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Robust Control Parameters Design of PBC Controller for LCL-Filtered Grid-Tied Inverter

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Abstract- **Owing to the strong robustness against system parameter changes and external perturbations, the Passivity-Based Control (PBC) has been widely adopted in Grid-Tied Inverter (GTI). However, for a PBC-based GTI with LCL-filter, there are three damping gains and two interactively-coupled-feedforward terms in the control loop, resulting in the controller design challenge for engineers. In order to help electrical engineers to well design the PBC controller for LCL-filtered GTI, a new design of the damping gains is proposed, by limiting the inherent steady-state error of grid-injected current. Furthermore, the state observer is also adopted to reduce the number of sensors. The robustness against the parameters shift and wide grid impedance variation is also addressed. The effectiveness of the proposed control design strategy will be verified through experimental results on a 3 kW/3-phase/110V experimental lab setup.** [1]

Index Terms**—Passivity-based control,** ***LCL*** **filter, Grid-tied inverter, Damping gain, Steady-state error.**

I. INTRODUCTION

In recent years, the development of renewable energy has received significant attention. In renewable generation systems, voltage source Grid-Tied Inverter (GTI) is a key device for linking the power generation equipment to the power grid [1]. In order to satisfy the harmonic standards of grid-injected current, generally, the output filter of GTI has to be adopted, where the LCL filter is utilized in most situations, owing to its good performance in the harmonic attenuation and low cost of metallic devices [2]. However, the LCL filter may cause possible resonance and great difficulty in the controller design of GTI, due to parameters shift as well as wide equivalent grid impedance variation [3].

Many conventional linear control methods, such as proportional-integral (PI) and proportional-resonant (PR) with passive [4]-[6], active [7]-[11] or hybrid [12]-[14] damping techniques, had been widely studied to suppress the possible resonance. Although these mentioned control methods can achieve stability under the condition of wide grid impendence variation, many shortcomings still exist. For instance, the passive damping method will lead to extra damping power losses, while the active damping (AD) method like the capacitor current feedback (CCF) will increase the number of sensors and the extra measure should be taken [3],[7],[8]. What's more, if the characteristic resonant frequency (f_r) of the LCL based system using conventional CCF AD control method is equal to 1/6 of the sampling frequency (f_s), the digitally controlled GTI with the total delay of 1.5/f_s (1/f_s for sampling delay and 0.5/f_s for PWM delay) can be hardly stable [15]. Although this problem can be solved by capacitor current proportional-integral positive feedback, more sensors should be utilized [8]. In other words, the parameters shift of the LCL filter should be limited in a reasonable range for the CCF AD control based GTI, if no extra measure is taken.

Due to the above shortcomings in conventional control method, a series of nonlinear control methods, which may provide a better solution for an essential nonlinear system such as GTI, have been also studied more and more frequently, including the adaptive control [16]-[19], the deadbeat control [20]-[22], the model predictive control [23]-[26], the slide model control [27],[28], and the Passivity-Based Control (PBC) [30]-[44], etc.

As one of the attractive nonlinear control methods, the PBC is a model-based control method, which contains the energy-shaping and damping-injection [45]. With the merits of clear physical significance, simple modeling process and strong robustness to system parameter changes, the PBC has become a powerful control strategy in power electronics [38], [42], [45]. It also can be seen as a hybrid control scheme, since it includes the instruction predicting feedforward control, the disturbance feedforward control, the decoupling control and the negative feedback control [39]. Currently, the PBC method had been successfully used in the switched reluctance based wind system [30], railway systems [31], [32], the energy storage systems [33],[34], the AC/DC converter [35], the islanded AC microgrid [36] and the GTI systems [37]-[43], etc. In the application of GTI systems, the PBC controllers have

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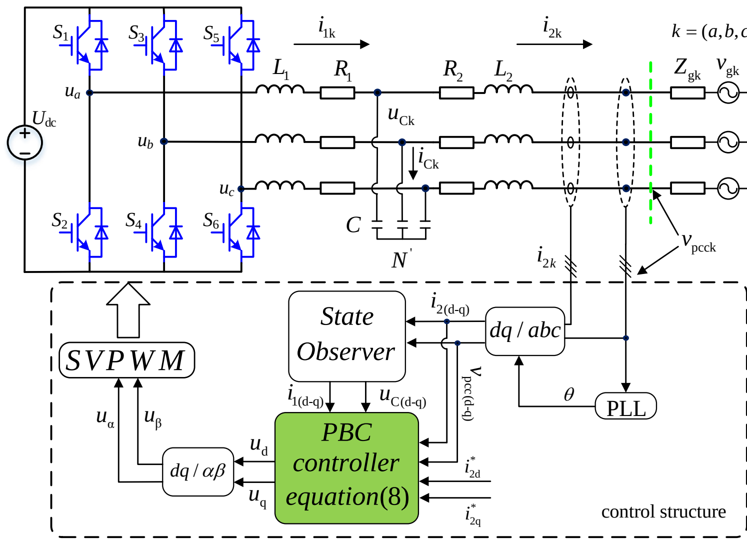

Fig. 1. Control diagram of LCL-filtered GTI using the PBC controller.

different structures with different power filters, such as L filter based systems [37]-[39], LC filter based systems [40],[41],[42] and LCL filter based system [43],[44]. Whatever, it is no doubt that the control parameters, named as the damping gains in the PBC controller, are a very important factor during the design.

In [39], the damping gain in the PBC controller was well designed through discrete root locus and unit step response, where the upper limit of damping gain was obtained through discrete root locus and then the suitable value was selected via unit step response. In [40], the damping gain was determined by the traditional analysis method, which is similar to the design of the proportional coefficient for a PI controller. In [41], [42] the damping gain was selected through attenuating the delay influence on the inverter to realize a passive system. Note that in [39]-[42], there is only one PBC control loop exists. Therefore, during designing this damping gain, the effect of feedforward term can be neglected since it has no effect on system stability.

However, different from the objects in [39]-[42], the PBC based LCL-filtered GTI system need three state variables to participate in the calculation, thus there are three damping gains and two interactively-coupled-feedforward terms in the control loop, which are shown in Fig. 2 with different colors. In this case, the conventional design methods using opened-loop design method as introduced in [39]-[42] are not available, since the feedforward terms cannot be directly neglected anymore, resulting in the controller design challenge for engineers. Although in [43], the PBC controller is designed for the LCL-filtered system, however, how to choose the damping gains had not been addressed. In [44], the damping gains design was introduced, nevertheless, the delay issue limiting the control bandwidth was ignored.

In order to help electrical engineers to well design the PBC controller for LCL-filtered GTI, a new step-by-step control parameters design strategy, which is based on the constraint condition of limiting the inherent steady-state error of grid-injected current, is proposed in this paper. Furthermore, in order to reduce the number of sensors as well as costs, the state observer [46], [47] is also adopted in the proposed PBC controller. At the same time, an additional integral regulator is finally adopted to achieve zero steady-state error of grid-injected current.

The rest of this paper is organized as follows. The mathematical model and the deduction of PBC control law of LCL-filtered GTI are first introduced in section II. Then, the design of three damping gains by limiting the inherent steady-state error of grid-injected current is proposed in section III. A brief introduction for the state observer and method to achieve the zero steady-state error of grid-injected current are presented in section IV. Next, a 3 kW / 3-phase / 110 V experimental device is constructed with dSPACE DS1202 to verify the effectiveness of the proposed control parameters design strategy in section V. Finally, there is a conclusion in section VI.

II. Mathematical Model and PBC Control Law Deduction of LCL-Filtered Grid-Tied Inverter

A. Mathematical Model

The topological structure diagram of LCL-filtered GTI using the PBC controller is shown in Fig. 1. The LCL filter is represented by L_1, C, and L_2, where R_1 and R_2 represent the line resistances and parasitic resistances of L_1 and L_2, respectively. The grid-injected current is represented by i_{2k} (k=a, b, c), which is sensed for the overcurrent protection and closed-loop feedback control. v_{pcck} (k=a, b, c) represents the voltage of Point of Common Coupling (PCC), which is sensed for the synchronization and input of the PBC controller. u_k, i_{1k} and i_{Ck} (k=a, b, c) are the inverter side voltage and current and capacitor current respectively. v_{gk} (k=a, b, c) and Z_g denote the ideal grid voltage and the equivalent grid impedance, respectively. Z_g is complex impedance with the inductance of L_g and the resistance of R_g. The DC bus voltage is denoted by U_{dc}. The control structure is given in the dash frame as shown in Fig. 1, where the sampling, the transformation and the

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control process are illustrated, and the Space Vector Pulse Width Modulation (SVPWM) technology is employed to obtain the driving signals.

Similar as introduced in [43], the mathematical model of the LCL-filtered GTI can be deduced by applying the Kirchhoff voltage and current laws. And then we obtain,

$$\begin{cases} L_1\dfrac{di_{1k}}{dt}+R_1 i_{1k}+u_{Ck}=u_k \\ C\dfrac{du_{Ck}}{dt}+i_{2k}-i_{1k}=0 \qquad ,(k=a,b,c) \\ L_2\dfrac{di_{2k}}{dt}+R_2 i_{2k}-u_{Ck}=-v_{pcck} \end{cases} \tag{1}$$

In order to get a better control performance, the *a-b-c* to *d-q* transformation is applied to (1), and the new equations in *d-q* coordinates can be described as,

$$\begin{cases} L_1\dfrac{di_{1d}}{dt}+R_1 i_{1d}-\omega L_1 i_{1q}+u_{Cd}=u_d \\ L_1\dfrac{di_{1q}}{dt}+R_1 i_{1q}+\omega L_1 i_{1d}+u_{Cq}=u_q \\ C\dfrac{du_{Cd}}{dt}+i_{2d}-\omega C u_{Cq}-i_{1d}=0 \\ C\dfrac{du_{Cq}}{dt}+i_{2q}+\omega C u_{Cd}-i_{1q}=0 \\ L_2\dfrac{di_{2d}}{dt}+R_2 i_{2d}-\omega L_2 i_{2q}-u_{Cd}=-v_{pccd} \\ L_2\dfrac{di_{2q}}{dt}+R_2 i_{2q}+\omega L_2 i_{2d}-u_{Cq}=-v_{pccq} \end{cases} \tag{2}$$

To simplify analysis and stability judgments, the Euler Lagrange (EL) model is adopted to describe the whole system. Define state variables as $x=\left(i_{1d} \quad i_{1q} \quad u_{Cd} \quad u_{Cq} \quad i_{2d} \quad i_{2q}\right)^T$, then the equation (2) can be rewritten in the EL form as,

$$M\dot{x}+Jx+Rx=u \tag{3}$$

where

$$M=\begin{pmatrix} L_1 & 0 & 0 & 0 & 0 & 0 \\ 0 & L_1 & 0 & 0 & 0 & 0 \\ 0 & 0 & C & 0 & 0 & 0 \\ 0 & 0 & 0 & C & 0 & 0 \\ 0 & 0 & 0 & 0 & L_2 & 0 \\ 0 & 0 & 0 & 0 & 0 & L_2 \end{pmatrix},$$

$$J=\begin{pmatrix} 0 & -\omega L_1 & 1 & 0 & 0 & 0 \\ \omega L_1 & 0 & 0 & 1 & 0 & 0 \\ -1 & 0 & 0 & -\omega C & 1 & 0 \\ 0 & -1 & \omega C & 0 & 0 & 1 \\ 0 & 0 & -1 & 0 & 0 & -\omega L_2 \\ 0 & 0 & 0 & -1 & \omega L_2 & 0 \end{pmatrix},$$

$$R=\begin{pmatrix} R_1 & 0 & 0 & 0 & 0 & 0 \\ 0 & R_1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & R_2 & 0 \\ 0 & 0 & 0 & 0 & 0 & R_2 \end{pmatrix},$$

and $u=(u_d \quad u_q \quad 0 \quad 0 \quad -v_{pccd} \quad -v_{pccq})^T$.

According to the passivity theory, the LCL-filtered system is strictly passive as introduced in [43]. Therefore, the PBC can be applied to design the controller [45].

B. PBC Control Law Deduction

Define the reference state variables as,

$$x^*=\left(i_{1d}^* \quad i_{1q}^* \quad u_{Cd}^* \quad u_{Cq}^* \quad i_{2d}^* \quad i_{2q}^*\right)^T \tag{4}$$

If the error vector is defined as $x_e = x^* - x$, then an error EL equation can be obtained as,

$$\begin{aligned} &M\left(\dot{x}^*-\dot{x}_e\right)+J(x^*-x_e)+R\left(x^*-x_e\right)=u \\ &M\dot{x}_e+Jx_e+Rx_e=M\dot{x}^*+Jx^*+Rx^*-u \end{aligned} \tag{5}$$

In order to accelerate the speed of the convergence, a damping matrix R_d can be added to the error system. Then, the injection damping matrix and new dissipation matrix are obtained as

$$\begin{aligned} &R_d=diag\{r_3 \quad r_3 \quad r_2 \quad r_2 \quad r_1 \quad r_1\}, \\ &R_{new}=R+R_d \end{aligned} \tag{6}$$

where r_1, r_2, r_3>0. Substitution (6) into (5), the new error equation can be obtained as,

$$M\dot{x}_e+Jx_e+R_{new}x_e=M\dot{x}^*+Jx^*+Rx^*+R_d x_e-u \tag{7}$$

According to (7), if x_e equals to zero, the left side of (7) also equals to zero. Then, expand (7), and the control law can be described in detail as,

$$\begin{cases} L_1\dfrac{di_{1d}^*}{dt}+R_1 i_{1d}^*-\omega L_1 i_{1d}^*+r_3\left(i_{1d}^*-i_{1d}\right)+u_{Cd}^*=u_d \\ L_1\dfrac{di_{1q}^*}{dt}+R_1 i_{1q}^*+\omega L_1 i_{1q}^*+r_3\left(i_{1q}^*-i_{1q}\right)+u_{Cq}^*=u_q \\ C\dfrac{du_{Cd}^*}{dt}-\omega C u_{Cq}^*+r_2(u_{Cd}^*-u_{Cd})+i_{2d}^*-i_{1d}^*=0 \\ C\dfrac{du_{Cq}^*}{dt}+\omega C u_{Cd}^*+r_2(u_{Cq}^*-u_{Cq})+i_{2q}^*-i_{1q}^*=0 \\ L_2\dfrac{di_{2d}^*}{dt}+R_2 i_{2d}^*-\omega L_2 i_{2d}^*+r_1(i_{2d}^*-i_{2d})-u_{Cd}^*=-v_{pccd} \\ L_2\dfrac{di_{2q}^*}{dt}+R_2 i_{2q}^*+\omega L_2 i_{2d}^*+r_1(i_{2q}^*-i_{2q})-u_{Cq}^*=-v_{pccq} \end{cases} \tag{8}$$

According to (8), the equivalent system diagram of LCL-filtered GTI using the PBC controller in the Laplace domain is plotted in Fig. 2.

As shown in Fig. 2, the parameters in the controller are marked with subscript "e" to distinguish control parameters and actual parameters. For example, sL_{1e} and R_{1e} are control parameters, while sL_1 and R_1 are actual parameters in the physical object. Due to the symmetrical structure, only the d-axis is drawn here, where $e^{-1.5sTs}$ represents the calculation and pulse width modulation delay which almost selected as 1.5 T_s (T_s is the switching or sampling cycle). Note that in order to further analyze the effect caused by the equivalent grid-impedance, the total inductance in the grid side is represented as L_t (L_2+L_g), while the equivalent resistor of L_t is still converted into R_2.

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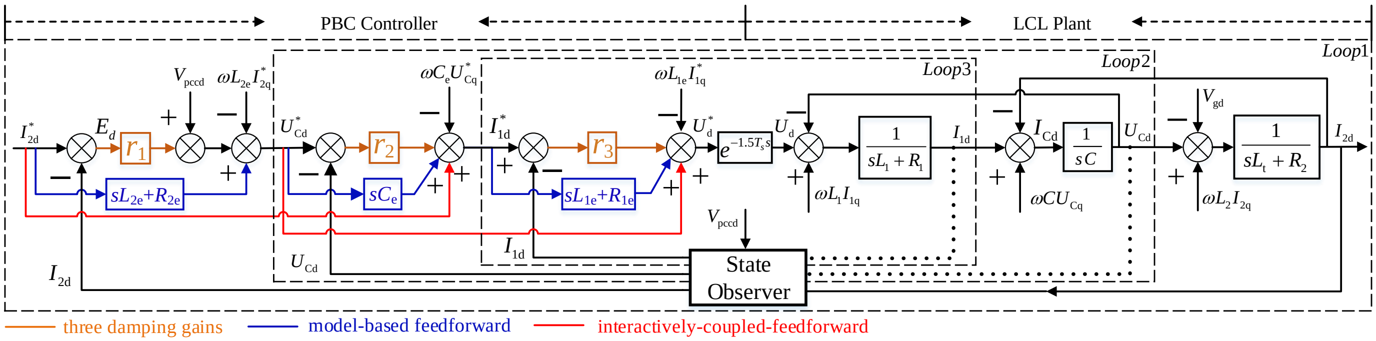

Fig. 2. Equivalent system diagram of LCL-filtered GTI using the PBC controller with state observer.

C. Design Challenge of Damping Gains

From Fig. 2, it can be found that due to the using of three state variables (i_1, U_C, i_2), there are two interactively-coupled-feedforward terms marked in red, three model-based feedforward terms marked in blue, and three damping gains terms (r_1, r_2, r_3) marked in brown. Therefore, the whole controller is a complex multi-loop controller.

According to the traditional control theory, a double-loop controller can be well designed, based on the relationship of control bandwidth between the inner loop and the outer one [49]. The similar method is adopted for a three-loop controller [50]. It should be pointed out that the GTI with the classic three-loop controller has poor dynamic performance under the weak grid condition in theory, especially when the switching frequency is not so high. Further, as shown in Fig. 2, there are two interactively-coupled-feedforward terms between the three control loops, and every loop is coupled connected with the interactively-coupled-feedforward terms in red. Therefore, there is a strong coupling relationship between these control loops, resulting in more design difficulty with the traditional three-loop design theory as introduced in [50].

Generally, the trial and error method is often adapted to design control parameters in a nonlinear system, but at the cost of computing resources, especially when there is no suitable guidance. Furthermore, it is also difficult to use the intelligent optimization algorithm to design these three damping gains, because the optimization cost function is difficult to be defined for this system, and there is not enough tangible constrains.

Therefore, it is very valuable to introduce a practical parameters design strategy for LCL-filtered GTI using the PBC controller, which does help for the industrial applications. Note that in theory, more sensors should be adopted to obtain the state variables of the PBC controller, resulting in extra sensing costs. In this paper, a separated loop control parameters design strategy will be proposed, and the state observer technology will be also utilized to reduce the number of sensors. The detailed description will be given in the next two sections.

III. Proposed Design of Damping Gains by Limiting the Inherent Steady-State Error of Grid-Injected Current

As shown in Fig. 2, the control system can be divided into three control loops from the inner to the outer, which are named as loop3, loop2 and loop1, respectively. Firstly, through the analysis for the inherent steady-state error of grid-injected current, the relationship of three damping gains can be roughly determined. Then, an efficient trial and error procedure can be obtained by using the stability criterion. Finally, due to the stable margin considered in the design procedure, robustness can be easily obtained.

A. Analysis on the Inherent Steady-State Error of Grid-Injected Current

From Fig. 2, it can be seen that the PBC based GTI is a Multi-Input Multi-Output (MIMO) system, when the coupling path between d- and q- axis is considered. In order to analyze the performance of the PBC controller, the closed-loop transfer function matrix of the whole system is deduced as,

$$\begin{aligned}\begin{bmatrix} I_{2\mathrm{d}} \\ I_{2\mathrm{q}} \end{bmatrix} &= \begin{bmatrix} G_{\mathrm{ddc1}} & G_{\mathrm{dqc1}} \\ G_{\mathrm{qdc1}} & G_{\mathrm{qqc1}} \end{bmatrix} \begin{bmatrix} I_{2\mathrm{d}}^* \\ I_{2\mathrm{q}}^* \end{bmatrix} \\ &= \begin{bmatrix} \dfrac{AD+BC}{C^2+D^2} & \dfrac{AC-BD}{C^2+D^2} \\ \dfrac{BD-AC}{C^2+D^2} & \dfrac{AD+BC}{C^2+D^2} \end{bmatrix} \begin{bmatrix} I_{2\mathrm{d}}^* \\ I_{2\mathrm{q}}^* \end{bmatrix}\end{aligned} \quad (9).$$

The expressions of A, B, C, D are in the appendix. As time $t \to \infty$, then $s \to 0$, in the steady-state, the G_{ddc1} and G_{dqc1} in (9) can be rewritten as,

$$\begin{aligned}\lim_{t\to\infty} G_{\mathrm{ddc1}}(t) &= \lim_{s\to 0} G_{\mathrm{ddc1}}(s) = \frac{MN+PQ}{N^2+Q^2} \\ \lim_{t\to\infty} G_{\mathrm{dqc1}}(t) &= \lim_{s\to 0} G_{\mathrm{dqc1}}(s) = \frac{MQ-NP}{N^2+Q^2}\end{aligned} \quad (10),$$

where

$$\begin{aligned}M &= r_1 r_2 R_{1\mathrm{e}} + r_2 r_3 R_{2\mathrm{e}} + r_2 R_{1\mathrm{e}} R_{2\mathrm{e}} - \omega^2 r_1 L_{1\mathrm{e}} C_{\mathrm{e}} - \omega^2 L_{1\mathrm{e}} r_2 L_{2\mathrm{e}} - \omega^2 r_3 C_{\mathrm{e}} L_{2\mathrm{e}} \\ &\quad - \omega^2 C_{\mathrm{e}} R_{1\mathrm{e}} L_{2\mathrm{e}} - \omega^2 L_{1\mathrm{e}} C_{\mathrm{e}} R_{2\mathrm{e}} + r_1 r_2 r_3 + r_1 + r_3 + R_{1\mathrm{e}} + R_{2\mathrm{e}} \\ N &= r_1 r_2 R_{1\mathrm{e}} + r_2 r_3 R_2 + r_2 R_{1\mathrm{e}} R_2 - \omega^2 r_1 L_{1\mathrm{e}} C_{\mathrm{e}} - \omega^2 r_2 L_{1\mathrm{e}} L_2 - \omega^2 r_3 C L_2 \\ &\quad - \omega^2 C R_1 L_2 - \omega^2 L_1 C R_2 + r_1 r_2 r_3 + r_1 + r_3 + R_1 + R_2 \\ P &= \omega r_1 C_{\mathrm{e}} R_{1e} + \omega r_2 R_{1\mathrm{e}} L_{2\mathrm{e}} + \omega r_2 L_{1\mathrm{e}} R_{2\mathrm{e}} + \omega r_3 C_{\mathrm{e}} R_{2\mathrm{e}} + \omega r_1 r_2 L_{1\mathrm{e}} + \omega r_2 r_3 L_{2\mathrm{e}} \\ &\quad + \omega r_1 r_3 C_{\mathrm{e}} + \omega C_{\mathrm{e}} R_{1\mathrm{e}} R_{2\mathrm{e}} - \omega^3 L_{1\mathrm{e}} C_{\mathrm{e}} L_{2\mathrm{e}} + \omega L_{1\mathrm{e}} + \omega L_{2\mathrm{e}} \\ Q &= \omega r_1 C_{\mathrm{e}} R_{1\mathrm{e}} + \omega r_2 R_{1\mathrm{e}} L_2 + \omega r_2 L_{1\mathrm{e}} R_2 + \omega r_3 C R_2 + \omega r_1 r_2 L_{1\mathrm{e}} + \omega r_2 r_3 L_2 \\ &\quad + \omega r_1 r_3 C_{\mathrm{e}} + \omega C R_1 R_2 - \omega^3 C L_1 L_2 + \omega L_1 + \omega L_2\end{aligned}$$

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From (10), it can be known that if and only if $L_1 = L_{1e}$, $R_1 = R_{1e}$, $C = C_e$, $L_2 = L_{2e}$, $R_2 = R_{2e}$, then G_{ddc1} equals to one and G_{dqc1} equals to zero, which indicates the system can track the reference value with the zero steady-state error. However, it is impracticable in a real physical system, since the parameter drifting always occurs and the line impedance is changing with the time. If neglecting the external disturbances, the inherent steady-state error of grid-injected current will only depend on r_1, r_2, r_3, $L_{1(e)}$, $R_{1(e)}$, $C_{(e)}$, $L_{2(e)}$ and $R_{2(e)}$. Since r_1, r_2 and r_3 are the damping gains can be artificially tuned, the inherent steady-state error can be controlled within a certain range. Note that the order of magnitude of capacitor (10^{-6}) and inductor (10^{-3}) is very small, many terms related to them, such as the product of $\omega^3 L_{1e} C_e L_{2e}$, can be neglected. Further, the equivalent resistance of inductor is also very small, if a very small value of r_2 is selected (suggest at least 100 times smaller than r_1 and r_3), many terms related to r_2, such as the products of $r_1 r_2 R_{1e}$, $r_2 r_3 R_{2e}$, and $r_2 R_{1e} R_2$, can be also neglected. And then the percentage of inherent steady-state error of grid-injected current can be simplified as

$$e\% \approx \frac{R_1 - R_{1e} + R_2 - R_{2e}}{r_1 r_2 r_3 + r_1 + r_3 + R_1 + R_2} \times 100\% \qquad (11).$$

From (11), it can be gotten that the larger values of r_1, r_3, the smaller inherent steady-state error. Note that, the outer loop damping gain of r_1, which determines the control bandwidth, had better be set as large as possible.

B. Step-by-Step Design of Damping Gains

As described in (9), the transfer function matrix is very complex, which causes trouble to select the control parameters and analyze the system stability. If the coupling path between *d*- and *q*- axis is not addressed, the situation will become much simpler. Note that, as introduced in [50], the coupling terms between *d*- and *q*- axis only have little effect on system stability, and they can be neglected during the controller design. Further, in order to simplify the calculation, the resistances in the controller and physical device are neglected, due to their small values. Based on Fig. 2, the closed-loop transfer function (neglecting coupling terms between *d*- and *q*- axis) of the whole system can be obtained as,

$$G_{c1}(s) = \frac{I_{2d}(s)}{I_{2d}^*(s)} = \frac{\begin{array}{c} s^3 C_e L_{1e} L_{2e} + s^2 (r_2 L_{1e} L_{2e} + r_3 C_e L_{2e} + r_1 C L_{1e}) \\ + s(r_2 r_3 L_{2e} + r_1 r_2 L_{1e} + r_1 r_3 C_e + L_{1e} + L_{2e}) + r_1 r_2 r_3 + r_1 + r_3 \end{array}}{\begin{array}{c} s^4 1.5 T_s C L_1 L_t + s^3 C L_1 L_t \\ + s^2 (r_1 C_e L_{1e} + r_3 C L_t + L_{1e} r_2 L_2 + 1.5 T_s L_1 + 1.5 T_s L_t) \\ + s(r_1 r_2 L_{1e} + r_1 r_3 C_e + r_2 r_3 L_2 + L_1 + L_t) + r_1 r_2 r_3 + r_1 + r_3 \end{array}} \qquad (12).$$

The PWM and sampling delay term of $e^{-1.5T_s s}$ is approximated as $1/(1.5sT_s + 1)$ [38]. According to the Routh stability criterion, the system will be stable, if $f_1(r_1, r_2, r_3)$ and $f_2(r_1, r_2, r_3)$ are greater than zero, which is described as,

$$f_1(r_1, r_2, r_3) = \frac{r_1 C_e L_{1e}}{1.5 T_s C L_1 L_t} + \frac{r_3}{1.5 T_s L_1} + \frac{r_2 L_{1e}}{1.5 T_s C L_1} - \frac{r_1 r_2 L_{1e}}{C L_1 L_t} - \frac{r_1 r_3 C_e}{C L_1 L_t} - \frac{r_2 r_3}{C L_1} \qquad (13a)$$

$$f_2(r_1, r_2, r_3) = \frac{r_1 r_2 L_{1e}}{C L_1 L_t} + \frac{r_1 r_3 C_e}{C L_1 L_t} + \frac{r_2 r_3}{C L_1} + \frac{1}{C L_t} + \frac{1}{C L_1} - \frac{r_1 r_2 r_3 + r_1 + r_3}{[r_1 C_e L_{1e} + r_2 L_{1e} L_t + r_3 C L_t - 1.5 T_s (r_1 r_2 L_{1e} + r_1 r_3 C_e + r_2 r_3 L_t)]} \qquad (13b)$$

Based on the analysis of the inherent steady-state error and the Routh stability criterion, the separated loop design strategy of the PBC controller for LCL-filtered GTI is proposed, where the detailed design procedure is illustrated in Fig. 3.

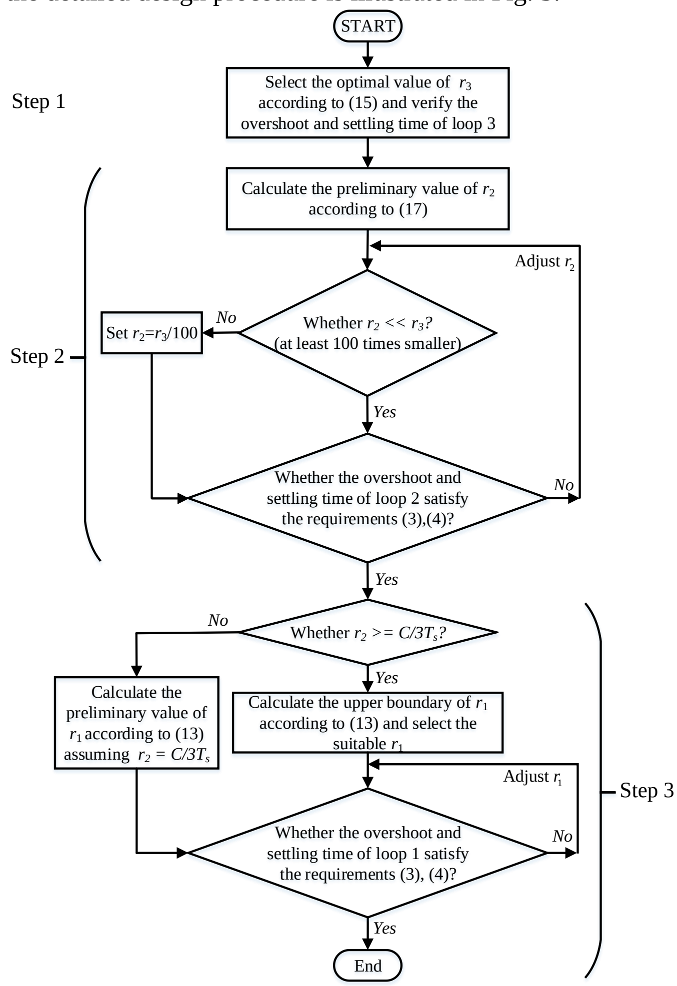


Fig.3. Proposed Design Strategy for PBC-based GTI with LCL-filter.

Some constraints are defined beforehand as,

1) The LCL filter is designed according to the current ripple, the rated power, and the IEEE harmonic standard [2].
2) The damping gain r_2 should be 100 times smaller than r_1 and r_3 at least.
3) The settling time of the nested inner loop should be at least 4 times faster than the related outer loop.
4) The overshoot of every closed-loop should be no more than 30 percent during the unit step response.

Before the control parameters design, the values of the LCL filter must be selected, where the filter design criteria can be found in [2], [48]. In our case, a 3 kW/3-phase/110V experimental setup with the sampling frequency of 10 kHz is constructed. According to the calculation formula in [2], [48], then L_1, L_2 and C are chosen as 1.2 mH, 1.2 mH, and 6 μF, respectively, where the resonant frequency is about 2654 Hz.

Step 1 Select the Optimal Damping Gain of Loop3 (r_3)

For three damping gains, the damping gain of loop3 (r_3) should be first determined, where the closed-loop transfer function of loop3 is,

$$G_{c3}(s)=\frac{I_{1d}(s)}{I_{1d}^{*}(s)}=\frac{sL_{1e}+r_3}{1.5s^2T_sL_1+sL_1+r_3}=\frac{s\dfrac{L_{1e}}{1.5T_sL_1}+\dfrac{r_3}{1.5T_sL_1}}{s^2+s\dfrac{1}{1.5T_s}+\dfrac{r_3}{1.5T_sL_1}} \quad (14).$$

From (14), it can be seen that the closed-loop transfer function of loop3 is similar to a second-order system when the total time delay is addressed. According to the control theory, when the damping ratio ξ is $\sqrt{2}/2$, the system can achieve the optimal model. Thus, r_3 can be calculated as,

$$r_3=\frac{L_1}{6\xi^2T_s} \quad (15).$$

Taking L_1=1.2 mH, $\xi=\sqrt{2}/2$, T_s=1/10000 S into (15), r_3 = 4 can be obtained in our case. The unit step response of G_{C3} with r_3 = 4 is plotted in Fig. 4. Fig. 4 shows that the settling time is about 1.03 ms and the overshoot is about 20 percent, which indicates that r_3 = 4 is the optimal selection.

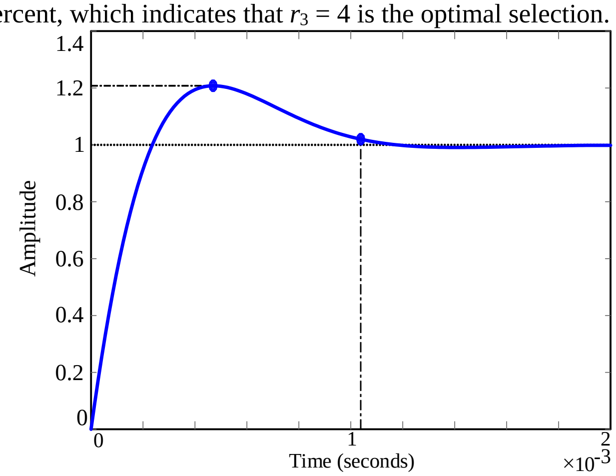


Fig.4 Unit step response of G_{c3} with r_3 = 4.

Step 2 Select the Suitable Damping Gain of Loop2 (r_2)

As r_3 is determined, r_2 can be selected next. At the same time, a large sufficient r_1 is expected to satisfy the bandwidth requirement. According to (13), the relationship curves between r_2 and the stable range of r_1 are plotted in Fig. 5, where the X-axis represents the value of r_1 which ranges from 0-30, and the Y-axis represents the calculated result of f_1 and f_2 respectively.

From Fig. 5, it can be seen that the stable range of r_1 is increased as r_2 decreases. Therefore, a larger r_1 can be selected when a smaller r_2 is adopted (suggest at least 100 times smaller than r_1 and r_3), which is consistent with the aim of reducing the inherent steady-state error of grid-injected current. However, if r_2 closes to 0 infinitely, the stable range of r_1 tends to be gigantic. For example, taking r_2 =0.01, the stable range of r_1 is $r_1 \geq 0$. And the precise stable range of r_1 can't be exactly found, due to the approximated expression of the time delay in (13).

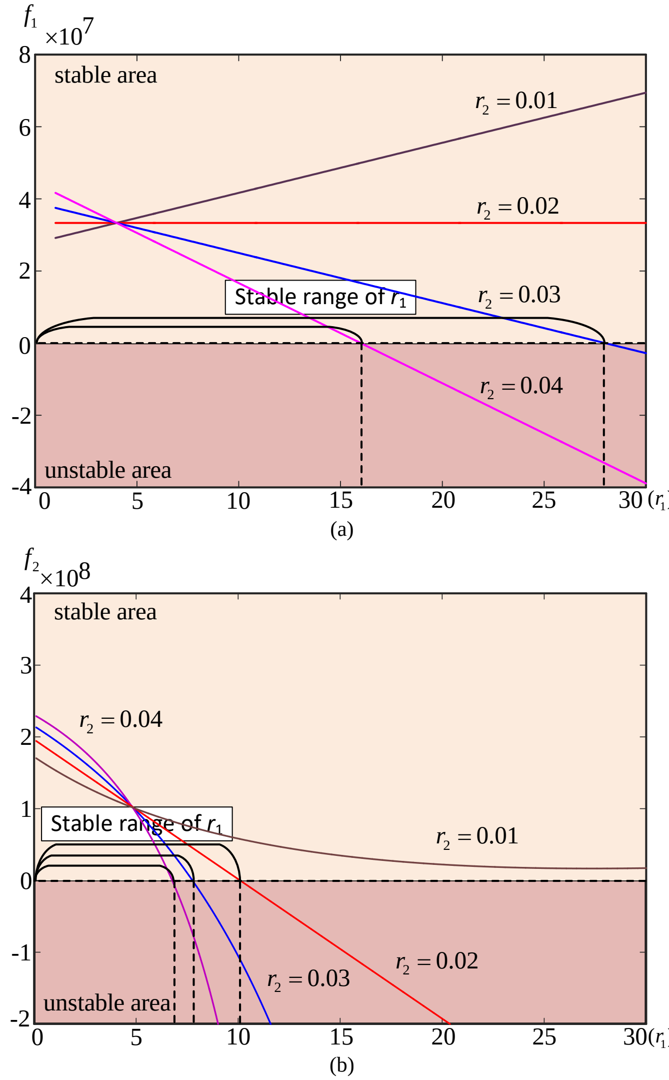


Fig. 5. Stable relationship between r_2 and r_1 according to (13): (a) f_1, (b) f_2.

Therefore, the critical r_2 is suggested to be defined as the preliminary value, which makes the slope of the line equal to 0 in Fig. 5(a). And it can be obtained via calculating the partial derivation of f_1 in (13a) as

$$\frac{\partial f_1}{\partial r_1}=0 \quad (16).$$

If the parameters of L_1 and C do not drift, the preliminary value of r_2 can be calculated as,

$$r_2=C/3T_s \quad (17).$$

It can be found from (17) that the critical value of r_2 is determined by C and T_s. In order to expend the calculable range of r_1, the small C and large T_s are preferred. Taking C = 6uF, T_s = 1/10000 S into (17), the preliminary value of r_2 is calculated as 0.02 in our case.

In order to analyze the performance of the middle loop and verify requirements of (3) and (4), the closed-loop transfer function of loop2 is deduced as,

$$G_{c2}(s)=\frac{U_{Cd}(s)}{U_{Cd}^{*}(s)}=\frac{s^2L_{1e}C_e+sr_3C_e+sL_{1e}r_2+r_2r_3+1}{1.5T_s s^3CL_1+s^2CL_1+s(1.5T_s+r_3C+L_{1e}r_2)+r_2r_3+1} \quad (18)$$

Under the condition that L_{1e}=L_1, C_e = C, r_2 = 0.02, r_3 = 4, the unit step response of G_{c2} is plotted in Fig. 6, where the overshoot is less than 20 percent and the setting time is about 4.46 ms. The requirements of (3) and (4) are both satisfied.

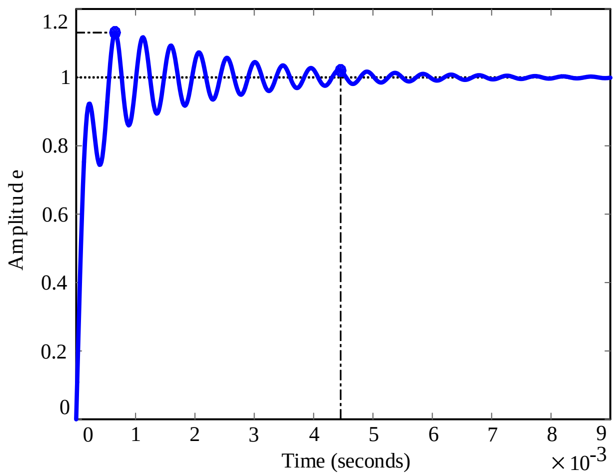


Fig.6. Unit step response of G_{c2} with r_3 = 4, r_2 = 0.02.

Step 3 Select the Suitable Damping Gain of Loop1(r_1)

If r_2 and r_3 have been determined, the stable range of r_1 can be obtained from (13). For instance, when r_2 = 0.02, r_3 = 4, the stable range of r_1 is greater than 0 according to f_1, while r_1 should be set between 0 and 10.1 according to f_2. If sufficient stable margins are addressed, r_1 had better be smaller than the upper boundary of 10.1 in our case, and r_1 = 8 is selected according to requirements of (3) and (4)，where the unit step response of loop1 is depicted in Fig. 7.

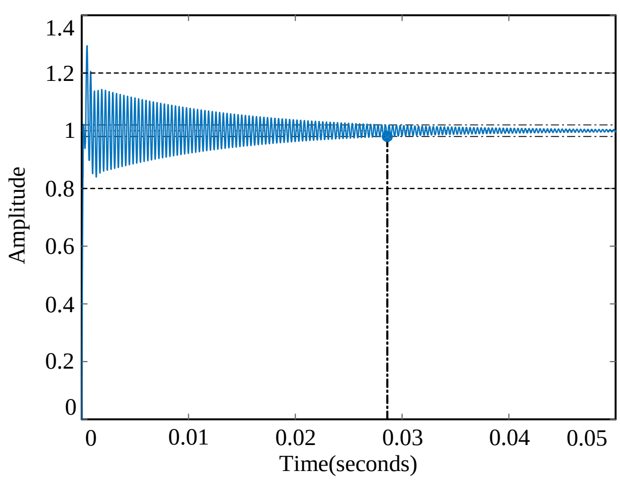
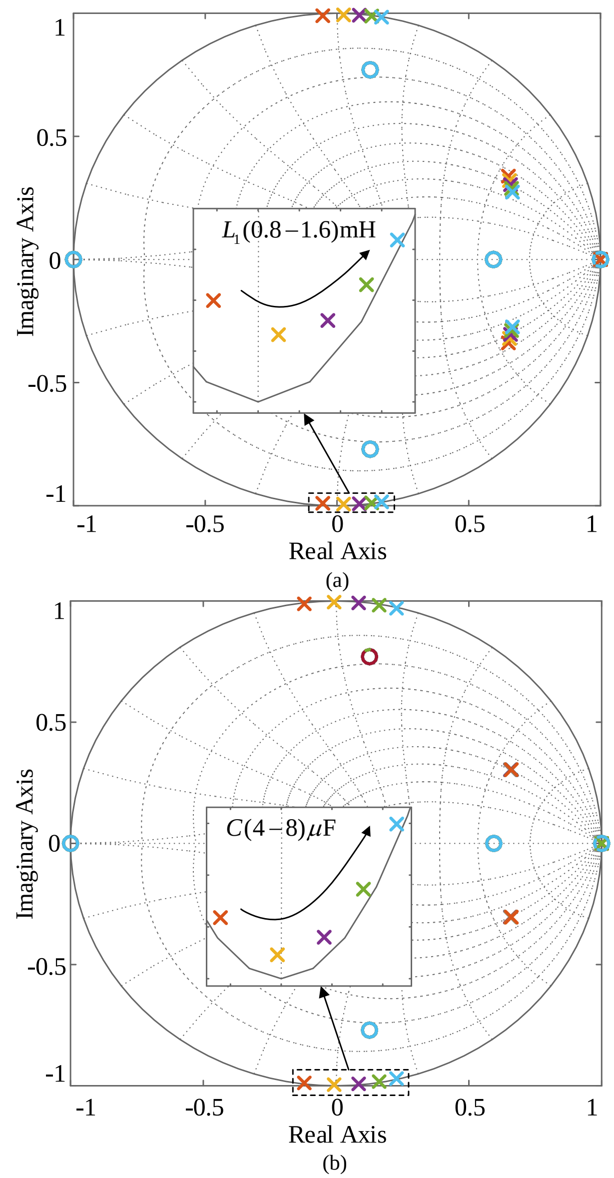


Fig.7. Unit step response of G_{c1} with r_1 = 8, r_2 = 0.02, r_3 = 4.

C. Robustness Analysis

The robustness of a controller is mainly manifested as its ability to suppress the adverse effect caused by the parameters drift of object. Assume that L_1 changes in the range of 0.8 mH ~ 1.6 mH (±33%), the variation of the capacitor is in the range of 4 μF~8 μF (±33%), and the total grid side inductance L_t = L_2 + L_g varies in the range of 0.8 mH ~ 6 mH (-33% ~ +400%). The closed-loop pole-zero maps in the discrete domain are drawn in Fig. 8 to investigate the robustness of the PBC controller under above situations. The control parameters used in the PBC controller are designed above, where r_1 = 8, r_2 = 0.02, r_3 = 4.

From Fig. 8, it can be seen that all the closed-loop poles are located in the unit circle, which indicates that the system is always stable. Further, as pointed in [15], the LCL-filtered GTI system with the conventional CCF AD method can be hardly stable, if f_r = 1/6 f_s when the total time delay is 1.5/f_s. Note that Fig. 8(c) proves that when f_r = 1/6 f_s, the proposed PBC controller can still stabilize the system well. Therefore, it can be concluded that compared with the conventional CCF AD method, the proposed PBC has higher robustness against the parameters drift.



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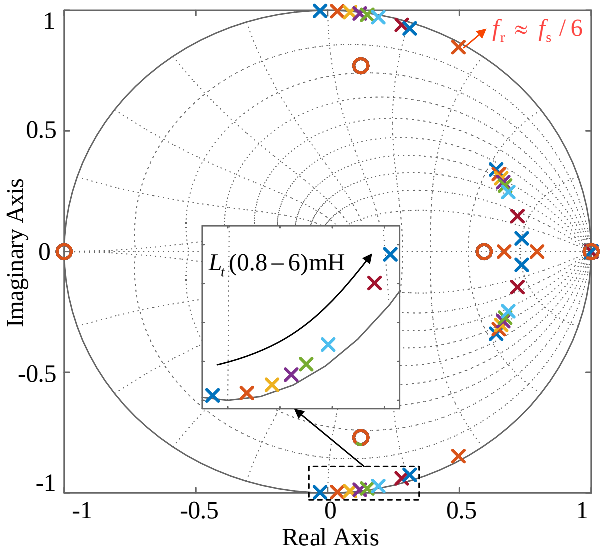

(c)

Fig. 8. Closed-loop pole-zero maps of PBC under parameters drift, (a) L_1 drifts, (b) C drifts, (c) L_t drifts and $f_r \approx f_s/6$.

III. State Observer to Save Sensors and Method to Achieve the Zero Steady-State Error

A. State Observer

In order to reduce the number of sensors, the state observer is adopted here, and a brief introduction of state observer is also given.

From (1), the state equation of the system can be rewritten as a matrix form in *d-q* axis,

$$\frac{dx'}{dt}=\underbrace{\begin{pmatrix} -\frac{R_1}{L_1}-j\omega & -\frac{1}{L_1} & 0 \\ \frac{1}{C} & -j\omega & -\frac{1}{C} \\ 0 & \frac{1}{L_2} & -\frac{R_2}{L_2}-j\omega \end{pmatrix}}_{A}x'+\underbrace{\begin{pmatrix} \frac{1}{L_1} \\ 0 \\ 0 \end{pmatrix}}_{B_1}u_k+\underbrace{\begin{pmatrix} 0 \\ 0 \\ -\frac{1}{L_2} \end{pmatrix}}_{B_2}v_{pccd\text{-}q},$$

$$x'=\begin{pmatrix} i_{1d-q} \\ u_{Cd\text{-}q} \\ i_{2d\text{-}q} \end{pmatrix}, i_{2d\text{-}q}=\underbrace{\begin{pmatrix} 0 & 0 & 1 \end{pmatrix}}_{C}x' \tag{19}$$

The state observer is designed with (19), and the needed state variables can be estimated using a full-order observer in practical applications, i.e.,

$$\frac{d\hat{x}'}{dt}=A\hat{x}'+B_1U_k+B_2V_{pccd\text{-}q}+L(i_{2d\text{-}q}-\hat{i}_{2d\text{-}q}), \hat{x}'=\begin{pmatrix} \hat{i}_{1d\text{-}q} \\ u_{Cd\text{-}q} \\ \hat{i}_{2d\text{-}q} \end{pmatrix} \tag{20},$$

$$\hat{i}_{2d\text{-}q}=C\hat{x}'$$

where $\hat{x}'$ is the estimated variable, and L is the observer gain vector. Because the i_{2k} and v_{pcck} are sensed and the converter voltage u_k is internally known, so i_{1k} and u_{Ck} can be obtained according to (20). Based on (19) and (20), the estimation error $x_e'=x'-\hat{x}'$ is calculated as,

$$\frac{dx_e'}{dt}=(A-LC)x_e' \tag{21}.$$

If the matrix of *(A-LC)* satisfies the Hurwitz condition, the estimated variables $\hat{x}'$ can converge to x' gradually, where the detailed analysis process can be found in [46], [47]. Fig. 9 shows the block diagram of the state observer, while its position in the whole system is depicted in Fig. 1.

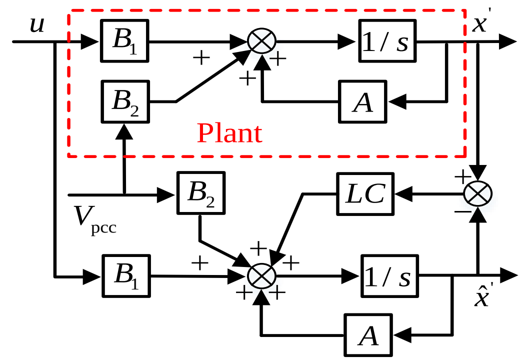

Fig. 9. Block diagram of state observer.

B. Method to Achieve Zero Steady-State Error

From the above analysis, we know that the PBC method will lead to an inherent steady-state error. And the damping gains can only reduce the value of inherent steady-state error of grid-injected current, but cannot eliminate it. However, the real steady-state error can be easily eliminated by using an additional integral regulator. In recent years, many modified PBC methods to eliminate or reduce the steady-state error had been proposed [38], [40]. In our case, the inherent steady-state error of grid-injected current also can be eliminated by using an additional integral regulator, where the damping gain of r_1 can be replaced by a PI controller with the proportion coefficient of $K_p = r_1 = 8$ and the integral coefficient of K_i = 800 in our case.

IV. Experimental Verifications

In this section, the effectiveness of the proposed control parameters design strategy is further studied. The 3 kW / 3-phase / 110 V GTI experimental lab setup is also developed. The three-phase grid is emulated with a Chroma 61830 three-phase grid simulator, the control algorithm is achieved via dSPACE DS1202 microlabbox, the DC voltage is given by a Chroma 62150H-600S DC power supply, a control desk project is developed to tune control parameters and reference value, and all the waveforms are captured from Yokogawa DL1640 digital oscilloscope. The experimental setup is shown in Fig. 10 and the parameters used for experiments are listed in Table I.

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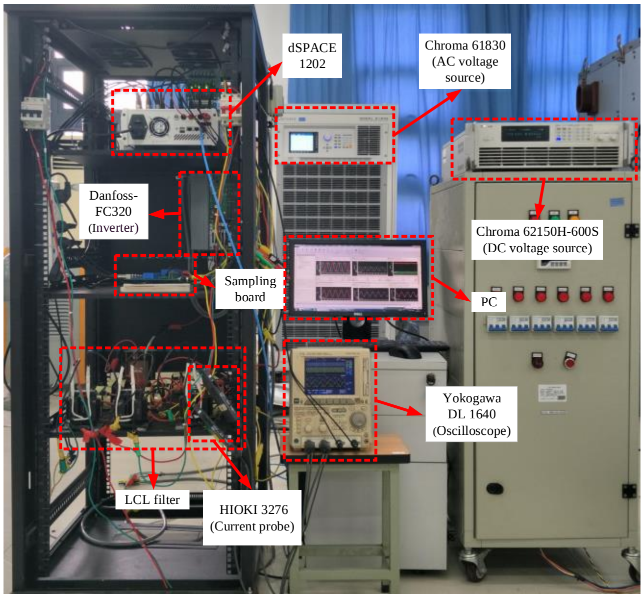

Fig.10. The experimental device of three-phase LCL-filtered GTI.

TABLE I
SIMULATION AND EXPERIMENT PARAMETERS FOR GTI

Symbols	*Description*	*Value*
V_g	Grid Voltage	110V(RMS)
f, f_s	Grid ,Switching and sampling frequency	50Hz,10kHz
L_1	Filter side inductor	1.2mH,2mH
C	Filter capacitor	6uF
L_2	Grid side inductor	1.2mH
L_g	Grid inductance	0mH, 4.8mH
U_{dc}	DC bus voltage	350V
L_{1e}, R_{1e}	estimated value of L_1 and R_1 in controller	1.2mH, 0.1Ω
L_{2e}, R_{2e}	estimated value of L_2 and R_2 in controller	1.2mH, 0.1Ω
r_1, r_2, r_3	Three damping gains	8, 0.02, 4
i_2^*	Reference current	12.86A(peak value)
K_p	Proportion coefficient	8
K_i	Integral coefficient	800

In order to verify the control performance of the PBC controller, many experiments are carried out next. Note that, the method to achieve zero steady-state error is not adopted at the beginning. The waveforms in Figs. 11 and 12 are the measured grid-injected currents and their dynamic responses under L_g = 0 and 4.8 mH, respectively. It can be seen that the grid-injected current is in perfect sinusoidal waveforms with the measured Total Harmonic Distortion (THD) of 1.62 % and 2.42 %, respectively. Furthermore, in Figs. 11 and 12, the peak values of grid-injected currents are both around 12 A when the references are set as 12. 86 A. Figs. 11 and 12 also show that the response time is less than a quarter of the cycle. From the data of Figs. 11 and 12, it can be deduced that the proposed PBC control has strong robustness against the wide variation of equivalent grid impedance.

Fig. 13 shows the grid-injected current under the condition of 50V voltage drop when L_g = 4.8 mH, where a very smooth transient process occurs. The dynamic response of grid-injected current under the condition of sharp phase variation when L_g = 4.8 mH is depicted in Fig. 14, which also indicates that the dynamic process is fast. All the dynamic results indicate that a satisfactory performance can be successfully achieved with the proposed PBC controller.

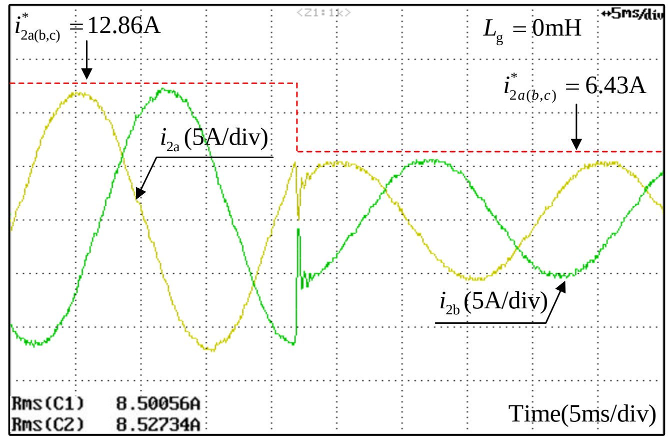

Fig.11. Measured grid-injected currents and their dynamic responses under L_g = 0 mH.

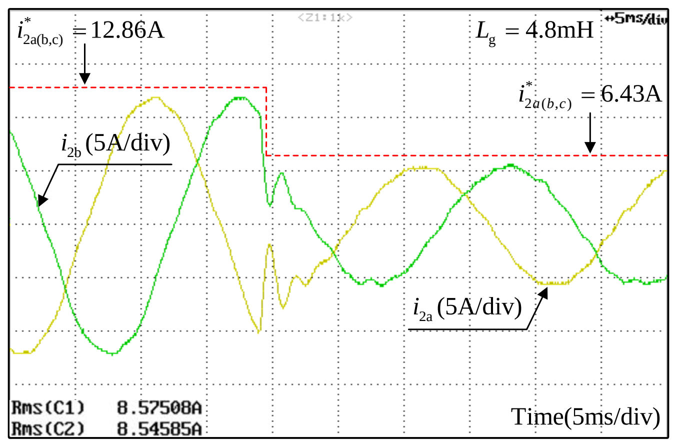

Fig.12. Measured grid-injected currents and their dynamic responses under L_g = 4.8 mH.

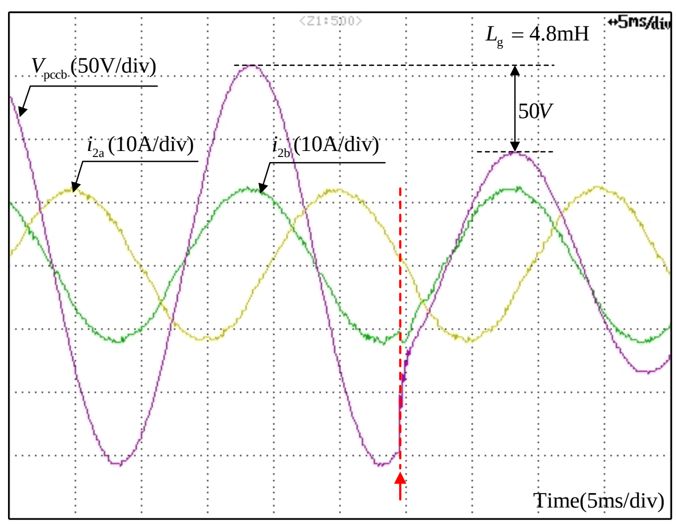

Fig.13. Grid-injected current under 50V voltage reduction under L_g = 4.8 mH.

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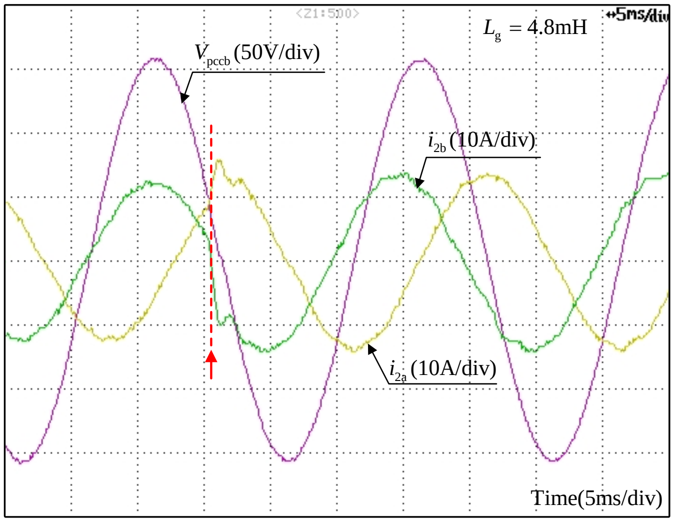


Fig.14. Measured grid-injected currents under the condition of sharp current phase variation when L_g = 4.8 mH.

The grid voltage background harmonics will affect the quality of the grid-injected current. Fig. 15 shows the measured grid-injected current together with the distorted grid voltage under L_g = 4.8 mH. Note that the grid voltage is distorted by the 3rd, 5th, 7th and 9th harmonic voltages, whose magnitudes with respect to the grid fundamental voltage are all 3%, and the THD of grid-injected current is about 3.37%. It can be seen that the proposed PBC controller has a strong ability to resist the adverse effect caused by the harmonic grid voltage.

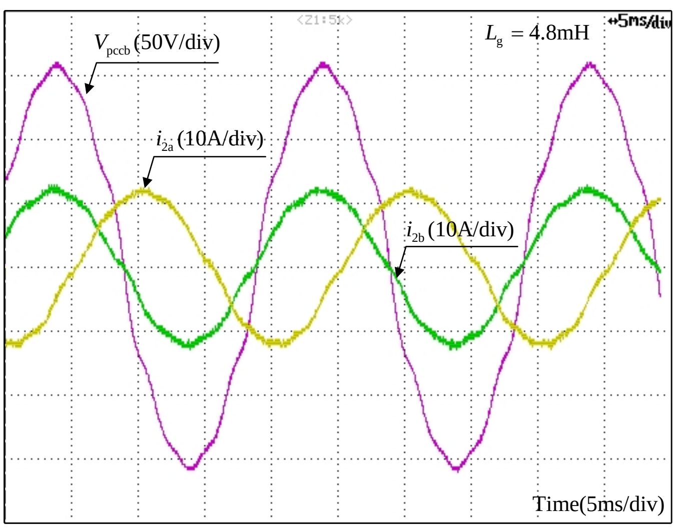


Fig. 15. Measured grid-injected currents under voltage distortion under L_g = 4.8 mH.

Fig. 16 shows the measured grid-injected current when the PI regulator is instead of r_1, i.e. the method to achieve zero steady-state error of grid-injected current. It can be seen that the RMS value is about 9.1 A (the reference RMS value is 9.09A), which means the zero steady-state error can be also easily realized for the proposed PBC.

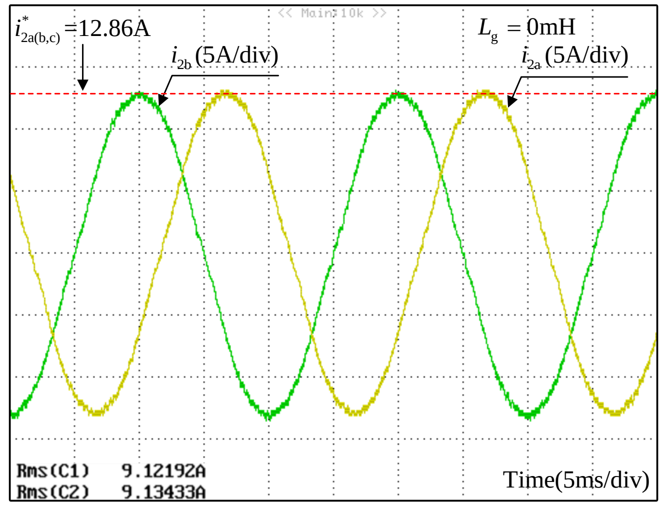
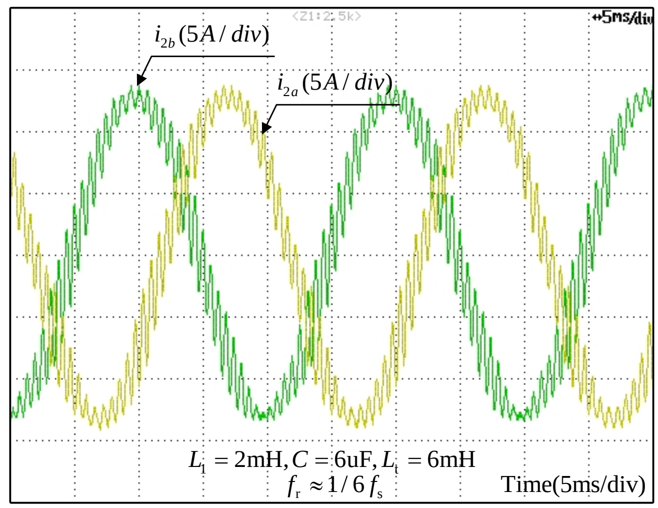
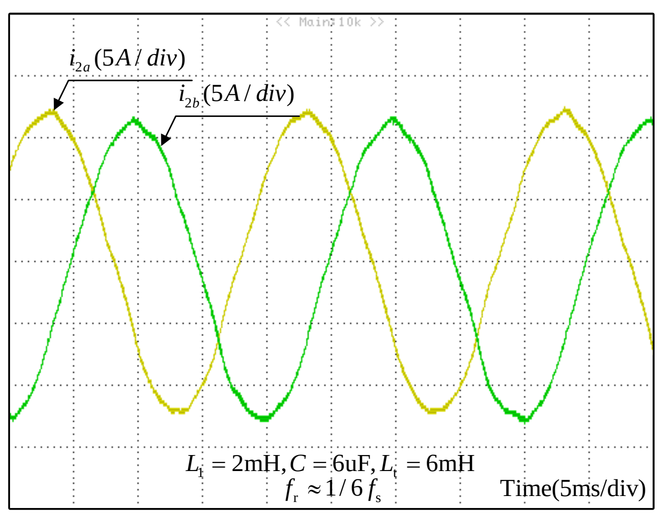


Fig.16. Grid-injected current with zero steady-state error with L_g = 0mH.



(a)



(b)

Fig.17. Measured grid-injected current under $f_r \approx f_s/6$, (a) conventional CCF control method, (b) proposed PBC control method.

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As analyzed above, the PBC controller can keep stable even if the resonant frequency equals 1/6 of the sampling frequency, while the conventional PR with CCF control method cannot. With $L_1 = 2$ mH, $C = 6$ μF and $L_t = 6$ mH, the characteristic resonant frequency of the system is calculated as 1678 Hz ≈ 1667 Hz (1/6 f_s), and the grid-injected current under this situation using the conventional CCF method and proposed PBC method is shown in Fig. 17. It can be seen from Fig. 17. (b) that the grid-injected current is stable without any oscillation when the proposed PBC controller is adopted, while the grid-injected current is serious oscillation with conventional CCF method as shown in Fig. 17(a). Therefore, compared with the conventional CCF method, the proposed PBC controller has the higher robustness against parameters drift.

Fig.18 shows the grid-injected current with $r_1 = 11$, $r_2 = 0.02$, $r_3 = 4$ when $L_g = 0$ mH. It can be seen that the grid-injected currents appear oscillation, when the value of r_1 higher than the calculated upper boundary of 10.1. Note that, due to the actual resistance in the experimental device, the critical value of r_1 to trigger off the oscillation is a little bigger than the theoretical calculation.

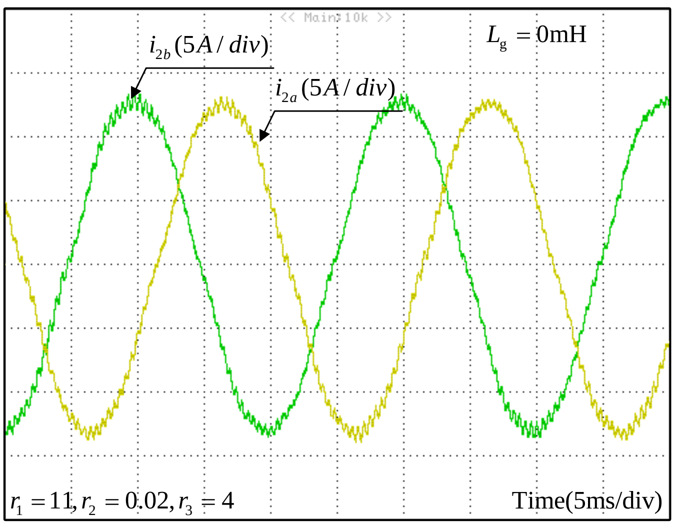

Fig.18. Grid-injected current with $r_1 = 11$, $r_2 = 0.02$, $r_3 = 4$ when $L_g = 0$ mH.

V. Conclusion

In this paper, we have analyzed why it is difficult to design the three damping gains in the PBC controller for the three-phase LCL-filtered GTI, and then put forward a step-by-step control parameters design strategy. The overall conclusion can be summarized as follows:

1) Based on the expectation of the minimized inherent steady-state error of grid-injected current, the three damping gains in the PBC controller for the LCL-filtered GTI can be effectively designed by using the proposed strategy, which provides a useful guideline for engineers.

2) The designed PBC controller can maintain the system stable, even when parameters of LCL filter vary in the range from -33 % to +33 % and the grid impedance varies in the range from 0% to +400 % of L_2. Compared with the conventional CCF AD control method for LCL-filtered GTI, the proposed PBC controller can achieve higher robustness against the parameter drifts of LCL filter and grid impedance.

3) State observer technology can be successfully applied in the proposed PBC controller, resulting in the saved sensors as well as costs.

The effectiveness of the proposed PBC controller has been fully verified via a 3 kW/3-phase/110V experimental lab setup based on dSPACE DS1202.

Appendix

The expressions of A, B, C and D

$$
\begin{aligned}
A = {} & (s^3 C_e L_{1e} L_{2e} + s^2 r_1 C_e L_{1e} + s^2 r_2 L_{1e} L_{2e} + s^2 r_3 C_e L_{2e} + s^2 C_e R_{1e} L_{2e} \\
& + s^2 C_e L_{1e} R_{2e} + s r_1 C_e R_{1e} + s r_2 R_{1e} L_{2e} + s r_2 L_{1e} R_{2e} + s r_3 C_e R_{2e} + s r_1 r_2 L_{1e} \\
& + s r_2 r_3 L_{2e} + s r_1 r_3 C_e + s C_e R_{1e} R_{2e} - 3 s \omega^2 L_{1e} C_e L_{2e} + s L_{1e} + s L_{2e} \\
& + r_1 r_2 R_{1e} + r_2 R_{1e} R_{2e} + r_2 r_3 R_{2e} - \omega^2 L_{1e} C_e R_{2e} - \omega^2 r_1 L_{1e} C_e - \omega^2 r_2 L_{1e} L_{2e} \\
& - \omega^2 r_3 C_e L_{2e} - \omega^2 C_e R_{1e} L_{2e} + r_1 r_2 r_3 + r_1 + r_3 + R_{1e} + R_{2e}) e^{-1.5 s T_s} \\
B = {} & (3 s^2 \omega L_{1e} C_e L_{2e} + 2 s \omega r_1 L_{1e} C_e + 2 s \omega r_2 L_{1e} L_{2e} + 2 s \omega r_3 C_e L_{2e} \\
& + 2 s \omega C_e R_{1e} L_{2e} + 2 s \omega L_{1e} C_e R_{2e} + \omega r_1 C_e R_{1e} + \omega r_2 L_{1e} R_{2e} + \omega r_2 R_{1e} L_{2e} \\
& + \omega r_3 C_e R_{2e} + \omega r_1 r_2 L_{1e} + \omega r_2 r_3 L_{2e} + \omega r_1 r_3 C_e + \omega C_e R_{1e} R_{2e} \\
& - \omega^3 L_{1e} C_e L_{2e} + \omega L_{1e} + \omega L_{2e}) e^{-1.5 s T_s} \\
C = {} & (2 s \omega r_1 L_{1e} C_e + 2 s \omega r_2 L_{1e} L_2 + 2 s \omega r_3 C L_2 + \omega r_1 C_e R_{1e} + \omega r_2 L_{1e} R_2 \\
& + \omega r_2 R_{1e} L_2 + \omega r_3 C R_2 + \omega r_1 r_2 L_{1e} + \omega r_2 r_3 L_2 + \omega r_1 r_3 C_e) e^{-1.5 s T_s} \\
& + (3 s^2 \omega L_1 C L_2 + 2 s \omega C R_1 L_2 + 2 s \omega L_1 C R_2 - \omega^3 L_1 C L_2 \\
& + \omega C R_1 R_2 + \omega L_1 + \omega L_2) \\
D = {} & (s^3 C L_1 L_2 + s^2 C R_1 L_2 + s^2 C L_1 R_2 + s C R_1 R_2 - 3 s \omega^2 L_1 C L_2 \\
& + s L_1 + s L_2 + R_1 + R_2 - \omega^2 C R_1 L_2 - \omega^2 L_1 C R_2) \\
& + (s^2 r_1 C_e L_{1e} + s^2 r_2 L_{1e} L_2 + s^2 r_3 C L_2 + s r_1 C_e R_{1e} + s r_2 R_{1e} L_2 \\
& + s r_2 L_{1e} R_2 + s r_3 C R_2 + s r_1 r_2 L_{1e} + s r_2 r_3 L_2 + s r_1 r_3 C_e \\
& + r_1 r_2 R_{1e} + r_2 R_{1e} R_2 + r_2 r_3 R_2 - \omega^2 r_1 L_{1e} C_e - \omega^2 r_2 L_{1e} L_2 \\
& - \omega^2 r_3 C L_2 + r_1 r_2 r_3 + r_1 + r_3) e^{-1.5 s T_s}
\end{aligned}
$$

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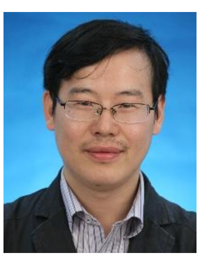

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