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STUDY ON SLIDING MODE VIRTUAL FLUX ORIENTED CONTROL FOR THREE-PHASE PWM RECTIFIERS

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Key words: PWM rectifier, Sliding mode control (SMC), Virtual flux oriented control (FOC), Proportional integral (PI) controller.

This paper presents the design of the sliding mode control technique for three-phase boost-type PWM rectifier based on virtual flux vector orientation. A sliding-mode control algorithm on synchronous rotating reference frame is employed for each current and dc voltage control loops. The Lyapunov direct method is used to ensure the reaching and sustaining of the sliding mode. The logic control signals (*Sa, Sb, Sc*) are calculated by using the space vector modulation (SVM) technique. To alleviate chattering phenomenon without compromising system robustness, a tangent hyperbolic function is used in the dc voltage loop instead of a conventionally sign function. The proposed sliding mode control system has been simulated using MATLAB SIMULINK Software and compared with traditional PI controller. The comparison shows that the sliding mode control scheme is more robust against load changes and power supply variations. Furthermore, it confers better dynamic performance.

1. INTRODUCTION

The three-phase boost-type PWM rectifier is used increasingly in a wide diversity of applications such as variable-speed drives, wind power generation and ac–dc power supplies for telecommunications equipment [1]. They are capable of viable advantages such as: bi-directional power flow, near unity power factor operation, input currents with low harmonic content, adjustment and stabilization of dc-link voltage, reduced dc filter capacitor size [2, 3].

Research interest in this type of PWM converter has grown rapidly over the last few years and various control strategies have been proposed in recent works. We can distinguish, at first, the voltage oriented control (VOC) [4] and virtual flux oriented control (VFOC) [5] which are close to vector-controlled ac motors. These methods adopt double closed loop PI controllers at synchronously rotating frame; an inner control loop for the currents and an outer one for the dc-link voltage. Two other methods, developed analogously with the well-known direct torque control (DTC) used for adjustable speed drives, are direct power control (DPC) [6, 7] and virtual flux based direct power control (VF-DPC) [8]. They are based on the instantaneous active and reactive power control loops.

It is obvious that in order to obtain better ac supply power quality and high performance, it is convenient to directly control the supply currents via internal control loops. So virtual flux oriented control (VFOC) seems to be the simplest technique used to forcing the supply current to follow its reference. Usually, the PI controller can offer good dynamic and steady state performances, whereas the PI regulator parameters are dependent on the load, system parameters and on the operating point. In addition this controller is not the most appropriate for the three phase PWM rectifier, which is a nonlinear and unstable system. In contrast with linear controllers, sliding mode control (SMC) is particularly interesting because of the merits of high speed response, insensitivity to parameters variations, disturbance rejection, and simple implementation [9, 10]. Thanks to these advantages, the SMC has being widely used in the non linear system [11, 12]. However, in spite of the advantages that can provide a SMC, its implementation may be obstructed by an undesirable harmful phenomenon which is known as “chattering”. To alleviate this drawback, numerous techniques have been proposed in the literature [13].

Some authors have treated the sliding mode control of these PWM rectifiers [14–16]. In ref. [14], a SMC algorithm on synchronous rotating reference frame for the external voltage loop was proposed while in inner current loops, classical PI controllers are used. This solution guarantees system robustness, but in opposite, the current tracking is not highly performed. In addition, the decoupled operation usually required with PI controllers makes the system more complex. In [15], the input currents are controlled in sliding mode and a PI controller is used to control the voltage loop. This approach yields no robust output voltage controllers. Paper [16] presents the design of the sliding-mode virtual flux oriented control for PWM rectifier, like in [15], only currents loops are controlled in sliding mode. The special stress in this work has been put on the designing of the sliding-mode observer for the virtual grid flux and grid current to provide a sensorless operation of the PWM rectifier without the ac side transducers.

This paper proposes a new control strategy for PWM rectifier using SMC associated with virtual flux oriented control in order to improve the system's robustness and dynamic response of output dc voltage. Instead of the conventional PI controllers, the SMC is employed for each current and voltage control loops. The Lyapunov direct method is used to ensure the reaching and sustaining of the sliding mode. The logic control signals (S_a, S_b, S_c) are calculated by using the space vector modulation (SVM) technique. This last one improves the current-steady state performance by reducing harmonics. The control strategy takes a cascaded structure; the reference of the quadratic current component is generated by the external dc voltage-SMC. Therefore, it depends on the discontinuous control of the outer controller which implies that this reference will have a discontinuous behavior. Hence, the waveforms of the currents are completely distorted. To overcome this problem, we will introduce an hyperbolic tangent function instead of a conventionally sign function. So, the transient of the hitting control will be smoothed without loosing the system robustness. In addition, the use of the SMC does not need the decoupling action usually adopted with classical PI controller. Finally, the simulation results show the efficiency and feasibility of the proposed approach.

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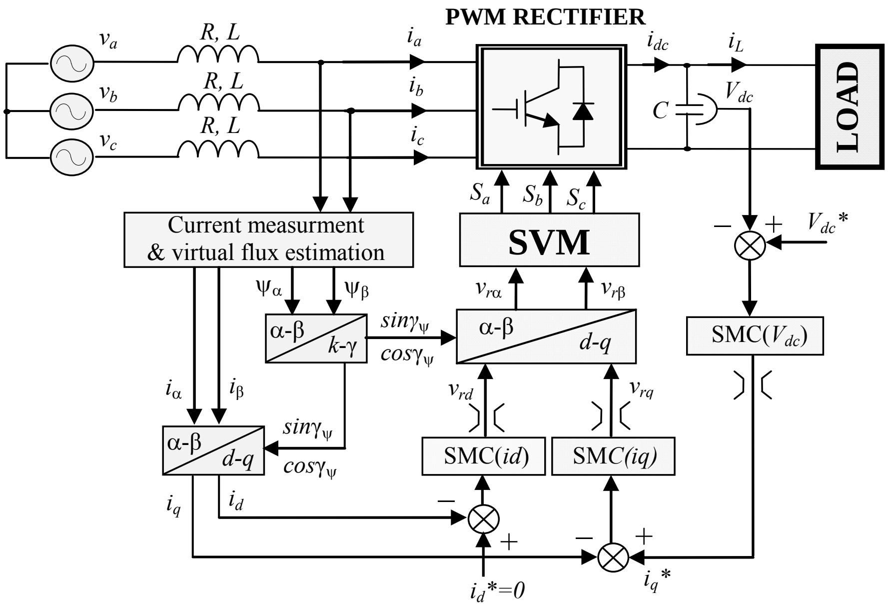


Fig. 1 – Block diagram of sliding mode VFOC PWM rectifier.

2. **CONTROL STRATEGY**

2.1. SYSTEM DESCRIPTION

Figure 1 shows the principle of the control system studied in this work. Two SMCs are used in the internal loops to ensure the current tracking, one for the direct component i_d set to zero for unity power factor condition, another for the quadratic component i_q which regulate the power transient between the ac and dc sides of the converter to keep V_{dc} to desired value. The control system takes a cascaded structure; the reference of the quadratic current is generated by the external dc voltage-SMC. The output signals from SMCs current controllers, after transformation to α-β coordinates system, are used for switching signals generation by space vector modulator.

2.2. VIRTUAL FLUX ESTIMATOR

Based on the measured dc-link voltage V_{dc} and converter switching states S_a, S_b and S_c, the flux components are calculated in α-β coordinates system as follows [17]:

$$\begin{aligned}\psi_\alpha &= \int\left(\sqrt{\frac{2}{3}}V_{dc}(S_a - \frac{1}{2}(S_b + S_c))\right)\mathrm{d}t + L.i_\alpha \\ \psi_\beta &= \int\left(\frac{1}{\sqrt{2}}V_{dc}(S_b - S_c)\right)\mathrm{d}t + L.i_\beta\end{aligned} \quad . \qquad (1)$$

2.3. PWM RECTIFIER MODEL

Electrical equations of the three-phase PWM rectifier are drawn based on the schematic power circuit of the converter shown in Fig. 2. Here v_x represents the source voltage, i_x the input current, and v_{rx} the rectifier input voltages, $x = a, b, c$ respectively. L and R denote the resistance and inductance of reactor, C is the dc link capacitor. R_L, i_L are load resistance and load current respectively, i_{dc} represent the dc side current.

S_a, S_b, S_c are the switching states of the converter. S_j ($j=a, b, c$) is defined as:

$$S_j = \begin{cases} 1, & Sj \quad \text{closed} \\ 0, & \overline{S}j \quad \text{closed} \end{cases}. \qquad (2)$$

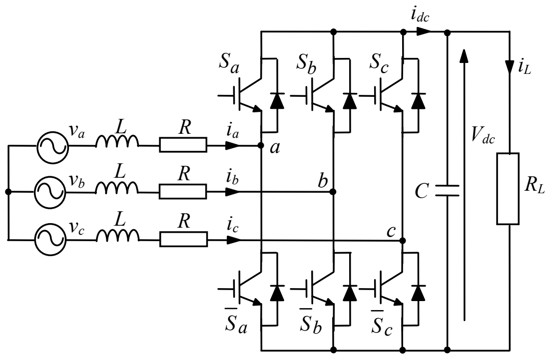


Fig. 2 – Circuit topology three-phase PWM rectifier.

The dynamic model of PWM rectifier in synchronous d-q reference frame can be described by equations (3–5) [5]:

$$L\frac{\mathrm{d}i_q}{\mathrm{d}t} = v_q - Ri_q - L\omega\ i_d - v_{rq}\,, \qquad (3)$$

$$L\frac{\mathrm{d}i_d}{\mathrm{d}t} = v_d - Ri_d + L\omega\ i_q - v_{rd}\,, \qquad (4)$$

$$C\frac{\mathrm{d}V_{dc}}{\mathrm{d}t} = i_{dc} - i_L = (S_d i_d + S_q i_q) - \frac{V_{dc}}{R_L}\,. \qquad (5)$$

3. **DESIGN CONTROLLER**

SMC is a robust nonlinear algorithm which uses discontinuous control to force the system trajectories to join some specified sliding surface; it has been widely used for its robustness to model parameter uncertainties and external disturbances.

In the case of this paper, three sliding surfaces related to control objectives are defined as follow:

$$S(i_q) = i_q^* - i_q, \tag{6}$$

$$S(i_d) = i_d^* - i_d, \tag{7}$$

$$S(V_{dc}) = V_{dc}^* - V_{dc}. \tag{8}$$

The sufficiency conditions for the existence and sustaining of the sliding mode is given as follow [18]:

$$S(x)\dot{S}(x) \le -\eta|S|, \tag{9}$$

where η is a positive constant.

A commonly used form of the control law is:

$$U = -K_x \operatorname{sign}(S(x)), \tag{10}$$

where K_x is the gain of the SMC, and:

$$\operatorname{sign}(S(x)) = \begin{cases} +1 & if \quad S(x) > 0 \\ 0 & if \quad S(x) = 0 \\ -1 & if \quad S(x) < 0 \end{cases}. \tag{11}$$

3.1. q AND d AXIS CURRENT CONTROL DESIGN

Differentiating (6) with respect to time and substituting corresponding relation from (3) yields:

$$\dot{S}(i_q) = \frac{\mathrm{d}i_q^*}{\mathrm{d}t} - \left(\frac{1}{L}v_q - \frac{R}{L}i_q - \omega\, i_d\right) + \frac{1}{L}v_{rq}. \tag{12}$$

Substituting (12) into (9) and using (10), the sliding mode control law of the q-axis current component is obtained as:

$$v_{rq}^* = -K_q \operatorname{sign}(S(i_q)) + \left(v_q - Ri_q - \omega L i_d - L\frac{\mathrm{d}i_q^*}{\mathrm{d}t}\right). \tag{13}$$

By repeating exactly the same procedure for the d-axis current component, we obtain the following control law:

$$v_{rd}^* = -K_d \operatorname{sign}(S(i_d)) + \left(v_d - Ri_d + \omega L i_q - L\frac{\mathrm{d}i_d^*}{\mathrm{d}t}\right). \tag{14}$$

The control laws (13) and (14) leads to limited performance due to the high control activity resulting in chattering. To reduce the latter, a boundary layer method substitutes the discontinuity of a SMC by a saturation function which results in a smooth control signal [19]. Then, we obtain:

$$v_{rq}^* = -K_q \operatorname{sat}\left(\frac{S_q}{Ø}\right) + \left(v_q - Ri_q - \omega L i_d - L\frac{\mathrm{d}i_q^*}{\mathrm{d}t}\right). \tag{15}$$

$$v_{rd}^* = -K_d \operatorname{sat}\left(\frac{S_d}{Ø}\right) + \left(v_d - Ri_d + \omega L i_q - L\frac{\mathrm{d}i_d^*}{\mathrm{d}t}\right). \tag{16}$$

$$\operatorname{sat}\left(\frac{S}{Ø}\right) = \begin{cases} \dfrac{S}{Ø} & \text{if} \quad |S| \le Ø \\ \operatorname{sign}(S) & \text{if} \quad |S| > Ø \end{cases}, \tag{17}$$

Ø represent the bandwidth of the saturation function.

3.2. DC VOLTAGE CONTROL DESIGN

In similar manner as in Section 1 and by using (5, 8, 9 and 10), the sliding mode control law of the dc voltage is obtained:

$$i_{dc}^* = K_{dc}\operatorname{sign}(S(V_{dc})) + \frac{V_{dc}}{R_L}. \tag{18}$$

By neglecting the converter losses and assuming that i_d =0, we obtain:

$$P_{dc} = V_{dc} i_{dc} = v_q i_q. \tag{19}$$

Therefore, the reference of the q axis current component is obtained as:

$$i_q^* = \frac{V_{dc}}{v_q}\left(K_{dc}\operatorname{sign}(S(V_{dc})) + \frac{V_{dc}}{R_L}\right). \tag{20}$$

To alleviate the chattering phenomenon, we will introduce a hyperbolic tangent function instead of sign function. So, the equation (20) becomes:

$$i_q^* = \frac{V_{dc}}{v_q}\left(K_{dc}\tanh\left(\frac{S(V_{dc})}{\varepsilon}\right) + \frac{V_{dc}}{R_L}\right), \tag{21}$$

where ε defines the thickness of the boundary layer.

4. SIMULATION RESULTS

To demonstrate the feasibility of the proposed SMC for three-phase PWM rectifier and to verify its superiority over classical PI controller, some simulations have been performed using MATLAB SIMULINK Software.

The simulations parameters are listed in table 1.

The SMC system regulators parameters are: K_{dc} = 20, K_d = 5 000, K_q = 8 000.

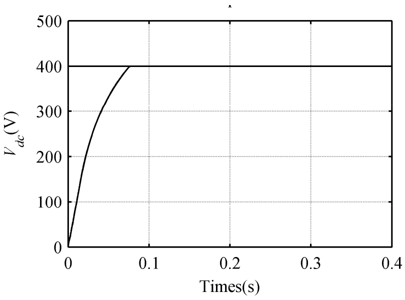
For the PI system, regulators parameters are calculated using the internal model control (IMC) method [20]; we obtain:

– Dc voltage loop: K_{vp} = 0.33, K_{vi} = 6.59.

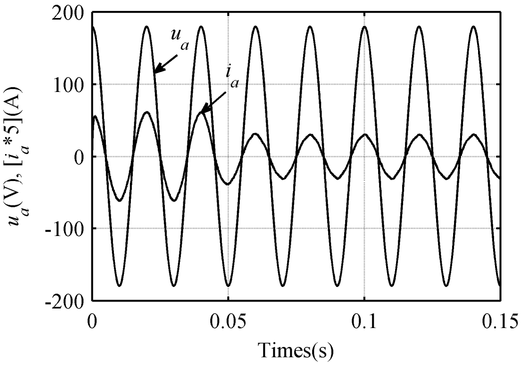
– d and q current loops: K_{ip} = 31.42, K_{ii} = 314.16.

The simulations results are shown in Fig. 3 to Fig. 8.

The response of the system to a step change in dc link resistive load (R_L = 100 Ω) is shown in Fig. 3a (SMC algorithm) and Fig. 4a (PI control scheme). It can be seen that the SMC ensure better dynamic property (no overshoot, smaller response time). Waves of phase voltage and phase current are shown in Fig. 3b (SMC) and Fig. 4b (PI control). We can see that input current is close to sine wave and has the same phase in respect with phase voltage which means that unity power factor (UPF) condition is achieved. It is necessary to note that the current waveform has been scaled by a factor of 5 for clearness.

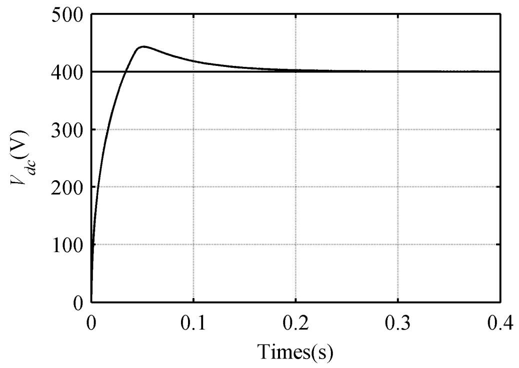


a) Dc voltage waveform

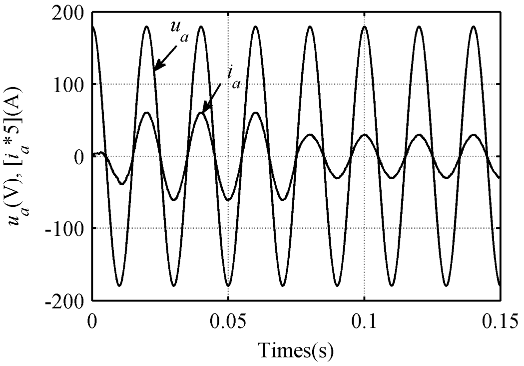


b) Ac side phase current and voltage waveforms

Fig. 3 – Simulation results for output dc voltage change (SMC system).



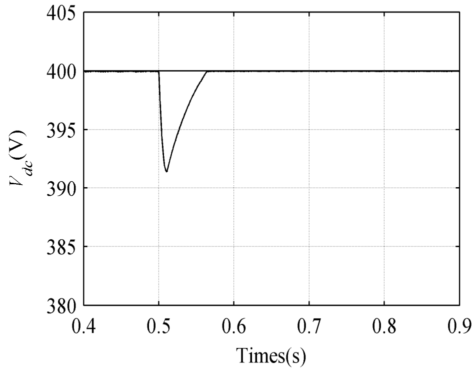
a) Dc voltage waveform



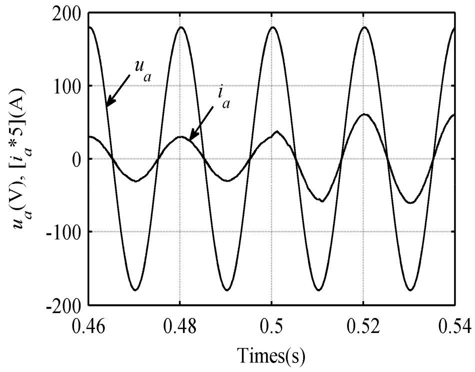
b) Ac side phase current and voltage waveforms

Fig. 4 – Simulation results for output dc voltage change (PI system).

The effect of dc load change (100 Ω to 50 Ω at $t = 0.5$ s) on the operation of the system for both control algorithms is illustrated by Figs. 5 and 6. As it is shown in Fig. 5, load variations has very limited effects on the dc voltage with SMC controller. On the other side, from Fig. 6, we can see that with PI controller, the dc voltage is largely affected by load change.

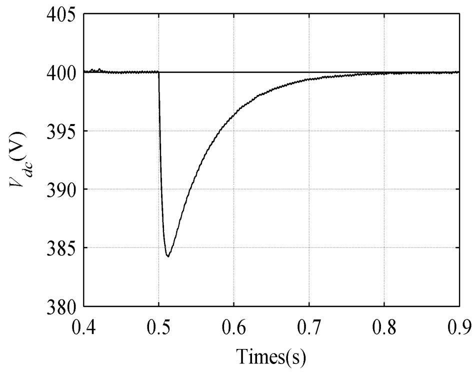


a) Dc voltage waveform

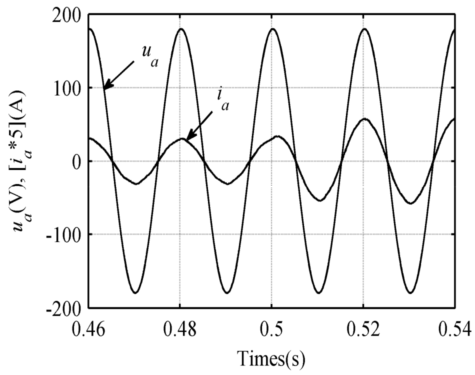


b) Ac side phase current and voltage waveforms

Fig. 5 – Simulation results for dc load change (SMC system).

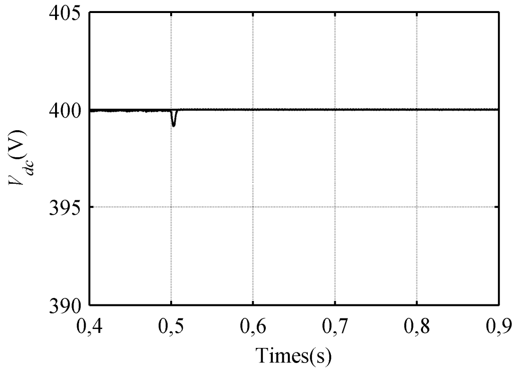


a) Dc voltage waveform

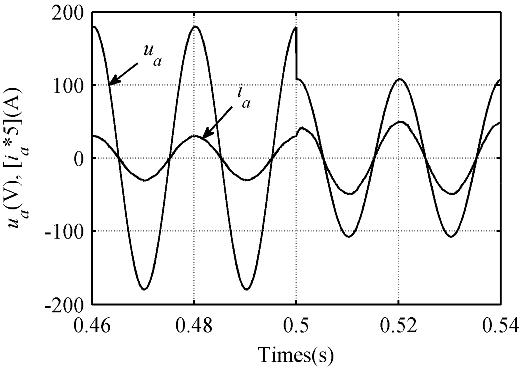


b) Ac side phase current and voltage waveforms

Fig. 6 – Simulation results for dc load change (PI system).

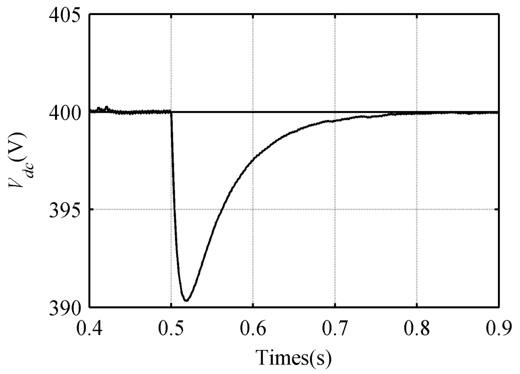


a) Dc voltage waveform

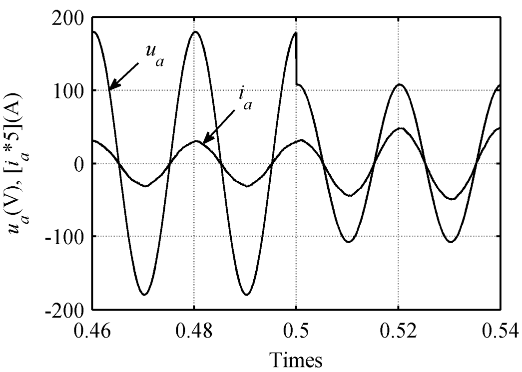


b) Ac side phase current and voltage waveforms

Fig. 7 – Simulation results for ac side voltage change (SMC system).



a) Dc voltage waveform



b) Ac side phase current and voltage waveforms

Fig. 8 – Simulation results for ac side voltage change (PI system).

Figures 7 and Fig. 8 are the simulation results for the both control algorithms when the ac side voltage falls suddenly of 40 percents at $t = 0.5$ s. As can be observed, the SMC system is clearly less sensitive to the applied disturbance comparatively with the PI control scheme.

5. CONCLUSION

Virtual flux oriented control based on sliding mode control theory for three- phase PWM rectifier has been studied in this paper. Regarding the simulation results, we can obtain the following conclusions:

- The sliding mode control algorithm proposed in this work exhibits better dynamic characteristics than conventional PI controller. It is also more robust against load changes and is strongly less sensitive to disturbances like power supply variations.
- Compared with traditional PI controller, sliding mode control scheme has some advantages in terms of few tuned parameters, Simple structure and easy implementation. Indeed, it is known that PI controller parameters calculated in practical applications still needs further adjustment to obtain good control system performance what makes difficult parameter designation. Moreover, decoupling operation needed for PI system controller add an additional degree of complexity.

APPENDIX

The parameters used in simulation are given in Table 1.

Table 1

Parameters used in simulation

Parameter		Value
Resistance of reactors	R	100 mΩ
Inductance of reactors	L	10 mH
Dc link capacitor	C	1 mF
Phase voltage	$V_{a,b,c}$	127 rms
Source voltage frequency	f	50 Hz
Switching frequency	f_{sw}	10^4 Hz
Dc-link voltage	V_{dc}	400 V

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