

BACKSTEPPING CONTROL FOR POWER QUALITY BASED ON A VIRTUAL FLUX OBSERVER

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Abstract— Micro-grids using renewable energy resources become a key solution to address both energy consumption and environmental restrictions in remote locations. For this reason, we try to find a new solution to develop different ways of distribution and energy use. The present paper proposes to combine the virtual flux oriented control with backstepping control to increase the filtering performance of three-phase shunt active power filter. We propose a new multi-function converter as an efficient solution to improve the power quality. The virtual grid flux vector estimated in the sliding-mode observer yields robustness against the line voltage distortions. For improving the quality of the energy transfer from the power supply to the load, and reducing the harmful effects of the harmonics generated by nonlinear load. The good dynamic and static performance under the proposed control strategy is verified by simulation.

Index Terms— PV, virtual flux oriented control; backstepping control; three-phase shunt active filter; sliding-mode observers SMO, sliding mode controller, DPC, VOC, FVOC.

1 INTRODUCTION

Photovoltaic is the most direct way to convert solar radiation into electricity. The generation of electricity is based on the photovoltaic effect, which was first observed by Henri Becquerel [1] in 1839. It is the technological symbol for a future sustainable energy supply system in many countries. A considerable amount of money is invested in research, development and demonstration; several governments set up substantial market introduction programs and industry invests in larger production facilities. This is a remarkable situation since at the same time photovoltaic (PV) electricity is regarded as much too expensive compared to conventional grid electricity.

Various control strategies have been proposed to control the shunt active power filter, such as hysteresis band current control [6], Voltage Oriented Control (VOC) [7], and Direct Power Control (DPC) [7]. Particularly the Voltage Oriented Control guarantees fast transient response and high performance in steady state operation. However, the VOC uses an internal current control loops and the final performance of the system strongly depends on applied current control tech-

niques [7].

The backstepping design recursively selects some appropriate functions of state variables as pseudo control inputs for lower dimension subsystems of the overall system. Each backstepping stage results in a new pseudo control design, expressed in terms of the pseudo control designs from preceding design stages. The final control input is derived from final Lyapunov function formed by summing up the Lyapunov functions associated with each individual design stage [9]. This method is a recursive procedure that skillfully interlaces the choice of a Lyapunov function with the control. Indeed, backstepping control can guarantee global stability, tracking and transient performance for a broad class of strict-feedback systems. Sliding mode observation and control schemes for both linear and nonlinear systems have caused considerable interest in recent times. Discontinuous nonlinear control and observation schemes, based on sliding modes, exhibit fundamental robustness and insensitivity to its properties of great practical value [1].

In this paper, a backstepping associated to VFOC is applied to three-phase shunt active filter. It is shown via simulation results that the proposed controller has high performance both

in the transient and in the steady state operations. The line currents are very close to sinusoidal waveforms, a good control of the DC-bus voltage is obtained, and unity power factor (UPF) operation is achieved

2 BACKSTEPPING CONTROL OF A GRID CONNECTED PHOTOVOLTAIC GENERATION SYSTEM WITH ACTIVE FILTERING FUNCTION

The motivation for this work was to design a digitally controlled, combination active filter and photovoltaic (PV) generation system. This work focuses on a proposed control scheme for the dual function system and on the effects of delay on the control of an active filter. The scheme of the proposed multi-function converter is shown in Fig.1

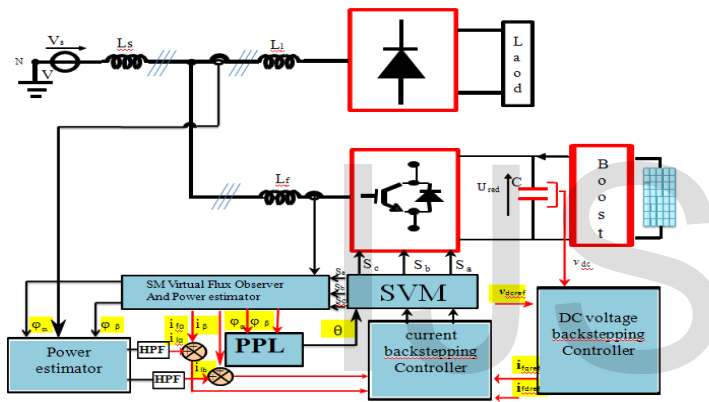


Fig 1: Scheme of the multi-function converter based on sliding-mode observers

2.1 Modelling of the PVG

The mathematical model of the PVG is given by model 1[20].

$$I = I_{sc} - I_0 \left[e^{\frac{(V+IR_s)}{nkT_c/q}} - 1 \right] - \frac{V + IR_{sr}}{R_{sh}} \quad (1)$$

With I and V are respectively the PV current and voltage, I_0 :leakage or reverse saturation current, q : electron charge, n : Ideality factor, K is the Boltzman's constant ($1.38.10^{-23}$ J/K), R_{sr} :series cell resistance, R_{sh} : shunt cell resistance

2.2 Boost converter

The Boost converter shown in Figure 2, it has step-up conversion ratio. Therefore the output voltage is always higher than the input voltage. The converter will operate throughout the entire line cycle, so the input current does not have distortions and continuous. It has a smooth input current because an inductor is connected in series in with the power source. In addition the switch is source-grounded; therefore it is easy to drive.

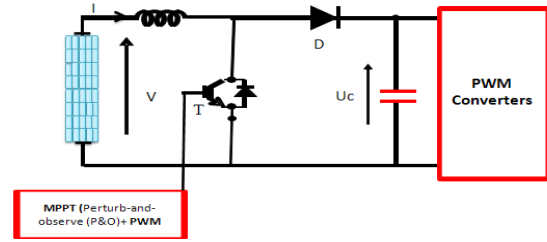


Fig.2. Boost converter

A. Mathematical Model of PWM Converters

A three phase voltage inverter is used to interface the PVG with the grid by converting the dc power generated by the PVG into AC power to be injected to the grid. The dynamic model of a PWM DC-AC Converter can be described in the well known (d-q) frame through the Park transformation as follows [1],:

$$\begin{cases} \frac{di_{fd}}{dt} = \frac{1}{L}(u_{cd} - e_{fd} - Ri_{fd}) - \omega i_{fq} \\ \frac{di_{fq}}{dt} = \frac{1}{L}(u_{cq} - e_{fq} - Ri_{fq}) + \omega i_{fd} \\ \frac{dv_{dc}}{dt} = -\frac{e_{fd}i_{fd} + e_{fq}i_{fq}}{Cv_{dc}} \end{cases} \quad (2)$$

The bi-directional characteristic of the converter is very important in this proposed photovoltaic system, because it allows the processing of active and reactive power from the generator to the load and vice versa, depending on the application. Thus, with an appropriate control of the power switches it is possible to control the active and reactive power flow

$$\begin{bmatrix} \hat{v}_d \\ \hat{v}_q \end{bmatrix} = \frac{1}{i_{fd}^2 + i_{fq}^2} \begin{bmatrix} i_{fd} & -i_{fq} \\ i_{fq} & i_{fd} \end{bmatrix} \begin{bmatrix} P \\ Q \end{bmatrix} \quad (3)$$

2.3 Sliding-mode current observer for virtual grid flux

The $(\alpha-\beta)$ components of the converter voltage vector u_{conv} can be estimated out of the dependence involving the DC-link voltage and the PWM pattern:

$$\begin{cases} v_{f\alpha} = \sqrt{\frac{2}{3}} \cdot U_{dc} \left(S_a - \frac{1}{2} (S_b + S_c) \right) \\ v_{f\beta} = \sqrt{\frac{2}{3}} \cdot U_{dc} (S_b - S_c) \end{cases} \quad (4)$$

The sliding mode observer uses the system model with model with the sign feedback function[22]. The continuous time version of the SMO is described by Equation (5).

$$\frac{d}{dt} \begin{bmatrix} i_{f\alpha} \\ i_{f\beta} \end{bmatrix} = \frac{1}{L_f} (\lambda \cdot \text{sign} \begin{bmatrix} i_{f\alpha} - i_{faest} \\ i_{f\beta} - i_{fbest} \end{bmatrix}) - R_f \begin{bmatrix} i_{f\alpha} \\ i_{f\beta} \end{bmatrix} - \begin{bmatrix} v_{f\alpha} \\ v_{f\beta} \end{bmatrix} \quad (5)$$

The estimated values of the grid voltage are obtained from the low-pass filter:

$$\begin{bmatrix} v_{saestSMO} \\ v_{s\beta estSMO} \end{bmatrix} = LPF(\lambda \cdot \text{sign} \begin{bmatrix} i_{f\alpha} - i_{faest} \\ i_{f\beta} - i_{fbest} \end{bmatrix}) \quad (6)$$

While the $(\alpha-\beta)$ components of the virtual grid flux are calculated as follows:

$$\begin{bmatrix} \varphi_{\alpha est} \\ \varphi_{\beta est} \end{bmatrix} = (\lambda \cdot \int \text{sign} \begin{bmatrix} i_{f\alpha} - i_{faest} \\ i_{f\beta} - i_{fbest} \end{bmatrix} dt) + \begin{bmatrix} \varphi_{\alpha est0} \\ \varphi_{\beta est0} \end{bmatrix} \quad (7)$$

Hence the structure of the virtual grid flux sliding-mode observer presented in Fig.3. The sliding surface representing the error between the measured and references currents are given by this relation :

$$\begin{bmatrix} \sigma_{\alpha} \\ \sigma_{\beta} \end{bmatrix} = \begin{bmatrix} i_{f\alpha} - i_{faest} \\ i_{f\beta} - i_{fbest} \end{bmatrix} \quad (8)$$

The sliding mode will exist only if the following condition[25]

$$\begin{cases} \dot{\sigma}_{\alpha} \sigma_{\alpha} < 0 \\ \dot{\sigma}_{\beta} \sigma_{\beta} < 0 \end{cases} \quad (9)$$

The sliding surface representing the error

3 VFOC-BACKSTEPPING CONTROL

The basic operation of the VFOC-backstepping control method is shown in Fig. 1. The capacitor voltage is compared with its reference value, $v_{dc ref}$, in order to maintain the energy stored in the capacitor constant. The backstepping controller is applied to regulate the error between the capacitor voltage and its reference. The output of backstepping controller presents the

reference of reactive current, while the reference active current value is set to zero for unity power factor.

The compensating currents are derived based on instantaneous p-q theory [7], the instantaneous active and reactive power can be computed in terms of transformed current signals and PCC virtual flux estimator [7]. The alternate value of active power is extracted using high-pass filters. The $\alpha-\beta$ harmonic components are computed using harmonic active and reactive powers. After that, using reverse $\alpha-\beta$ transformation, the d-q compensating currents are derived in terms of $\alpha-\beta$ currents [7]. Then higher harmonics compensating signals

\tilde{i}_{Ld} and \tilde{i}_{Lq} are added with an opposite sign to the standard VFOC reference signals $i_{fd ref}$ and $i_{fq ref}$ as follow:

$$\begin{aligned} i_{d_err} &= i_{fd ref} - \tilde{i}_{Ld} - i_{fd} \\ i_{q_err} &= i_{fq ref} - \tilde{i}_{Lq} - i_{fq} \end{aligned} \quad (10)$$

The output signals from backstepping currents controllers are used for switching signals generation by Space Vector Modulator (SVM) [10].

3.1 Backstepping controller synthesis

From equations (2), it is obvious that the converter is a nonlinear and coupled system. So a nonlinear controller based on the backstepping method is developed in this section.

The system (1) is subdivided in three subsystems as follows:

Subsystem 1:

$$\frac{dv_{dc}}{dt} = -\frac{e_{fd} i_{fd} + e_{fq} i_{fq}}{C v_{dc}} \quad (11)$$

The first subsystem is characterized by only one state $x = v_{dc}$ and only one control input $u = i_{fq}$.

The equation (11) can be written as follow:

$$\begin{cases} \dot{x}_1 = L_f h_1 + L_g h_1 u \\ y_1 = h_1(x) = v_{dc}, \quad y_{1d} = v_{dc ref} \end{cases} \quad (12)$$

Where:

$$x_1 = v_{dc}, \quad u = i_{fq}, \quad L_f h_1 = -\frac{e_{fd} i_{fd}}{C v_{dc}}, \quad L_g h_1 = -\frac{e_{fq}}{C v_{dc}}$$

Subsystem 2:

$$\frac{di_{fd}}{dt} = \frac{1}{L}(u_{cd} - e_{fd} - Ri_{fd}) - \omega i_{fq} \quad (13)$$

$$i_{fq\ ref} = \frac{-k_1 z_1 - L_f h_1 + y_{1d}}{L_g h_1} \quad (21)$$

The second subsystem is also characterized by only one state $x = i_{fd}$ and only one control input $u = u_{cd}$. The equation (13) can also be written as follow:

$$\begin{cases} \dot{x}_2 = L_f h_2 + \bar{u}_{cd} \\ y_2 = h_2 = i_d \end{cases} \quad (14)$$

Where

$$x_2 = i_{fd}, \quad L_f h_2(x) = -R/L i_{fd} - \omega i_{fq} - e_{fd}/L, \quad \bar{u}_{cd} = u_{cd}/L$$

Subsystem 3

$$\frac{di_{fq}}{dt} = \frac{1}{L}(u_{cq} - e_{fq} - Ri_{fq}) + \omega i_{fd} \quad (15)$$

The third subsystem is characterized by one state $x = i_{fq}$ and one control input $u = u_{cq}$. The equation (15) can also be written as follow:

$$\begin{cases} \dot{x}_3 = L_f h_3 + \bar{u}_{cq} \\ y_3 = h_3 = i_q \end{cases} \quad (16)$$

Where:

$$x_3 = i_{fq}, \quad L_f h_3(x) = -\frac{R}{L} i_{fq} + \omega i_{fd} - \frac{1}{L} e_{fq}, \quad \bar{u}_{cq} = \frac{1}{L} u_{cq}$$

3.1 DC voltage controller synthesis

The synthesis of the DC voltage controller is based on the first subsystem.

The first tracking error is defined as:

$$z_1 = x_1 - y_{1d} \quad (17)$$

Its derivative is:

$$\dot{z}_1 = L_f h_1 + L_g h_1 u - \dot{y}_{1d} \quad (18)$$

The Lyapunov function is chosen as:

$$V_1 = z_1^2 / 2 \quad (19)$$

The derivative of (19) is given by:

$$\dot{V}_1 = z_1 \dot{z}_1 = z_1(L_f h_1 + L_g h_1 u - \dot{y}_{1d}) \quad (20)$$

The current component $i_{q\ ref}$ represent the control law of the first subsystem. It is selected such as the Lyapunov function \dot{V}_1 should be definite negative [13], as follow:

Where: k_1 is positive constant.

3.1.2 Current controller synthesis

The active and reactive power can be indirectly controlled via the control of the both outputs $y_2 = i_{fd}$ and $y_3 = i_{fq}$. To achieve unity power factor operation, the active current command is set to zero. The active current command is delivered from the outer DC voltage controller [13].

The errors z_2 and z_3 are defined as:

$$\begin{cases} z_2 = x_2 - y_{2d} \\ z_3 = x_3 - y_{3d} \end{cases} \quad (22)$$

The Lyapunov functions are given by the following expression:

$$\begin{cases} V_2 = \frac{1}{2} z_2^2 \\ V_3 = \frac{1}{2} z_3^2 \end{cases} \quad (23)$$

And consequently, their derivatives are given by:

$$\begin{cases} \dot{V}_2 = z_2 \dot{z}_2 = z_2(L_f h_2 + \bar{u}_{cd} - \dot{y}_{2d}) \\ \dot{V}_3 = z_3 \dot{z}_3 = z_3(L_f h_3 + \bar{u}_{cq} - \dot{y}_{3d}) \end{cases} \quad (24)$$

To make $\dot{V}_2 < 0$ and $\dot{V}_3 < 0$, we must choose:

$$\begin{cases} \bar{u}_{cd} = -k_2 z_2 - L_f h_2 + \dot{y}_{2d} \\ \bar{u}_{cq} = -k_3 z_3 - L_f h_3 + \dot{y}_{3d} \end{cases} \quad (25)$$

Where: k_2 and k_3 are positive constants.

We also have:

$$\begin{bmatrix} \bar{u}_{cd} \\ \bar{u}_{cq} \end{bmatrix} = D \begin{bmatrix} u_{cd} \\ u_{cq} \end{bmatrix} \quad (26)$$

Where:

$$D = \begin{bmatrix} \frac{1}{L} & 0 \\ 0 & \frac{1}{L} \end{bmatrix}$$

The D matrix determinant is different to zero, and then the control law is given as:

$$\begin{bmatrix} u_{cd} \\ u_{cq} \end{bmatrix} = D^{-1} \begin{bmatrix} \bar{u}_{cd} \\ \bar{u}_{cq} \end{bmatrix} \quad (27)$$

5. SIMULATION RESULT

In simulation part, power system is modeled as 3wired 3-

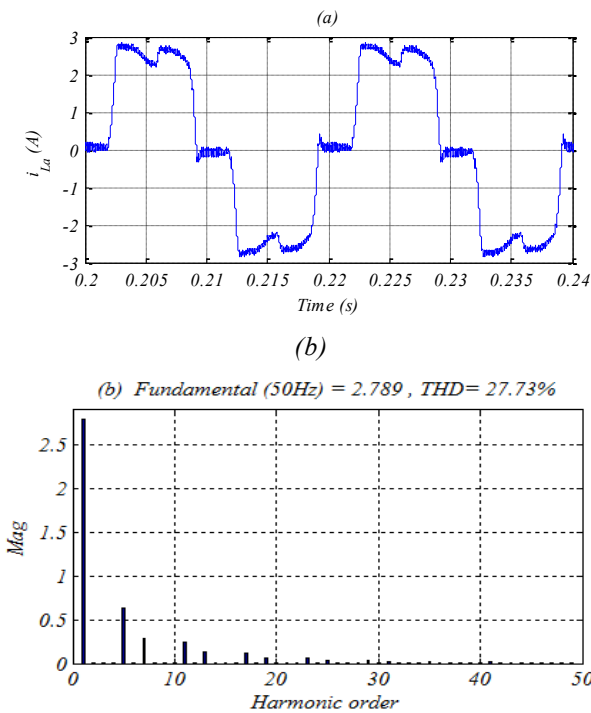


Fig 6 (a) Supply current before harmonics compensation, (b) its harmonic spectrum

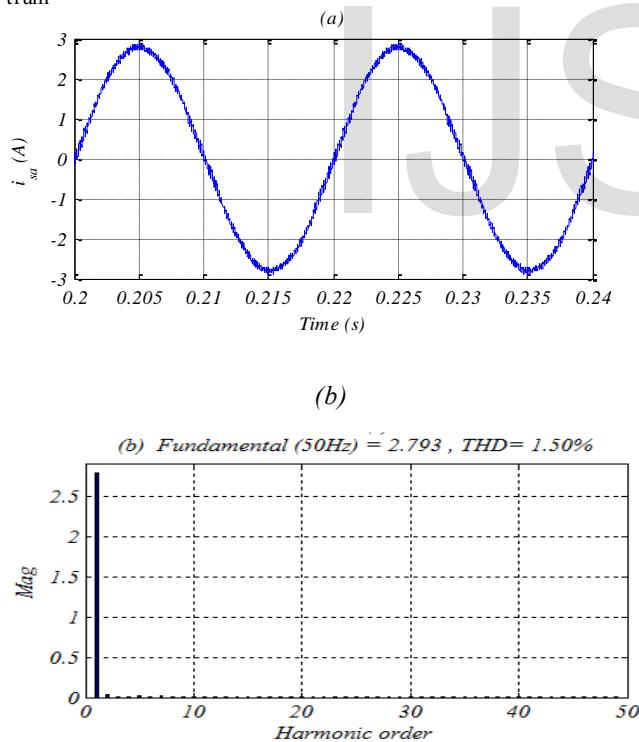


Fig 7 (a) Supply current after harmonics compensation, (b) its harmonic spectrum phase system by an RL load with uncontrolled diode rectifier.

In the circuit, the ac source with frequency of 50Hz. The grid side line voltage is 220V. The line resistor is 0.25Ω. The line inductance of each phase is 1mH. The dc capacitor is 5000μF;

the dc voltage is set to be 750V. The switching frequency for three-phase is 15 kHz. The Pv model applied in simulation is as Fig .1. (250 C Temp. and sun radiation G=1kW/m)

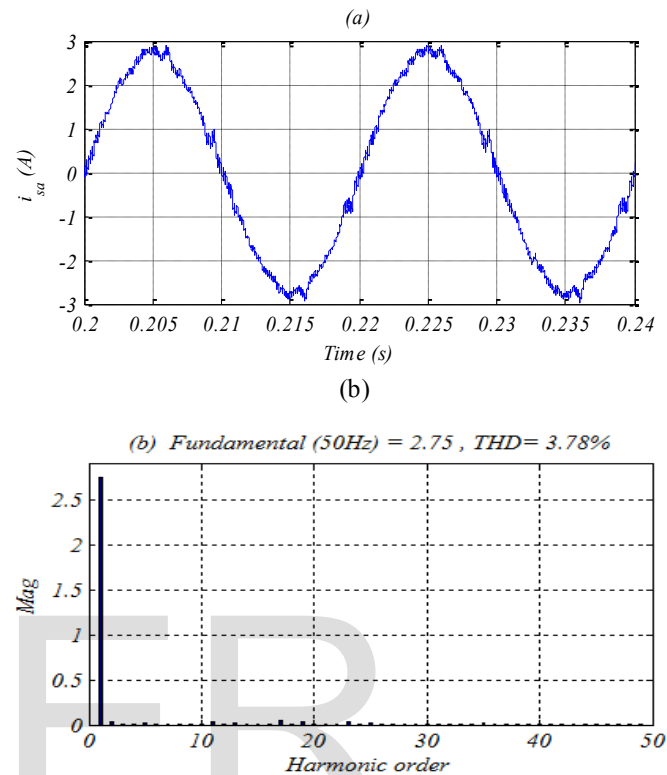


Fig 8 (a) Supply current after harmonics compensation, (b) its harmonic spectrum with PI controller

5.1. Current harmonic compensation

The AC supply current of the first phase and its harmonic spectrum before and after compensation are illustrated in Figs. 6, 7 and 8. It results that the active filter decreases the total harmonic distortion (THD) in the supply currents from 27.73% to 3.78% with PI controller. However, with backstepping controller, the THD is increased to 1.50% which proves the effectiveness of the proposed nonlinear controller.

5.2. Dynamic response performance

In this section, the performance of the backstepping controller is analyzed under 100% step change in the load resistance at t = 0.5s.

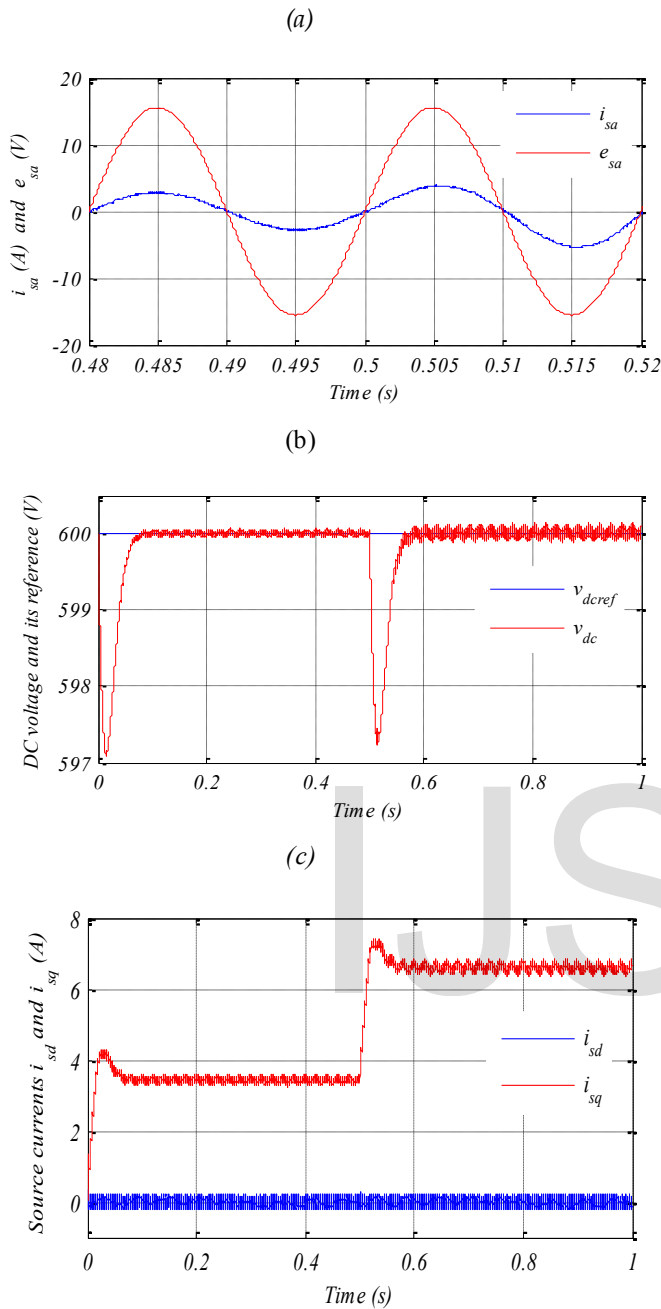


Fig 9 Simulation results of proposed VFOC-backstepping control

The dynamic behavior under a step change of the load is presented in Figs. 9 and 10. It can be observed that the unity power factor operation is successfully achieved, even in this transient state. Notice that, after a short transient, the dc-bus voltage and reactive current are maintained close to their reference values and the active current is maintained zero.

The absence of an overshoot in DC voltage response during load change, and low current ripples, as shown in Fig. 9, demonstrates the superiority of the backstepping controller compared to

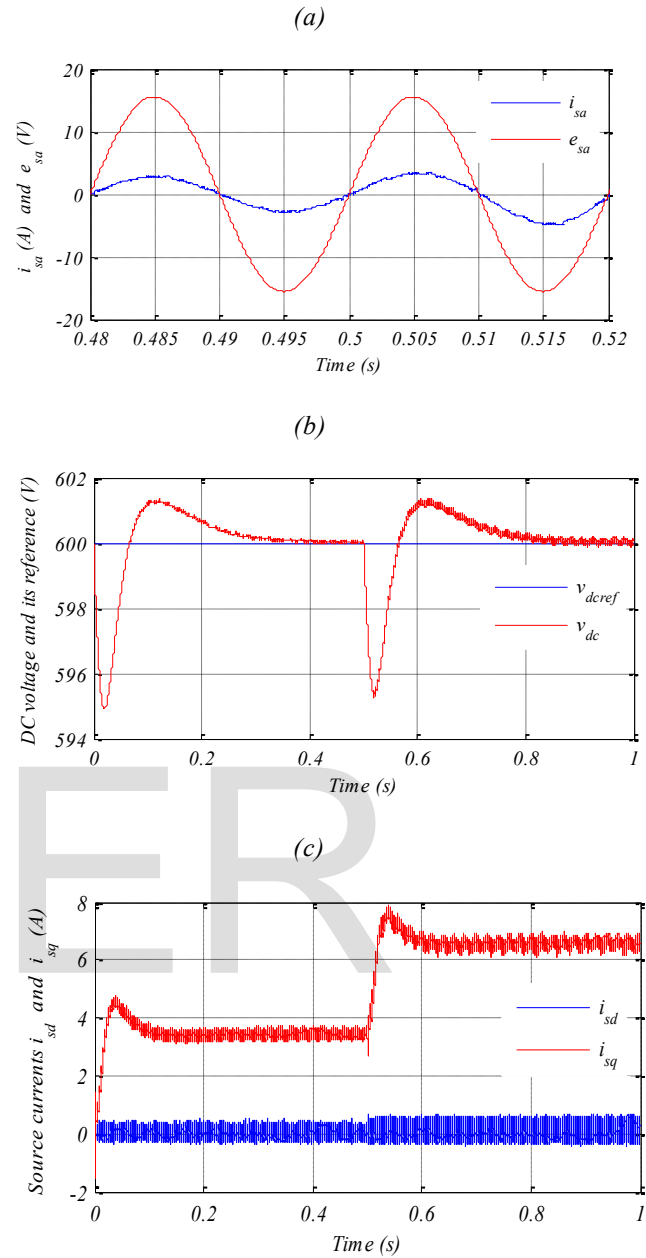


Fig 10 Simulation results of VFOC with PI controller

its counterpart traditional PI controller.

4 CONCLUSION

This paper presents the design of virtual flux oriented control using backstepping approach for three-phase shunt active filter based on a sliding mode observer. The main characteristic of the proposed control method is the orientation of line filter current vector toward the PCC virtual flux vector instead of PCC voltage vector. This manner to proceed increases the ro-

bustness of active filter control system to grid voltages distortions. Simulation results show that the proposed active filter system can compensate effectively both harmonic currents and reactive power of nonlinear load. Furthermore, its performance is not much affected by the distortion of the supply. Thus, the proposed SAF system is found effective to meet IEEE 519 recommended harmonic standard limits.

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