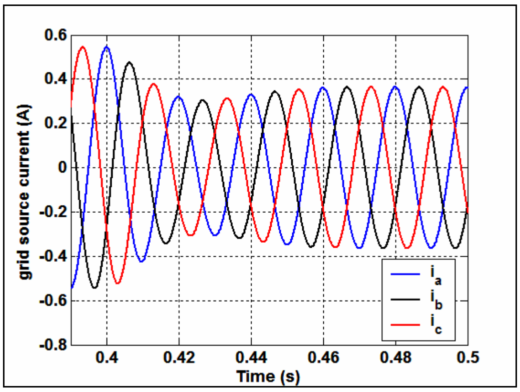
Figure 7 shows the simulation results when the load changes from 500 Ω to 750 Ω at t = 0.4 s. **Figure 7(a)** shows the output DC voltage and current waveforms when the load and input voltage fluctuates, the system can adjust to the desired value of the voltage in a short period of time. **Figure 7(b)** shows the waveforms of the grid source side current, the current always maintains unity power factor in the dynamic process.

Moreover, in view of the actual operating conditions, there are more or less fluctuations of the three-phase input voltage especially three-phase input voltage unbalance in the operation of the circumstances. **Figure 8(a)** shows unbalanced three-phase input voltage in the system, single-phase unbalance is up to 20%, DC output voltage fluctuations is less than 0.2 V as shown in **Figure 8(b)**.

To validate the superiority of SMC over conventional PI controller, comparative simulations are conducted and the results are shown in **Figure 9**. **Figure 9(a)** shows the waveform of the dc-bus voltage under conventional PI controller while **Figure 9(b)** is that of SMC. It can be seen clearly that the overshoots of the dc-bus voltage for the rectifier with PI controller is much higher than that with SMC, and the dynamic performance of the system is improved significantly.

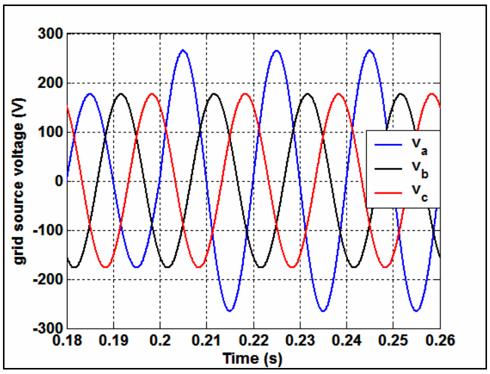


(a) Output voltage and current

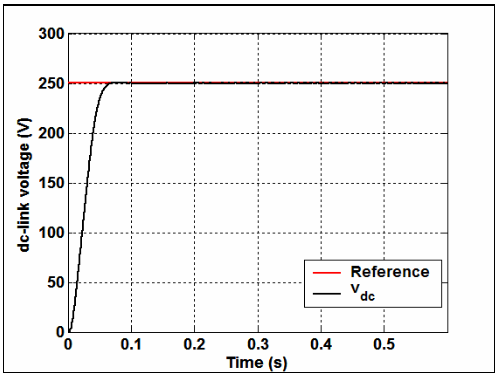


(b) Grid source side current

Figure 7. Simulation waveforms at load changes.

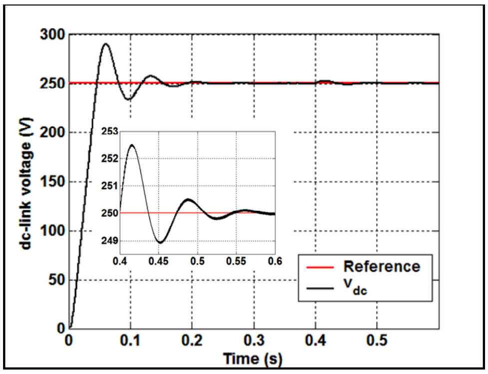


(a) Three-phase input voltage fluctuations

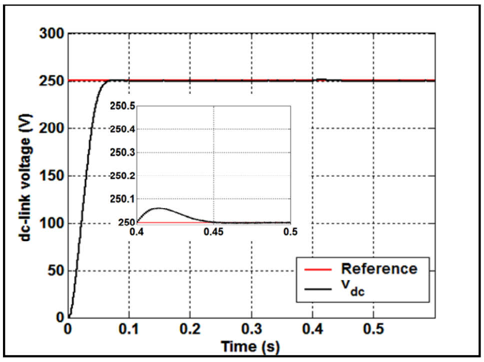


(b) DC output voltage waveform

Figure 8. Voltage unbalance at the DC output waveforms.



(a) dc-bus voltage with conventional PI



(b) dc-bus voltage with SMC

Figure 9. DC output voltage waveforms.

5. Conclusions

The problem of the voltage control system of the three-level rectifier is thoroughly analyzed and presented in this paper. Through the study on the voltage equation of the rectifier, the nonlinear characteristic of the voltage control is carefully discussed and detailed based on a new strategy which uses sliding mode control (SMC). The proposed control strategy is adopted for the dc bus voltage control to obtain better dynamic performance based on the presented mathematical model. Simulation results which are included in this paper, indicate that the unity power factor is achieved and the proposed scheme exhibits better dynamic and steady state performance than conventional controller.

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