

Article

Second-Order Repetitive ADRC with Frequency Robustness and Harmonic Suppression for Single-Phase Grid-Connected Inverters

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Abstract

Conventional active disturbance rejection control (ADRC) schemes for grid-connected inverters (GCIs) suffer from performance degradation under grid frequency fluctuations and voltage distortions, primarily due to the limited bandwidth of conventional extended state observers (ESOs) and their reliance on fixed-frequency internal models. To address these issues, this paper proposes a second-order repetitive ADRC (SRC-ADRC) strategy for LCL-type single-phase GCIs. The proposed method integrates a second-order repetitive extended state observer (SRC-ESO) and a frequency-adaptive proportional-quasi resonant (FPQ) controller. By cascading an additional internal model into the repetitive control-based ESO, the SRC-ESO achieves higher internal model gain and wider resonant bandwidth, significantly improving periodic disturbance estimation accuracy under off-nominal grid frequencies. The FPQ controller ensures reference current tracking without phase error and amplitude attenuation, while the control law and observer are decoupled for independent parameter design. The robustness and frequency adaptability of the SRC-ADRC are verified through theoretical stability analysis and frequency-domain evaluation. Experiments on a 1.5 kW LCL-type single-phase GCI platform show that SRC-ADRC provides better frequency robustness, stronger harmonic mitigation, and improved current-tracking performance than the comparison methods.

Keywords: active disturbance rejection control; second-order repetitive control; grid-connected inverter; frequency robustness; harmonic suppression; extended state observer



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1. Introduction

With the continuous increase in the deployment scale of renewable energy generation and its rising penetration in power systems [1], the operating characteristics of modern power systems are undergoing significant changes [2], gradually exhibiting features such as frequency fluctuations [3], voltage distortion [4], and line impedance uncertainty [5]. Grid-connected inverters (GCIs) are the interface between renewable-energy conversion units and the utility grid, and the quality of their injected current directly affects grid stability as well as the integration capability of renewable energy [6]. Therefore, it is essential to develop injected current control strategies with strong anti-disturbance performance and frequency robustness so as to improve the stability of power systems.

Active disturbance rejection control (ADRC) is independent of precise plant modeling and is capable of eliminating both internal and external disturbances, exhibiting strong

robustness [7,8]. Nevertheless, its closed-loop performance is highly dependent on the estimation accuracy of the extended state observer (ESO) [9]. Conventional ESO estimates disturbances using an integrator [10]. According to the internal model principle, the integrator can only accurately estimate DC disturbances, but cannot accurately estimate numerous high-frequency periodic harmonics [11]. In inverter systems, disturbances mainly consist of numerous high-frequency periodic components [12]. Therefore, to address the issue that conventional ESO cannot estimate numerous high-frequency periodic disturbances, existing literature has improved the ESO structure to observe high-frequency periodic signals. Examples include high-order generalized ESO [13] and ESO based on an internal model controller [14–19]. Increasing the order of the ESO enhances the precision of disturbance estimation, but its effective bandwidth is still limited by measurement noise, which makes accurate observation of high-frequency harmonics difficult.

The internal-model-based ESO introduces a periodic signal internal model, thereby enhancing the estimation capability for specific frequency disturbances, while resulting in higher system structural complexity and greater parameter tuning difficulty [17]. On this basis, some studies have introduced repetitive control into the ESO structure, forming a repetitive control-extended state observer (RC-ESO) to enhance the observation capability for periodic disturbances. For example, embedding a repetitive control link into the ESO can effectively suppress harmonic disturbances in motor or converter systems [11,19]. However, because the poles of repetitive control are located on the unit circle, the convergence rate of the RC-ESO is relatively slow. To address this issue, some studies have proposed an improved RC-ESO, which improves the convergence speed of disturbance estimation by adjusting the pole distribution or introducing a compensation mechanism [15]. Although the above methods can effectively improve the observation accuracy of periodic disturbances under rated frequency conditions, they inherently rely on a fixed-frequency internal model [20]. When the grid frequency deviates, inconsistency arises between the internal model and the real disturbance frequency, causing a shift in the resonance characteristics of the observer, thereby significantly reducing the estimation accuracy for periodic disturbances and consequently deteriorating the overall control performance of the entire system.

To address the frequency-sensitivity problem of RC-ESO-based ADRC, this study proposes a second-order repetitive ADRC (SRC-ADRC) method. The main contributions are as follows:

- (1) An SRC-ADRC scheme is developed to overcome the frequency-sensitivity issue, enabling effective periodic-signal observation under grid frequency variations.
- (2) A second-order repetitive extended state observer (SRC-ESO) is formed by cascading a new internal model into the RC-ESO, improving periodic disturbance observation.
- (3) Compared with traditional repetitive control, the proposed second-order repetitive control (SRC) features higher internal model gain and wider resonance bandwidth, achieving effective current harmonic suppression under grid frequency fluctuations.

2. Modeling of Plant

The topology of the plant system is shown in Figure 1. E_d denotes the DC bus voltage, u_i is the inverter output voltage, L_1 and L_2 represent the filter inductors, C is the filter capacitor, R is the damping resistor, L_g is the grid impedance, u_g is the grid voltage, i_g is the injected current, i_{ref} is the reference current, and PCC denotes the point of common coupling. The phase and frequency information of the grid voltage are extracted through the phase-locked loop (PLL), and the reference current i_{ref} is generated based on the extracted information and the reference current magnitude I_{ref} .

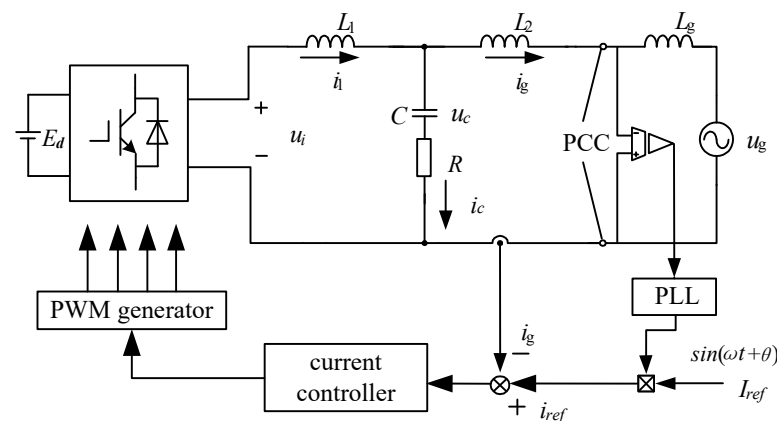


Figure 1. LCL-type single-phase grid-connected inverter.

In this paper, a passive damping configuration employing a capacitor-resistor series branch is utilized in the filter, while the parasitic resistances R_1 and R_2 of the inductors are very small and can be neglected. The time-domain expression of the injected current is

$$\begin{aligned}
 i_g'''(t) = & -\frac{(L_1 + L_2)R}{L_1 L_2} i_g''(t) - \frac{L_1 + L_2}{L_1 L_2 C} i_g'(t) \\
 & + \frac{1}{L_2} u_g''(t) + \frac{R}{L_1 L_2} u_g'(t) + \frac{1}{L_1 L_2 C} u_g(t) \\
 & + \frac{R}{L_1 L_2} u_i'(t) + \frac{1}{L_1 L_2 C} u_i(t).
 \end{aligned} \quad (1)$$

High-order ADRC suffers from structural complexity and multiple control parameters, which restricts its engineering applicability. Since ADRC regards modeling errors as internal disturbances that can be estimated and compensated in real time by the observer, the model in Equation (1) can be reduced through the Padé approximation [15,16]. The corresponding reduced-order time-domain model is expressed as

$$\begin{aligned}
 i_g'(t) = & -\frac{L_1 L_2 C}{L_1 + L_2} i_g'''(t) - RC i_g''(t) \\
 & + \frac{CL_1}{L_1 + L_2} u_g''(t) + \frac{CR}{L_1 + L_2} u_g'(t) + \frac{1}{L_1 + L_2} u_g(t) \\
 & + \frac{RC}{L_1 + L_2} u_i'(t) + \frac{1}{L_1 + L_2} u_i(t).
 \end{aligned} \quad (2)$$

According to the reduced-order time-domain model derived from Equation (2), the coefficient associated with the control input u_i is naturally expressed as $1/(L_1 + L_2)$. Therefore, in this paper, the nominal input gain is defined as $b_0 = 1/(L_1 + L_2)$, which corresponds to the nominal values of the filter inductances.

In practical applications, however, the actual plant parameters may differ from their nominal values owing to parameter uncertainties, component tolerances, aging effects, and variations in operating conditions. To accommodate these uncertainties, the actual input gain is denoted by b , and the input-gain mismatch $(b - b_0)$ is lumped into the generalized disturbance $f(t)$. Consequently, Equation (2) can be rewritten as

$$i_g'(t) = f(t) + b_0 u_i(t). \quad (3)$$

where

$$\begin{aligned} f(t) = & -\frac{L_1 L_2 C}{L_1 + L_2} i_g'''(t) - RC i_g''(t) \\ & + \frac{CL_1}{L_1 + L_2} u_g''(t) + \frac{CR}{L_1 + L_2} u_g'(t) + \frac{1}{L_1 + L_2} u_g(t) \\ & + \frac{RC}{L_1 + L_2} u_i'(t) + (b - b_0) u_i(t). \end{aligned} \quad (4)$$

It can be observed from the above equation that the generalized disturbance contains substantial periodic disturbances, and the accuracy of disturbance estimation of the ESO directly affects the control performance of ADRC.

3. Design and Analysis of the SRC-ADRC

According to Equation (2), the discrete structure of the reduced-order SRC-ADRC can be derived, in which the original third-order system is simplified into a first-order integrator model. The reduced-order plant corresponds to the block $P(z)$ in Figure 2. The remaining components in Figure 2, including the control law and the SRC-ESO, are not obtained from the plant reduction process. Instead, they constitute the proposed enhancements to the NRC-ADRC in [15]. Specifically, the conventional RC-ESO is replaced by the proposed SRC-ESO, and a frequency-adaptive quasi-resonant (FPQ) control law is introduced to improve disturbance rejection capability and reference-tracking performance under grid frequency variations.

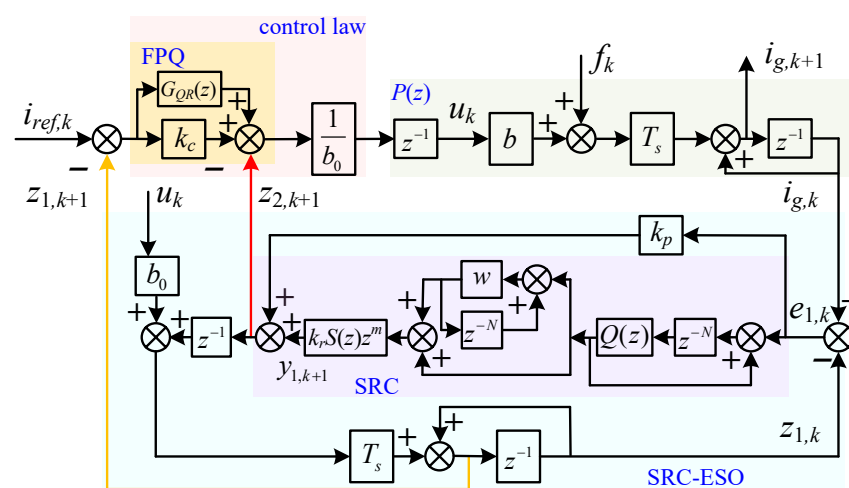


Figure 2. The discrete diagram of SRC-ADRC.

According to [19], Figure 2 can be transformed into an equivalent unity-feedback system through an equivalent block-diagram transformation, as shown in Figure 3.

$$G_c(z) = \frac{G_{SRC} + k_p}{b_0}, \quad (5)$$

$$G_h(z) = \frac{(G_{QR}(z) + k_c)[z^2 - z + T_s(G_{SRC}(z) + k_p)]}{(G_{SRC}(z) + k_p)[(G_{QR}(z) + k_c)T_s z + z^2 - z]}. \quad (6)$$

Assuming that no input gain mismatch is present ($b = b_0$), the transfer functions from the reference signal and the generalized disturbance to the injected grid current can be derived as

$$G_{i_{ref},k(z)}^{i_{g,k+1}(z)}(z) = \frac{(G_{QR}(z) + k_c)T_s}{(G_{SRC}(z) + k_c)T_s + z - 1}, \quad (7)$$

$$G_{f_k(z)}^{i_{g,k+1}(z)}(z) = \frac{T_s z^2}{z^2 - z + T_s(G_{SRC}(z) + k_p)}. \quad (8)$$

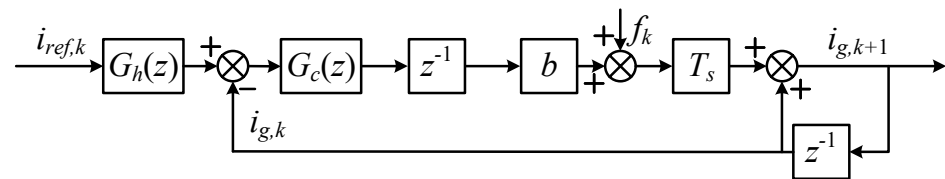


Figure 3. Equivalent unity feedback diagram of SRC-ADRC.

Equation (7) denotes the closed-loop transfer function from the reference signal $i_{ref,k}(z)$ to the grid current $i_{g,k+1}(z)$, describing the tracking characteristics of the proposed controller. Equation (8) denotes the transfer function from the disturbance input $f_k(z)$ to the grid current $i_{g,k+1}(z)$, which characterizes the disturbance attenuation performance of the system.

$$G_{QR}(z) = \frac{2T_s k_r w_c z^2 - 2T_s k_r w_c z}{z^2 + (T_s^2 w_r^2 + 2w_c T_s - 2)z - 2T_s w_c + 1}. \quad (9)$$

$$G_{SRC}(z) = k_{rc} S(z) z^m \frac{Q(z) z^{-N}}{1 - Q(z) z^{-N}} \frac{1 + w - w z^{-N}}{1 - w z^{-N}}. \quad (10)$$

$G_{QR}(z)$ is the transfer function of the frequency-adaptive quasi-resonant control, k_c is the proportional control gain, k_r is the resonant gain, ω_r is the resonant angular frequency, ω_c is the resonant bandwidth, and T_s is the sampling time. k_p is the proportional gain, which enhances the dynamic performance of SRC. $G_{SRC}(z)$ is the transfer function of the second-order repetitive control, where k_{rc} is the repetitive control gain and N is the number of delay steps in the repetitive control. $S(z)$ is a compensator used to attenuate high-frequency disturbance and further improves the stability of the ESO. The term z^{-m} is a phase-lead compensation element used to compensate for the phase lag introduced by the controlled plant, thereby reducing the steady-state error. $Q(z)$ is a zero-phase-shift low-pass filter that improves the stability margin of the controller. w is a second-order weighting factor used to adjust the frequency sensitivity of SRC.

Equations (7) and (8) show that the reference tracking and disturbance rejection performances of SRC-ADRC are independent and decoupled, allowing them to be designed independently without mutual interference [17].

3.1. Design of the Control Law

In grid-connected current control systems, the reference signal is a periodic signal synchronized in both frequency and phase with the grid voltage, while the grid frequency continuously fluctuates in real time. Therefore, the tracking performance of the control system under grid frequency fluctuations must be carefully considered. In this paper, the control law of the proposed SRC-ADRC adopts a frequency-adaptive quasi-proportional resonant control whose resonant frequency can dynamically track the grid frequency, thus ensuring accurate tracking of the periodic reference signal.

3.2. Design of the SRC-ESO

The design concept of the SRC-ESO is to construct a second-order repetitive extended state observer by cascading an additional internal model with the RC-ESO. Compared with the RC-ESO, the proposed SRC-ESO provides higher internal model gain and a wider resonant bandwidth. Consequently, it achieves more accurate observation of periodic disturbances while exhibiting enhanced frequency robustness.

For ease of analysis, set $Q(z) = 1$ and $k_{rc} = 1$. The transfer function of the ideal internal model of the SRC is expressed as Equation (11).

$$G_{ISRC} = \frac{z^{-N}(1+w-wz^{-N})}{1-(1+w)z^{-N}+wz^{-2N}} = \frac{A(z)}{1-B(z)}. \quad (11)$$

According to [18], the structure of $B(z)$ is consistent with that of the second-order repetitive control. Therefore, the proposed ESO is a second-order repetitive extended state observer.

The SRC-ESO can be modeled as an independent closed-loop system, in which $i_{g,k}(z)$ and $z_{1,k}(z)$ are the reference and feedback signals, respectively. The plant $P(z)$ is given as follows:

$$P(z) = \frac{T_s}{z^2 - z}. \quad (12)$$

The transfer function $P_0(z)$ from $y_{1,k+1}$ to $z_{1,k}(z)$ is the new plant of the SRC proportional control connected in parallel [19],

$$P_0(z) = \frac{P(z)}{1+k_p P(z)} = \frac{T_s}{z^2 - z + k_p T_s}. \quad (13)$$

3.3. Design of the Internal Model Filter

For a discrete-time system, closed-loop stability is guaranteed only when every root of the characteristic equation is located within the unit circle in the complex domain.

According to Equation (7), the SRC system contains N poles corresponding to $z^{-N} = w$ and N poles corresponding to $z^{-N} = 1$. To ensure system stability, these poles must remain inside the unit circle. For the poles associated with $z^{-N} = w$, stability can be guaranteed by selecting an appropriate second-order weighting factor w such that its absolute value is less than 1,

$$-1 < w < 1. \quad (14)$$

According to Figure 2 that the error transfer function $E(z)$ of SRC-ESO is given by

$$E(z) = \frac{i_{g,k}(z)}{1 + [G_{SRC} + k_p]P(z)}. \quad (15)$$

According to Equation (15), the characteristic polynomial of SRC-ESO is as follows:

$$[1 + k_p P(z)][1 + G_{SRC}(z)P_0(z)]. \quad (16)$$

According to Equation (16), the stability of the SRC-ESO can be guaranteed if the following conditions are satisfied:

- (1) The roots of $1 + k_p P(z)$ lie inside the unit circle.
- (2) $|1 + G_{SRC}(z)P_0(z)| \neq 0$.

Substituting the Equation (10) into stability condition (2) yields

$$\left| \frac{1 - Q(z)z^{-N} + k_{rc}S(z)z^m Q(z)z^{-N}G_1P_0(z)}{1 - Q(z)z^{-N}} \right| \neq 0, \quad (17)$$

$$G_1 = \frac{1 + w - wz^{-N}}{1 - wz^{-N}}. \quad (18)$$

When both the reference signal and the disturbance signal have frequencies that are integer multiples of the fundamental frequency, $|z^{-N}| = 1$,

$$|Q(z)(1 - k_{rc}S(z)z^m G_1 P_0(z))| < 1. \quad (19)$$

Through frequency-domain characteristic analysis, $P_0(z)$ exhibits frequency characteristics $P_0(j\omega) = N_p(\omega) \exp(j\theta_p(\omega))$, where $N_p(\omega)$ and $\theta_p(\omega)$ are amplitude-frequency and phase-frequency characteristics, respectively. Similarly, the compensator $S(z)$ has frequency characteristics $S(j\omega) = N_s(\omega) \exp(j\theta_s(\omega))$, and $N_s(\omega) > 0$. According to reference [11] and Euler's formula, substituting the frequency-domain expressions of $P_0(z)$ and $S(z)$ into Equation (19) yields

$$|\theta_p(\omega) + \theta_s(\omega) + m\omega| < 90^\circ, \quad (20)$$

$$0 < k_{rc} < \frac{2(1+w) \min[\cos(\theta_p(\omega) + \theta_s(\omega) + m\omega)]}{(1+2w) \max[N_p(\omega) + N_s(\omega)]}. \quad (21)$$

When w , m , and k_{rc} satisfy Equations (14), (20) and (21), the SRC-ESO is stable.

3.4. Performance Analysis of SRC-ESO

When $f_s = 10$ kHz and $\omega = 0, \pm 0.4$, and ± 0.8 , respectively, the magnitude-frequency response curves of G_{ISRC} are shown in Figure 4.

From Equation (11), when $w = 0$, the SRC reduces to the CRC, and the SRC-ESO reduces to the RC-ESO. Therefore, SRC-ESO is compatible with RC-ESO. As shown in Figure 4a, when $0 < w < 1$, the SRC exhibits higher internal model gain and wider resonant bandwidth at integer multiples of the grid frequency compared with the CRC. Moreover, as the value of w increases, the higher the gain and the wider the resonant bandwidth of the SRC. Consequently, the observation error of the SRC-ESO becomes smaller, the frequency adaptability improves, and the disturbance rejection performance of SRC-ADRC becomes better. As shown in Figure 4b, when $-1 < w < 0$, the SRC exhibits lower internal model gain and narrower resonant bandwidth than the CRC. The smaller the value of w , the lower the gain and the narrower the resonant bandwidth of the SRC. A smaller w leads to weaker gain and reduced bandwidth, causing larger observation error and greater sensitivity to grid-frequency deviation. Under this condition, the disturbance-rejection ability of SRC-ADRC deteriorates.

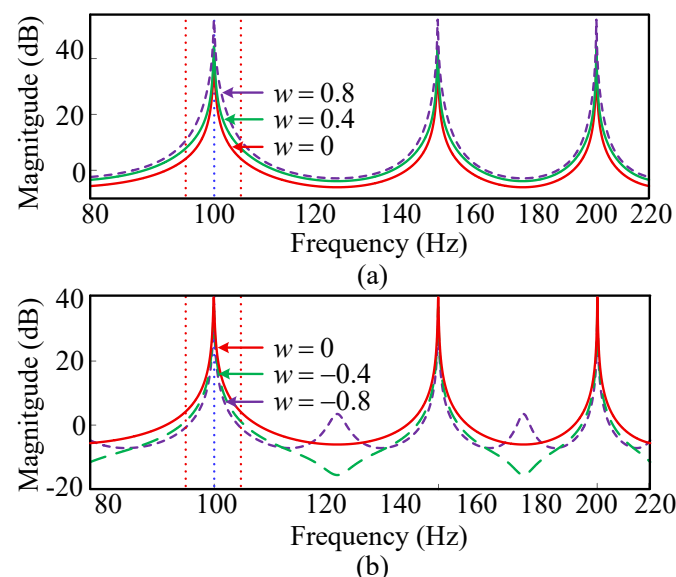


Figure 4. Magnitude-frequency response curves of G_{ISRC} . (a) $w = 0, 0.4, 0.8$; (b) $w = 0, -0.4, -0.8$.

4. Parameter Design

4.1. Control Law Parameter Design

The design of FPQ involves three key parameters, and the corresponding design guidelines are as follows.

- (1) Select an appropriate cutoff frequency ω_c to ensure sufficient bandwidth centered at the resonant frequency.
- (2) Choose a suitable proportional gain k_c to achieve the desired transient performance of the system.
- (3) Select an appropriate resonant gain k_r to effectively eliminate the steady-state error.

To facilitate the parameter design of the control law, let $k_r = k_{r1} \times b_0$ and $k_c = k_{c1} \times b_0$.

The pole maps when k_{c1} varies are shown in Figure 5. The stable range of k_{c1} is (0, 60). An appropriate adjustment of the proportional gain k_{c1} can improve the dynamic performance of the system, but an excessively large k_{c1} may lead to system instability. According to the theoretical poles in Figure 5 and the iterative experimental results, $k_{c1} = 20$ is selected in this paper.

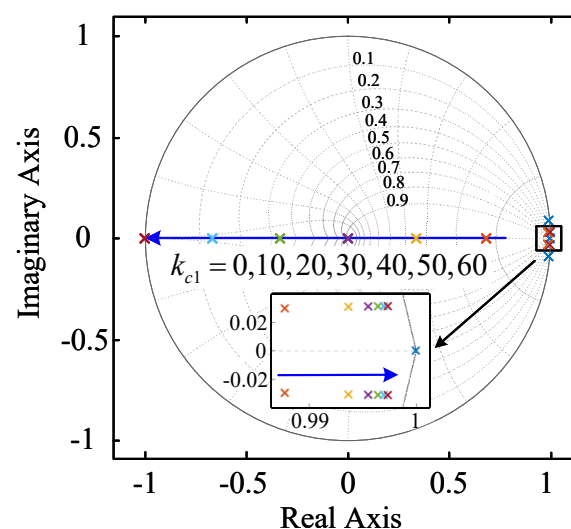


Figure 5. Pole map of the SRC-ADRC system when k_{c1} varies.

The pole maps and bode plots when k_{r1} varies are shown in Figures 6 and 7, respectively. It can be seen from Figure 7 that increasing k_{r1} reduces the phase lag of SRC-ADRC at the grid frequency, thereby improving the tracking performance of the reference current. However, as shown in Figure 6, as k_{r1} increases, the system stability gradually decreases. Considering the theoretical pole distributions in Figures 6 and 7 together with the iterative experimental evaluations, k_{r1} is selected as 350.

In the proposed SRC-ADRC scheme, the frequency-adaptive quasi-resonant control is employed in the control law, where the resonant frequency ω_r is continuously adjusted according to the grid frequency, thereby ensuring precise frequency synchronization with the grid. Because this frequency adaptation mechanism ensures that the resonant frequency of the controller remains consistent with the grid frequency, the requirement on the bandwidth ω_c of the resonant term is correspondingly reduced. On this basis, $\omega_c = 3.14$ rad/s is selected in this paper to ensure that the system still exhibits good control performance under frequency fluctuation conditions.

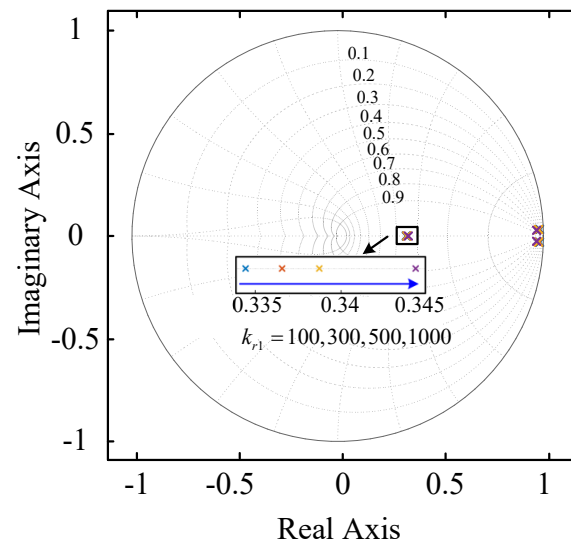


Figure 6. Pole map of the SRC-ADRC system when k_{r1} varies.

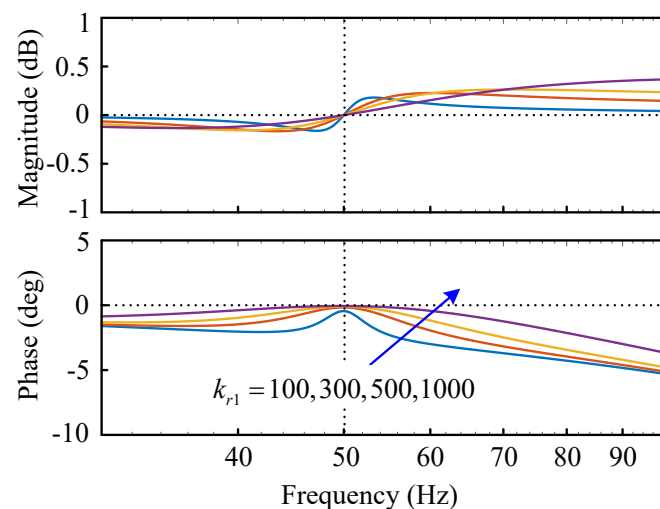


Figure 7. Bode plots of the SRC-ADRC system when k_{r1} varies.

4.2. Parameter Design of SRC-ESO

The design of SRC-ESO involves six key parameters, and the corresponding design guidelines are as follows.

- (1) Design k_p to satisfy the stability criterion.
- (2) Design an appropriate $Q(z)$ to ensure the observer bandwidth.
- (3) Select a suitable $S(z)$ to suppress high-frequency harmonics.
- (4) Determine z^m by taking into account the phase delay caused by $S(z)$ and $P_0(z)$.
- (5) Design an appropriate w to improve the frequency robustness of the SRC-ESO.
- (6) Select an appropriate k_{rc} according to w and z^m , so as to further improve the accuracy of disturbance estimation.

4.2.1. Design of k_p

According to the stability criterion, the stability of the SRC-ESO can be guaranteed only when the roots of the characteristic polynomial are located within the unit circle, namely, all poles of $P_0(z)$ should remain inside the unit circle. To verify this requirement, Figures 8 and 9 present the pole distribution and the Bode plots of $P_0(z)$ under different values of k_p , respectively. As illustrated in Figure 8, all poles are confined within the unit circle as k_p changes from 0 to 10000. Hence, the selected range of k_p satisfies the

stability condition given in (1). Further analysis of Figure 9 shows that as k_p gradually increases from 500 to 3500, the passband of $P_0(z)$ at the fundamental frequency gradually widens, while the attenuation of the magnitude response within the 1 kHz band gradually decreases. From the perspective of phase response, a larger k_p results in smaller phase lag within 1 kHz, which is more favorable for system phase compensation. Based on the above analysis, $k_p = 2500$ is selected in this paper.

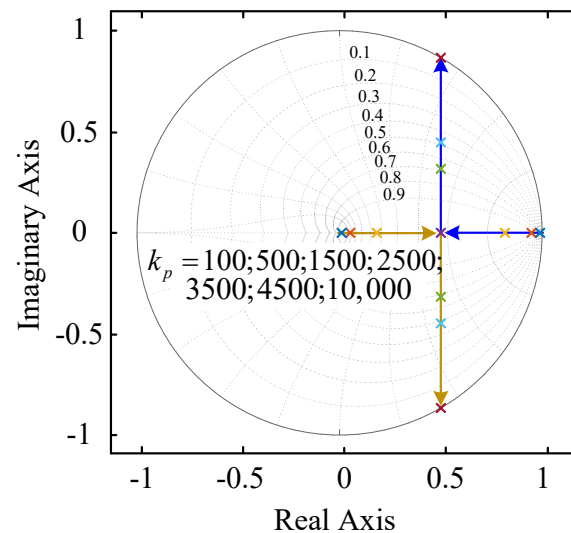


Figure 8. Pole map when k_p varies.

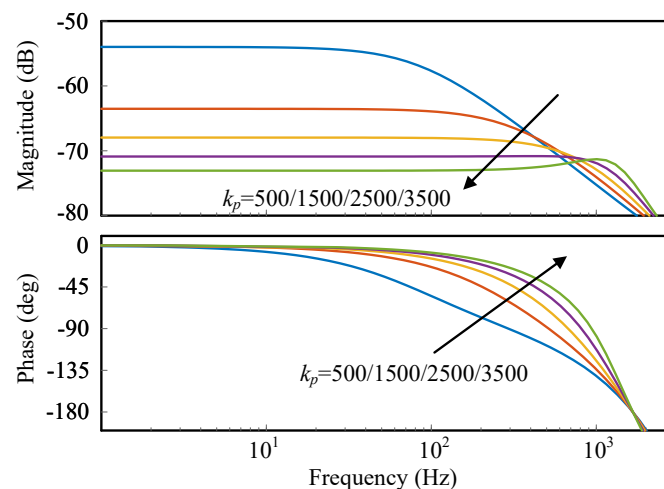


Figure 9. Bode plots when k_p varies.

4.2.2. Design of $Q(z)$

According to [20], this paper selects a zero-phase-shift low-pass filter for the inner membrane filter. The transfer function of $Q(z)$ is as follows:

$$Q(z) = \sum_{i=0}^M \alpha_i z^i + \sum_{i=1}^M \alpha_i z^{-i}, \quad (22)$$

$$\alpha_0 + 2 \sum_{i=1}^M \alpha_i = 1, \quad \alpha_i > 0,$$

where M is the order of the zero-phase-shift low-pass filter, α_0 and α_i are the filter coefficients.

The main harmonic components of the grid current are concentrated below 2 kHz; Therefore, $Q(z)$ needs sufficient bandwidth to track disturbances within this range. At

the same time, using a high-order zero-phase filter would increase computational burden. Consequently, a first-order filter is adopted. When $k = 1$ and α_0 takes the values 0.4, 0.6, and 0.8, the magnitude-frequency response curves of $Q(z)$ is shown in Figure 10. The bandwidths of $Q(z)$ are 1643 Hz, 2065 Hz, and 3261 Hz, respectively. Therefore, $k = 1$ and $\alpha_0 = 0.6$ are selected in this paper.

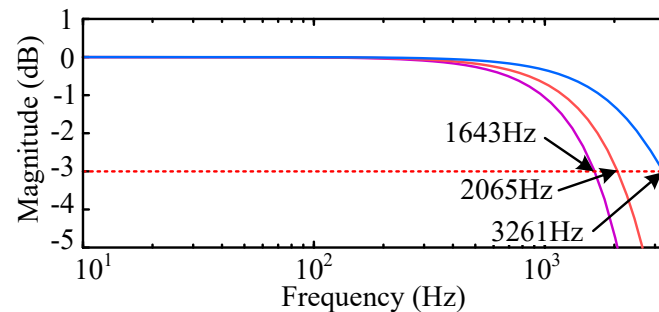


Figure 10. Magnitude-frequency response curves of $Q(z)$ when α_0 varies.

4.2.3. Design of $S(z)$

The compensator $S(z)$ is designed as a fourth-order Butterworth low-pass filter with a cutoff frequency of 1 kHz, as recommended in [20].

$$S(z) = \frac{0.004824z^4 + 0.0193z^3 + 0.02895z^2 + 0.0193z + 0.004824}{z^4 - 2.37z^3 + 2.314z^2 - 1.055z + 0.1874}. \quad (23)$$

4.2.4. Design of z^m

To compensate for the phase lag introduced by the low-pass filter $S(z)$ and $P_0(z)$, a phase compensation element z^m is designed such that $S(z)P_0(z)z^m$ exhibits zero gain and zero phase within 1 kHz as much as possible. The phase-frequency response curves of $S(z)P_0(z)z^m$ for different values of m are shown in Figure 11. When $m = 8$, the gain of $S(z)P_0(z)z^m$ is closest to 0 dB. Therefore, $m = 8$ is selected in this paper.

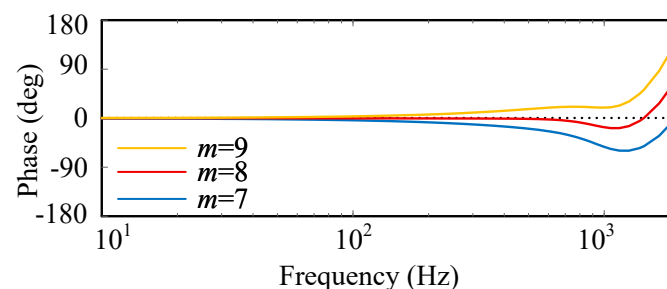


Figure 11. Phase-frequency response curves of $S(z)P_0(z)z^m$ when m varies.

4.2.5. Design of w and k_{rc}

According to Figure 11, when $m = 8$, the angle of $\cos(\theta_p(\omega) + \theta_s(\omega) + m\omega)$ varies from 0° to -16.5° within the 1 kHz frequency range. Since $S(z)$ is a low-pass filter, $\min[\cos(\theta_p(\omega) + \theta_s(\omega) + m\omega)] = 0.958$, and the value of $\max[N_p(\omega) + N_s(\omega)]$ is determined by the gain of $N_p(\omega)$. The phase-frequency response curve of $S(z)P_0(z)$ is shown in Figure 12. Within the 1 kHz frequency range, the gain of $S(z)P_0(z)$ varies from -67.9 dB (0.0004027) to -75.9 dB (0.00016037). According to Equation (21), the allowable range of k_{rc} is obtained:

$$0 < k_{rc} < \frac{2 \times 0.958(1 + w)}{0.0004027(1 + 2w)}. \quad (24)$$

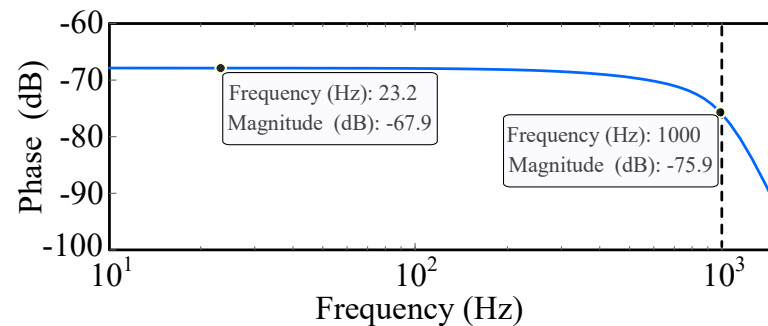


Figure 12. Phase-frequency response curve of $S(z)P_0(z)$.

Since the second-order weighting coefficient w affects the sensitivity of SRC-ESO to grid frequency, based on Figure 4, $w = 0.8$ is selected as the weighting factor in this paper. Therefore, the range of k_{rc} is obtained as $0 < k_{rc} < 3386$. The magnitude-frequency response curves of the compensated repetitive control plant when k_{rc} varies is shown in Figure 13. As k_{rc} increases, the gain of $k_{rc}S(z)P_0(z)z^m$ also increases. Since the compensated repetitive control plant is expected to exhibit unity-gain characteristics, $k_{rc} = 2500$ is selected.

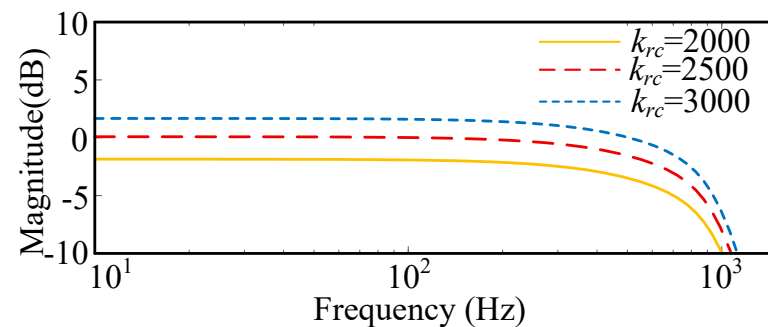


Figure 13. Magnitude-frequency response curves of $k_{rc}S(z)P_0(z)z^m$.

Although the SRC-ESO contains multiple design parameters, they are determined based on a fixed internal equivalent system and can be directly reused without further adjustment. Therefore, practical tuning is mainly limited to the resonant gain k_r and proportional gain k_c of the frequency-adaptive quasi-proportional resonant controller, resulting in a low tuning complexity.

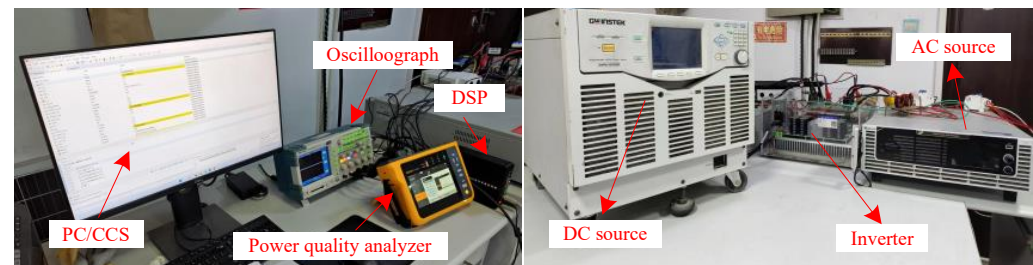
5. Experimental Verification

The effectiveness of the proposed SRC-ADRC strategy was further validated through experiments conducted on the LCL-type single-phase GCI platform depicted in Figure 14. In the experimental setup, the DC-link voltage was supplied by a programmable DC source, while the grid voltage u_g together with the background harmonic components was emulated using a programmable grid simulator. The THD of the grid-connected current was obtained by means of a power quality analyzer. The main experimental parameters adopted in the tests are listed in Table 1.

Three control strategies are compared: NRC-ADRC, TDF-ADRC, and SRC-ADRC. The experimental parameters for NRC-ADRC are: $k_c = 10,000$, $k_p = 2500$, and $b_0 = 333.3333$. For TDF-ADRC: $b_0 = 333.3333$, $k_c = 10,000$, $k_p = 10,000$, $\omega_0 = 1500$ rad/s, and $k_{rc} = 2500$. For SRC-ADRC: $b_0 = 333.3333$, $k_p = 2500$, $k_{rc} = 2500$, $k_c = 20 \times b_0$, $k_r = 350 \times b_0$, $\omega_c = 3.14$ rad/s, $m = 8$, and $w = 0.8$.

Table 1. Parameters of the experimental platform.

Parameter	Value
Inverter-side filter inductor L_1	2 mH
Grid-side filter inductor L_2	1 mH
Filter capacitor C	10 μ F
Damping resistor R	10 Ω
DC bus voltage E_d	400 V
Grid rated frequency f_g	50 Hz
Sampling frequency f_s	10 kHz

**Figure 14.** Experimental platform of the LCL-type single-phase GCI.

5.1. Steady-State Performance Verification

When the grid current is 10 A, the grid frequency f_g is 50 Hz, and the grid background harmonic distortion is 5.7%, the grid voltage, injected current, current error waveforms, and harmonic spectra of the injected current under the three control strategies (NRC-ADRC, TDF-ADRC, and SRC-ADRC) are compared in Figure 15.

The experimental results show that all three control strategies achieve stable grid-connected operation, but their current-tracking performances differ significantly. In terms of time-domain performance, TDF-ADRC exhibits a larger current error amplitude, with a steady-state error peak of approximately ± 3.25 A. In contrast, NRC-ADRC and SRC-ADRC significantly improve current-tracking performance, with error amplitudes of about ± 0.4 A. Regarding frequency-domain characteristics, the THD of the injected current under TDF-ADRC is 2.7%, while those under NRC-ADRC and SRC-ADRC are 1.0%. Both are markedly better than TDF-ADRC and meet the IEEE Std. 519-2022 requirements. A further comparison shows that SRC-ADRC achieves lower overall amplitudes at low-order harmonics and a more concentrated harmonic spectrum, demonstrating superior harmonic suppression performance and better injected current quality.

When the grid harmonic distortion is 0.2%, 5.7%, and 9.2%, the THD of the injected current under the three control strategies is listed in Table 2. As the grid voltage distortion increases, the THD of the injected current under all three strategies rises to some extent, but their suppression effects differ significantly. Under mild grid voltage distortion, TDF-ADRC achieves a grid current THD of 1.9%, indicating limited harmonic suppression capability. In contrast, both NRC-ADRC and SRC-ADRC exhibit good harmonic suppression performance, with SRC-ADRC achieving the lowest THD of only 0.4%. When the grid voltage distortion becomes more severe, the THD of TDF-ADRC increases notably, reaching 2.2% and 2.5%, respectively. Meanwhile, the THD values of NRC-ADRC and SRC-ADRC remain around 1.0%, showing little sensitivity to grid voltage distortion. A comprehensive comparison shows that SRC-ADRC maintains a low injected current THD under various grid voltage distortion conditions, demonstrating stronger harmonic suppression capability and robustness. This indicates that the proposed SRC-ADRC can effectively mitigate the impact of grid voltage harmonics on the injected current and maintains superior current quality control performance even under distorted grid voltage conditions.

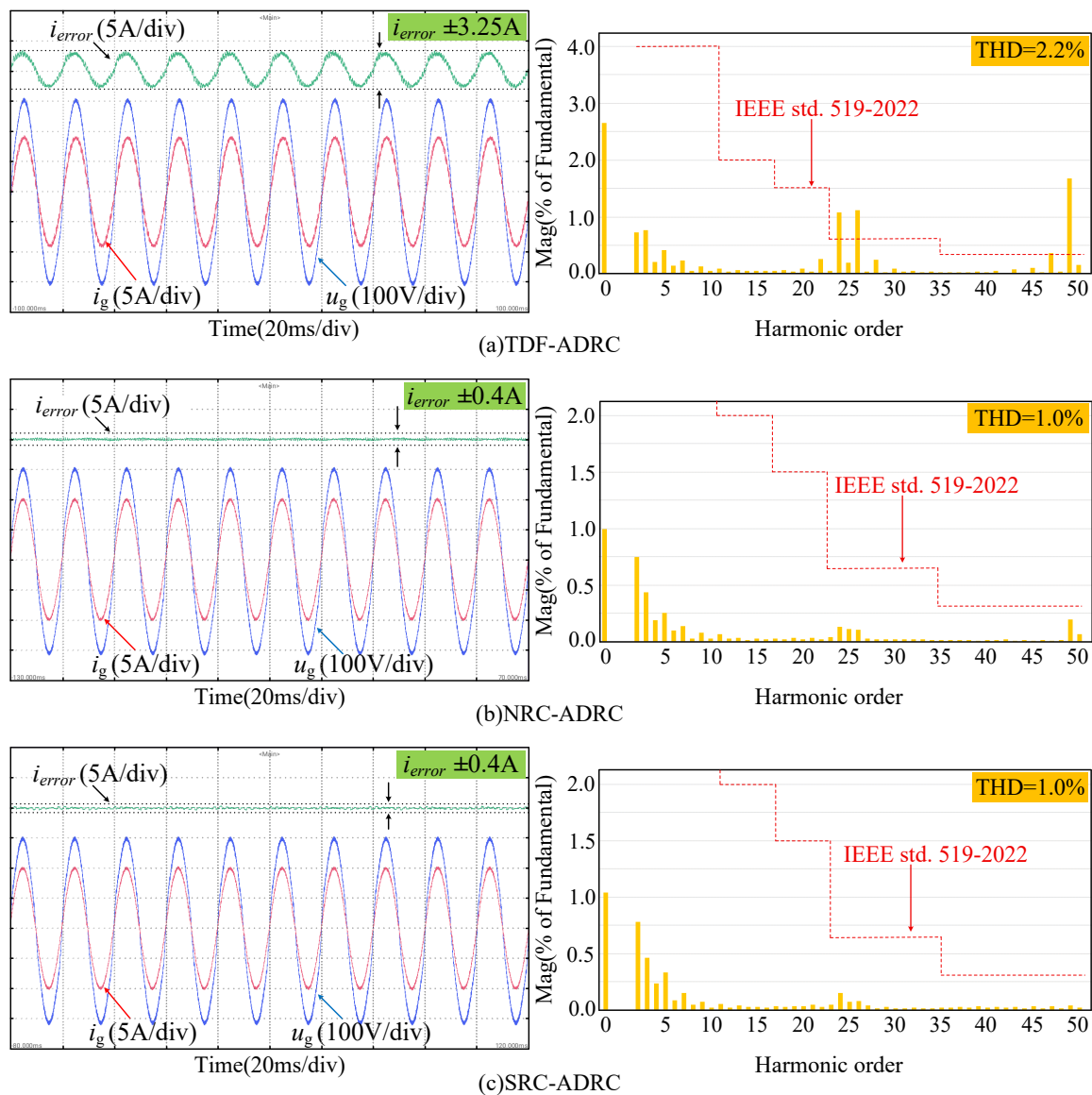


Figure 15. Waveforms and harmonic spectra of injected current under different control strategies.

Table 2. THD of injected current under different grid voltage distortions.

Control Schemes	THD of $u_g = 0.2\%$	THD of $u_g = 5.7\%$	THD of $u_g = 9.2\%$
TDF-ADRC	1.9%	2.2%	2.5%
NRC-ADRC	0.6%	1.0%	1.1%
SRC-ADRC	0.4%	1.0%	1.0%

5.2. Frequency Robustness Verification

When the grid current is 10 A, the grid background harmonic distortion is 5.7%, and the grid frequency f_g is 49.6 Hz, the grid voltage, injected current, current error waveforms, and THD of the injected current under the three control strategies (NRC-ADRC, TDF-ADRC, and SRC-ADRC) are shown in Figure 16.

Under the condition that the grid frequency is 49.6 Hz, the injected current waveform of TDF-ADRC exhibits significant distortion, with a current-tracking error peak of approximately ± 3.2 A, and the THD of the injected current reaches 4.9%. This indicates that the disturbance rejection capability of this control strategy for grid voltage harmonics is limited under grid frequency fluctuations. NRC-ADRC improves the current waveform quality to some extent, reducing the current error peak to about ± 3.0 A and the current

THD to 3.6%, but harmonic suppression is still affected by frequency variations. In contrast, SRC-ADRC maintains superior control performance under grid frequency fluctuations, with the current-tracking error peak further reduced to ± 0.9 A and the injected current THD lowered to 1.6%, significantly outperforming both TDF-ADRC and NRC-ADRC. The above results demonstrate that SRC-ADRC exhibits good harmonic suppression and reference tracking performance even under grid frequency fluctuation conditions.

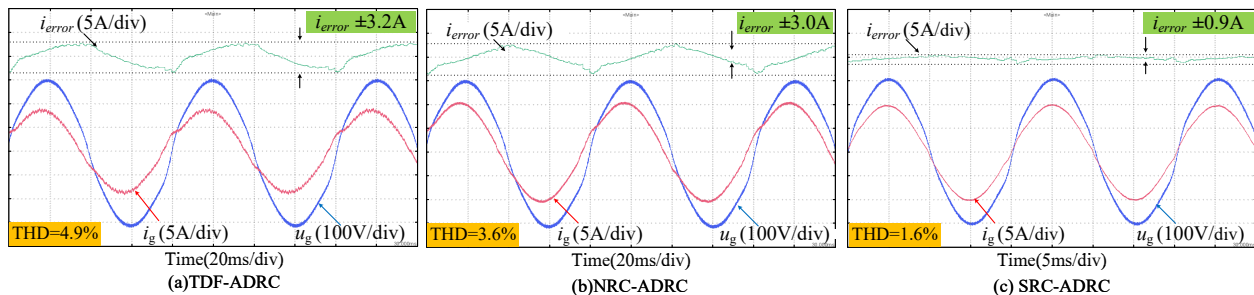


Figure 16. Waveforms of injected current under different control strategies when the grid frequency $f_g = 49.6$ Hz.

To better visualize the harmonic suppression performance of SRC-ADRC under grid frequency fluctuations, Figure 17 shows the line charts of the injected current THD for the three control strategies when the grid frequency within the range of 50 ± 0.4 Hz.

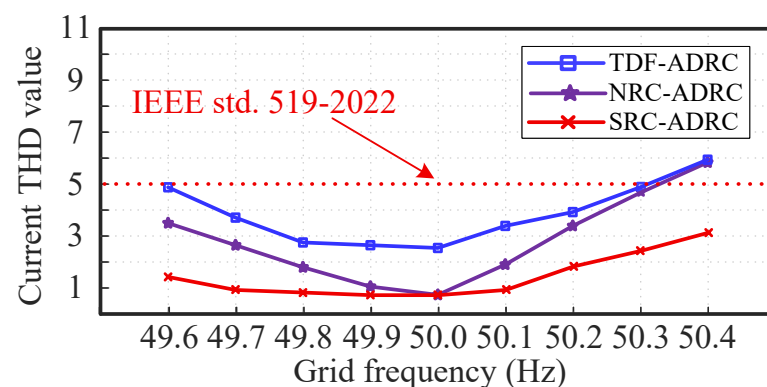


Figure 17. Injected current THD under different control strategies during grid frequency fluctuations.

When the grid frequency fluctuates within 50 ± 0.4 Hz, TDF-ADRC consistently exhibits the highest injected current THD, while NRC-ADRC shows a slightly lower THD than TDF-ADRC. When the grid frequency deviates to 50.4 Hz, both TDF-ADRC and NRC-ADRC exceed the limit specified by IEEE Std. 519-2022. In contrast, when the grid frequency varies within ± 0.4 Hz of the fundamental frequency, SRC-ADRC maintains the lowest injected current THD, which always complies with the IEEE Std. 519-2022 requirement (below 5%), demonstrating better frequency robustness.

5.3. Comparative Analysis with Contrast Methods

To provide a clear evaluation of the proposed SRC-ADRC scheme, Table 3 presents a comprehensive comparison between SRC-ADRC and ADRC, TDF-ADRC [19], and NRC-ADRC [15]. The comparison is conducted in terms of harmonic suppression capability, frequency adaptability, steady-state error, parameter tuning complexity, average computation time, and memory consumption. The first four performance metrics are assessed experimentally. The average computation time is evaluated based on the number of executions of each control algorithm within an identical observation interval, whereas the memory consumption is determined by the total memory occupied by all associated variables.

As summarized in Table 3, the proposed SRC-ADRC scheme achieves superior harmonic suppression capability, enhanced frequency adaptability, and reduced steady-state error while maintaining a relatively low parameter tuning complexity, thereby outperforming the existing approaches. Although SRC-ADRC requires additional computational and memory resources compared with the benchmark methods, the associated overhead is modest and remains well within the available resources of the adopted DSP platform, thereby ensuring practical real-time implementation.

Table 3. Performance comparison of different control schemes.

Method	Harmonic Suppression Capability	Frequency Adaptability	Steady-State Error	Parameter Tuning Complexity	Average Computation Time	Memory
ADRC	Weak	Weak	Large	High	2.17 μ s	0.05 KB
TDF-ADRC [19]	Moderate	Weak	Large	High	2.54 μ s	1.64 KB
NRC-ADRC [15]	Strong	Weak	Small	Low	2.73 μ s	1.65 KB
SRC-ADRC	Strong	Strong	Small	Low	3.96 μ s	2.63 KB

6. Conclusions

To address the performance degradation of injected current control under grid frequency fluctuations and grid voltage distortion, this paper proposes an SRC-ADRC control strategy. In this method, a new internal model is cascaded into the RC-ESO structure to construct a second-order repetitive extended state observer, thereby increasing the internal model gain of RC and widening the resonant bandwidth. Experimental results show that, compared with conventional control methods, the proposed SRC-ADRC achieves better overall performance in terms of frequency robustness and harmonic suppression capability. It should be noted that under larger grid frequency fluctuations, the total harmonic distortion (THD) of the injected current increases, which indicates a gradual degradation in the harmonic suppression performance of the proposed strategy. Therefore, further improvements are still required to maintain satisfactory harmonic mitigation under severe frequency variations. Future work will further investigate the control structure and parameter design methods to enhance its adaptability to a wider range of grid frequency fluctuations and maintain good harmonic suppression performance under large frequency variations.

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Abbreviations

The following abbreviations are used in this manuscript:

ADRC	Active Disturbance Rejection Control
SRC-ADRC	Second-Order Repetitive Active Disturbance Rejection Control

SRC-ESO	Second-Order Repetitive Extended State Observer
RC-ESO	Repetitive Control Extended State Observer
FPQ	Frequency-Adaptive Proportional-Quasi Resonant
GCI	Grid-Connected Inverter
ESO	Extended State Observer
SRC	Second-Order Repetitive Control
PLL	Phase-Locked Loop
THD	Total Harmonic Distortion
LCL	Inductor-Capacitor-Inductor
PCC	Point of Common Coupling
NRC-ADRC	Novel Repetitive Control Active Disturbance Rejection Control
TDF-ADRC	Two-Degree-of-Freedom Active Disturbance Rejection Control

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