

Stationary Frame Current Control Evaluations for Three-Phase Grid-Connected Inverters with PVR-based Active Damped LCL Filters

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Abstract

Grid-connected inverters (GCIs) with an LCL output filter have the ability of attenuating high-frequency (HF) switching ripples. However, by using only grid-current control, the system is prone to resonances if it is not properly damped, and the current distortion is amplified significantly under highly distorted grid conditions. This paper proposes a synchronous reference frame equivalent proportional-integral (SRF-EPI) controller in the $\alpha\beta$ stationary frame using the parallel virtual resistance-based active damping (PVR-AD) strategy for grid-interfaced distributed generation (DG) systems to suppress LCL resonance. Although both a proportional-resonant (PR) controller in the $\alpha\beta$ stationary frame and a PI controller in the dq synchronous frame achieve zero steady-state error, the amplitude- and phase-frequency characteristics differ greatly from each other except for the reference tracking at the fundamental frequency. Therefore, an accurate SRF-EPI controller in the $\alpha\beta$ stationary frame is established to achieve precise tracking accuracy. Moreover, the robustness, the harmonic rejection capability, and the influence of the control delay are investigated by the Nyquist stability criterion when the PVR-based AD method is adopted. Furthermore, grid voltage feed-forward and multiple PR controllers are integrated into the current loop to mitigate the current distortion introduced by the grid background distortion. In addition, the parameters design guidelines are presented to show the effectiveness of the proposed strategy. Finally, simulation and experimental results are provided to validate the feasibility of the proposed control approach.

Key words: Active damping, LCL filter, PVR-AD scheme, Stability, Synchronous frame equivalent PI

I. INTRODUCTION

Recently, distributed generation (DG) systems based on renewable energy systems, such as photovoltaics and wind turbines, have been attracting more and more attention. These renewable energy sources are usually installed in a distributed way. The grid-connected inverter (GCI), as an interface between DGs and the network, plays an important role in ensuring that high-quality power is injected to the grid [1]-[4].

In a GCI, an L filter or an LCL filter is usually used as an

interface between the inverter and the grid. However, if only an inductor is used, high-frequency (HF) switching is needed to ensure that no excessive switching ripples are generated from the pulse-width modulation (PWM) process, which is accompanied by the undesirable problems of excessive switching losses and electromagnetic interference (EMI), especially in high-power applications [5]. Compared to the classical L filter, the LCL filter has better attenuation capacity of the switching harmonics and better dynamic characteristics, which usually results in a lower volume and cost [6]. However, the LCL filter is a three-order system and instability problems may occur at the resonant frequency, thus proper damping solutions are mandatory to stabilize the whole system [7].

The ways to damp resonance problems can be classified into passive damping (PD) and active damping (AD) methods.

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PD is achieved by inserting an additional resistor in series or in parallel with the filter inductor or filter capacitor [8]. The PD scheme of adding a resistor in series with the filter capacitor has been widely adopted due to its simplicity and high reliability. However, the additional resistor will result in power loss and decrease the attenuation of the LCL filter [9]. In order to overcome these drawbacks, the concept of a virtual resistor was proposed. This concept is called the AD method [10], [11]. An interesting control strategy based on the feedback of the splitting capacitor current was proposed in [12]. With this method, the injected current is not controlled directly, and the damping capability relies on the LCL parameters. Therefore, among the AD methods, the methods involving the feedback of the capacitor current have attracted a considerable amount attention due to their simple implementation and wide applicability [2], [13]-[17].

Another interesting approach, which is the main focus of this paper, is to consider the current controller implementation so that better performance is ensured. It is well-known that the proportional integral (PI) controller has an infinite gain for the dc component. This guarantees precise tracking of the dc references without steady-state errors. However, for ac references, a PI controller leads to steady-state errors due to the finite gain at the selected frequency [18], [19]. Proportional resonant (PR) controllers can provide an infinite gain at the selected resonant frequency to suppress the effects of unwanted harmonics, which ensures zero steady-state errors when tracking an ac reference at the selected frequency [20]-[22]. On the other hand, applying a PI controller in the dq synchronous frame can ensure zero steady-state errors, since the ac signal is transformed into a dc signal, which insures that an infinite gain is achieved by the PI controller at the dc component [23], [24]. However, the synchronous frame PI control scheme requires accurate phase synchronization of the grid voltage by using a phase-locked-loop (PLL), which may deteriorate the tracking performance under grid disturbances [24]. Moreover, the straightforward analysis method, referred to as the PI model, replace the PI controller in the dq synchronous frame with a PI plus resonant controller in the $\alpha\beta$ stationary frame. However, the approximation is not accurate and the coupling terms are ignored [25]. In [26], an accurate synchronous frame equivalent PI (SRF-EPI) in the stationary frame with an L filter was presented. This method shows robust performance under a wide frequency range. However, systems with LCL resonance are not considered and the effects of power-stage parameter deviations and grid background distortion were not taken into account. In addition, the discrete model and the control delay effect were also neglected.

This paper proposes a novel parallel virtual resistance (PVR)-based AD method integrated with the SRF-EPI control strategy in the $\alpha\beta$ stationary frame for three-phase

grid-connected inverters with an LCL filter. Multiple PR controllers and a grid voltage feed-forward loop for improving tracking precision under background grid distortion conditions are investigated when the proposed strategy is used. Moreover, the SRF-EPI, PI and PR controllers are analyzed, designed and compared to show the feasibility and effectiveness of the proposed strategy.

This paper is organized as follows. In Section II, the system description and the proposed control scheme are presented. The damping characteristics of the various PD schemes, the series virtual resistance (SVR) and PVR-based AD strategies are analyzed, and the use of the SRF-EPI, PI and PR controllers are compared in the stationary frame. Section III presents parameter design guidelines. The impact of LCL-parameter variations based on the SRF-EPI controller in the stationary frame is also studied. Section IV presents simulation and experimental results of a 2.2 kVA three-phase DG system. Finally, Section V concludes this paper.

II. SYSTEM DESCRIPTION AND CONTROL SCHEME

Fig. 1 shows the topology of a three-phase grid-connected voltage source inverter (VSI) based on an LCL filter, where the inductor parasitic resistances are neglected. In addition, a switch is used to choose the current controllers, and $G_{PI}(s)$, $G_{eq}(s)$, and $G_{PR}(s)$ correspond to the PI controller, the SRF-EPI controller, and the PR controller, respectively. To achieve a good filtering performance, the resonance frequency of the LCL filter should be in the range of $10f_0 < f_{res} < (f_s/2)$, where f_0 denotes the fundamental frequency, f_{res} represents the resonance frequency, and f_s is the switching frequency [7], [18], [27]. Based on these guidelines, the parameters of the power-stage are given in Table I.

A. Stationary Frame SRF-EPI, PI, and PR Controllers

A major objective for the ac current regulators is to achieve zero phase and magnitude errors. Here, the synchronous reference frame equivalent PI (SRF-EPI) controller in the stationary frame was presented as an alternative solution for fundamental frequency reference tracking purposes [26], [28].

The synchronous reference frame equivalent integral controller is $F\angle\theta$, and $n < 0$ refers to the signal is in negative sequence.

$$F = \begin{cases} \infty & n = 1 \\ \frac{2k_i}{\omega_0 |n-1|} & n \neq 1 \end{cases}, \quad \theta = \begin{cases} 0^\circ & n = 1 \\ -90^\circ & (n > 1) \cup (n < 0) \\ 90^\circ & 0 < n < 1 \end{cases} \quad (1)$$

The proportional (P) controller in the dq synchronous frame is the same as the P controller in the $\alpha\beta$ frame. Then, the SRF-EPI controller transfer function is:

$$G_{eq}(s) = k_p + F\angle\theta = k_p + F\cos\theta + jF\sin\theta \quad (2)$$

Compared with the SRF-EPI controller, the integral and

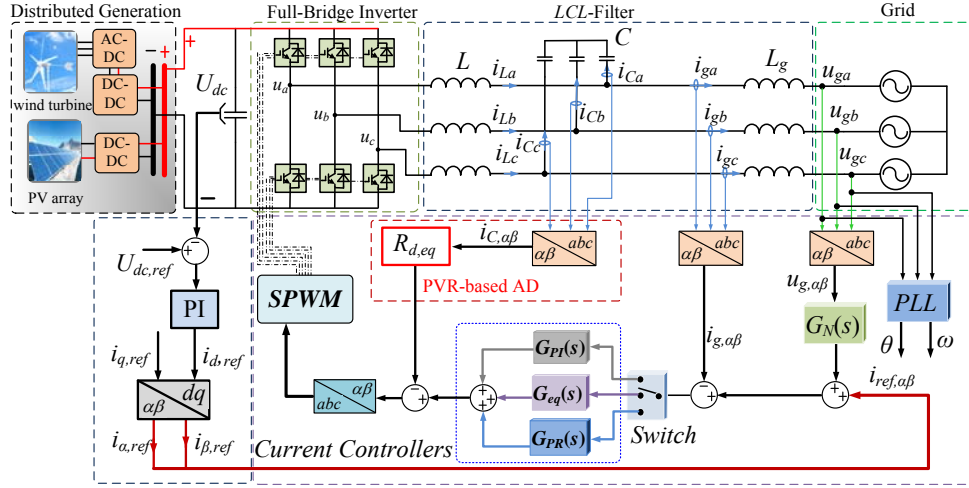


Fig. 1. System topology and control strategy of the three phase grid-connected inverters with LCL filter.

 TABLE I
 SYSTEM PARAMETERS OF THE LCL GRID-TIED INVERTER

Symbol	Quantity	Value
U_{dc}	Input voltage amplitude	650 V
u_g	Grid voltage	311 V
T_s	Sampling and switching period	100 μ s
ω_0	Fundamental angular frequency	100 π rad/s
C	Capacitor	10 μ F
L	Converter-side inductor	1.8 mH
L_g	Grid-side inductor	1.8 mH

$$F_R = \begin{cases} \infty & n = 1 \\ \frac{2k_i}{\omega_0^2(n^2 + 1)} & n \neq 1 \end{cases}, \theta_R = \begin{cases} 0^\circ & n = 1 \\ -90^\circ & n > 1 \\ 90^\circ & n < 1 \end{cases} \quad (4)$$

The PI and PR current controllers are obtained by substituting (3) and (4) into (2), respectively.

Fig. 2 shows bode plots of the equivalent model G_{eqb} , and the resonant and integral controllers for the fundamental frequency signal tracking. It can be seen that the models show remarkable differences from each other in the low frequency range. Thus, direct implementation of the PR or PI models to predict system performance leads to inaccurate results. As for the phase characteristic, the equivalent PI controller G_{eqi} in the positive sequence shows the same phase characteristic as the resonant model, while G_{eqi} in the negative sequence shows the same phase-frequency characteristic as the integral model.

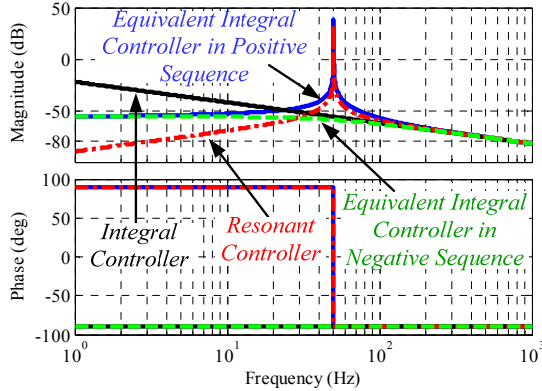


Fig. 2. Bode plots of the integral controller, resonant controller, and equivalent integrator controller in positive and negative sequences.

B. Passive Damping under Grid Current Feedback

As shown in Fig. 3, the grid current is conventionally used as a feedback variable of the current controller to regulate the current injected into the grid. An amplitude peak exists at the resonant frequency of the LCL filter, which limits the design procedure of the current control-loop [6], [29].

A direct way to damp the LCL filter resonance is to insert a passive resistor in the inductor or capacitor of the LCL filter. In this case, the open-loop transfer functions of the grid current feedback control algorithm with various passive damping (PD) methods can be obtained as:

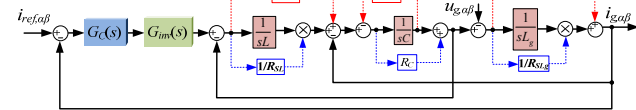


Fig. 3. Block diagram of various PD methods under grid current control for the three-phase inverter with LCL output filter.

resonant models of the PI and PR models are given by:

$$F_I = \begin{cases} \infty & n = 0 \\ \frac{k_i}{\omega_0 n} & n \neq 0 \end{cases}, \theta_I = -90^\circ \quad (3)$$

$$\begin{cases} G_{Series-L}(s) = \frac{G_c(s)G_{im}(s)}{s^3LL_gC + s^2CL_gR_{st} + s(L + L_g) + R_{st}} \\ G_{Series-L_g}(s) = \frac{G_c(s)G_{im}(s)}{s^3LL_gC + s^2CLR_{st_g} + s(L + L_g) + R_{st_g}} \\ G_{Parallel-L}(s) = \frac{G_c(s)G_{im}(s)(sL + R_{pl})}{s^3LL_gCR_{pl} + s^2LL_g + sR_{pl}(L + L_g)} \\ G_{Parallel-L_g}(s) = \frac{G_c(s)G_{im}(s)(sL_g + R_{pl_g})}{s^3LL_gCR_{pl_g} + s^2LL_g + sR_{pl_g}(L + L_g)} \end{cases} \quad (5)$$

$$\begin{cases} G_{Series-L}(s) = \frac{G_c(s)G_{im}(s)}{s^3LL_gC + s^2CLR_{st_g} + s(L + L_g) + R_{st_g}} \\ G_{Parallel-L_g}(s) = \frac{G_c(s)G_{im}(s)(sL_g + R_{pl_g})}{s^3LL_gCR_{pl_g} + s^2LL_g + sR_{pl_g}(L + L_g)} \end{cases} \quad (6)$$

$$G_{Series-C}(s) = \frac{G_c(s)G_{im}(s)(sCR_c + 1)}{s^3LL_gC + s^2CR_c(L + L_g) + s(L + L_g)} \quad (7)$$

$$G_{Parallel-C}(s) = \frac{G_c(s)G_{im}(s)R_d}{s^3LL_gCR_d + s^2LL_g + sR_d(L + L_g)} \quad (8)$$

where $G_{im}(s) = U_{dc}/U_{tri}$ is the gain of the PWM inverter, U_{dc} is the inverter dc voltage, and $U_{tri} = 1$ is the carrier amplitude. $G_{Series-L}$ and $G_{Series-Lg}$ represent the open-loop transfer functions from $i_{ref,\alpha\beta}$ to $i_{g,\alpha\beta}$ when the damping resistor is connected in series with L and L_g , respectively. $G_{Parallel-L}$ and $G_{Parallel-Lg}$ represent the open-loop transfer functions when the damping resistor is connected in parallel with L and L_g , respectively. $G_{Series-C}$ and $G_{Parallel-C}$ are the open-loop transfer functions when the damping resistor is connected in series and in parallel with C , respectively.

Fig. 4 shows bode plots of the open-loop transfer functions with various PD schemes using the parameters in Table I when a PI current regulator is used. As shown in Fig. 4(a), an LCL filter with the PD methods has less attenuation in the low frequency (LF) region when a series resistor is connected with L or L_g . As shown in Fig. 4(b), there is less attenuation in the high frequency (HF) region when a parallel resistor is connected with L or L_g . As shown in Fig. 4(c), it can be seen that less attenuation is also caused when a series resistor is connected with C . As shown in Fig. 4(d), there is no impact on the LF and HF regions when a parallel resistor is connected with C . However, an excessive power loss always exists among the various PD schemes.

C. Virtual Resistance based Active Damping (AD) Strategy

Compared with the PD methods, the active damping (AD) strategy is more flexible and more efficient. This method adopts a virtual resistor to eliminate power loss and can be easily incorporated into the existing control algorithm. The block diagram transformation of a traditional virtual resistor in series with a capacitor is derived in Fig. 5. However, it has been reported that the series virtual resistance based AD (SVR-based AD) needs to introduce differential element, that it causes noise amplification problems, and that the system can be unstable due to HF interference [17].

Therefore, an effective parallel virtual resistance based active damping (PVR-based AD) method in parallel with the capacitor for an LCL filter is proposed in this paper. This method avoids the LCL resonance problems, and satisfactory stability and robustness can be achieved. The principle diagram and the block diagram transformations of the PVR-based AD are shown in Fig. 6.

The transfer function of the PVR-based AD describing the grid current $i_{g,\alpha\beta}$ as a function of the voltage $u_{s,\alpha\beta}$ is given by:

$$G_{active-i_g u_s}(s) = \frac{i_{g,\alpha\beta}(s)}{u_{s,\alpha\beta}(s)} = \frac{1}{s(LL_gCs^2 + R_{d,eq}L_gCs + (L + L_g))} \quad (9)$$

where $R_{d,eq}$ is the equivalent proportional term, $R_{d,eq} = L/(CR_d)$.

According to Fig. 6(b), the open-loop transfer function of

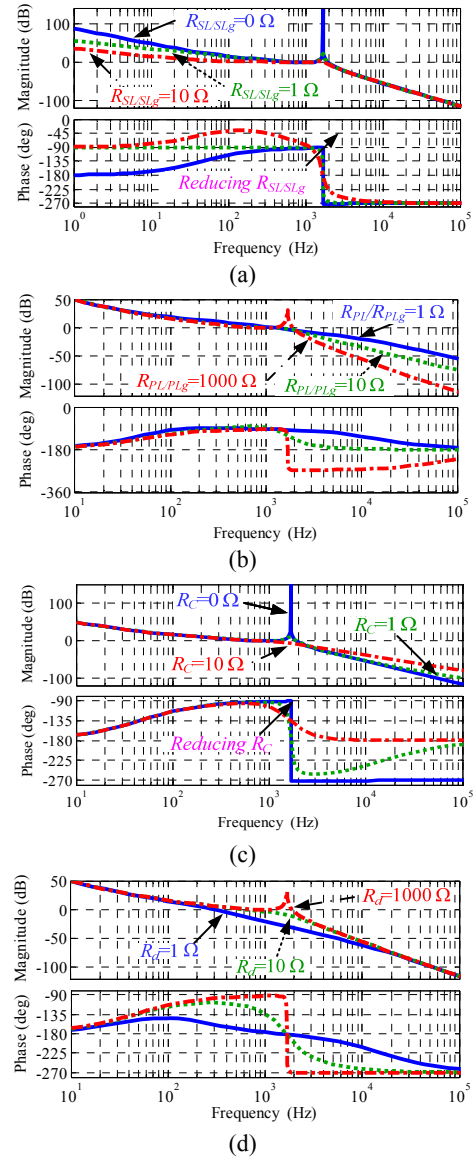


Fig. 4. Bode plots of the open-loop transfer function under grid current feedback control with various PD methods (Magnitude, phase are abbreviated as *Mag*, *Ph*): (a) The resistor in series with L or L_g , (b) The resistor in parallel with L or L_g , (c) The resistor in series with C , and (d) The resistor in parallel with C .

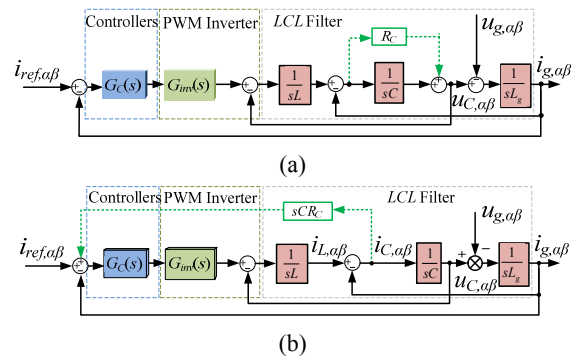


Fig. 5. Block diagram transformation of virtual resistor in series with capacitor: (a) Passive damping of capacitor series with resistor and (b) Equivalent SVR-based AD.

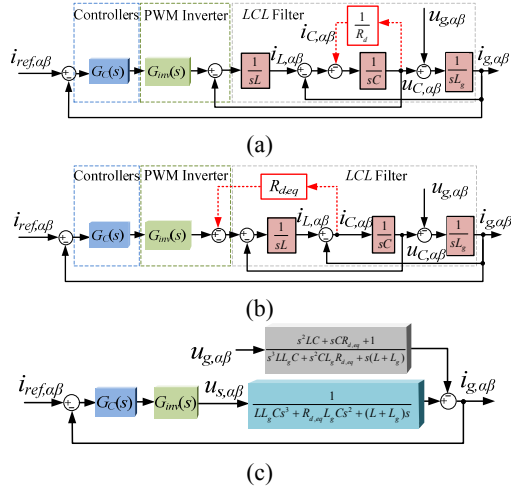


Fig. 6. Block diagram transformations of the PVR-based AD method: (a) PD scheme using parallel-connected resistor across capacitor, (b) Block transformation of PD control strategy in (a), and (c) The block diagram of parallel virtual resistance active damping.

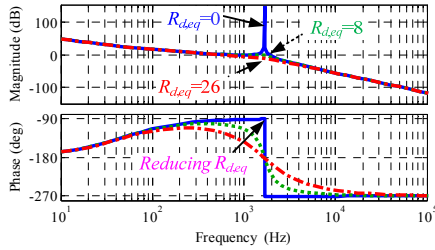


Fig. 7. Bode plots of the open-loop transfer function with PVR-based AD strategy.

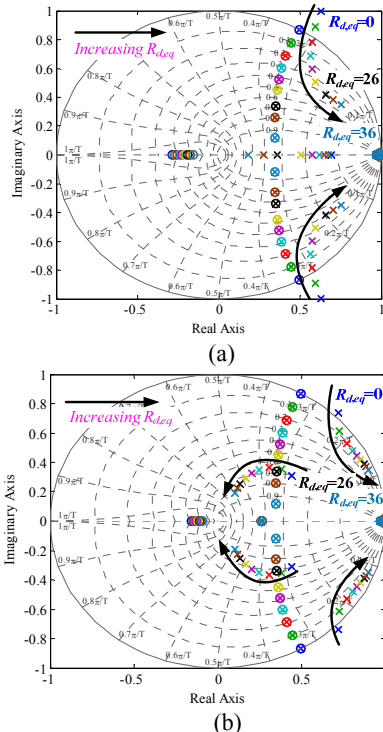


Fig. 8. Pole-zero map for close-loop transfer function with PVR-based AD strategy: (a) No delay and (b) $3/2T_s$ delay.

the PVR-based AD is given by:

$$G_{open}(s) = \frac{G_C(s)G_{inv}(s)}{LL_g Cs^3 + R_{d,eq} L_g Cs^2 + (L + L_g)s} \quad (10)$$

From (10), it can be seen that there are no unstable open-loop poles, i.e., $P=0$. If the PI current controller is used, using the parameters in Table I, the bode plots of the open-loop transfer function $G_{open}(s)$ with the AD method, when $R_{d,eq}$ is 0, 8 and 26, are illustrated in Fig. 7, respectively. The corresponding parallel virtual resistor R_d is ∞ , 22.5 Ω , and 6.9 Ω , respectively. According to the Nyquist stability criterion [7], the system is stable if $Q=0$.

$$Q = P + N_+ - N_- \quad (11)$$

where P is the number of unstable open-loop poles, and N_+ and N_- are the number of times the path crosses the line in the clockwise and counter-clockwise directions. Hence, there is a one-to-one correspondence between the positive half of the Nyquist plot and the open-loop bode plot. N_+ and N_- are two times the number of positive (from lower to upper) and negative (from upper to lower) crossings of $(2k+1)\pi$ (k is an integer) in the open-loop bode plot in the frequency range with gains above 0 dB. For minimum phase system, i.e., where $P=0$, the system is stable if $N_+=N_-$ [29], [30].

- 1) When $R_{d,eq}=0$, the $PH-F$ contour cross -180° from upper to lower at f_{res} (in $Mag>0$ dB region). Thus, when $P=0$, $N_-=0$, $N_+=2$, and $Q=2$, the system is unstable.
- 2) When the feedback gain $R_{d,eq}$ increases, the $PH-F$ contour cross -180° outside the $Mag>0$ dB region if the value of $R_{d,eq}$ is appropriate. Thus, when $P=0$, $N_-=0$, $N_+=0$, and $Q=0$, the system is stable. The cases for $R_{d,eq}=0, 8, 26$ are shown in Fig. 8.

In order to intuitively show the variation tendency of the system stability when $R_{d,eq}$ varies, a discrete pole-zero map when the feedback gain $R_{d,eq}$ increases from 0 to 36 is illustrated in Fig. 8 by using the ZOH method. It can be seen that the system stability changes along with the delay time. The discrete open-loop transfer function of the current controller using the PVR-based AD scheme is:

$$G_{open}(z) = G_C(z) \times Z \left[G_d(s)G_{inv}(s) \frac{1 - e^{-sT_s}}{s} \times \frac{1}{LL_g Cs^3 + R_{d,eq} L_g Cs^2 + (L + L_g)s} \right] \quad (12)$$

$$G_d(s) = e^{-sT_d} \quad (13)$$

where $G_d(s)$ represents the delay in the digital control, T_d is the delay time which it is related to the analog-digital sampling process, the PWM generation process, and the hardware filtering [6], [26], [30]. The PWM switching frequency and the sampling frequency of the digital current controller are both selected to be 10 kHz.

Fig. 9 shows a frequency domain comparison between the PVR-based AD strategy, the PD method and the no damping scenario. It can be observed from Fig. 7 and Fig. 9 that the PVR-based AD method ensures effective damping

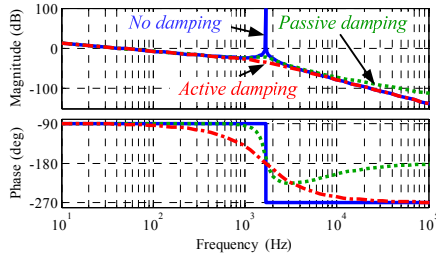


Fig. 9. Bode plots of PVR-based AD scheme compared with PD and no damping strategies.

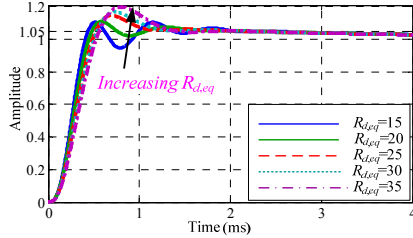


Fig. 10. Step response of the PVR-based AD scheme with different $R_{d,eq}$.

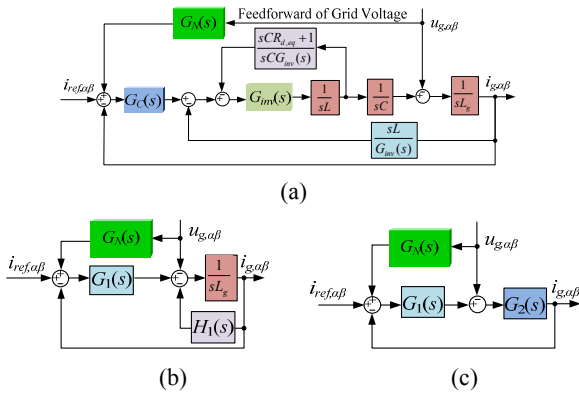


Fig. 11. Block diagram transformation of the PVR-based AD control algorithm with the grid voltage feed-forward loop.

characteristics. These figures also show that the attenuations in the LF and HF regions are not affected, which is an advantage over the other PD methods.

Once again referring to Fig. 8, it can be seen that values of $R_{d,eq}$ that are either too large or too small will affect the stability of the system. By gradually increasing $R_{d,eq}$, the poles can be moved close to the center of the unit circle. However, the poles will be moved away from the circle when $R_{d,eq}$ continues to increase. This has the effects of reducing the stability margin and increasing instability. Hence, it is necessary to optimize the damping coefficient $R_{d,eq}$, which ensures that the system has a sufficient stability margin and good dynamic performance.

Equation (9) is equivalent to a first order model combined with a second order plant model. In addition, the resonance is caused by the second order system. Notably, the resonance of the second order system is determined by the damping ratio ζ . The damping ratio ζ in (9) is shown in (14). To eliminate the resonance, let $\zeta=0.707$. Thus, the feedback gain is calculated

as $R_{d,eq}=26.8$, i.e., the parallel virtual resistor R_d is 6.7Ω .

$$\zeta = \frac{R_{d,eq}}{2L\omega_{res}} = \frac{R_{d,eq}}{2} \sqrt{\frac{L_g C}{(L + L_g)L}} \quad (14)$$

Fig. 10 shows the step responses of the PVR-based AD method with different $R_{d,eq}$ scenarios. It can be seen that the dynamic response is influenced by the value of $R_{d,eq}$. The overshoots as well as the oscillations are greatly reduced with an increasing $R_{d,eq}$ and the settling time is the shortest when $R_{d,eq}=26.8$. In summary, optimal performance is achieved when $R_{d,eq}$ equals 26.8, and it changes along with the delay time.

D. Grid Voltage Disturbance Rejection

To mitigate the effect of grid voltage distortion on the quality of the output currents of the LCL-filter, the proportional resonant (PR) was introduced to provide infinite gain at the selected harmonic frequencies to guarantee sinusoidal grid currents [22], [31]. This method based is on the concept of harmonic impedance, which is to incorporate the grid voltage harmonics in the control loop through an additional feed-forward path to suppress the effects of unwanted harmonics. This does not affect the phase margin. However, the compensation accuracy of this approach is not satisfactory [32].

Block diagram transformations of the PVR-based AD algorithm with the grid voltage feed-forward loop are shown in Fig. 11. The current injected into the grid can be derived as:

$$\begin{aligned} i_{g,\alpha\beta}(s) &= \frac{W(s)}{1+W(s)} i_{ref,\alpha\beta}(s) - \frac{1-G_N(s)G_1(s)}{1+W(s)} G_2(s) u_{g,\alpha\beta}(s) \\ &= i_{g1}(s) + i_{g2}(s) \end{aligned} \quad (15)$$

where:

$$G_1(s) = \frac{G_C(s)G_{mv}(s)}{s^2 LC + sCR_{d,eq} + 1} \quad (16)$$

$$G_2(s) = \frac{s^2 LC + sCR_{d,eq} + 1}{s^3 LL_g C + s^2 CL_g R_{d,eq} + s(L + L_g)} \quad (17)$$

and $W(s)=G_1(s)G_2(s)$.

The transfer function from the grid voltage to the grid side current with the PVR-based AD is derived as:

$$\frac{i_{g,\alpha\beta}(s)}{u_{g,\alpha\beta}(s)} = \frac{s^2 LC + sCR_{d,eq} + 1}{G_C(s)G_{mv}(s) + s^3 LL_g C + s^2 CL_g R_{d,eq} + s(L + L_g)} \quad (18)$$

Fig. 12 shows bode plots of (18) under no damping, and the PD and PVR-based AD cases without feed-forward of the grid voltage. It can be observed that a voltage disturbance near the resonance frequency results in large input currents in the no damping case. However, the PD and PVR-based AD schemes provide efficient resonance damping characteristics, and the PVR-based active damping scheme shows additional damping performance without causing additional power losses.

In order to mitigate the effect of the grid voltage on the

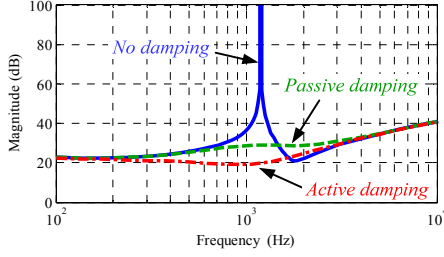


Fig. 12. Bode plots comparison of disturbance voltage to grid current of the harmonic admittance without the feed-forward of grid voltage.

quality of the LCL-filter currents, a feed-forward loop of the grid voltage and the multiple resonant regulators at the harmonic frequencies of the grid voltages are applied in the current loop. According to (15), setting $1-G_N(s)G_1(s)$ to zero and adopting an approximate full compensation approach, the feed-forward function is derived as:

$$G_N(s) = \frac{1}{G_1(s)} = \frac{s^2 LC + sCR_{d,eq} + 1}{G_C(s)G_{inv}(s)} \quad (19)$$

Considering that derivative functions are difficult to implement and mainly for middle- and higher-frequency harmonics, and that the noise amplification problems reduce the stability margin, (19) is approximated by a first order low pass filter, and the higher order differential terms are neglected to avoid high frequency instabilities [29], [33]. As shown in Fig. 11(a), if the feed-forward control lies behind the current controller, the feed-forward factor has a very simple form, the proportional feed-forward is adopted, and $G_N(s)=1/G_{inv}(s)$. Moreover, multiple resonant controllers are achieved by paralleling several resonant blocks tuned to the harmonic frequencies to be compensated. This is expressed as:

$$G_h(s) = \sum_{h=5,7,11,\dots} \frac{2K_{ih}s}{s^2 + (h\omega_0)^2} \quad (20)$$

where h is the harmonic order to be compensated, ω_0 represents the fundamental frequency, and K_{ih} represents the respective resonant gain, which should be tuned relatively high, but within the stability limits, for minimizing the steady-state error. The dynamics of the fundamental PI, PR and SRF-EPI controllers in the stationary frame are unaffected when multiple resonant controllers are added, since the resonant controllers only compensate for frequencies that are very close to the selected resonant frequencies.

III. DESIGN GUIDELINES

A. Stability of the SRF-EPI, PI and PR Controllers

The resonance damping and dynamic response have been taken into consideration for the aforementioned active damping design procedure. Thus, the current regulator $G_C(s)$ is designed to maintain suitable values for the PM, GM and

cutoff frequency f_c . Generally, a PM in the range of $30^\circ \sim 60^\circ$ and a $GM \geq 3 \sim 6$ dB are adopted for a compromise between the stability, dynamic response, and robustness [10], [14].

The cutoff frequency f_c is typically kept lower than f_s , considering that the effect of attenuating the HF noise and f_c generally can be chosen lower than $1/10$ of f_s . Therefore, considering the frequency response of the system below the cutoff frequency, the influence of the capacitor of the LCL filter can be ignored and $G_{open}(s)$ can be approximated as:

$$G_{open}(s) \approx \frac{G_C(s)G_{inv}(s)}{(L+L_g)s} \quad (21)$$

Here, a PI controller is adopted to design the parameters for the PR and SRF-EPI controllers for comparison. At the cutoff frequency, the magnitude-frequency characteristic of the open-loop system is zero. As a result:

$$20 \lg |G(j2\pi f_c)| \approx 20 \lg \left| \frac{k_p G_{inv}(s)}{(L+L_g) * j2\pi f_c} \right| = 0 \quad (22)$$

Thus, the controller gain k_p is approximately represented as:

$$k_p \approx \frac{2\pi f_c (L+L_g)}{G_{inv}(s)} \quad (23)$$

which shows that f_c is approximately proportional to k_p . Therefore, a larger k_p means a faster dynamic response and a larger loop gain at low frequencies.

The steady-state error of $i_{g,\alpha\beta}$ at the fundamental frequency $\omega_0 = 2\pi f_0$ is an important index of the controller parameter design. As given by (15), assuming the adoption of the grid feed-forward control, the steady-state tracking error of the grid current $i_{g,\alpha\beta}$ can be calculated as:

$$E_g(s) = \frac{i_{g1}(s) - i_{ref}(s)}{i_{ref}(s)} = \frac{W(s)}{1+W(s)} - 1 \quad (24)$$

The fundamental components of i_{g2} and $u_{g,\alpha\beta}$ are denoted by i_{g2}^* and u_g^* , respectively. Since the influence of the filter capacitor is negligible at the fundamental frequency f_0 , considering (18), i_{g2}^* can be approximated as:

$$i_{g2}^* \approx -\frac{u_g^*}{G_{inv}G_C(j2\pi f_0)} \quad (25)$$

Considering equation (21) and (25), I_{g2} can be derived as:

$$I_{g2} \approx \frac{U_g}{G_{inv}|G_C(j2\pi f_0)|} \approx \frac{U_g}{2\pi f_0 |G_{open2}(j2\pi f_0)|} \quad (26)$$

where I_{g2} and U_g are the root mean square (RMS) values of i_{g2}^* and u_g^* , respectively.

According to (21) and the transfer function of the PI controller (26) can be rewritten as:

$$G_{open2}(f_0) = 20 \lg |G_{open2}(j2\pi f_0)| = 20 \lg \left| \frac{G_{inv}(k_p + \frac{k_i}{j2\pi f})}{j2\pi f_0(L+L_g)} \right| \quad (27)$$

Substituting (23) into (27), the integration gain is derived as:

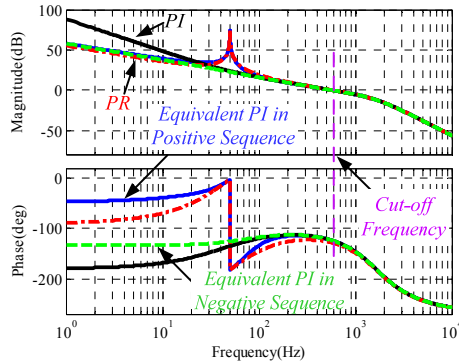


Fig. 13. Bode plots of the open-loop transfer function with PVR-based AD strategy.

$$k_i = \frac{4\pi^2 f_0 (L + L_g)}{G_{inv}} \sqrt{\left(10^{\frac{G_{open2}(f_0)}{20}} f_0\right)^2 - f_c^2} \quad (28)$$

According to (17) and (18), the PM of the system can be expressed as:

$$PM = 180^\circ + \angle \frac{G_c(s)G_{inv}(s)}{LL_g Cs^3 + R_{d,eq} L_g Cs^2 + (L + L_g)s} \Big|_{s=j2\pi f_c} \quad (29)$$

Substituting the transfer function of the PI into (29) yields:

$$PM = \arctan \frac{2\pi L(f_{res}^2 - f_c^2)}{G_{inv} f_c} - \arctan \frac{k_i}{2\pi f_c k_p} \quad (30)$$

Then, (30) can be rewritten as:

$$k_i = 2\pi f_c k_p \frac{2\pi L(f_{res}^2 - f_c^2) - G_{inv} f_c R_{d,eq} \tan PM}{2\pi L(f_{res}^2 - f_c^2) \tan PM + G_{inv} f_c R_{d,eq}} \quad (31)$$

Using the parameters of Table I, the resonance frequency f_{res} of the LCL filter is equal to 1.68 kHz, and the value of the cutoff frequency f_c is selected as 600Hz. In addition, the PM is selected as 50 degrees. With the system parameters given in Table I, the SRF-EPI, PI and PR controller parameters are selected as $k_p=0.02$ and $k_i=5.77$. Bode plots of the open-loop transfer function with the PVR-based AD, when $R_{d,eq}=26.8$, is shown in Fig. 13. When the PI compensator is adopted, the GM and PM are 11.1 dB and 52.2 degrees, respectively. When an equivalent model of the synchronous-frame PI compensator in the stationary frame is adopted, the GM and PM are 11 dB and 44.9 degrees, respectively. When the PR compensator is adopted, the GM and PM are 9.56 dB and 31.5 degrees, respectively.

As shown in Fig. 13, except for the LF range, the open-loop bode plots of the stationary frame equivalent model of the PI controller in the positive sequence and the PR model are the same at the fundamental frequency, and are similar in the HF range. This indicates that the SRF-EPI controller in the stationary frame has similar performance to the PR controller in the stationary frame when tracking the fundamental frequency reference. Thus, the SRF-EPI controller has better performance than the PI and PR controllers at all frequencies when they are used in the stationary frame. Except for reference tracking at the

fundamental frequency, the SRF-EPI controller in the stationary frame is not equivalent to the PI or PR controllers in the $\alpha\beta$ frame.

B. Impact of LCL-Parameter Variations

Filter parameters drift away from the rated values due to the effects of parasitic parameters variations, operating conditions, temperature and grid impedance [10], [34]. To examine the robustness of a system with the PVR-based AD scheme of the SRF-EPI controller in the $\alpha\beta$ frame, bode plots of the compensated loop gain considering variations of the LCL filter parameters are given in Fig. 14. The grid impedance is considered to be a part of L_g . It is found that although C varies from 7 to 13 μF ($10 \mu\text{F} \pm 30\%$), L varies from 1.26 to 2.34 mH ($1.8 \text{ mH} \pm 30\%$), or L_g varies from 1.26 to 3.6 mH ($1.8 \text{ mH} -30\% \sim +100\%$), the lowest cutoff frequency is still higher than 417 Hz, the PM is larger than 37.7 degrees, and the GM is larger than 9.84 dB. The frequency response characteristics in Fig.14 verify that the designed system has a strong robustness.

IV. SIMULATION AND EXPERIMENTAL RESULTS

In order to verify the effectiveness of the proposed control strategies, simulation and experimental results of the proposed control strategy are presented and compared. The simulation studies were implemented using Matlab/Simulink. As shown in Fig. 15, an experimental setup was built and tested. The setup consists of two 2.2 kW Danfoss inverters, one working as a grid-connected inverter and the other controlled in the voltage control mode to emulate the grid with distortion. The dSPACE1006 platform was used to implement the control algorithms. In order to provide effective comparisons, all the control parameters are the same for the simulation and experiments. The system parameters are given in Table I and the SRF-EPI, PI and PR controller parameters are selected as $k_p=0.02$ and $k_i=5.77$.

Fig. 16 demonstrates the performance of the PVR-based AD strategy when the SRF-EPI controller is adopted. When the PVR-based AD strategy is enabled, the system is stable without resonance. However, when the active damping is disabled, resonance appears and the system becomes unstable. This shows that the PVR-based AD method is necessary for the SRF-EPI controller to maintain stability and to ensure harmonic-free grid currents.

Steady-state waveforms of the simulation and experimental results for the different controllers in the stationary frame with the proposed PVR-based AD method are shown in Fig. 17 and Fig. 21, respectively. As shown in Fig. 17, the fundamental RMS values of i_{ga} (A phase) in Fig. 17(a), (b), and (c) are 1.63 A, 1.422 A, and 1.42A with a reference value of 1.414 A (RMS). The measured steady-state errors in Fig. 17(a), (b), and (c) are 15.3%, 0.57%, and 0.42% respectively.

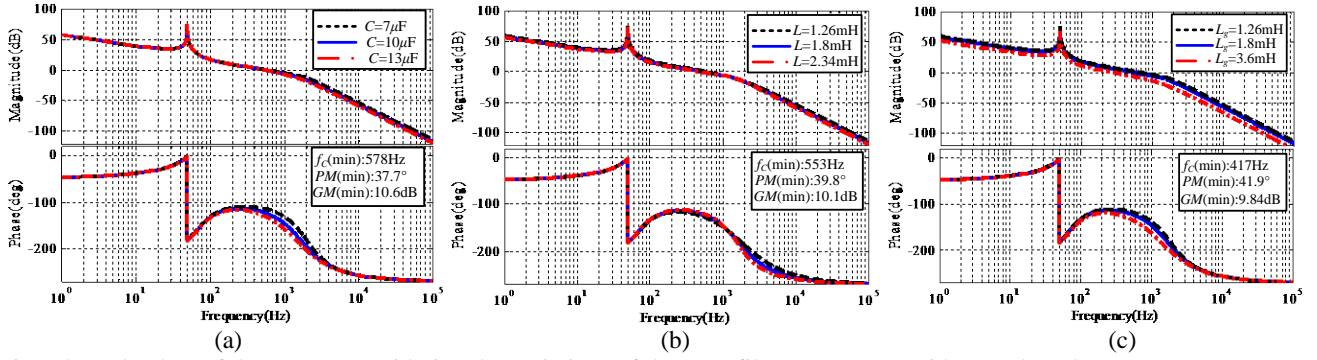


Fig. 14. Bode plots of the system considering the variations of the LCL filter parameters with PVR-based AD strategy: (a) C : $10 \mu\text{F} \pm 30\%$, (b) L : $1.8 \text{ mH} \pm 30\%$, and (c) L_g : $1.8 \text{ mH} -30\% \sim +100\%$.



Fig. 15 Experimental setup.

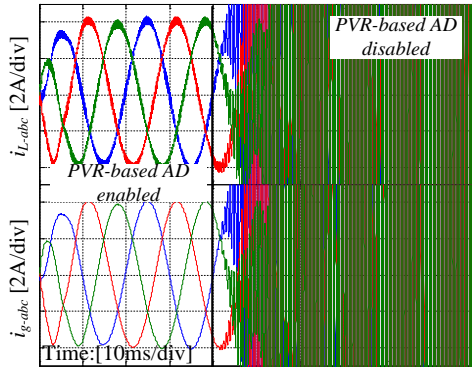


Fig. 16. Simulation result of the SRF-EPI control strategy when PVR-based AD strategy is disabled at 40ms.

In addition, the total harmonic distortions (THD) are 1.22%, 1.16%, and 1.12%, respectively. As shown in Fig. 21, the measured steady-state errors in Fig. 21(a), (b), and (c) are consistent with the simulation results. However, the current THDs are 1.28%, 4.8%, and 4.75%, respectively. It is noted that the THDs of the experimental results with the PR and SRF-EPI controllers are worse than that of the PI controller because the current of the PI controller has a higher amplitude due to the effect of steady state errors. Therefore, sufficient accuracy can be achieved by using the SRF-EPI controller, which is slightly better than the PR controller.

Simulation and experimental results of the transient responses when the reference steps up from 1.0 kW to 2.0 kW for the different controllers in the stationary frame

with the PVR-based AD method are shown in Fig. 18 and Fig. 22, respectively. As shown in Fig. 18, good dynamic performances are achieved when the PR and SRF-EPI controllers in the $\alpha\beta$ frame are adopted. Fig. 22 shows that the transient response of the SRF-EPI controller is slightly better than that with the PR controller. The experimental results are consistent and in good agreement with the theoretical analysis and simulation results.

In order to evaluate the robustness of the SRF-EPI control scheme with the PVR-based AD method, simulation and experimental results under different virtual parallel damping resistance in the case of a control delay and the effect of grid voltage harmonics are presented.

Fig. 19 and Fig. 23 show the effects of the feedback gain $R_{d,eq}$ of the PVR-based AD method on the output current quality with a control delay time of $T_d=1.5T_s$, when $R_{d,eq}$ of the SRF-EPI controller is chosen to be 26.8, 18.8, and 6.8. The simulation and experimental results under different values of $R_{d,eq}$ with $T_d=1.5T_s$ are shown in Fig. 19 and Fig. 23.

Although the system is stable, the harmonic content increases dramatically when the value of $R_{d,eq}$ increases. In addition, the current THDs (phase A) in Fig. 19 (a), (b), and (c) are 35.82%, 6.04%, and 2.95%, respectively. The current THDs (phase A) of the experimental results in Fig. 23(a), (b), and (c) are 45%, 6.5%, and 3.6%, respectively.

Therefore, it is necessary to optimize the value of $R_{d,eq}$ to ensure a sufficient stability margin and a good dynamic response when a control delay is introduced.

To validate the performance of the control algorithm in the case of distorted grid conditions, the grid-emulator was distorted with 5th and 7th harmonics, with voltage THDs of 3% and 2%, respectively. Simulation and experimental results of the grid currents with the SRF-EPI controller in the stationary frame under different control scenarios are shown in Fig. 20 and Fig. 24. The multiple resonant controllers tuned at the 5th and 7th harmonics with $K_{rh}=5$ where $h=5$ and 7 are added to the SRF-EPI controller in the stationary frame with the proposed active damping method. The THDs of the simulated grid current i_{ga} under different control scenarios are shown in

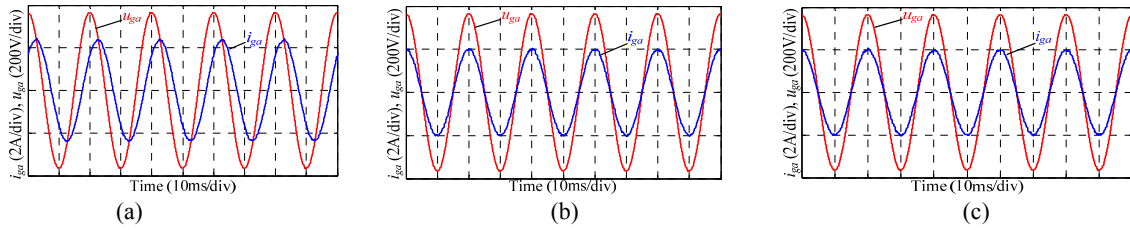


Fig. 17. Steady-state simulation results under 50% load condition with the proposed PVR-based AD strategy. (a) The PI controller. (b) The PR controller. (c) The SRF-EPI controller.

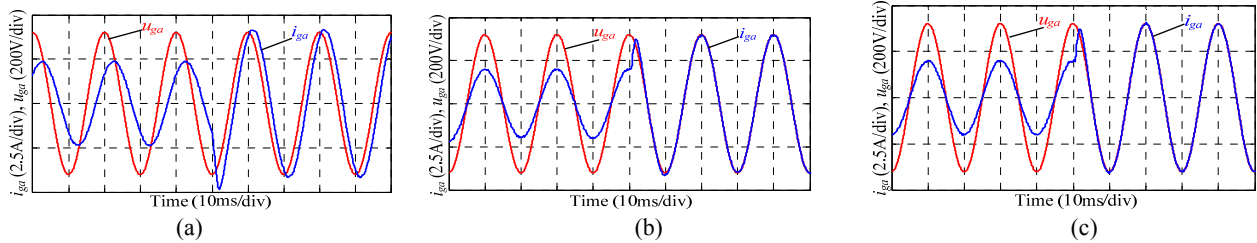


Fig. 18. Transient responses when the reference of i_{ga} steps up from half-load to full-load with the proposed PVR-based AD strategy: (a) The PI controller, (b) The PR controller, and (c) The SRF-EPI controller.

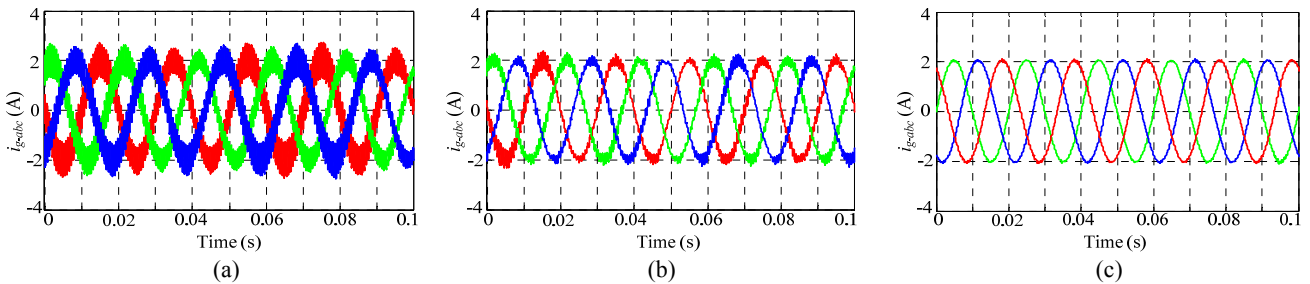


Fig. 19. Simulation results of different $R_{d,eq}$ with the delay time $T_d=1.5T_s$ when the SRF-EPI controller is used in the proposed PVR-based AD strategy: (a) $R_{d,eq}=26.8$, (b) $R_{d,eq}=18.8$, and (c) $R_{d,eq}=6.8$.

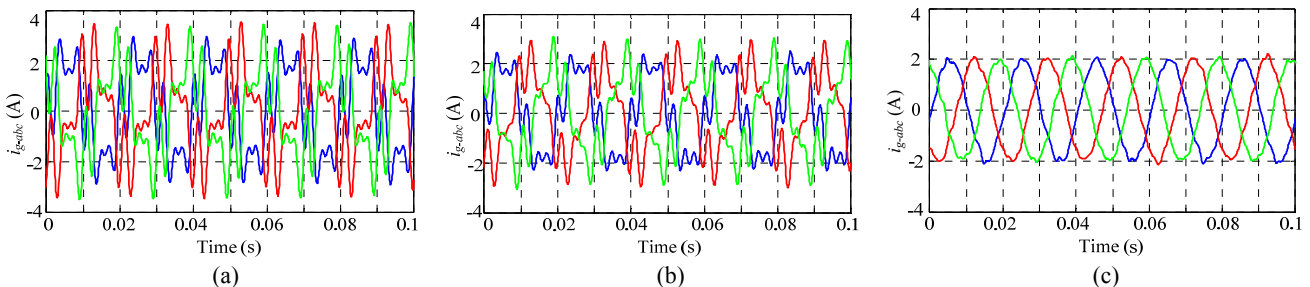


Fig. 20. Simulation results of the SRF-EPI controller with PVR-based AD strategy when grid voltages are highly distorted: (a) Without grid voltage feed-forward control, (b) With grid voltage feed-forward control, and (c) Combine with resonant controllers and grid voltage feed-forward control.

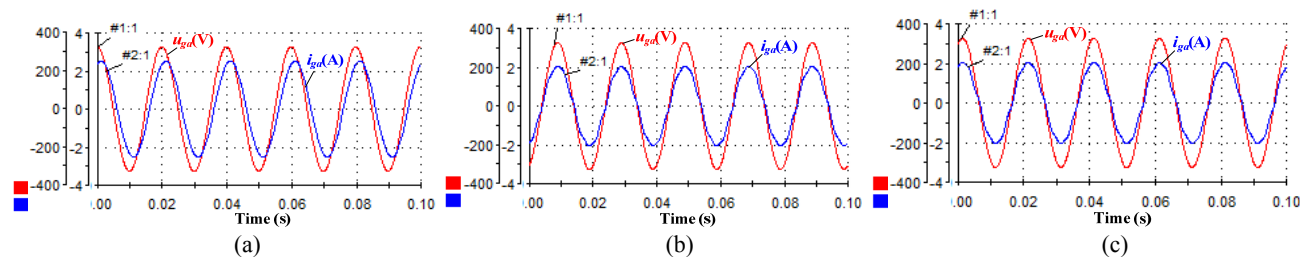


Fig. 21. Steady-state experimental results under half-load condition with the proposed PVR-based AD strategy: (a) The PI controller, (b) The PR controller, and (c) The SRF-EPI controller.

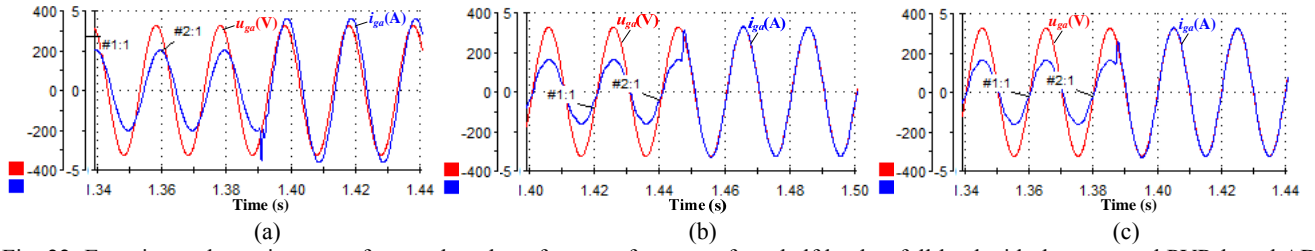


Fig. 22. Experimental transient waveforms when the reference of i_{ga} steps from half load to full load with the proposed PVR-based AD strategy: (a) The PI controller, (b) The PR controller, and (c) The SRF-EPI controller.

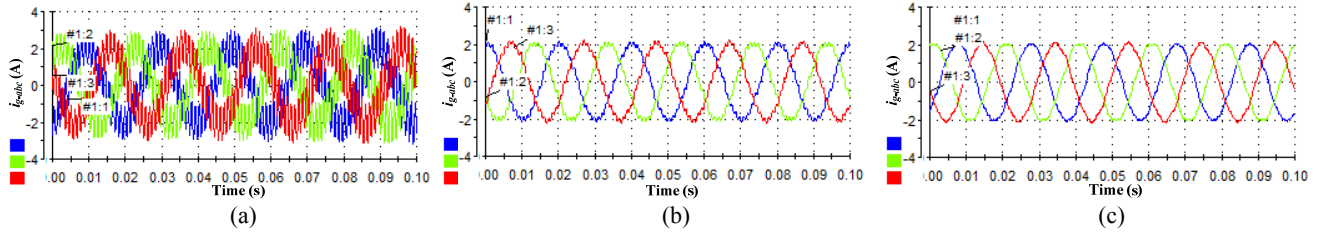


Fig. 23. Experimental results of different $R_{d,eq}$ with the delay time $T_d=1.5T_s$ when the SRF-EPI controller is used in the proposed PVR-based AD strategy: (a) $R_{d,eq}=26.8$, (b) $R_{d,eq}=18.8$, and (c) $R_{d,eq}=6.8$.

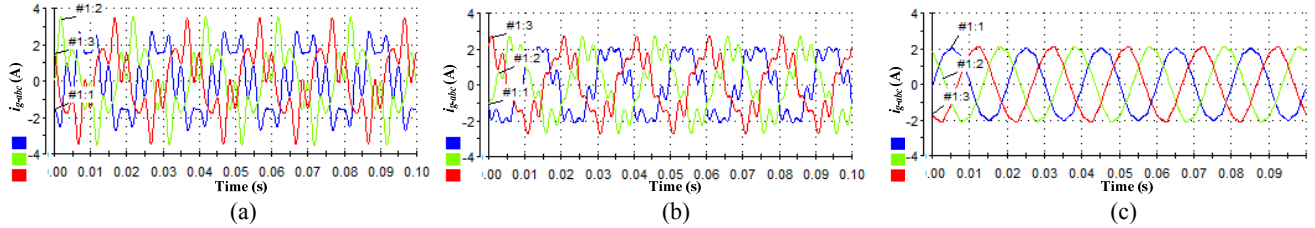


Fig. 24. Experimental results of the SRF-EPI controller in stationary frame with the PVR-based AD strategy when the grid voltages are highly distorted: (a) The SRF-EPI controller without grid voltage feed-forward control, (b) The SRF-EPI controller with grid voltage feed-forward control, and (c) The SRF-EPI controller combines with the resonant controllers and grid voltage feed-forward control.

Fig. 24, the current THDs of the experimental results are about 54.38%, 29.12% and 3.92%. This shows that when only the SRF-EPI controller is used, the grid currents are highly distorted due to harmonic distortion of the grid voltages. When the grid voltage feed-forward scheme is used in combination with the proposed control strategies, the current distortion can be effectively suppressed. However, the THD of the grid current is still high and the distortion in the grid currents cannot be rejected by the voltage feed-forward loop alone. However, when the SRF-EPI controller is adopted in combination with the grid voltage feed-forward loop and the multiple resonant controllers in the current loop, sinusoidal waveforms are guaranteed in the grid currents.

V. CONCLUSION

In this paper, a stationary frame SRF-EPI control algorithm using the PVR-based AD method is proposed for three-phase *LCL* type grid-connected DG systems. It achieves a compromise between the resonance damping and the dynamic performance and makes it easier to stabilize the whole system. The accurate stationary frame of the SRF-EPI controller is introduced to achieve a high closed-loop

bandwidth and good robustness. It is found that the SRF-EPI controller is equivalent to a PR controller in the stationary frame at the fundamental frequency. However, it shows different phase characteristics at other frequency ranges compared to the PR and PI controllers in the stationary frame.

In order to guarantee the quality of grid currents under non-ideal grid conditions, grid voltage feed-forward control with multiple resonant controllers in the harmonic frequencies is adopted, and the performance under the no damping, PD, and AD methods are compared. Furthermore, design guidelines of the systems of the SRF-EPI, PI and PR controllers are presented. Finally, simulation and experimental results of a three-phase *LCL*-based grid-connected inverter are presented to validate the proposed control approach.

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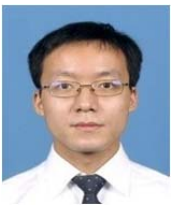
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