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# A Simple Structure and Fast Dynamic Response for Single-Phase Grid-Connected DG Systems

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**Abstract- This project presents a conventional control technique to regulate the active and reactive powers in single-phase systems to eliminate 90° phase-shift operation used in conventional methods and greatly improves the tracking speed of the responses. In this method directly acts on the instantaneous power and it eliminates the need for calculating dq components. In a single-phase system, the instantaneous power comprises a dc component and a double-frequency ac component as opposed to a three-phase balanced system where it only comprises a dc component. The proposed control system is successfully applied to a photovoltaic system and performance evaluation results using computer simulation.**

**Index terms- Distributed generation (DG), enhanced phase-locked loop (PLL), Instantaneous power control, photovoltaic, renewable energy, single -phase power control.**

## 1 .Introduction

Distributed generatation (DG) systems have become considerably attractive due to several reasons. DG systems: 1) are often based on renewable energy resources such as sun, wind, and water and thus reduce the amount of greenhouse gasses and other air pollutants; 2) improve the reliability of the system against possible shortages of power and outages;3) provide a more economic solution for remote areas due to transmission costs; 4) reduce transmission system losses and upgrade rates; 5) can offer combined heat and power (CHP)solution to customers; and 6) reduce troublesome dependence on the ever-growing fossil fuel price.

The new method for controlling the exchange of power between a single-phase distributed generation system and the grid . The power electronic converters are widely used to interface the DGs with the utility grid . In a three-phase system, the active and reactive powers can be conveniently controlled using the concept of dq rotating synchronous reference frame. The dq components of the current signals are dc variables that are proportional to active/reactive powers.

Thus, simple proportional-integrating (PI) controllers accompanied with decoupling terms can be used to control those variables. The concept of instantaneous power control has also been developed and extensively studied in the technical literature for three-phase systems. In single-phase applications, the current dq components can also be generated using aß-dq transformation where the same three-phase current control strategy can be applied. In these approaches, however, the ß component is not externally available and needs to be synthesized using a 90◦phase-shift operation at the fundamental frequency. The 90phase shift operation can be performed by different methods such as time-delay, all-pass filter, Hilbert transform, second-order generalized integrator (SOGI), or the phase-locked loop (PLL). In addition to the challenges involved in accurate and efficient realization of the phase-shift operation, its dynamics strongly contribute to decrease the speed and the stability margins of the control system. Another class of power control strategies for single-phase applications is based on performing the control at the fundamental frequency using the proportional-resonant (PR) controllers. The current reference is generated as a pure sinusoidal signal whose amplitude and phase angle are controlled. In one approach, which is the one widely used in multistage topologies, to balance the input power extraction with output power injection, the dc link voltage is regulated to a desired value, which results in a reference for the magnitude of the output current. The angle of the current is synchronized with the grid voltage using a PLL. In a single-phase system, the instantaneous power comprises a dc component and a double-frequency ac component as opposed to a three-phase balanced system where it only comprises a dc component. This makes it impossible to use the instantaneous power as a control variable within a linear time invariant (LTI) system because an LTI system operating at the fundamental frequency cannot comprise double-frequency signals. Therefore, such a concept does not fall within the whole concept of LTI systems and, to the best of our knowledge; this issue has not been addressed in the available literature. This paper addresses this issue within a control loop that inevitably comprises non-LTI components. The problem is first formulated from an optimization perspective that results in a control loop comprising a nonlinear component as well as an LTV component. In order to simplify the system and the design, a linear counterpart is derived for the nonlinear component.

This reduces the system to an LTV system and a complete stability analysis and design method is presented for this system. The proposed method directly acts on the instantaneous power and it eliminates the need for calculating dq components. It, thus, obviates the need for a 90◦ phase-shift operation. This eliminates the aforementioned problems that are caused by the dynamics of such operation for the control system. Thus, a very fast and stable control operation is achieved despite the highly simple structure of the whole control loop.

## II. Proposed Technique

In this section, the proposed instantaneous power controller structure is derived and presented. The derivation initially results in a nonlinear structure and then by some simplifications, an equivalent linear time varying (LTV) structure is developed that allows easier stability analysis and parameter design.

### A. Derivation of the Controller Equations

The grid voltage is denoted by $v_g$ (t) and the grid current is $i_g(t)$ .The grid current is controlled in such a way that the appropriate active P and reactive Q powers are injected to the grid. In a sinusoidal situation where $v_g(t) = v_0 \cos(t)$ and $i_g(t) = I_0 \cos(\omega t - \emptyset)$

the instantaneous power is $P(t) = v_g\ (t)\ i_g(t)$

$$= \frac{1}{2}V_0 I_0 \cos\phi[1+\cos(2\omega t)] + \frac{1}{2}I_0 \sin\phi \sin(2\omega t)$$

$$= \mathrm{P}[1+\cos(2\omega t)] + Q \sin(2\omega t) \text{.................... (1)}$$

Where P= $\frac{1}{2}V_0 I_0 \cos\phi$ and Q = $\frac{1}{2}V_0 I_0 \sin\phi$

The instantaneous power comprises a dc component and a double-frequency component.

Assume that the commands for active and reactive powers are denoted by $P^*$and $Q^*$, respectively. Then, the command for the instantaneous power is given by

$$p^*(t) = p^*[1+\cos(2\omega t)] + Q^* \sin(2\omega t) \text{........... (2)}$$

Define the cost function

$$J[i_g(t)] = [p^*(t) - p(t)]^2 = [p^*(t) - v_g(t) i_g(t)]^2$$

Which is the instantaneous square error between the instantaneous power and its reference? The objective is to find an appropriate current $i_g(t)$that minimizes J $[i_g(t)]$ for given values of $v_t(t)$, $P^*$ and $Q^*$. To address a solution to this problem, write the current as $i_g\ (t) = i_g \cos(\phi_i)$where $i_g$ and $\phi_i$ are to be found. The cost function will then be a function of smooth unknown variables $\theta = (i_g, \phi_i)$. The gradient descent method is used to derive the dynamics of according to

$$\dot{\theta} = -\mu \frac{\partial J(e)}{\partial e} \text{.......................................................... (3)}$$

In which $\mu$ is a positive-definite 2 × 2matrix. Assuming a diagonal structure as $\mu$ = diag $\{\mu_1, \mu_2\}$, the resulted equations can be summarized as

$$i_g(t) = \mu_1 e(t) v_g(t) \cos(\emptyset_i)$$

$$\dot{\emptyset}_I(t) = -\mu_2 e(t)\ v_g(t) \sin(\emptyset_i) + \omega \text{........................... (4)}$$

Where $e(t) = p^*(t) = p^*(t) - v_g(t) i_g(t)$

Equation set (5) shows how the desired variables $Ig$ and $\Phi i$ must be changed to ensure minimum error between the actual power and the desired power. There still remain two issues to be addressed: 1) the instantaneous power reference $p^*(t)$ must be synthesized from the active and reactive reference values $p^* and\ Q^*$) the system frequency $\omega$ must be available (assuming that it can have variations from the nominal value). The instantaneous power reference can be synthesized from (3) if the voltage phase-angle $\emptyset_e = \omega t$ is available. To address both issues, PLL can be employed on the voltage signal to obtain the phase-angle and frequency. The most important advantage of the PLL, in this context, is its ability to is a control system that generates an output signal whose phase is related to the phase of an input "reference" signal. It is an electronic circuit consisting of a variable frequency oscillator and a phase detector eliminate the double-frequency harmonics in single-phase applications that makes it specifically attractive for grid-connected single-phase applications. Equations of the PLL are provided as follows:

$$\dot{v_g}(t) = \mu_3 e_v(t) \cos(\emptyset_v)$$

$$\Delta\dot{\omega}(t) = -\mu_4 e_v(t) \sin(\emptyset_v)$$

$$\dot{\emptyset}_v(t) = -\mu_4 \mu_5 e_v(t) \sin(\emptyset_v) + \omega \text{...................... (5)}$$

Where $e_v\ (t) = v_g\ (t) - V_g \cos(\emptyset_v)$ and $\mu_3$ to $\mu_5$ are real positive numbers. The overall control structure consists of the dynamics represented by (5) and (6) whose block diagram is shown in Fig. 2. The instantaneous power controller structure is shown

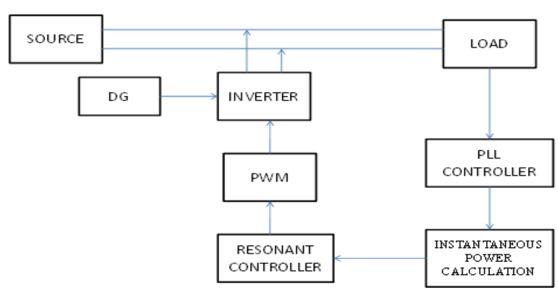


Fig 1 Block diagram of proposed system

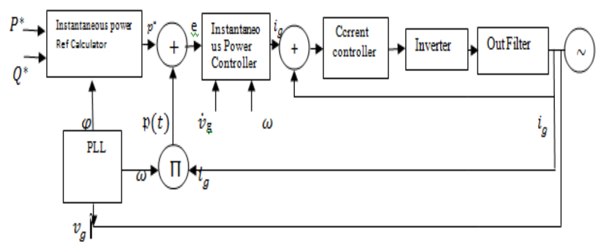


Fig. 2. Block diagram of the loop connecting the DG to the grid through the converter.

In the fig 2 .The PLL estimates $\phi_e$ from the measured voltage signal. A simplified local stability analysis is the nonlinear equations (5) and (6) are presented .A based on the assumption that the current control is very fast and its dynamic is neglected.

## IV. Linear Equivalence of the Proposed Control System without Current Control Loop:

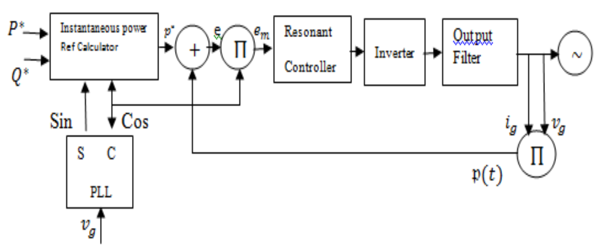


Fig.3. Block diagram of the proposed simplified instantaneous power Controller

Since the PLL tracks a sinusoidal input signal with no steady-state error, it can be concluded that the gain of the loop is infinity at the input frequency. This means that the loop gain from z to i*g is also infinity at this frequency similar to a conventional resonant controller of the form µs/s$^2$+$\omega^2$

The idea of controlling the instantaneous power makes the current control redundant because the current control is automatically achieved when the instantaneous power is perfectly controlled. In the proposed, simplified structure of Fig. 5, the current control loop is removed and only the instantaneous power control loop exists. The stability analysis of the proposed simplified control loop of Fig. 5 is necessary since the loop components have changed and this will be performed in next section. However, to give more insight into the operation of the system of Fig. 5, the following simple analysis is presented. Multiplying the signal p (t) with the normalized grid voltage cosωt results in

$$P_m(t) = p(t)\cos(\omega t)$$
$$= P\ (1 + \cos 2\omega t) + \cos(\omega t) + Q\sin 2\omega t\cos\omega t$$
$$= \frac{3}{2}\ P\cos\omega t + \frac{1}{2}\ Q\sin\omega t + \ldots + \frac{1}{2}\ P\ \cos 3\omega t + \frac{1}{2}\ Q\ \sin 3\omega t + \ldots \quad (6)$$

Therefore

$$e_m(t) = [p^*(t) - p(t)\sin\omega t$$
$$= (p^* - p)(1 + \cos 2\omega t)\cos\omega t + \ldots + \ldots (Q^* - Q)\sin 2\omega t\cos\omega t$$
$$= \frac{3}{2}(p^* - p)\cos\omega t + \frac{1}{2}(Q^* - Q)\sin\omega t + \ldots + \ldots \frac{1}{2}(P^* - P)\cos\omega t + \frac{1}{2}(Q^* - Q)\sin 3\omega t \ldots \quad (7)$$

The following points are observed form

1) The modulated error signal $e_m$ $(t)$ has two ac terms at fun- damental and at third harmonics;
2) The coefficients of both terms are related in the sense that if one of the terms is controlled to zero the other one will also approach zero;
3) Since the sine and cosine functions are orthogonal, if $e_m$is regulated to zero, all the terms will be regulated to zero;
4) by regulating $e_m$ to zero, both active and reactive power components will be regulated to their reference values.

Thanks to the infinite gain of the resonant controller at the fundamental frequency, the signal $e_m$ approaches to zero and the power control (as well as current control) is achieved.

## V Stability Analysis of the Proposed System

The stability analysis of this section is performed for a converter with inductive output filter. In a similar way, the method can be extended for other types of filter.

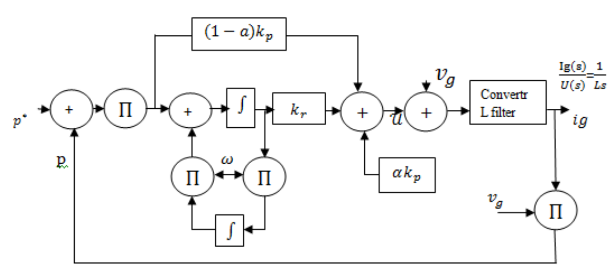


Fig. 6. Detailed control loop of the proposed system (without the EPLL) for an L-type output filter.

The detailed control structure consists of a resonant controller kc e/s$^2$+w$^2$ in the forward path and an internal state current feedback with gain kp as shown in Fig. 6 where the PLL is not depicted because it does not involve in the stability of the control loop. The constant a ∈ [01] does not have any impact on the closed-loop stability of the linearized system. It is used as an extra tuning gain in order to further adjust the speed with which the power commands are transferred to the output.2pThe state space equations of the closed-loop system (for a =1) can be written as

$$\dot{x_1} = -\omega x_2 + k_r \cos \omega t\,(p^* - p)$$
$$\dot{x_2} = \omega x_1$$
$$\dot{x_3} = \frac{1}{L} x_1 - \frac{k_p}{L} x_3 \qquad (8)$$

where $x_1$ and $x_2$are the state variables of the resonant controller and x3 is the grid current. This set of equations represents a LTV system with a time-varying reference input. The equation set (10) has a solution at

$$\dot{x_1}(t) = \sqrt{L^2\omega^2 k_p^2}\, I^* \sin(\omega t - \emptyset + \beta)$$
$$\dot{x_2}(t) = -\sqrt{L^2\omega^2 k_p^2}\, I^* \cos(\omega t - \emptyset + \beta)$$
$$x_3^*(t) = I_g^* \sin(\omega t - \emptyset) \qquad (9)$$

Define the new set of variables as $z = x - x^*$ and derive the state space equations for z as

$$\dot{x_1^*} = -\omega x_2^*$$
$$\dot{x_2} = \dot{\omega} x_1^*$$
$$\dot{x_3} = \frac{1}{L} x_1^* - \frac{k_p}{L} x_3^* \qquad (10)$$

The equation set (12) represents an LTV system with no reference input. Moreover, the state variables $z$have a dc nature as opposed to $x$ that are sinusoidal. The system equations shown in (12) can be represented as $\dot{z} = A(t)z$ where $A(t)$ is a time varying matrix. It can further be decomposed into

$$\dot{z} = A_0 z + k_r A_1(t) z \qquad (11)$$

Where the matrices $A_0$ and $A_1$ are given by

$$A_0 = \begin{pmatrix} 0 & -\omega & 0 \\ \omega & 0 & 0 \\ \frac{1}{L} & 0 & -\frac{k_p}{L} \end{pmatrix}$$

$$A_1(t) = -V_g \cos^2 \omega t \begin{pmatrix} 0 & 0 & 1 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{pmatrix}$$

Theorem C stated in Appendix C is used here to prove the stability of (14). The correspondence between the variables used in Theorem C and the aforementioned variables are summarized as: $x = z, f(x) = A_0 x, c = k_r, g(t, x, \epsilon) = A_1(t)x$. In order to use this theorem, the zero equilibrium point of the system $\dot{x} = f(x)$must be exponentially stable. To satisfy this requirement, the matrix $A_0$ is modified to

$$A_0 = \begin{pmatrix} -2\zeta\omega & -\omega & 0 \\ \omega & 0 & 0 \\ \frac{1}{L} & 0 & -\frac{k_p}{L} \end{pmatrix}$$

This corresponds to a characteristic polynomial of $s^2+2£ws+w^2$ for the resonant controller where the value of ζ is a small positive number. This modification is also justified from the fact that in implementations of the resonant controller for practical applications it is required to add the damping factor ζ in the digital implementation due to numerical limitations. Having applied this modification,$A_0$ will be an exponentially stable matrix, g is bounded and T-periodic, and thus the system of(11) holds all conditions of Theorem C. Existence, uniqueness, and exponential stability of a T-periodic solution for the system is then concluded from this theorem.

The constant £ defined in Theorem C is obtained from the continuity condition of the matrix ∂f/∂x at the origin. Since this matrix is constant for the system of (11), the constant becomes arbitrary and can have any large positive value. This results in that the existence, uniqueness, and exponential stability of the periodic solution is guaranteed for all positive values of the resonant controller gain $k_r$.

D. Design of the Controller Gains for the System of Fig. 6

An alternative state space description for the system which is more suitable for design purposes, is given by

$$x_1 = -\omega x_2 + \dot{k_r} \cos \omega t (p^* - p)$$
$$= -\omega x_2 - V_g \cos^2 \omega t x_3 + p^* \cos \omega t$$
$$\dot{x_2} = \omega x_1$$
$$\dot{x_3} = \frac{k_r}{L} x_1 - \frac{k_p}{L} x_3 \qquad (12)$$

Assuming that $x^*$ is the steady-state solution and defining$z = x - x^*$,, the equations for z are given as

$$\dot{z_1} = -\omega x_2 - V_g \cos^2 \omega t\, z_3$$
$$\dot{z_2} = \omega z_1$$
$$z_3 = \frac{\dot{k_r}}{L} z_1 - \frac{k_p}{L} z_3 \qquad (13)$$

Since the $z$ variables have dc nature, the high-frequency term can be neglected for design purposes. Then, the (13) can be rewritten as

$$\dot{z} = A_z + B_u \qquad (14)$$

A and B are defined as

$$A = \begin{pmatrix} 0 & -\omega & 0 \\ \omega & 0 & 0 \\ 0 & 0 & 0 \end{pmatrix} \qquad ,B = \begin{pmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & \frac{1}{L} & 0 \end{pmatrix}$$

Design of the state feedback gains can be done using different methods such as the Bode diagram, root-locus, and optimal control methods. In other words, the same current control design used in conventional methods can be employed to design the proposed controller's gains.

## VII. Simulation Result

Performance of the proposed controller is evaluated in this section by computer simulations performed using MATLAB software. The power circuit 400-V dc link, connected to the grid through a full-bridge VSI and an L-filter with L = 10 mH. The grid voltage rms value is 240 V and its frequency is 60 Hz. The inverter operates using unipolar PWM technique with a switching frequency of 20 kHz that amounts to an actual value of 40 kHz seen by the filter.

The conventional method of Fig. 1(a) and the proposed method of Fig. 5 are implemented and compared. The PI controllers are set at Kp=0.5and Ki= 500. The resonant controller for the proposed method is designed using the optimal technique of in digital domain w ith a sampling frequency of 40 kHz synchronized with the switching frequency. Moreover.
The constant $\alpha$ is selected as 0.5. For the conventional method, there is a tradeoff between the speed and the harmonic response

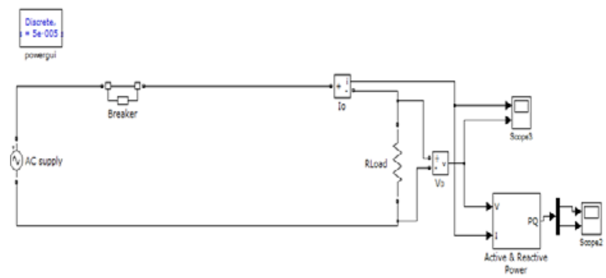


Fig 8 Simulink diagram of proposed without DG

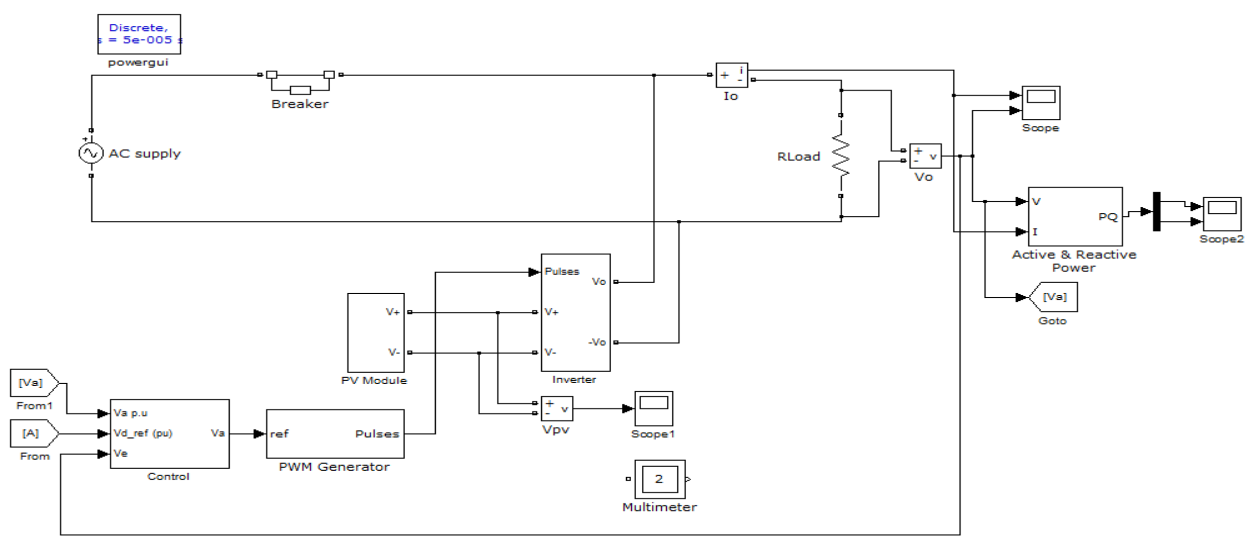


Fig 9 Simulink diagram of proposed system with DG

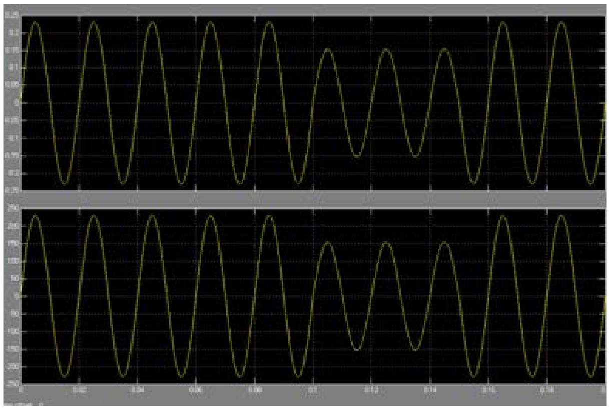

Fig10 inject voltage and current in without DG

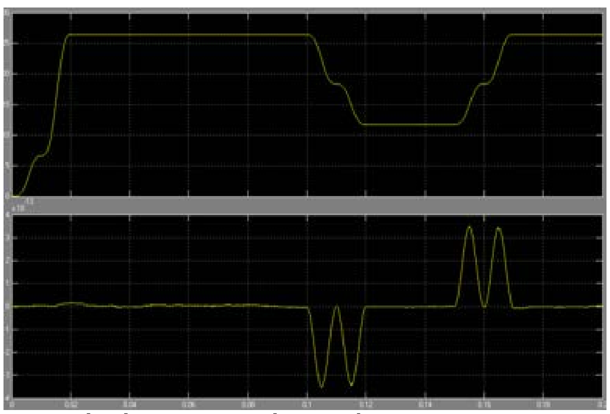

Fig11Real and Reactive power losses without

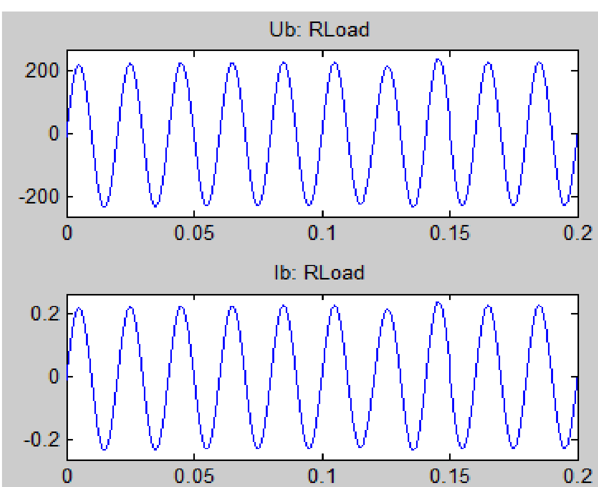


Fig12 performance of Injection voltage and current with DG

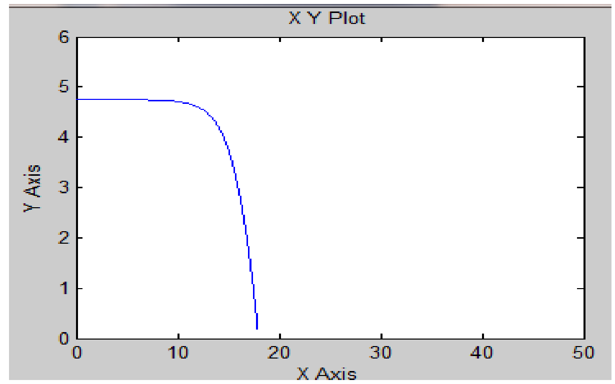


Fig13 Performance of VI characteristics of Photo voltaic

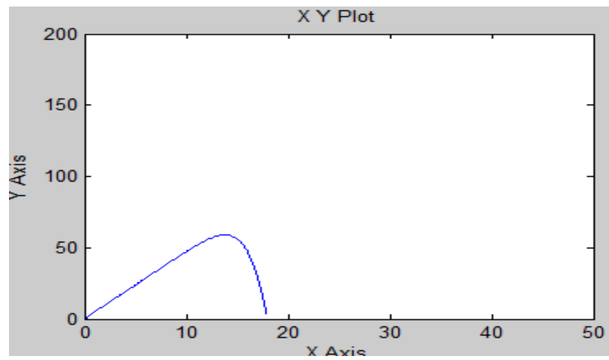


Fig14 Performance of PV characteristics of Photo voltaic

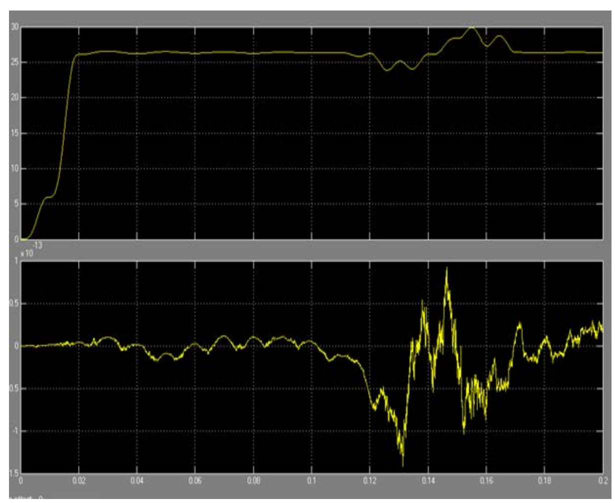

Fig 15 performance of reactive power compensation

## VIII Conclusion

A method for controlling exchange of power between a single phase DG and the utility grid is proposed and discussed. Contrary to the conventional techniques that separately control active and reactive powers, the proposed controller directly acts on the instantaneous power signal. This eliminates the need for 90◦phase-shift operation used in conventional methods and thus, greatly improves the tracking speed of the responses. The proposed has a simple yet robust structure and can be further improved to compensate harmonics. The requirements of the stability of a second-order polynomial and the aforementioned discussion, show that the necessary and sufficient condition for the local stability of the system is that all five design parameters µ1to µ5re positive.

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