Robust Delay Compensation Strategy for LCL-Type Grid-Connected Inverter in Weak Grid

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ABSTRACT Capacitor-current-feedback active damping has been widely adopted in digitally controlled LCL-type grid-connected inverters (GCIs). However, their digital control delay causes the system stability to deteriorate within a negative damping range. Particularly, in a weak grid, such GCIs are prone to be unstable against the grid impedance variation. To address the above problem, a delay compensation strategy is proposed for an LCL-type GCI, in which a second-order low-pass filter (LPF) is inserted into the capacitor-current-feedback path. This increases the virtual-positive-resistance-region from one-sixth of the sampling frequency to almost within the Nyquist frequency. Therefore, the robustness of the grid-connected inverter can be significantly enhanced against the grid impedance variation. Furthermore, to ensure the system stability and improve the injecting current tracking performance, a proportional complex integral (PCI) strategy is adopted. More specifically, a robustness analysis is conducted under discrete domain. Finally, a 2-kW prototype is constructed, and the experimental results confirm the effectiveness of the theoretical expectations.

INDEX TERMS LCL-type grid-connected inverter, delay compensation, active damping, robustness, weak grid

NOMENCLATURE

| LCL | Inductor-capacitor-inductor |
|----|----------------------------|
| GCIs | Grid connected inverters |
| CCFAD | Capacitor-current-feedback active damping |
| VPRR | Virtual-positive-resistance-region |
| RHP | Right-half-plane |
| PWM | Pulse width modulator |
| PCI | Proportional complex integral |
| $U_{dc}$ | Input DC voltage |
| $S_1, S_4$ | Switch tubes of the inverter |
| $U_{in}$ | Output voltage of the inverter |
| $L_1, L_2, C_1$ | Inverter side / grid side filter inductor, filter capacitor |
| $L_g$ | Grid impedance |
| $U_g$ | Grid voltage |
| $U_{PCC}$ | Voltage at the point of common coupling |
| SOGI | Second-order generalized integrator |
| PLL | Phase locked loop |

$K_{ad}$ Feedback coefficient of the capacitor current
LPF Low-pass filter
SCR Short-circuit ratio
$S_N$ Power capacity of the grid
$K_{PWM}$ Transfer function of the inverter
$i_1, i_C$ Inverter-side current, filter capacitor current
$u_C$ Grid-connected current
$f_0$ Frequency of the grid voltage
$\omega_0$ Fundamental angular frequency of the grid voltage
$u_{PCC}$ Filter capacitor voltage
$T_s$ Sampling frequency, sampling period
$T_{r}$ Resonant angular frequency, resonant frequency of the LCL-type
filter

\( T_d \)  
Control delay  

\( k_f \)  
Gain coefficient of the second-order low-pass filter  

\( \tau \)  
Filter time coefficient of the second-order low-pass filter  

\( \omega_s \)  
Cur-off frequency of the second-order low-pass filter  

\( \omega_r \)  
Constraint angular frequency  

\( \omega_d \)  
External disturbance angular frequency  

\( k_p, k_i \)  
Proportional coefficient, integral coefficient of the PCI  

PI  
Proportional integral  

PR  
Proportional resonant  

QPR  
Quasi-proportional-resonant  

\( x_{\alpha}, x_{\beta} \)  
Grid-connected current error in \( \alpha\beta \) coordinate frame  

\( m_{\alpha}, m_{\beta} \)  
Intermediate variables in \( \alpha\beta \) coordinate frame  

\( y_{\alpha}, y_{\beta} \)  
Out signal of PCI in \( \alpha\beta \) coordinate frame  

\( f_c \)  
Cutoff frequency  

PM  
Phase margin  

GM1, GM2  
Gain margin at \( f_c \), gain margin at \( f_c/2 \)  

THD  
Total harmonic distortion  

DSP  
Digital signal processor  

IGBT  
Insulated gate bipolar transistor  

I. INTRODUCTION

Inductor-capacitor-inductor (LCL)-type grid connected inverters (GCIs) are commonly used in the distributed generation power systems owing to their excellent harmonic suppression capability [1-3]. Generally, the inherent resonance phenomenon of an LCL-type filter can challenge system stability, and the corresponding phase may pass through \(-180^\circ\) at the resonance frequency \( f_r \), causing a negative crossing. To suppress the resonance spike, conventional solutions include passive damping and active damping methods [4][5]. Compared with passive damping schemes, active ones are free of power losses and additional resistance, and achieve a better suppression effect [6][7]. Capacitor-current-feedback active damping (CCFAD) has been more interesting than other algorithms, owing to its effectiveness and simplicity [8].

In a digitally controlled system, the conventional CCFAD is no longer equivalent to a virtual pure resistor, instead it manifested as a virtual frequency-dependent impedance connected in parallel with a filter capacitor [9-13]. As shown in [9], the boundary frequency of the conventional CCFAD is one-sixth of the sampling frequency \( f_s/6 \), i.e., the virtual-positive-resistance-region (VPRR) is \( f_s < f_c/6 \) and the negative damping region is \( f_s > f_c/6 \). When \( f_s \) is in the negative damping region, two right-half-plane (RHP) poles are introduced in the open-loop gain, leading a non-minimum phase system, which may result in system instability [15][16]. In addition, the low amplitude margin causes the system to oscillate as \( f_c \) approaches \( f_s/6 \). Particularly, when \( f_s = f_c/6 \), regardless of the value of the feedback coefficient, the system is in an unstable state. In many applications, due to the increasing penetration of the new distributed resources and long distance transportation, the grid impedance can no longer be ignored, and the grid exhibits a weak characteristic instead of normal grid conditions. Meanwhile, the short circuit ratio (SCR) is used to divide stiff grid and weak grid. The grid with SCR>10 belongs to a stiff grid, and SCR≤10 a weak grid [17]. In a weak grid, the actual resonance frequency \( f_c \) may pass through the boundary frequency \( f_s/6 \) as the grid impedance increases, resulting in system instability [20]. To address this problem, system stability can be ensured by setting the inherent resonant frequency \( f_c \) of an LCL-type filter far away from \( f_s/6 \), however, a low \( f_c \) requires a larger filter parameter, and a high \( f_c \) is difficult to realize in hardware. Therefore, it is necessary to increase the boundary frequency, for facilitating the implementation and enhancing the system robustness.

Recently, there have been many efforts to extend the VPRR. There are mainly categorized into two types of approaches. Considering the control delay originating from the digitally controlled modulation, the first simple approach is to directly reduce the control delay. As shown in [18], the computation delay is modified by shifting the capacitor current sampling instant to the pulse width modulation (PWM) reference update instant. However, this sampling scheme may cause aliasing of the modulation wave, and it is sensitive to switching noise. To deal with this drawback, a real-time computation scheme with a dual sampling mode is proposed, which removed the delay from the inner and outer control loops simultaneously, while maintaining the the control performance and system robustness [19]. However, it made the existing noise resistance insufficient. The second type of technique to extend the VPRR is to compensate the control delay. In [20][21], the state prediction delay compensation method was used to optimize the virtual impedance function, the phase lag caused by the control delay was compensated, and the modulus of the output impedance was increased. However, in practical implementation, the model error may cause prediction deviation and affect the system stability. In [22], a notch filter was proposed to improve the delay compensation, and the VPRR was extended to \((0, f_c/3)\), however, its implementation requires accurate resonant frequency determination. Compared to a state observer and a notch filter, phase compensation is much more commonly used because it does not need specific model information. Typical phase compensations include lead lag compensation, first-order lead compensation and second-order integrator [23-27]. In [27], a proportional integral (PI)-based CCFAD was used to extend the VPRR to almost within the Nyquist frequency \( f_c/2 \), significantly improving the system stability and robustness against grid impedance.
variations. However, the integral term may continuously accumulate the dc bias under a weak grid.

Based on previous studies, this paper proposes an improved delay compensation strategy with a second-order low-pass filter (LPF) connected in series in the capacitor current feedback channel, effectively extending the VPRR to almost within the Nyquist frequency \(0, f_s/2\), without modifying the sampling-instant. Specifically, the resonant frequency forbidden region of the LCL filter is eliminated, breaking the limitation of the resonant frequency. Simultaneously, a proportional complex integral (PCI)[28] is used as a grid-connected current controller, realizing zero steady-state error tracking. Furthermore, through carefully parameter design in weak grid, the grid-connected inverter system can achieve high reliability, strong robustness, and high noise immunity adapt to the grid impedance variation.

The main contributions of this study are summarized as follows:
1) Analyze the influence of the control delay, and propose a easily implemented delay compensation for LCL-type grid-connected inverter system.
2) Show how the proposed delay compensation strategy extends the positive damping region and permits the inverter to have strong robustness against the wide grid impedance variation.
3) Design the closed-loop parameters after compensation based on the requirements of the gain and amplitude margins, and conduct the robustness analysis of a zero-pole distribution diagram in a discrete domain

The remainder of this paper is organized as follows: In Section II, a digitally controlled LCL-type GCI with the conventional CCFAD are described, and the influence of the control delay on the stability is discussed. In Section III, the delay compensation design is presented. In Section IV, the parameter selection and robustness analysis to ensure the robustness against grid impedance variations are discussed. In Section V, the simulation and experimental results are provided to validate the proposed strategy. Finally, the conclusions are drawn in Section VI.

II. INFLUENCE OF CONTROL DELAY ON LCL-TYPE GRID-CONNECTED INVERTER

Fig. 1 shows the configuration of an LCL-type GCI system[22]. In the figure, \(U_{dc}\) is the input DC voltage provided by distributed generation, and switch tubes \(S_1-S_4\) form an inverter bridge. The LCL filter is composed of \(L_1, L_2\) and \(C\), and \(L_g\) is the grid impedance. \(U_g\) is the grid voltage and \(U_{PCC}\) is the voltage at the point of common coupling (PCC). The grid angular frequency \(\omega_0\) and the phase \(\theta\) are obtained using a phase locked loop (PLL), to realize synchronization between the injecting current and the power grid. A delay compensation is connected in series in the capacitor current path. The grid-connected current in the \(ab\) frame are obtained using a second-order generalized integrator (SOGI).

\[
L_g = \frac{U_g^2}{2f_0\cdot SCR\cdot S_N} \tag{1}
\]

Where \(f_0\) is the grid frequency, \(S_N\) is the rated capacity of the grid-connected inverter. Therefore, \(L_g\) can be used to simulate the weak grid. It can be seen from (1) that with the grid impedance increases, SCR becomes smaller, meaning the grid becomes much weaker.

A. MODELING OF LCL-TYPE GRID-CONNECTED INVERTER

The mathematical model of the digitally controlled LCL-type inverter based on Fig. 1, is shown in Fig. 2, where \(K_{ad}\) is the feedback coefficient of the capacitor current, \(K_{PWM}\) is the transfer function of the inverter, and the grid-connected current control \(G_d(s)\) is achieved by a PCI controller.

\[
G_d(s) \text{ represents the control delay, including the computation delay and modulation delays, where the calculation delay is } e^{-sT_s}, \text{ and the modulation delay generated by a zero-order holder (ZOH) is } e^{-0.5sT_s} (T_s \text{ is the sampling period of the system}). G_d(s) \text{ is expressed as}
\]
Referring to Fig. 2, an equivalent diagram is drawn by replacing the feedback of the capacitor current \( i_C \) with the capacitor voltage \( u_C \), and moving its output to the output of \( 1/sL_1 \). The red dotted and red solid lines represent the original and transformed forms of the model, respectively. Therefore, the conventional CCFAD is equivalent to a virtual impedance \( Z_{eq1}(s) \), which can be expressed as follows:

\[
Z_{eq1}(s) = \frac{L_1}{CK_{ad}K_{PWM}} e^{1.5sT_s}
\]

(3)

Substituting \( s = j\omega \) into (3) yields \( Z_{eq1}(j\omega) \), which can be expressed as a parallel connection of a resistor \( R_{eq1}(\omega) \) and a reactor \( X_{eq1}(\omega) \), as shown in Fig. 3.

\[
Z_{eq1}(\omega) = R_{eq1}(\omega) / jX_{eq1}(\omega)
\]

(4)

where

\[
\begin{align*}
R_{eq1}(\omega) &= \frac{L_1}{CK_{ad}K_{PWM}} \cos(1.5\omega T_s) \\
X_{eq1}(\omega) &= \frac{L_1}{CK_{ad}K_{PWM}} \sin(1.5\omega T_s)
\end{align*}
\]

(5)

According to (5), a frequency characteristic analysis of the \( R_{eq1}(\omega) \) and \( X_{eq1}(\omega) \) can be conducted, as shown in Fig. 4. It is noticed that the resistor \( R_{eq1}(\omega) \) damps the resonance of the LCL filter, and that the reactor \( X_{eq1}(\omega) \) changes the resonance frequency. The frequency boundary of \( R_{eq1}(\omega) \) is one-sixth of the sampling frequency \( f_s/6 \), thus, \( R_{eq1}(\omega) \) presents positive impedance characteristics in range \((0, f_s/6)\) and presents negative impedance in range \((f_s/6, f_s/2)\). When \( f_s \) exceeds \( f_s/6 \), two right half-plane poles are introduced into the open-loop gain, threatening the system stability. In addition, the boundary frequency of \( X_{eq1}(\omega) \) is one-third the sampling frequency \( f_s/3 \), thus, \( X_{eq1}(\omega) \) presents inductive characteristics in range \((0, f_s/3)\), whereas capacitive characteristics in range \((f_s/3, f_s/2)\).
FIGURE 6. Bode diagrams of loop gain under various grid impedance values with conventional CCFAD.

Fig. 5 shows the Bode diagrams of the open-loop gain $T_D(s)$ under different control delays $T_d$. It can be seen that with the increase in the control delay, the phase lag becomes increasingly large, and the possibility of crossing $-180^\circ$ increases, causing the system to be unstable.

Fig. 6 shows the Bode diagrams of the loop gain $T_D(s)$ under the conventional CCFAD with varying grid impedance values. It is noticed that when the resonant frequency $f_r$ exceeds $f_s/6$, with the increase of the grid impedance $L_g$, the resonant frequency decreases and shifts to the left, gradually approaching the critical frequency $f_s/6$. Thus, it is difficult for the resulting amplitude margin to meet the requirements. To enhance the stability and robustness of the system, it is necessary to improve the boundary frequency and expand the effective damping range by a delay compensation.

III. IMPROVED DELAY COMPENSATION ACTIVE DAMPING STRATEGY

To reduce the influence of the control delay, a second-order LPF is selected to meet the above system requirements. In the s-domain, the transfer function $G_f(s)$ of the second-order LPF can be expressed as follows

$$G_f(s) = \frac{k_f}{(\tau s)^2 + \tau s + 1}$$ \hspace{1cm} (8)

where $\tau$ is the filter time coefficient and $k_f$ is the gain coefficient. The block diagram of the LCL-type GCI considering the second-order LPF is obtained as shown in Fig. 7, where the dotted line represents the capacitor-current path in series with the second-order LPF, and the solid line represents the transformed forms of equivalent model, respectively.

FIGURE 7. Equivalent block diagram of delay compensation strategy.

In addition, we define $\omega_b$ as the cut-off frequency of the second-order LPF. To ensure the system stability and suppress the influence of an external disturbance, $\omega_b$ should be satisfied as follows:

$$0.1\omega_s > \omega_b > 10\omega_d$$ \hspace{1cm} (9)

where $\omega_s$ is the constraint angular frequency to ensure the stability of the control system and $\omega_d$ is the external disturbance angular frequency. Substituting $\omega_b = 0.5098\tau^{-1}$ into (9) yields

$$0.05098\omega_s > \tau > 5.098\omega_d$$ \hspace{1cm} (10)

Because the influence of other noises is not be considered in this study, so let $\omega_d = 0$, fundamental frequency $\omega = 100\pi$ rad/s, and $\omega_s = \omega_r$.

As shown in Fig. 7, with the delay compensation, the virtual impedance $Z_{eq2}(s)$ is expressed as

$$Z_{eq2}(s) = \frac{L_i((\tau s)^2 + \tau s + 1)}{CK_{ad}k_fK_{PWM}}e^{1.5s\tau}$$ \hspace{1cm} (11)

Substituting $s = j\omega$ into (11) yields

$$Z_{eq2}(j\omega) = R_{eq2}(\omega) / jX_{eq2}(\omega)$$ \hspace{1cm} (12)

where

$$R_{eq2}(\omega) = \frac{L_i[(1 - \tau^2\omega^2) + \tau^2\omega^2]}{CK_{ad}k_fK_{PWM}g_f(\omega)}$$

$$X_{eq2}(\omega) = \frac{L_i[(1 - \tau^2\omega^2) + \tau^2\omega^2]}{CK_{ad}k_fK_{PWM}g_x(\omega)}$$ \hspace{1cm} (13)

where

$$g_f(\omega) = \sqrt{(1 - \tau^2\omega^2)^2 + \tau^2\omega^2}\sin(1.5\omega\tau_s + \phi)$$

$$g_x(\omega) = -\sqrt{(1 - \tau^2\omega^2)^2 + \tau^2\omega^2}\cos(1.5\omega\tau_s + \phi)$$ \hspace{1cm} (14)

$$\phi = \arcsin(\frac{1 - \tau^2\omega^2}{\sqrt{(1 - \tau^2\omega^2)^2 + \tau^2\omega^2}})$$

Based on (13) and (14), the sign of $R_{eq2}(\omega)$ depends on the sign of $g_f(\omega)$; the inductive and capacitive characteristics of $X_{eq2}(\omega)$ depend on the sign of $g_x(\omega)$. Substituting $\omega = 2\pi f$ into (14) yields
Based on (15), \( g(r) \) and \( g_s(r) \) are periodic functions related to the sampling frequency \( f_s \). When \( R_{eq2}(\omega) > 0 \), \( g(r) \) is positive in quadrants 1 and 2, and because \( 1.5\omega T_s + \varphi = 3\pi f / (f_s + \varphi) = (k\pi, (k + 1)\pi) \) \( (k = 0, 2, 4,...) \), we obtain \( f = (k\pi / 3 - \varphi f_s / (3\pi)), (k + 1)\pi / 3 - \varphi f_s / (3\pi) \). Because the effective damping region of the system considered in this study is the main region of the periodic function \( g(r) \) \( (k = 0) \) and \( f \) is always greater than 0, the frequency range when \( R_{eq} > 0 \) is \( f \in (0, f_s) \), where \( f_s = f_s / 3 - \varphi f_s / (3\pi) \) and \( f_s = f_s / 3 - \varphi f_s / (3\pi) \). Summarizing, the variation curves of \( g(r) \) and \( g_s(r) \) with the increase in \( f \) are in the range of \( (0, f_s / 2) \).

According to (13), the frequency characteristic of \( R_{eq} \) with \( r \) changes is described in Fig.8. It is indicated that, with \( r \) decreases, the boundary frequency of \( R_{eq2}(\omega) \) moves to right. When \( r = 0.001s \), the boundary frequency is equal to Nyquist frequency \( f_s / 2 \), therefore, we select \( r = 0.001s \).

**Fig. 9** shows the frequency characteristics of virtual impedance. In figure, \( R_{eq1}(\omega) \) and \( X_{eq1}(\omega) \) represent virtual impedance with the conventional CCFAD; \( R_{eq2}(\omega) \) and \( X_{eq2}(\omega) \) represent virtual impedance with the phase-lead compensator in [27]; \( R_{eq3}(\omega) \) and \( X_{eq3}(\omega) \) represent virtual impedance with the second-order LPF compensator in this paper. It is noticeable that the frequency boundary of the equivalent resistor is shifted to \( f_s / 4 \) by \( R_{eq3}(\omega) \), and shifted to \( f_s / 2 \) by \( R_{eq2}(\omega) \). Thus, the proposed compensation widens the effective damping region and increases the robustness. In addition, based on the comparison analysis of \( X_{eq1}(\omega), X_{eq2}(\omega) \) and \( X_{eq3}(\omega) \), the frequency boundary of the proposed strategy approaches \( f_s / 2 \), which is higher than those of the former methods.

**IV. PARAMETER SELECTION AND ROBUSTNESS ANALYSIS**

**A. GRID-CONNECTED CURRENT BASED ON PCI CONTROLLER**

To reduce the steady-state error, a PCI is adopted in the grid-connected current regulator [28], and the expression is

\[
G_c(s) = k_p + \frac{k_i}{s} = k_p + \frac{k_i (s + j\omega_0)}{s^2 + \omega_0^2}
\]

(16)

where \( k_p \) is the proportional coefficient, \( k_i \) is the integral coefficient, and \( \omega_0 = 2\pi f_0 \) is the fundamental angular frequency, where \( f_0 \) is the fundamental frequency of the grid voltage. When \( \omega_0 = 0 \), \( G_c(s) \) is equivalent to a PI controller, and when \( s = j\omega_0 \), \( G_c(s) \) is equivalent to a proportional resonant (PR) controller. A quasi-proportional-resonant (QPR) controller is the same as a PR one.

**Fig. 10** compares the frequency characteristics of several controllers. It can be seen that under the same parameters, the phase frequency characteristics of the PCI controller are better than those of the PR controller, and the resonant bandwidth of the PCI controller is smaller than that of QPR controller. For the GCI, the above suggests that the noise bands amplified by the high gain of the PCI controller are...
narrow, and the number of harmonics entering the system is reduced.

Based on the physical relevance of complex number j, a PCI controller can be implemented by a static two-phase αβ coordinate, as shown in Fig. 11.

![PCI controller diagram](image)

**FIGURE 11.** Structure diagram of PCI controller in αβ frame

In Fig. 11, $x_a$ and $x_b$ represent grid-connected current error in αβ coordinate frame, respectively. The intermediate variables $m_a$ and $m_b$ are orthogonal variables, which meet $m_a = j m_b$, $m_b = -j m_a$. Then, through the PCI controller, the output signal $y_a$ and $y_b$ can be obtained. According to the theory of complex variable function, bilinear transformation (i.e. Tustin transformation) is used to convert continuous time system and discrete-time system, which is written as follows

$$G_c(z) \Leftrightarrow G_c(s), s = \frac{2}{T} \frac{1 - z^{-1}}{1 + z^{-1}} \quad (17)$$

According to (17), the discrete equation of the PCI controller is expressed as follows

$$m_a(k) = m_a(k-1) + \frac{kT}{2} [x_a(k) - \frac{\alpha_0}{k_i} m_p(k)]$$

$$+ \frac{kT}{2} [x_a(k-1) - \frac{\alpha_0}{k_i} m_p(k-1)] \quad (18-1)$$

$$m_b(k) = m_b(k-1) + \frac{kT}{2} [x_p(k) + \frac{\alpha_0}{k_i} m_a(k)]$$

$$+ \frac{kT}{2} [x_p(k-1)+ \frac{\alpha_0}{k_i} m_a(k-1)] \quad (18-2)$$

$$y_a(k) = k_p x_a(k) + m_a(k) ; y_b(k) = k_p x_p(k) + m_b(k) \quad (19)$$

where $x_a(k)$, $x_b(k)$, $m_a(k)$, $m_b(k)$, $y_a(k)$ and $y_b(k)$ represent the values at the moment k, and $x_a(k-1)$, $x_b(k-1)$, $m_a(k-1)$, $m_b(k-1)$, $y_a(k-1)$ and $y_b(k-1)$ represent the values at the moment k−1. This realizes an discrete control of the PCI controller and reduces the harmonics of the grid-connected current. Therefore, the overall control performance of the system is improved.

### B. PARAMETER DESIGN

Considering the delay compensation proposed in this paper, the system loop gain $T_{D2}(s)$ can be rewritten as

$$T_{D2}(s) = \frac{1}{sL_1(L_2 + L_g)C} \frac{G_c(s)K_{PWM}}{s^2 + \frac{1}{CZ_{eq2}}(s) + \omega_i^2} \quad (20)$$

As presented by Fig. 9, the effective damping range of the proposed delay compensation strategy is $(0, f_s/2)$. Based on the requirements of gain margin GM1 at $f_c$, gain margin GM2 at $f_s/2$ and phase margin (PM) at cutoff frequency $f_c$, the value range of capacitor-current-feedback coefficient $K_{ad}$ can be determined.

$$\begin{align*}
GM_1 &= -20 \log |T_{D2}(j2\pi f_c)| \\
GM_2 &= -20 \log |T_{D2}(j2\pi f_s/2)|
\end{align*} \quad (21)$$

The gain of the current controller $G_c(s)$ is approximately a proportional component, expressed as $|G_c(j2\pi f_c)| \approx |G_c(j2\pi f_s/2)| = k_p$, when $f_c$ is less than $f_s$ and $f_s/2$. The system loop gain in this frequency range can be expressed as

$$|T_{D2}(j2\pi f_s/2)| \approx \frac{K_{PWM}G_c(j2\pi f_c)}{j2\pi f_s/2} \quad (22)$$

$K_{ad}$ GM1 and $K_{ad}$ GM2 are respectively the constraint conditions of active damping coefficient $K_{ad}$ under the requirements of GM1 and GM2, which are derived as follows:

$$K_{ad} - GM_1 = \frac{10^{GM_{1/20}}2\pi f_c L_1}{K_{PWM}(1 - 4\pi^2 f_c^2 + 2\pi f_c)} \quad (23)$$

$$K_{ad} - GM_2 = \frac{2\pi f_c}{K_{PWM}(1 - 4\pi^2 (f_s/2)^2 + 2\pi f_s/2)} \quad (24)$$

The constraint condition of the PM on the feedback coefficient $K_{ad}$ PM is recorded as $K_{ad}$ PM, which is obtained as

$$K_{ad} - PM = \frac{2\pi L_1(f_s^2 - f_c^2)}{f_c K_{PWM} \cos(3\pi T_s)} \quad (25)$$

$$\left[\pi f_c^2 - 2\pi f_c(10^{GM_{2/20}}f_0 - f_c) \tan(3\pi T_s) \right] \left[\tan(3\pi T_s) + PM \right]$$

$$\left\{ +1 + \pi f_c^2 \left[\tan(3\pi T_s) + PM \right] - \tan(3\pi T_s) \right\}$$

Fig. 12 shows the constraint region of the active damping coefficient $K_{ad}$. To ensure sufficient dynamic performance and steady-state accuracy of the system, the PM is set as 45°, gain $T_{D0}$ at the fundamental loop is set as 50dB, and $f_c = 1200$ Hz. The upper limit of $K_{ad}$ depends on GM1 and PM, and the lower limit depends on GM2, making $K_{ad} = 0.02$. 

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In order to verify the robustness and adaptability of the proposed delay compensation strategy in weak grid, according to (20), we can obtain the Bode diagrams of the system loop gain $T_{D2}(s)$, as seen in Fig. 13. From Fig. 13(a), it can be seen that when $L_g = 0$ mH, the actual resonant frequency $f_r < f_s/2$, than with the increase in $L_g$, $f_r$ decreases and shifts to the left, always in the positive damping range. The phase frequency curve only crosses once at $-180^\circ$, and the corresponding amplitude margin is greater than 0, meeting stability requirements. Therefore, with the delay compensation strategy, the system shows strong robustness against the grid impedance variation.

Fig. 13 (b) shows the Bode diagrams of the system loop gain when $L_g = 0$ mH, in range of $f_r > f_s/2$. It can be seen that when $f_r > f_s/2$, the system is in the negative damping range, and two right half-plane poles are introduced into the loop gain, causing the phase frequency curve to cross $-180^\circ$ twice. According to the Nyquist stability criterion, the stability of system needs to be satisfied: $P = 2(N^+ - N^-)$, where $N^+$ and $N^-$ are the numbers of positive and negative crossings of $-180^\circ$, respectively. With the increase in $L_g$, $f_r$ decreases and shifts to the left of $f_s/2$, always in the positive damping range, meeting stability requirements. The above analysis shows that using the proposed delay compensation, the damping range is effectively improved and the robustness is enhanced.

C. SYSTEM ROBUSTNESS ANALYSIS

From (6), the loop gain $T_{D2}(s)$ contains a time-delay component $e^{-1.5sT_z}$; therefore, it is impossible to directly calculate the closed-loop poles of the system. Thus, the expression of the system loop gain $T_{D2}(z)$ in the $z$-domain can be obtained first [33-36], which is expressed as follows

$$T_{D2}(z) = T_{D2}(s) \left| \frac{K_{PWM}G_c(z)}{\omega_r L_1(z-1)} \right|$$

$$\frac{\omega_r T_z[z^2 - 2z \cos(\omega_r T_z) + 1] - (z-1)^2 \sin(\omega_r T_z)}{z[z^2 - 2z \cos(\omega_r T_z) + 1] + (z-1)^2 \sin(\omega_r T_z)}$$

where $G_c(z)$ and $G_d(z)$ are the discrete expressions of PCI the current controller and delay compensation, respectively. Consequently, the closed-loop transfer function $G_C(z)$ of the system can be obtained as

$$G_C(z) = \frac{G_d(z)K_{PWM}}{\omega_r L_1(z-1)}$$

$$\frac{\omega_r T_z[z^2 - 2z \cos(\omega_r T_z) + 1] - (z-1)^2 \sin(\omega_r T_z)}{z[z^2 - 2z \cos(\omega_r T_z) + 1] + (z-1)^2 \sin(\omega_r T_z)}$$

$$\left[ \frac{z((z^2 - 2z \cos(\omega_r T_z) + 1) - (z-1)^2 \sin(\omega_r T_z))}{\omega_r L_1} \right]$$

$$+ G_c(z)K_{PWM}[\omega_r T_z[z^2 - 2z \cos(\omega_r T_z) + 1] - (z-1)^2 \sin(\omega_r T_z)]$$

(29)
In order to analyze the stability of the grid-connected system in detail, using (29), pzmap diagrams of $G_c(z)$ can be obtained, and the result is shown in Fig. 14.

![Pzmap Diagrams](image)

**FIGURE 14.** Pzmap diagrams of discrete system. (a) conventional CCFAD. (b) with phase-lead compensator (c) with second-order LPF compensator.

Fig. 14 shows the closed-loop pole distributions when the grid impedance changes as $L_g = 0-5mH$, where the directions of the arrows represent the directions of the $L_g$ increases. As seen from Fig. 14(a), before the delay compensation, with the increase in $L_g$, the closed-loop poles gradually move from the unit circle to the boundary, even moving outside the unit circle when $L_g = 2mH$. Comparatively, Fig. 14(b) shows that with the phase-lead compensator, the closed-loop poles approach the unit boundary with the increase in $L_g$, and move outside the unit circle when $L_g = 4mH$.

Fig. 14(c) shows that with the proposed second-order LPF compensation, the poles are farther away from the unit circle boundary even when $L_g = 5mH$. It can be seen that under different grid impedance values, the poles of the closed-loop system after compensation are in the unit circle, and the system has good stability. Therefore, the delay compensation method proposed in this paper can effectively improve the stability of grid connected current under weak grid and improve the robustness of the system.

V. SIMULATION AND EXPERIMENTAL RESULTS

To verify the effectiveness and correctness of the delay compensation strategy proposed in this paper, an LCL-type GCI was modeled in the MATLAB/Simulink environment and built experimentally. The grid voltage was 110 V/50 Hz, the grid-connected current reference full load $I_{gref}$ was 20 A, and the half load is 10 A. In addition, the other main parameters as listed in Table I were used.

| Symbol | Quantity | Value          |
|--------|----------|----------------|
| $f_s$  | Switching frequency | 10 kHz         |
| $L_g$  | Grid impedance       | 1 mH, 3 mH, 5mH |
| $u_{dc}$ | DC voltage           | 311 V          |

A. PERFORMANCE ANALYSIS BY SIMULATION

In this part of the study, a steady-state simulation was conducted using the conventional CCFAD described in section II, the phase-lead compensator in [27] and the second-order LPF compensator in this paper, where the $L_g$ values were set at as 1mH, 3mH and 5mH. The simulation waveforms are shown in Fig. 15, Fig. 16 and Fig. 17.

The reasonable parameter design enables the system to achieve good control performance in a strong grid before compensation, as can be seen from Fig.15(a). When $L_g = 1mH$, both $i_g$ and $U_{pcc}$ have smooth sinusoidal waveforms, and the GCI system can operate stably. As shown in Fig. 15(b), when $L_g = 3mH$, large ripples are generated in both $i_g$ and $U_{pcc}$, therefore, their waveforms oscillate. As $L_g$ increases, the waveforms degrade, and when $L_g = 5mH$, the GCI system diverges and is in an unstable state as shown in Fig. 15(c).

Fig. 16 shows the steady-state waveforms obtained with the phase-lead compensator under different grid impedance values. Because the compensated system expands the positive resistance range of active damping to $f_s/4$, the GCI system can work stably when $L_g$ changes over a wide range. However, when $L_g = 5mH$, large ripples are generated in both $i_g$ and $U_{pcc}$, therefore, their waveforms oscillate. As $L_g$ increases, the waveforms degrade, and when $L_g = 5mH$, the GCI system diverges and is in an unstable state as shown in Fig. 15(c).

Fig. 17 shows the steady-state waveforms obtained with the proposed delay compensation under different grid impedance values. Because the compensated system expands the positive resistance range of active damping, the GCI system can work stably when $L_g$ changes over a wide range. Moreover, the system presents high grid-connected...
current quality, regardless of $L_g$ being set as 1mH, 3mH, and 5mH, as shown in Fig. 15. In conclusion, the steady-state simulation results are consistent with the above theoretical analysis.

![Steady-state simulation waveforms with conventional CCFAD.](image1)

![Steady-state simulation waveforms with phase-lead compensator.](image2)

![Steady-state simulation waveforms with the second-order LPF compensator.](image3)

![THDs of $i_g$ with different $L_g$ values.](image4)

Referring to Fig.18, the total harmonic distortion (THD) results of the $i_g$ simulated waveforms versus $L_g$ values are presented. It presents that the THD values with the proposed delay compensation are 2.72%, 2.82%, 2.93% and 3.56%, respectively, which are far lower than those with the traditional CCFAD and the phase-lead compensator. In addition, when $L_g = 5$mH, the THDs with no delay compensation largely exceeds the grid-connected standard.

**B. EXPERIMENTAL RESULTS**

To verify the effectiveness of the proposed delay compensation scheme and the correctness of the simulation results, a 2-kW LCL-type single-phase GCI prototype was constructed, which is shown in Fig.19. A TMS320F28335 floating-point digital signal processor (in the figure, DSP)
(Texas Instruments) was used as the controller processing chip, and an insulated gate bipolar transistor (IGBT) of Infineon Technologies with $V_{CES} = 1200V$ and $I_{C} = 100A$ was selected. In addition, a voltage sensor LV 25-P of LEM company was adopted, whose measurement accuracy is 0.9%. A current sensor LA 100-P was adopted, with a measurement accuracy of 0.45%. The main experimental parameters are listed in Table I.

The experimental verification was conducted using $L_g$ values of 1mH, 3mH, and 5mH, respectively. Considering the practical scenario, the experimental parameters of the GCI system could be slightly different from those in Table I. The steady-state experimental results are show in Fig. 20, Fig. 21 and Fig. 22, respectively.

Fig. 20 shows the steady-state experimental results with the traditional CCFAD under the full-load condition. It can be seen that in Fig. 20(a), when $L_g = 1mH$, the GCI system can operate stably in a stiff grid and the measured THD is 3.1%. When the system works in a weak grid, both $i_g$ and $U_{pcc}$ show many ripples, as presented in Fig. 20(b) and (c), respectively. It can be seen that the grid impedance has a severe impact on both $i_g$ and $U_{pcc}$, which is consistent with the simulation waveforms in Fig. 15.

Fig. 21 shows the steady-state experimental results with the phase-lead compensator under the full-load condition. It can be seen that in Fig. 21(a) and Fig. 21(b), the grid-connected system can operate stably when $L_g = 1mH$ and $L_g = 3mH$. However, when $L_g = 5mH$, both $i_g$ and $U_{pcc}$ show many ripples, as presented in Fig. 21(c), which is consistent with the simulation waveforms in Fig. 16.

Fig. 22 shows the steady-state experimental results with the proposed delay compensation scheme under the full-load condition. Because the compensated system expands the positive resistance range of active damping, the system can work stably when $L_g$ changes over a wide range. Moreover, this is owing to the grid-connected current PCI regulator which has a sufficiently high fundamental gain and ensures the tracking accuracy. In conclusion, the steady-state experimental results suggests that, using the proposed delay compensation, the LCL-type inverter remains in a satisfactory steady-state and presents strong robustness against the grid impedance variation.

Fig. 23 shows that grid-connected current THDs for $L_g = 5mH$ with different strategies. It can be obviously noticed that with the proposed second-order LPF delay compensation, the THD meet the grid-connected requirements, even in a weak grid.
To verify the dynamic performance of the delay compensation strategy, dynamic experiments were conducted, and the results are shown in Fig. 24. Fig. 24(a) presents that as full load 20A switches to half load 10A, the grid-connected current \( i_g \) tracks the reference \( I_{g_{ref}} \) smoothly, and the adjustment time is only 1ms. Fig. 24(b) shows that as half load 10A switches to full load 20A, the adjustment time is only 2ms. Moreover, the THD results of the corresponding analysis of the current \( i_g \) under full load 20A are shown in Fig. 24(c), exhibiting that the requirements are met. It can be seen that using the proposed delay compensation strategy, the system can achieve good dynamic performances and exhibits high robustness against the grid impedance variation.

VI. CONCLUSION

Aiming at the problem of finding a method to expand the equivalent positive damping region of active damping in an LCL-type GCI system, an improved delay compensation strategy based on traditional capacitor-current active damping is proposed. Detailed procedures of the scheme design are provided, and a 2-kW experimental prototype is also constructed. The major conclusions can be summarized as follows

1) When the capacitor-current-feedback channel is connected in series with the second-order LPF, the positive damping region is expanded to \((0, f_s/2)\). In addition, the grid-connected current channel is controlled by a PCI, which realizes zero steady-state error tracking.

2) The proposed strategy improves the steady-state and dynamic performance of the system under a weak grid, exhibiting strong robustness against the grid impedance variation over a wide range.
The stability margin of the system is guaranteed by optimizing the parameters, and the system parameter design is more flexible.

The simulation and experimental results are consistent with the theoretical analysis, verifying the effectiveness of the proposed delay compensation scheme.

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