A DIAGNOSTIC APPROACH FOR CALCULATION OF THRESHOLD POINTS OF A PR CONTROLLER FOR SINGLE-PHASE GRID CONNECTED PV SYSTEM

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Abstract— Proportional plus Resonant controller is presented as the injected grid current regulator applied to a single-phase grid connected inverter. This paper establishes a systematic approach to design the gains of the PR controller by determining its minimum and maximum threshold points, to achieve improved transient and steady-state performance. Extensive simulations are carried out by varying system parameters and their effects on magnitude and bandwidth of the amplitude-frequency response are explained.

Keywords— Current control, resonant controller, Grid-connected

1. INTRODUCTION

With the rapid increase in energy consumption, greener technologies like wind, photovoltaic, fuel cells are in widespread demand [1]. Thus, various topologies of power converters in Distributed Power Generation have been of great research interest [2][3]. The purpose of this converter is to synchronize and transfer the generated variable power to the grid based on a control algorithm. While selecting the controller to meet these requirements, an optimal balance between cost, computational complexity of the control and quality of waveform must be made [4]-[8].

For single-phase system, among the various controllers available [9][10], though PI controllers are simple, easy to realize and linear, they are susceptible to steady-state errors while tracking alternating input quantities [11][12]. Any variations have been suggested [13]-[16], but they deteriorate the bandwidth, stability limits of the system and introduction of harmonics into line currents [17]-[20]. Amongst the current control methods in voltage source converters, accuracy in the system modelling plays a vital role

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in Model-predictive control [21][22]. Whereas in hysteresis current control [23], switching rate and high frequency cause hindrance to achieve zero steady-state error. Thus, to achieve precise control of a single-phase converter in reference tracking, with reduced computational burden and same frequency response characteristics, a PR controller was introduced in [24] and several applications of PR controller for voltage source converters is detailed in [25]. In [26], PR controller is implemented using Infinite Impulsive Response (IIR) digital filter and details about the design of filter coefficients, to make the controller adaptive to the frequency variations.

Design of the gains any single-phase or three-phase PR controller may be performed by anyone of the methods internal model control, Bode plots or nyquist plots. In [27], gains of the resonant controller are generalized with respect to the closed loop bandwidth of the system. The analysis is carried out in $\alpha\beta$ frame and is applied to design the controller for a weak grid. Shift-operator digital implementation based on Tustin Transform is discussed in [28]. a and b coefficients involved in the z -domain equation of the controller are derived. In [29], PR controller is implemented for current control in a single-phase PV system based on a pseudo synchronous d - q transformation technique. Gains of the controllers are selected based on theory. In [30], standard transfer functions for PR controllers is proposed and methods to obtain the gains are presented. These are applied to some test cases and performance is compared to the tuning methods of PI controller. In [31], PR controller is realized in $\alpha\beta$ reference frame, to study the control of positive and negative sequence currents during fault conditions. Here the K_p and K_r values are given directly as 71 and 6000 respectively. It is observed that tuning of the controller gains while maintaining a trade-off between the various control parameters can be achieved when an appropriate range of the gains is provided. In this paper a systematic approach to design the gains of the PR controller by determining its minimum and maximum threshold points is explained in detail.

2. DESIGN OF PR CONTROLLER

Fig. 1 Single-stage Circuit of Single-Phase VSC with PR controller connected to Grid

Current and voltage of solar panel are measured and fed to MPPT algorithm, to track the maximum point of the PV array. Based on the output of the MPPT and Phase Locked Loop (PLL), DC-AC inverter produces a current signal to be injected into the grid. Outer DC link voltage control loop generates the reference current for the inner current control loop. As reference current signal is a sinusoidal quantity, so there is a need for an ac compensator. Main objective of an ac compensator is to obtain zero phase and magnitude error. For an input Dc signal, PI controller meets this objective. Whereas, to meet the same objective with an input ac signal, dc compensator (transfer function $H_{DC}(s)$) must be converted to an equivalent ac compensator (transfer function $H_{AC}(s)$) maintaining the

same bandwidth and frequency response [32]. This is accomplished by shifting the frequency to ω_0 , as :

$$H_{AC}(s) = H_{DC}(s) \left(\frac{s^2 + \omega_1^2}{2s} \right) \quad (1)$$

When applied to a PI controller, it can be represented as:

$$H_{AC}(s) = K_P + \frac{2K_I s}{s^2 + \omega_1^2} \quad (2)$$

where K_P : Proportional gain, K_I : Resonant gain. Its frequency response is shown in Fig. 2. This lead to the to an ideal P+ Resonant controller. It is observed that, it can be realized with a proportional term and an integrator. Ideal dc integrator is approximated to a high gain low-pass filter with ω_b as the lower breakpoint frequency, gains K_r and K_p govern the amplitude gain and controller dynamics respectively. Accordingly, the response modifies to as shown in Fig. 3 and ac PI compensator in Eq. (2) is modified to give a finite gain, represented as:

$$H_{AC}(s) = K_P + \frac{2K_r \omega_b s}{s^2 + 2\omega_b s + \omega_1^2} \quad (3)$$

Fig. 2 Frequency response of an ideal PR controller

Fig. 3 Frequency response of a non-ideal PR controller

Fig. 4 Overall scheme for control of injected current

At fundamental frequency, magnitude-frequency response of the system is zero and is expressed as:

$$20 \log \left| \left(\frac{2K_r\omega_b(j\omega_c)}{-\omega_c^2 + 2\omega_b(j\omega_c) + \omega_1^2} \right) \frac{1}{j\omega_c L} \right| = 0 \quad (4)$$

As $\omega_c > \omega_1$, by successive approximations Eq. (4) becomes:

$$20 \log \left| \frac{K_P}{j\omega_c L} \right| = 0 \quad (5)$$

$$K_P = |j\omega_c L| \quad (6)$$

$$\frac{i_g^* - i_g}{i_g^*} = 1 - \frac{G(s)}{1+G(s)} \quad (7)$$

At fundamental frequency, steady-state error in injected grid current is considered as η and is expressed as:

$$\eta = \frac{i_g^* - i_g}{i_g^*} \quad (8)$$

$$\text{Then, } \eta = \frac{1}{1+G(s)} \quad (9)$$

So, at ω_1 , open loop transfer function and phase-response of the PR controller will be given by (11) and (13) respectively as:

$$|G(j\omega_1)| = \frac{1}{\eta} - 1 \quad (10)$$

$$PR(s) = \frac{s^2 + 2\omega_b s \left(1 + \frac{K_r}{K_P}\right) + \omega_1^2}{s^2 + 2\omega_b s + \omega_1^2} \quad (11)$$

$$\Phi(\omega) = K_p + \frac{2K_r\omega_b s}{s^2 + 2\omega_b s + \omega_1^2} \quad (12)$$

$$\Phi(\omega) = \tan^{-1} \left(\frac{2\omega_b \omega \left(1 + \frac{K_r}{K_P}\right)}{\omega_1^2 - \omega_c^2} \right) - \tan^{-1} \left(\frac{2\omega_b \omega}{\omega_1^2 - \omega_c^2} \right) \quad (13)$$

For stability, $\Phi(\omega_c) > -\Phi$

$$\text{Or, } \left\{ \tan^{-1} \left(\frac{2\omega_b \omega_c \left(1 + \frac{K_r}{K_P}\right)}{\omega_1^2 - \omega_c^2} \right) - \tan^{-1} \left(\frac{2\omega_b \omega_c}{\omega_1^2 - \omega_c^2} \right) \right\} > -\Phi \quad (14)$$

$$|L(\omega_1)G(j\omega_1)| - \omega_c \leq K_r \leq \left| K_P \left\{ \left(\frac{\omega_1^2 - \omega_c^2}{2\omega_b \omega_c} \right) \tan \left\{ \tan^{-1} \left(\frac{2\omega_b \omega_c}{\omega_1^2 - \omega_c^2} \right) - \Phi \right\} - 1 \right\} \right| \quad (15)$$

According to Eq. (15) minimum and maximum threshold limits of PR controller can be defined by Eq. (16) and (17) respectively:

$$K_{r_{min.}} = |L(\omega_1)|G(j\omega_1)| - \omega_c| \quad (16)$$

$$K_{r_{max.}} = \left| K_P \left\{ \left(\frac{\omega_1^2 - \omega_c^2}{2\omega_b\omega_c} \right) \tan \left\{ \tan^{-1} \left(\frac{2\omega_b\omega_c}{\omega_1^2 - \omega_c^2} \right) - \Phi \right\} - 1 \right\} \right| \quad (17)$$

3. RESULTS AND DISCUSSION

System parameters taken for simulation are given in Table I. Cut-off frequency is always less than one-tenth of the switching frequency [33], thus cut-off frequency is selected to be 1000Hz.

Table I. System Parameters

Name	Value
Switching Frequency	10 KHz
Fundamental frequency	50 Hz
Grid Phase voltage	230V
L filter inductor	6 mH

Using Eq. (6) and the values given in Table 1, K_p is calculated to be approximately 38. Consider steady-state error in injected grid current, $\eta=1\%$.

From Eq. (11),

Open-loop transfer function of PR controller at fundamental frequency:

$$|G(j\omega_1)| = 99.$$

Evaluating (16) and (17) yields to:

$$K_{r_{min.}} = 149 \text{ and } K_{r_{max.}} = 2000.$$

(a)

Fig. 5 Effect of varying K_r on open-loop bode plots for PR regulator by varying f_c Values

Cut-off frequency allows the degree of freedom in the design of the frequency response of resonant controllers. Therefore, an increase in its value broadens the bandwidth, as illustrated in Fig. 5(a), Fig. 5(b) and Fig. 5(c). Bandwidth is further increased with the increase in K_r value from 400 to 2000 in Fig. 5(a) to Fig. 5(b). For a particular resonant term, the peak amplitude remains the same irrespective of the changes in cut-off frequency, whereas this magnitude increases with increase in K_r value. For $K_r=400$, peak amplitude is around 53db and it is increased to about 66 db for $K_r=2000$. The same is illustrated in Fig. 6(a), Fig. 6(b) and Fig. 6(c).

Fig. 6(a)

(b)

(c)

Fig. 6 Effect of varying K_p on open-loop bode plots for PR regulator by varying L values

Fig. 7. Effect of varying K_r on Open-loop bode plots for PR regulator

Fig. 7 highlights that amplitude at a particular frequency is increased by increasing the value of resonant term, K_r . K_r also controls the bandwidth of the system around this particular frequency.

Fig. 8 Grid voltage and inverter output voltage

Fig. 9 and Fig. 10 show that for a pure resistive load the PR controller enables the perfect tracking of the injected grid current to the variations in the reference current at $t=0.1$ sec. It tracks the grid current completely in 4ms for step change in grid current from 5 to 10 A at $t=0.1$ sec, as illustrated in Fig. 9. For better understanding of the grid voltage and current, grid voltage is scaled down by 30 in Fig. 10 and scaled down by 10 in Fig. 10.

Fig. 9 Grid voltage (scaled down) and injected current when reference current is increased from 5 A to 10A at $t=0.1$ sec.

Fig. 10. Grid voltage (scaled down) and injected current when reference current is increased from 10 A to 20A at $t=0.1$ sec

The current tracking ability by the designed PR controller is compared with PI controller described in [34]. For a step change in grid current, ripple of 16% in the converter current is obtained with the implementation of [34]. Whereas the PR controller presented in this paper demonstrates a superior performance with reduced ripple in converter current and faster tracking of grid current.

Fig. 9 Current tracking by PR controller for step change in grid current

4. CONCLUSION

This paper presents the need for a Proportional resonant controller in a single-phase grid connected system and also details diagnostic approach the steps to design the PR controller. Using the numerical results from the design process, simulations are performed and extensive studies are carried out using bode diagrams by varying the system parameters. It was demonstrated that K_r also controls the bandwidth of the system around a particular frequency and amplitude at this frequency increases with increasing K_r value. Moreover, for a particular resonant term, the peak amplitude remains the same irrespective of the changes in cut-off frequency. Simulation results show that the injected current follows the variations in the reference current

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