

A Proportional Resonant Controller Tuning Method for Grid connected Power Converters with LCL+trap Filter

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Abstract— Three-phase power converters are largely employed in power generation facilities based on renewable energy sources, and these converters commonly working in grid connection mode are frequently connected to the grid through passive filters. Nowadays the need for larger power converter-based generation plants increases continuously, and LCL+trap filter arises as a solution when switching frequency is reduced due to the power increase. However, the tuning of current controllers becomes more complex as the LCL+trap filter lead to a considerably high order system. In this work a method for tuning the gains of proportional resonant current controllers for three-phase high power converters linked to the grid through a LCL+trap filter is proposed and analyzed. Simulation and experimental results taken in different cases are shown to validate the proposed method.

Keywords- DC-AC power converters; Distributed power generation; Automatic control; Closed loop systems; Control design

I. INTRODUCTION

The installed capacity of the distributed generation systems based on renewable energy systems have been increasing hugely during last years [1]-[2], and the remarkable increase is continuing without any sign of pausing. Along with this, the power of the grid connected converters is also experiencing an impressive boom. Two-level three-phase converters connecting to utility grid through LC or LCL filter, which are typically adopted in modern generation systems, are no longer the best option due to that the high switching frequency (typically over 10kHz) will lead to a relatively big switching power loss which is difficult to ignore.

As reduced switching frequency is necessary in high power implementations, high order grid link filters arise as an alternative option complementary to Multi-level converters (MLC) to deal with the issue of reduced switching frequency. In the family of high order grid link filters, LCL+trap filter gradually becomes more adopted [3]-[5]. This type of filters contains LCL branches, but additionally has a branch specially designed for minimizing the harmonics at switching frequency. With this configuration, the resonant frequency of the filter can be increased remarkably, which leaves enough freedom for the design of current controller. However, LCL+trap filter which is 5th order system itself, will consequently lead to a higher order system compared with LCL filters. As a result, the design and

tuning of controllers of the converter becomes more challenging because the dynamics of the filter should be considered in a fine tuning of the controllers.

Among all the control designing tasks, the design and tuning of current controller is a critical issue and even the first issue appearing in many cases. Proportional Resonant (PR) controllers in stationary α - β reference frames and proportional Integral (PI) controllers in rotating d-q reference frames are both widely employed for grid connected power converters, and both strategies are able to provide a zero steady state error. However, rotating d-q frames should be transformed to stationary α - β frames for examining stability of a system, then there will occur cross-coupling terms between stationary frame α - β components [6]-[7]. Additionally, comparing with PI controllers, removal of the steady state error in single phase systems, no need for voltage feed-forward and easy tuning stand as the main advantages of PR controllers [8].

Tuning methods for PR and PI controllers are both seen in literatures [9]-[11], but control tuning for high power converters with LCL+trap filters is not completely discussed. This work proposes a method for tuning the PR current controllers for two-level three-phase high power converters with LCL+trap filter. Description of the applied systems is shown in II, the control tuning method is elaborated in III, and simulation and experimental results in different cases are presented in IV to validate the proposed method

II. DESCRIPTION OF THE SYSTEMS

A. Setup Structure

The proposed tuning method is oriented to setups shown in Fig. 1, which consist of a two-level 3-phase voltage source inverter (VSI) and a LCL+trap filter. The inverter is a two-level 3-phase VSI, which has a high reliability compared to MLC, and the filter is made up of a branch tuned to eliminate the harmonics at the switching frequency and a LCL filter to eliminate harmonics with higher frequency. For typical LCL filters, the resonant frequency is designed low enough to realize a small gain at the switching frequency. And when the switching frequency is reduced in high power implementations, the design of the filter is in a trade-off between the dynamics and power quality. On the purpose of increasing the resonant

frequency without weakening the filtering capability, LCL+trap filter is designed.

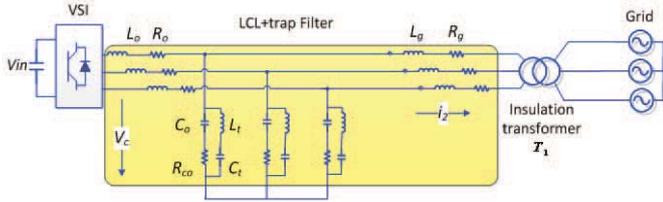


Fig. 1. VSI connecting to grid through LCL+trap filter.

B. Control Scheme

The α - β transformation (Clark Transformation) is used for implementing the control on the α - β stationary reference frame. And the PR controller, which is the only type of controller that can ensure a theoretical zero steady state error is adopted in the control loop. The controller output is processed by an inverse α - β transformation and then goes into the PWM block. The grid side current i_2 flowing through grid side inductors L_g (in Fig. 1) is adopted as the feedback variable to realize an accurate control of power injection without theoretical delay. The scheme of the current control loop expressed by transfer functions in continuous-time domain is shown in Fig. 2. i^* is provided by outer loops such as power controllers. ω_0 is the grid angular frequency which is commonly measured by a Frequency Locking Loop (FLL) synchronizing system, K_p and K_r are the proportional gain and resonant gain, T_{sw} is the switching period and one switching period is modeled as the modulation delay in the loop. The transfer function of the LCL+trap filter $G(s)$ is the expression of I_2/V_c (shown in Fig. 1). And,

$$\alpha_1 = C_o C_l L_o L_g L_t,$$

$$\alpha_2 = C_o C_l L_o L_g R_{co} + C_o C_l L_o L_t R_{co} + C_o C_l L_g L_t R_{co} + C_o C_l L_o L_t R_g + C_o C_l L_g L_t R_{co},$$

$$\alpha_3 = C_o L_o L_g + C_l L_o L_g + C_l L_o L_t + C_l L_g L_t + C_o C_l L_o R_{co} R_g + C_o C_l L_g R_{co} R_o + C_o C_l L_t R_{co} R_o + C_o C_l L_t R_{co} R_g + C_o C_l L_t R_o R_g,$$

$$\alpha_4 = C_o L_o R_{co} + C_o L_g R_{co} + C_o L_o R_g + C_o L_g R_o + C_l L_o R_g + C_l L_g R_o + C_l L_t R_o + C_l L_t R_g + C_o C_l R_{co} R_o R_g,$$

$$\alpha_5 = L_o + L_g + C_o R_{co} R_o + C_o R_{co} R_g + C_o R_o R_g + C_l R_o R_g,$$

$$\alpha_6 = R_o + R_g.$$

III. CONTROLLER TUNING METHOD

A. Analytical Solving of Controller Gains

The control is implemented digitally, and the control loop

must be discretized in control tuning. In order to model the control loop to make it as close as possible to the real system, the whole control loop is discretized by sampling frequency f_s . Assuming that the switching frequency is lower than 5 kHz, which is usually the case for 100 kW converter systems, the sampling frequency of the control system f_s can be adopted to be four times of f_{sw} in order to ensure the synchronizing system working effectively and also realize an accurate discrete-time domain control. The transfer function of PR controller in continuous-time domain $G_C(s)$ is shown in (1). Employing the backward Euler method in the direct branch of the resonant controller and forward Euler method in the feedback branch of the resonant controller, the transfer function of PR controller is discretized as $G_C(z)$ and shown in (2), where T_s is the sampling period. Therefore the open-loop transfer function in discrete-time domain $G_{OL}(z)$ can be expressed in (3), where $G(z)$ is the discrete-time domain transfer function of the filter. As all the values of the filter components are available, $G(z)$ can be converted from the transfer function $G(s)$ (shown in Fig. 2) by software like MATLAB. Through this multiple sampling implementation, the modulation delay of one switching period is equal to 4

sampling periods and is expressed by $\frac{1}{z^4}$.

$$G_C(s) = K_p + \frac{\omega_0 K_r s}{s^2 + \omega_0^2} \quad (1)$$

$$G_C(z) = K_p + K_r \frac{\omega_0 \times T_s \times z \times (z-1)}{(z-1)^2 + (\omega_0 \times T_s)^2 z} \quad (2)$$

$$G_{OL}(z) = G_C(z) \times G(z) \times \frac{1}{z^4} \quad (3)$$

So the issue of determining the values of K_p and K_r turns into the issue of solving the formula of (3). As it is shown, (3) is a complex number equation, which can be converted to 2 real number equations. So the degree of freedom is equal to the unknown numbers as long as the complex number z and the value of $G_{OL}(z)$ are specified.

In order to specify the values of z and $G_{OL}(z)$, it's easier to consider the transfer function in continuous-time domain. When s is specified to cross-over frequency ω_c , the module of $G_{OL}(s)$ is equal to 1, and the phase of $G_{OL}(s)$ is related to the phase margin. These relationships are shown in (4) and (5), where phm is the system phase margin. Fixing the phase margin phm and cross-over frequency ω_c in open-loop, the

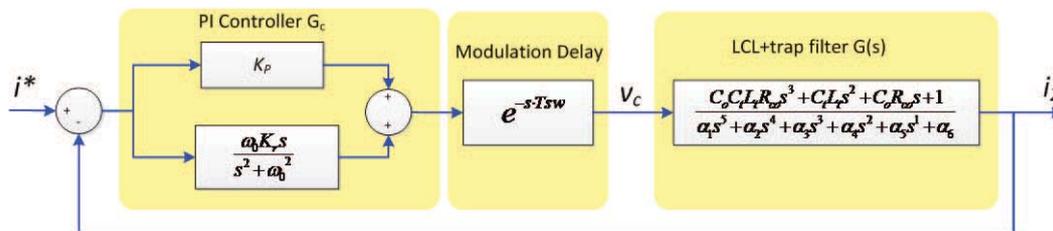


Fig. 2. Current control scheme where grid side current i_2 is adopted as the feedback variable.

values of controller gains K_p and K_i can be calculated through (6).

$$\left|G_{ol}(j\omega)\right|_{\omega=\omega_c} = 1 \quad (4)$$

$$\angle G_{ol}(j\omega)\big|_{\omega=\omega_c} = -(180^\circ - phm) \quad (5)$$

$$G_{OL}(z)\big|_{z=e^{j\omega_c T_s}} = 1 \angle -(180^\circ - phm) \quad (6)$$

B. Control Tuning

In order to obtain a group of controller gains that can make the system works properly with good dynamic response, ω_c and phm should be fixed reasonably. On this purpose, the tuning method is developed. By this method, ω_c and phm are specified with different values, then different solutions of controller gains are obtained through solving (6). By comparing the dynamics of the numerous systems defined by each solution, an optimal solution is selected.

Therefore, as the first step of the tuning procedure, tuning range should be determined. A range of ω_c and phm are determined in this step.

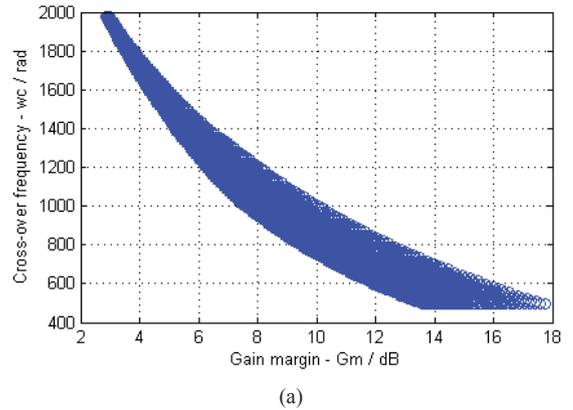
A 100 kW setup is taken here as an example to illustrate the tuning method. The associated parameters of the setup are shown in TABLE I. The nominal power of the converter S is 100 kW.

TABLE I. PARAMETERS OF THE 100kW CONVERTER

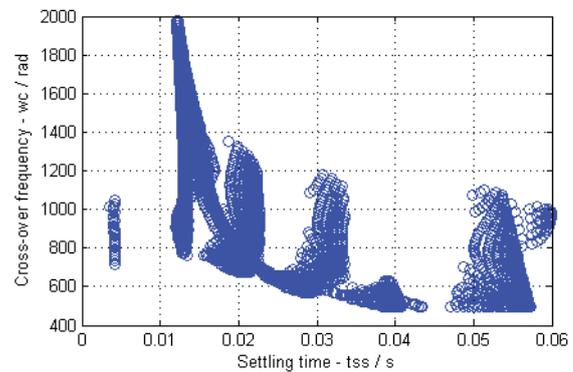
Parameters	Values
V_{in} / V	780
$L_o / \mu H$	778
$L_g / \mu H$	402
$C_o / \mu F$	66
$C_i / \mu F$	30
$L_1 / \mu H$	85
R_o / Ω	0.0073
R_g / Ω	0.0021
R_{co} / Ω	0.5
S / kW	100
f_{sw} / Hz	3150
f_s / Hz	12600
Utility grid / V/Hz	400 / 50

Fig. 3 is plotted by specifying different ω_c and phm , and for each solution, the stability margin, system settling time and overshoot to unitary step input are checked. Fig. 3a shows the relationship between ω_c and gain margin Gm . As it is shown, with ω_c above 1600 rad, the gain margin has no chance to be above 5 dB. In order to ensure the stability of the system, systems with gain margin lower than 5dB is not accepted in the tuning. Therefore, the upper bound of the tuning range of ω_c is selected to be $\frac{\omega_{sw}}{12}$ (1649 rad), where ω_{sw} is the switching angular frequency shown in (9). Fig. 3b shows the relationship between ω_c and unitary step response settling time tss . As it is shown, with ω_c below 600 rad, tss has no chance to be smaller than 25 ms. In order to ensure the system's speed of response, solutions with tss larger than 25 ms are rejected in the tuning.

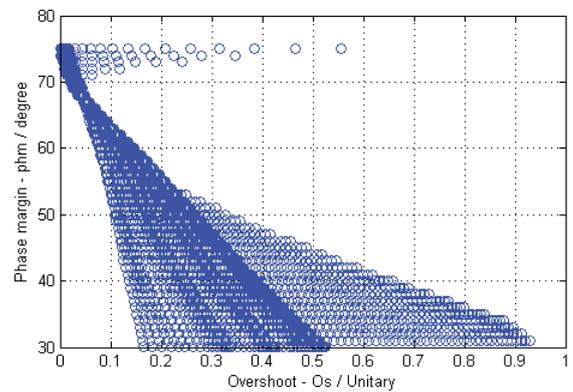
So the lower bound of the tuning range of ω_c is selected to be $\frac{\omega_{sw}}{33}$ (599.8 rad). To sum up, the tuning range of ω_c is shown in (7), where $d\omega$ is the tuning step length shown in (10).



(a)



(b)



(c)

Fig. 3. Interaction among different variables of the system: a) Interaction between ω_c and gain margin Gm ; b) Interaction between ω_c and unitary step response settling time tss ; c) Interaction between phm and unitary step response overshoot Os .

As shown in Fig. 3c, solutions with phm smaller than 40° all has a unitary step response overshoot Os larger than 15%. Since a larger overshoot will make the system have dangerous transient response which could even cause tripping or damage to the converter, the lower bound of the tuning range of phm is selected 40° . And the upper bound is selected 70° . The tuning range of phm is shown in (8).

$$\omega_c = \left\{ \frac{\omega_{sw}}{33}, \frac{\omega_{sw}}{33} + d\omega, \frac{\omega_{sw}}{33} + 2d\omega, \dots, \frac{\omega_{sw}}{12} \right\} \quad (7)$$

$$p_{hm} = \{40^\circ, 41^\circ, 42^\circ, \dots, 70^\circ\} \quad (8)$$

$$\omega_{sw} = 2 \times \pi \times f_{sw} \quad (9)$$

$$d\omega = \frac{\omega_{sw}/12 - \omega_{sw}/33}{100} \quad (10)$$

With the tuning range shown in (7) ~ (8), all the solutions that doesn't exceed the limits shown in TABLE II will be restored and one optimal solution can be selected. To select the optimal solution, different criteria can be used to rank the solutions and select the most desired one. In this tuning case, the solution with the largest bandwidth is selected, which is shown in TABLE III.

TABLE II. TUNING LIMITS FOR CURRENT CONTROLLER OF THE 100kW CONVERTER

Variables	Limits
Settling time(ms)	<25
Overshoot (%)	<15
Gain margin (dB)	>5
Phase margin (°)	>40

TABLE III. SELECTED SOLUTION FOR 100kW CONVERTER

Variables	Values
Kp	1.2163
Kr	0.5601
Settling time(ms)	22.5
Overshoot (%)	14.65
Gain margin (dB)	7.1276
Phase margin (°)	59.9995
Band width (rad)	6085
Cross-over frequency (rad)	1080

To have a clear view of the values shown in TABLE II, open-loop BODE plot is shown in Fig. 4, and the closed-loop system's response to a unitary step is shown in Fig. 5.

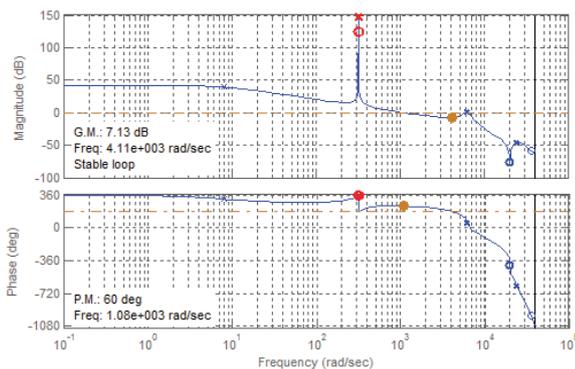


Fig. 4. BODE diagram of the open-loop system defined by the selected solution.

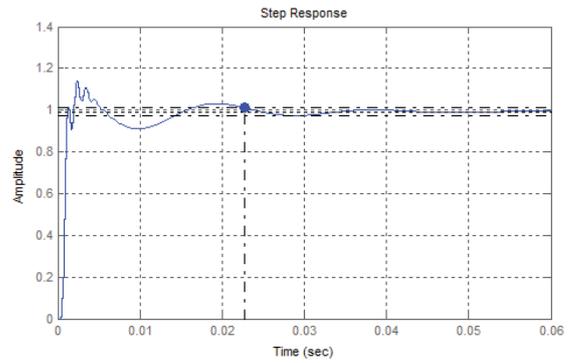


Fig. 5 The unitary step response of the system defined by the selected solution.

IV. RESULTS

A. Simulation Results

A simulated plant with the topology of Fig. 1 and the parameters of TABLE I is built in PSIM (Fig. 6), and the control is implemented by referring to a DLL file generated by C code. The controller structure shown in Fig. 2 with the gains shown in TABLE III is implemented. SVM algorithm is adopted for modulation, and the modulation signals are generated by the control block and sent to the pulses generation block. The pulses generation block consists of three comparators with triangular wave as the input of the negative port. The triangular wave has a peak to peak value of 1, and the frequency is 3150Hz (switching frequency). The simulation results are shown in Fig. 7.

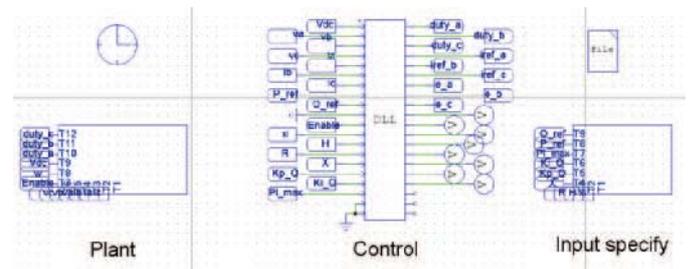


Fig. 6. Simulation model of 100 kW converter with LCL+trap filter.

Fig. 7 shows the waveforms of three-phase current with their references and grid injection power during a transient state. Initially power reference of 50 kW and 37.5 kVar (inductive) is given, and at 0.3s, a step in power reference is given with the new reference of 100 kW and 75 kVar. It is seen that the current are well regulated under both steady and transient. The power regulation settling time is below 25 ms, which meet the design requirements. And the current tracks the reference without any static error.

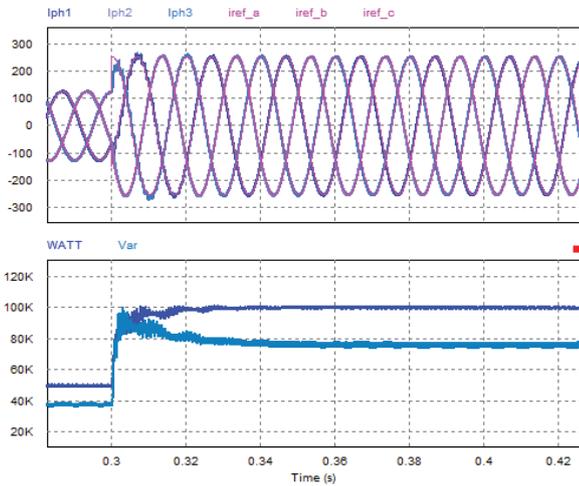
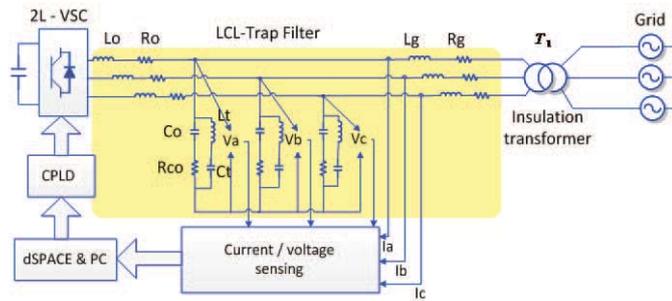


Fig. 7. Simulation results of 100 kW converter with LCL+trap filter.

B. Experimental results

In order to further validate the proposed tuning method, tests are conducted on an experimental plant (shown in Fig. 8). In this case, the setup has the same structure as Fig.1, with different parameters shown in TABLE IV. The selected control parameters are shown in TABLE V, and the experimental results are shown in Fig. 9.



(a)



(b)

Fig. 8. Experimental system: a) Scheme of the experiments; b) Picture of the experimental setups.

Fig. 9 shows the waveforms of three-phase current, the current reference in phase A and grid injection power. A step of power reference is given increasing from 4 kw and 3 kvar to

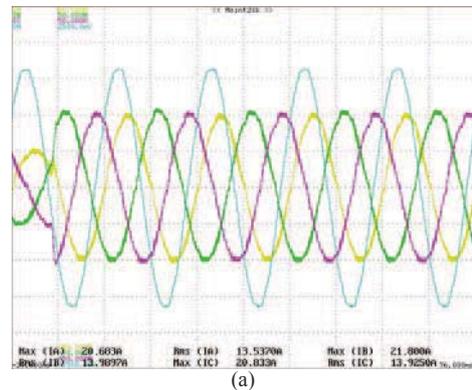
8 kw and 6 kvar. The current are well regulated under both steady and transient state, and the controller has a fast speed of response as desired, with very limited overshoot.

TABLE IV. PARAMETERS OF 10 KVA CONVERTER

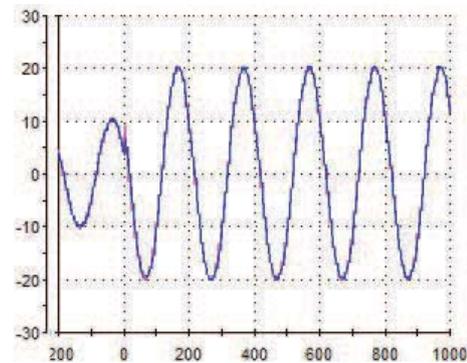
Parameters	Values
V_{in} / V	640
L_o / mH	2.6
$L_g / \mu H$	662
$C_o / \mu F$	5.5
$C_t / \mu F$	1
$L_t / \mu H$	244
R_o / Ω	0.025
R_g / Ω	0.094
R_{co} / Ω	1
S / KVA	10
f_{sw} / Hz	10050
f_s / Hz	10050
Utility grid / V/Hz	400 / 50

TABLE V. SELECTED CONTROL PARAMETERS FOR 10 KVA CONVERTER

Variables	Values
Kp	8.7818
Kr	7.7968
Settling time(ms)	11.6
Overshoot (%)	12.47
Gain margin (dB)	8.0197
Phase margin (°)	60.9998
Band width (rad)	7816
Cross-over frequency (rad)	2810



(a)



(b)

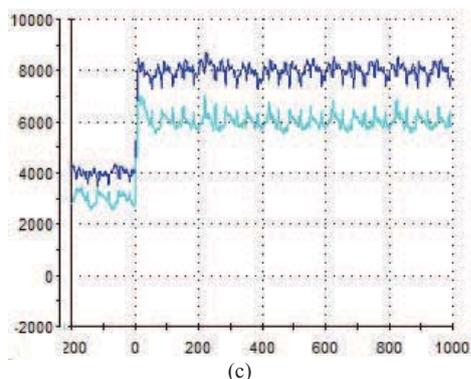


Fig. 9. Experimental results of 10 kVA converter: a) Three-phase grid injection current and grid voltage in phase A; b) Grid injection current in phase A and its reference; c) Active and reactive power injection.

V. CONCLUSIONS

This work proposed a Proportional Resonant current controller tuning method for grid connected power converters with grid linked LCL+trap filter. By analyzing the analytical relationship among the gains of the controller, the open-loop cross-over frequency and the phase margin. A formula for calculating the gains of the controller is found. By analyzing the interaction among different system variables, control tuning range is determined. Combining the formula and the tuning range, different solutions of controller gains are obtained, among which an optimal one is selected according to the design requirement.

This tuning method has been applied to multiple cases, not all presented in this paper, and more results will be presented in an extended version. The simulation results of 100 kW system and experimental results of 10 kVA system are presented, and both results have demonstrated the method to be effective.

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