

Article

Improved Control Strategy for Dual-PWM Converter Based on Equivalent Input Disturbance

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Abstract: Aiming at the problems of jittering waveforms and poor power quality caused by external disturbances during the operation of a dual-pulse-width-modulation (PWM) converter, an improved terminal sliding mode control and an improved active disturbance rejection control (ADRC) are investigated. The method is based on mathematical models of grid-side and machine-side converters to design the controllers separately, and the balance between the two sides is maintained by the capacitor voltage. An improved terminal fuzzy sliding mode control and equivalent input disturbance (EID)-error-estimation-based active disturbance rejection control are presented on the grid side to improve the voltage response rate, and an improved support vector modulation (SVM)-direct torque control (DTC)-ADRC method is developed on the motor side to improve the robustness against disturbances. Finally, theoretical simulation experiments are built in MATLAB R2023a/Simulink to verify the effectiveness and superiority of this method.

Keywords: dual-PWM converter; terminal fuzzy sliding mode control; active disturbance rejection control; equivalent input disturbance; SVM-DTC control



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1. Introduction

The high-level development of science and technology cannot neglect energy, which plays an important role in transportation [1], industrial production [2], electricity generation [3], and other fields. But with energy constraints, environmental pollution, and other problems, there is an urgent need for energy efficient, clean, and environmentally friendly green energy [4]. Electricity is an important part of energy. In order to improve the utilization of electric energy that is derived from a number of power electronic devices, these devices are being gradually miniaturized, intelligent, and integrated in current developments [5]. However, when a power electronic device is connected to the power grid, an insufficient control strategy will cause the power grid to produce a high degree of harmonics, resulting in the impeded operation of the power grid, and the utilization of electric energy will be reduced [6]. Therefore, control strategy optimization and stability improvements to power electronic devices are particularly important [7]. A converter, as a common power electronic device, has applications that widely cover many fields such as new energy generation [8,9], power transmission [10,11], and frequency conversion speed control systems [12,13]. At present, a common converter device has three parts: a grid measurement rectifier, a direct current (DC)-side capacitor, and an inverter, where the inverter circuit is usually connected to an alternating current (AC) motor as a nonlinear load. And permanent-magnet synchronous motors, which have significant advantages, such as a simple structure, small size, light weight, high rated power, high efficiency, etc., are widely used in AC frequency conversion speed control systems, and scholars at home and abroad have conducted in-depth research on them [14,15]. In recent years, PWM

controlled rectifiers have been used instead of diodes to form dual-PWM converters, which ensures the stability and controllability of the DC-side voltage, reduces current harmonics on the rectifier side, realizes highly efficient unit power factor operation, and supports the bidirectional flow of energy [16].

This study takes the dual-PWM converter as the research object. The dual-PWM converter topology is typically divided into three parts—a PWM rectifier, a DC-side capacitor, and a PWM inverter—and the balance between the rectifier and inverter is realized by ensuring the voltage stability of the capacitor [17]. On the one hand, it is important to ensure that the DC bus voltage has good resistance to sudden load variations and remains stable during parameter changes; on the other hand, the control strategy of the dual-PWM converter is improved, and algorithms are used to realize the unified control of the rectifier and inverter, which reduces the fluctuation of the DC-side voltage, ensures stable operation at high power levels, provides fast responses, reduces system faults, and optimizes its energy transfer efficiency [18]. To address these two problems, scholars have proposed the concepts of “DC-side voltage fluctuation” and “energy transfer efficiency” [19]. Aiming at these two problems, numerous scholars have conducted in-depth research and have proposed a variety of intelligent control methods. For example, Carvalho et al. proposed a technique involving asymmetric modulation of the pulse width for control in the current-fed dual-active bridge converters that are used in energy storage systems (ESSs), and an ESS was applied using high-capacity batteries operating at low voltage and used in low-power scenarios, such as data centers, residential photovoltaic systems, and uninterruptible power supplies. Based on the ESS application, this converter operates at a stable DC bus voltage, regardless of whether it is in the charging or discharging mode [20]. Guo et al. proposed a PWM strategy using carrier waveforms for the space vector pulse-width modulation problem of converters, taking a dual-output two-stage matrix converter as an example. They solved the complexity of the modulation strategy and enabled flexible adjustment of the output voltage ranges of both inverters [21]. Chen et al. presented a dual-DC-port DC–AC converter using virtual space vector pulse-width modulation for a hybrid photovoltaic–battery power train system [22]. Lin et al. proposed a method of combining Model predictive control (MPC) with nonlinear inductor (NI) that is effective at dealing with system parameter uptake and perturbation energy problems during power conversion in microgrids [23]. Hu et al. presented a novel asymmetrical pulse-width modulation control approach for improving the integral efficiencies of double-bridge resonant converters by applying it on the high-pressure sides of the converter’s bridges and combining it with phase-shifting controls [24]. For a dual-PWM converter, the keys are to enable it to operate with a high power factor and fast response and to ensure that the system is able to respond effectively to external perturbations and has high stability during parameter changes [25].

In order to effectively solve the perturbation immunity and power quality problems of dual-PWM converters, this paper proposes a new control strategy. First, for the grid-side system, the sliding mode control affects the system, and we propose a terminal fuzzy sliding mode control method based on EID error estimation for analyzing the effect on the current. In the voltage loop, we optimize the self-resilient control, which improves the response rate of the voltage and suppresses voltage jitter. Secondly, for the motor-side system, we improve the system’s disturbance immunity and power quality by introducing the EID error estimation method into the classical SVM-DTC control method. Finally, we build a model in Simulink to verify the validity and superiority of the proposed method based on the system response, jitter vibration, and power quality under different external conditions. The experimental results meet the expected goals.

The structural framework of the content of this paper is as follows:

In Section 2, we explain the modeling and perform a characterization of the model.

In Section 3, we propose a new control strategy for the fast-balancing control network of the dual-PWM converter.

In Section 4, we experimentally validate the dual-PWM converter for different operating scenarios. Finally, the work of the study paper is summarized.

2. Modeling and Analysis

In this section, the operating method of the system is first analyzed, and the mathematical model of the dual-PWM converter is established by using the state space equations; finally, the energy flow model of the dual-PWM converter is established under steady-state operation.

2.1. PWM Rectifier Mathematical Model

Define the switch function as

$$S_{K(K=a,b,c)} = \begin{cases} 0, & \text{Upper bridge arm conduction} \\ 1, & \text{Lower bridge arm conduction} \end{cases} \quad (1)$$

The PWM rectifier is modeled based on Kirchhoff's law of voltage and current.

$$\begin{cases} L \frac{di_a}{dt} = e_a - Ri_a - u_0 s_a - \frac{1}{3}(s_a + s_b + s_c)u_0 \\ L \frac{di_b}{dt} = e_b - Ri_b - u_0 s_b - \frac{1}{3}(s_a + s_b + s_c)u_0 \\ L \frac{di_c}{dt} = e_c - Ri_c - u_0 s_c - \frac{1}{3}(s_a + s_b + s_c)u_0 \\ C \frac{du_0}{dt} = i_a s_a + i_b s_b + i_c s_c - i_L \end{cases} \quad (2)$$

where $e_{a,b,c}$, $i_{a,b,c}$ are the AC voltage and current, respectively; L , R are the AC-side inductor and resistance, respectively; C is the DC-side capacitor; i_L is the load current; u_0 is the DC voltage. Equation (2) can be further simplified as:

$$\begin{cases} \frac{du_0}{dt} = \frac{1}{C} \left(\sum_{n=a,b,c} S_n i_n - i_L \right) \\ L \frac{di_s}{dt} = e_s - Ri_s - [u_0 (S_n - \frac{1}{3} \sum_{n=a,b,c} S_n)] \end{cases} \quad (3)$$

where $i_s = [i_a \ i_b \ i_c]^T$ is the three-phase current of the grid flowing into the VSR; $e_s = [e_a \ e_b \ e_c]^T$ is the three-phase grid voltage; $S_n = [S_a \ S_b \ S_c]^T$ is the switching function.

After a Clark coordinate transformation, $[e_d \ e_q] = C_{abc/dq}[e_a \ e_b \ e_c]$, $[i_d \ i_q] = C_{abc/dq}[i_a \ i_b \ i_c]$ and $[S_d \ S_q] = C_{abc/dq}[S_a \ S_b \ S_c]$. Here, $C_{abc/dq}$ is the matrix of the Clark coordinate transformation. Using Kirchhoff's current theorem, we can obtain $i_a + i_b + i_c = 0$. Simplifying the above obtains the final mathematical model of the PWM rectifier as:

$$\begin{cases} L \frac{di_d}{dt} = e_d - Ri_d + \omega Li_q - S_d u_0 \\ L \frac{di_q}{dt} = e_q - \omega Li_d - Ri_q - S_q u_0 \\ C \frac{du_0}{dt} = S_d i_d + S_q i_q - i_L \end{cases} \quad (4)$$

where e_d, e_q is the grid voltage dq-series component, i_d, i_q is the grid current dq-series component, and S_d, S_q is the switching function dq-series component.

2.2. PWM Inverter Mathematical Model

A mathematical model based on the design of the PWM inverter can provide an analysis of the expected switching signals. We take a permanent-magnet synchronous motor (PMSM) as the main body of the PWM inverter in this paper. In order to facilitate

the modeling analysis, it is assumed that the PMSM in the PWM inverter structure satisfies the following:

Assumption 1. Core saturation is neglected;

Assumption 2. Eddy currents and hysteresis loss are neglected;

Assumption 3. The rotor has no damping effect;

Assumption 4. The three-phase stator windings are perfectly symmetrical, and the axes of the phases are 120° apart;

Assumption 5. The rotor's permanent magnetic field is sinusoidally distributed in the air-gap space.

Considering a table column PMSM and assuming that the change in rotational speed is zero for a short period of time, the mathematical model of the PMSM in the stationary coordinate system is:

$$\begin{cases} \frac{di_\alpha}{dt} = -\frac{R_s}{L_s}i_\alpha + \frac{u_\alpha}{L_s} + \frac{\omega_e\psi_f}{L_s}\sin\theta \\ \frac{di_\beta}{dt} = -\frac{R_s}{L_s}i_\beta + \frac{u_\beta}{L_s} - \frac{\omega_e\psi_f}{L_s}\cos\theta \\ \frac{d\omega_e}{dt} = 0 \\ \frac{d\theta}{dt} = \omega_e \end{cases} \quad (5)$$

where u_α , u_β and i_α , i_β are the stator voltage and stator current $\alpha\beta$ -series components, respectively; R_s is the stator resistance; L_s is the stator inductance; ω_e is the rotor's electric angular speed; ψ_f is the rotor magnetic chain; θ is the rotor position angle.

3. Dual-PWM Converter Fast-Balancing Voltage Control

A new ADRC control strategy is proposed to improve the response rate of ADRC control by introducing the fast and accurate error estimation characteristic of EID control to the traditional ADRC control method while improving the system's disturbance resistance. The control strategy structure is shown in Figure 1.

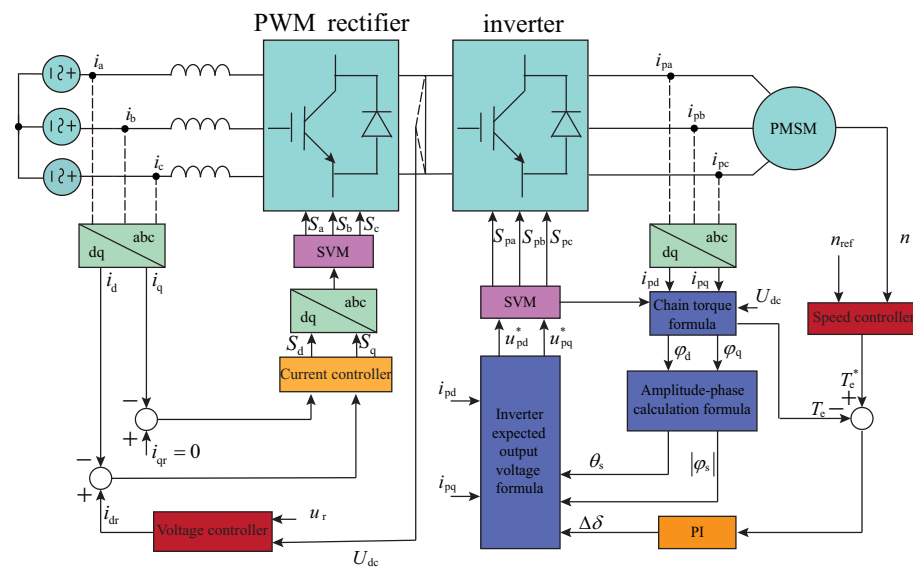


Figure 1. Control strategy structure.

3.1. PWM Rectifier Current Controller Design

The current loop control system is described as

$$\begin{cases} L \frac{di_d}{dt} = -Ri_d + \omega Li_q - s_d u_0 + e_d \\ L \frac{di_q}{dt} = -\omega Li_d - Ri_q - s_q u_0 + e_q \end{cases} \quad (6)$$

The design slip surfaces are:

$$\begin{cases} s_1 = e_1 + \beta_1 e_1^{p/q} \\ s_2 = e_2 + \beta_2 e_2^{p/q} \end{cases} \quad (7)$$

where e_1 and e_2 denote the d- and q-axis current error values, and β_1 and β_2 are positive odd integers. Because the current control system is a first-order system, there is no need to consider the singularity problem.

Derivation of Equation (7) yields:

$$\begin{cases} \dot{s}_1 = \frac{Ri_d - \omega Li_q + s_d u_0 - e_d}{L} + \beta_1 \frac{p}{q} e_1^{\frac{p}{q}-1} \dot{e}_1 \\ \dot{s}_2 = \frac{\omega Li_d + Ri_q + s_q u_0 - e_q}{L} + \beta_2 \frac{p}{q} e_2^{\frac{p}{q}-1} \dot{e}_2 \end{cases} \quad (8)$$

We define the Lyapunov function as

$$\begin{cases} V_1 = \frac{1}{2} s_1^2 \\ V_2 = \frac{1}{2} s_2^2 \end{cases} \quad (9)$$

The derivative of Equation (9) yields:

$$\begin{cases} \dot{V}_1 = s_1 \left(\frac{Ri_d - \omega Li_q + s_d u_0 - e_d}{L} + \beta_1 \frac{p}{q} e_1^{\frac{p}{q}-1} \dot{e}_1 \right) \\ \dot{V}_2 = s_2 \left(\frac{\omega Li_d + Ri_q + s_q u_0 - e_q}{L} + \beta_2 \frac{p}{q} e_2^{\frac{p}{q}-1} \dot{e}_2 \right) \end{cases} \quad (10)$$

According to the system stability principle, $\dot{V}_1 \leq 0$ and $\dot{V}_2 \leq 0$, i.e., the design control rate is

$$\begin{cases} s_d = \frac{-Ri_d + \omega Li_d + e_d - Lk_1(t) \text{sgn}(s_1) - L\beta_1 \frac{p}{q} e_1^{\frac{p}{q}-1} \dot{e}_1}{u_0} \\ s_q = \frac{-Ri_q - \omega Li_d + e_q - Lk_2(t) \text{sgn}(s_2) - L\beta_2 \frac{p}{q} e_2^{\frac{p}{q}-1} \dot{e}_2}{u_0} \end{cases} \quad (11)$$

where $k_1(t) = \max|d_1(t)| + \eta_1$, $k_2(t) = \max|d_2(t)| + \eta_2$, η_1 and η_2 are small normal numbers, and sgn is a symbolic function.

Substituting Equation (11) into Equation (10) for a current system with external disturbances yields

$$\begin{cases} \dot{V}_1 = s_1 \dot{s}_1 = -k_1(t)|s_1| - s_1 d_1(t) \leq -\eta_1 |s_1| \\ \dot{V}_2 = s_2 \dot{s}_2 = -k_2(t)|s_2| - s_2 d_2(t) \leq -\eta_2 |s_2| \end{cases} \quad (12)$$

When $\dot{V}_1 \equiv 0$ and $\dot{V}_2 \equiv 0$, $\dot{s} \equiv 0$: the system is asymptotically stable according to LaSalle's invariance principle [26].

In order to ensure that the system can be better stabilized under disturbances, the fuzzy control theory is used to optimize the problem that the upper bounds of disturbances $k_1(t)$ and $k_2(t)$ are set too large. In the terminal sliding mode control rate Equation (11), we make the disturbance upper bounds $k_1(t)$ and $k_2(t)$ vary with the disturbance $d_1(t)$ and $d_2(t)$ and satisfy the $K_{1,2}$ at every moment everywhere slightly larger than the disturbance.

In order to make $k_{1,2}(t)$ as close as possible to the real $d_{1,2}(t)$, we first set up an EID strategy to estimate the system error, thereby obtaining the current loop error observations $\hat{d}_1(t)$ and $\hat{d}_2(t)$. From the above analysis, it can be concluded that the switching gain $k_{1,2}(t)$ needs to be slightly larger than the error $d_{1,2}(t)$, so this paper adds a very small normal value of η to the above-mentioned switching gain design. But if the error is too large, the error measurement of the observer plus a very small η cannot guarantee that the switching gain $k_{1,2}(t)$ maintains good results. So this paper uses fuzzy control to design an adaptive η to solve the problem.

The expression of the switching gain is

$$\begin{cases} \tilde{k}_1(t) = \hat{d}_1(t) + H_1 \int_0^t \Delta\eta_1 dt \\ \tilde{k}_2(t) = \hat{d}_2(t) + H_2 \int_0^t \Delta\eta_2 dt \end{cases} \quad (13)$$

where H_1 and H_2 are empirical coefficients.

Firstly, EID error estimation is used to estimate the current system errors $d_1(t)$ and $d_2(t)$. The system control equation is

$$\begin{cases} \frac{di_d}{dt} = \frac{-Ri_d + \omega Li_q - s_d u_0 + e_d}{L} + v_1 \\ \frac{di_q}{dt} = \frac{-\omega Li_d - Ri_q - s_q u_0 + e_q}{L} + v_2 \end{cases} \quad (14)$$

To better describe the system, let $x = [x_1 \ x_2]^T = [i_d \ i_q]^T$, $u = [u_1 \ u_2]^T = [s_d \ s_q]^T$, define the external perturbation as $v = [v_1 \ v_2]^T$, and set the grid voltage to $e = [e_d \ e_q]^T$, with external disturbance v as the total disturbance $d = [d_1 \ d_2]^T$.

Equation (14) is expressed as

$$\begin{cases} \dot{x} = Ax + Bu + d \\ y = Cx \end{cases} \quad (15)$$

$$\text{where } A = \begin{bmatrix} -\frac{R}{L} & \omega \\ -\omega & -\frac{R}{L} \end{bmatrix}, B = \begin{bmatrix} -\frac{u_0}{L} & 0 \\ 0 & -\frac{u_0}{L} \end{bmatrix}, C = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}.$$

We design observers as

$$\dot{\hat{x}} = A\hat{x} + Bu_f + L[y - C\hat{x}] \quad (16)$$

where $L = [L_1 \ L_2]^T$ are the observer gains.

According to the principle of EID error estimation described in [27], the interference estimate is obtained as

$$\hat{d} = B^+ LC(x - \hat{x}) + u_f - u \quad (17)$$

where $B^+ = (B^T B)^{-1} B^T$.

Second, the following fuzzy system is designed using η , $s\dot{s}$ as system inputs and $\Delta\eta_1$, $\Delta\eta_2$ as system outputs. The fuzzy set is set as

$$s\dot{s} = \{NB \ NM \ NS \ ZO \ PS \ PM \ PB\} \quad (18)$$

$$\Delta\eta_{1,2} = \{NB \ NM \ NS \ ZO \ PS \ PM \ PB\} \quad (19)$$

When $s\dot{s}$ is PM and PB , to restore the system to a steady state quickly, the switching gain needs to be increased; that is, $\Delta\eta$ needs to be increased. Then the fuzzy value of $\Delta\eta$ is PB ; when $s\dot{s}$ is PS , the switching gain does not need to be too large. Thus, $\Delta\eta$ needs to be fine-tuned so that the fuzzy value of $\Delta\eta$ is PM . Similarly, when $s\dot{s}$ is NM and NB , to let the system quickly eliminate jitter, the switching gain needs to be quickly reduced. Then the fuzzy value of $\Delta\eta$ can be PB , and we can let the fuzzy value of $\Delta\eta$ be NB . When $s\dot{s}$ is NS , the switching gain at this time does not need to be too large; then we can let the fuzzy value of $\Delta\eta$ be NM . When $s\dot{s}$ is ZO , $\Delta\eta$ is ZO based on the above analysis, the fuzzy rule design table of this paper is shown in the Table 1.

Table 1. Fuzzy inference rule table.

$s\dot{s}$	NB	NM	NS	ZO	PS	PM	PB
$\Delta\eta_{1,2}$	NB	NB	NM	ZO	PM	PB	PB

To solve the problem that sign functions are not continuous at the zero point or not smooth at the transition point, we use a hyperbolic tangent function instead of a sign function. Figure 2 shows that at the inflection point, the sgn function has a sudden change, which will cause the system control force to not be smooth enough, resulting in chattering and current distortion. However, the $\tanh$ function is a smooth and continuous nonlinear function everywhere, which avoids the current buffeting problem.

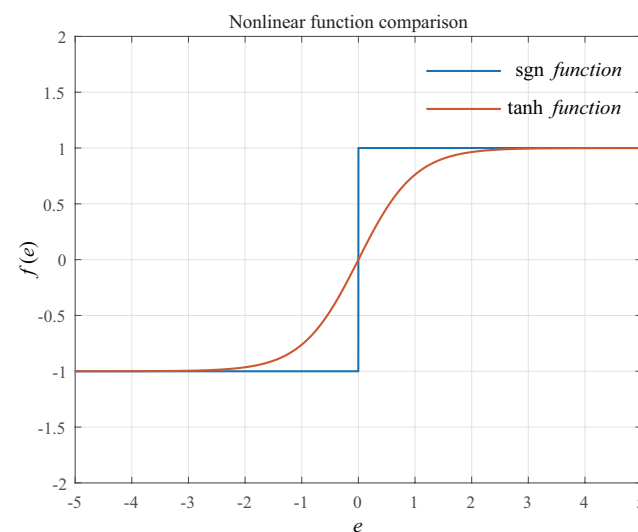


Figure 2. Nonlinear function comparison.

The mathematical expression is

$$\tanh(x) = \frac{e^{2ax} - 1}{e^{2ax} + 1} \quad (20)$$

where a denotes a positive real number.

The final controller function is

$$\begin{cases} s_d = \frac{-Ri_d + \omega Li_d - \tilde{L}k_1(t) \tanh(s_1) - L\beta_1 \frac{p}{q} e_1^{q-1} \dot{e}_1}{u_0} \\ s_q = \frac{-Ri_q - \omega Li_q - \tilde{L}k_2(t) \tanh(s_2) - L\beta_2 \frac{p}{q} e_2^{q-1} \dot{e}_2}{u_0} \end{cases} \quad (21)$$

3.2. PWM Rectifier Voltage Controller Design

When PWM rectifiers are used with conventional load power feedforward controllers, it is found that the response speed of the system voltage is slow and the anti-interference capability is weak. To solve these problems effectively, an EID/ADRC controller is designed instead. As shown in Figure 3, EID is used to improve the response speed of the system and to reduce the steady-state error, and ADRC is used to improve the interference immunity and robustness of the system.

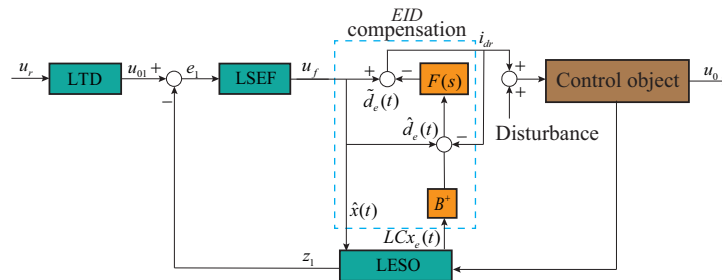


Figure 3. EID/ADRC controller.

We utilize the equivalent resistance R_L instead of the motor side to better describe the power expression on the PWM rectifier side. After conversion in equal power coordinates, the instantaneous power balance relation of the PWM rectifier can be expressed as

$$e_d i_d = L(i_d \frac{di_d}{dt} + i_q \frac{di_q}{dt}) + R(i_d^2 + i_q^2) + Cu_0 \frac{du_0}{dt} + \frac{u_0^2}{R_L} \quad (22)$$

Let $\psi = u_0^2$, $\psi^* = u_r^2$, $i_d = i_{dr}$, $i_q = 0$. Equation (22) can be expressed as

$$\frac{1}{2}C\dot{\psi} = -\frac{1}{R_L}\psi + e_d i_{dr} - R(i_{dr})^2 \quad (23)$$

We recharacterize Equation (23) and define $f(\psi, d_e(t))$ as the sum of the internal dynamics, unknown disturbances, and unmodeled dynamics of the system. These can be included in Equation (23) to obtain

$$\dot{\psi} = -\frac{2}{R_L C}\psi + \frac{2e_d}{C}i_{dr} + f(\psi, d_e(t)) \quad (24)$$

where the error $d_e(t)$ is unknown. Therefore, EID error compensation is utilized to estimate its magnitude.

Due to the functional relationship between u_0 and i_d , the state space expression for both can be written as

$$\begin{cases} \dot{x}(t) = A_1 x(t) + B_1 u_0(t) + d_e(t) \\ i_d(t) = x(t) \end{cases} \quad (25)$$

where $x = i_{dr} + \frac{R_L Cu_r}{Li_{dr}}u_0$, $A_1 = \frac{R_L(e_d - 2Ri_{dr})}{Li_{dr}}$, $B_1 = -\frac{R_L^2 Cu_r + 2u_r Li_{dr}}{L^2 i_{dr}^2}$.

The state observer is designed as follows.

$$\begin{cases} \hat{\dot{x}}(t) = A_1 \hat{x}(t) + B_1 u_f(t) + L_1 [i_d(t) - \hat{i}_d(t)] \\ \hat{i}_d(t) = \hat{x}(t) \end{cases} \quad (26)$$

where $\hat{x}(t)$ and $\hat{i}_d(t)$ are observer outputs, L_1 is observer gain, and u_f is the nonlinear error function output.

The final estimated value of $d_e(t)$ can be summed up from the above equation as

$$\hat{d}_e(t) = B_1^+ L_1 [x(t) - \hat{x}(t)] + u_f(t) - i_{dr}(t) \quad (27)$$

where $B_1^+ = (B_1^T B_1)^{-1} B_1^T$.

In order to decrease the jitter at certain amplitudes caused by the non-smooth characteristic of the traditional S-function, this paper adopts $\text{fatg}(e, \alpha) = |e|^\alpha \arctan e$ and $\text{sat}(e)$ as the S-functions. Therefore, the final model of the EID/ADRC controller can be obtained as

$$\begin{cases} \dot{u}_{01} = u_{02} \\ \dot{u}_{02} = -k_1 \text{sat}(u_{01} - r + \frac{u_{02}|u_{02}|}{2k_1}) \\ e_0 = z_1 - \psi \\ \dot{z}_1 = \hat{d}_e(t) - \beta_1 \text{fatg}(e_0, \alpha_1) + i_{dr} \\ e_1 = \psi^* - z_1 \\ u_f = k_2 \text{fatg}(e_1, \alpha'_1) \\ \dot{i}_{dr} = u_f - \hat{d}_e(t) \end{cases} \quad (28)$$

3.3. PWM Inverter Controller Design

The PWM inverter adopts a new SVM-DTC-ADRC control strategy to improve the waveforms of rotational speed, magnetic chain, and torque to enhance the control effect of the system by introducing the EID control method instead of the PI control method. This enables the PMSM to track the load torque and speed command values better while reducing the fluctuations in the stator magnetic chain. The desired component of the inverter AC output voltage is obtained.

$$\begin{cases} u_{p\alpha}^* = R_{ps} i_{p\alpha} + \frac{|\varphi_s^*| \cos(\theta_s + \Delta\delta) - |\varphi_s| \cos \theta_s}{T_s} \\ u_{p\beta}^* = R_{ps} i_{p\beta} + \frac{|\varphi_s^*| \sin(\theta_s + \Delta\delta) - |\varphi_s| \sin \theta_s}{T_s} \end{cases} \quad (29)$$

where θ_s is the stator chain; φ_s is the phase angle of the stator chain; φ_s^* is the desired value of the stator chain. To keep the stator chain amplitude constant, let $|\varphi_s^*| = |\varphi_s|$. In this case, θ_s and amplitude $|\varphi_s|$ are obtained as

$$\begin{cases} |\varphi_s| = \sqrt{(\varphi_\alpha)^2 + (\varphi_\beta)^2} \\ \theta_s = \arctan\left(\frac{\varphi_\beta}{\varphi_\alpha}\right) \end{cases} \quad (30)$$

where the stator chain components φ_α , φ_β are obtained from Equation (5), and the motor stator terminal voltage components $u_{p\alpha}$, $u_{p\beta}$ are obtained from the inverter switching states S_{pa} , S_{pb} , S_{pc} , and U_{dc} . The appropriate inverter input voltage vector is obtained according to Equation (30), and then, the stator magnetic chain rotation speed is controlled by the SVM technique to achieve motor speed control.

Meanwhile, we introduce the control method of EID error estimation in the SVM-DTC-ADRC control technique to effectively improve the control effect on the motor side. The EID error observer here is the same as in the previous section on observer design in voltage-loop control of PWM rectifiers.

The equation of motion of the PWM inverter can be expressed as

$$J \frac{d\omega_r}{dt} + B\omega_r = T_e - T_L \quad (31)$$

where T_L is the load torque; T_e is the electromagnetic torque of the motor; ω_r is the mechanical angular velocity; J is the rotational inertia of the motor; B is the friction coefficient.

Organizing the above equation gives:

$$\frac{d\omega_r}{dt} = \frac{T_e}{J} - \frac{T_L}{J} - \frac{B\omega_r}{J} \quad (32)$$

Based on the characteristics of ADRC, the load torque and friction coefficient in Equation (32) can be regarded as the disturbance $w(t)$. Taking the given torque signal as the output signal of the speed ring, Equation (32) can be rewritten as:

$$\frac{d\omega_r}{dt} = \frac{T_e^*}{J} + w(t) \quad (33)$$

Equation (33) has the same structure as Equation (23), so the EID/ADRC controller design for the rotational speed is consistent with the PWM rectifier voltage controller, and the final controller is

$$\begin{cases} e_0 = z_1 - n \\ \dot{z}_1 = \hat{d}_e(t) - \beta_1 fal(e_0, \alpha_1, \delta_0) + T_{e0}^* \\ e_1 = n_{ref} - z_1 \\ T_{e0}^* = k_2 fal(e_1, \alpha'_1, \delta_1) \\ T_e^* = T_{e0}^* - \hat{d}_e(t) \end{cases} \quad (34)$$

4. Computer Simulation

In this section, the control method proposed in this paper is simulated and verified in MATLAB R2023a. By comparing and analyzing the control effect of the system in different cases, the following simulation results are obtained.

4.1. Steady Experiment

We analyze the system under steady-state conditions with a given speed of 500 r/min and a given voltage of 200 V.

Figure 4 shows the steady-state voltage waveform, and the proposed method achieves voltage stabilization at 0.03 s. During the subsequent stabilization, there is no significant overshoot, and the voltage jitter is essentially zero. Meanwhile, the comparison method has obvious overshooting, and the stabilization of the voltage jitter is larger and reaches ± 0.15 V.

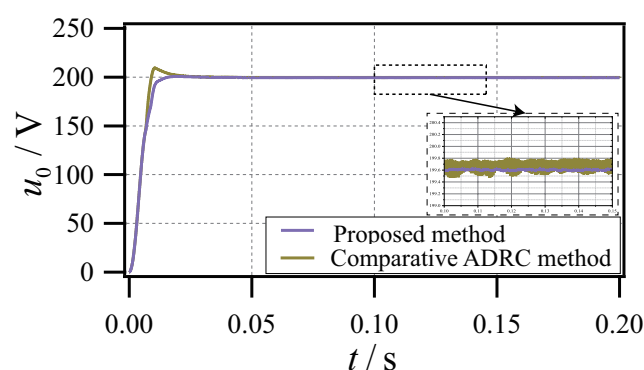


Figure 4. Voltage waveform.

Figure 5 represents the steady-state rotational speed waveform, and it can be seen that although both methods have no obvious overshooting, the proposed method has a shorter stabilization time and the rotational speed waveform has no obvious jitter, while the comparison method has a longer stabilization time and the rotational speed waveform jitter reaches ± 1 r/min.

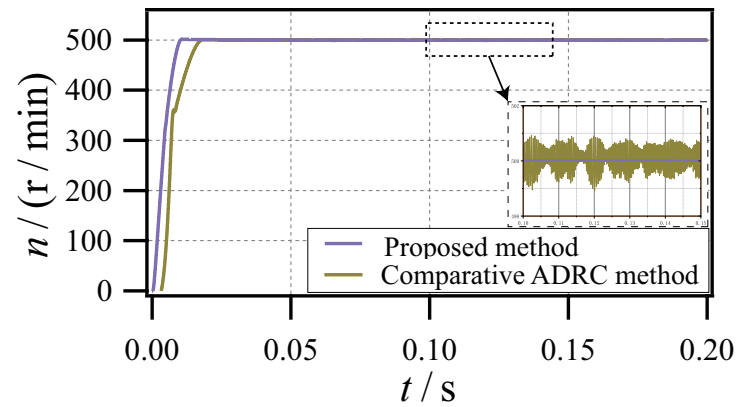


Figure 5. Speed waveform.

Figures 6 and 7 show the voltage and current waveforms of an a-phase grid, and it is thus clear that the current waveform jitter of the two methods is basically the same, but the proposed method achieves current sinusoidalization stabilization in half a voltage cycle, while the comparison method has to achieve the current sinusoidalization stabilization in two voltage cycles.

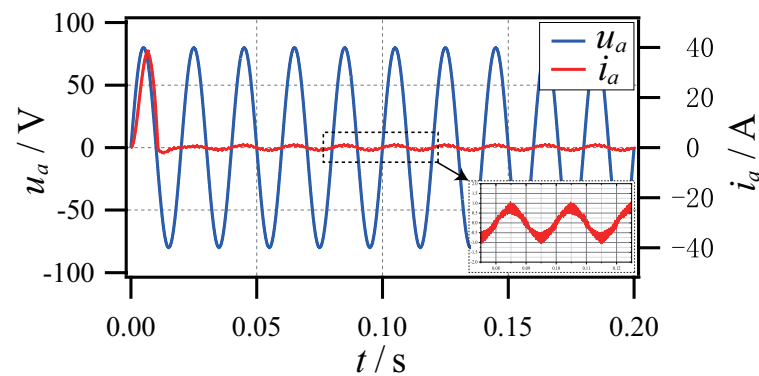


Figure 6. A-phase grid voltage and current waveforms in steady state (proposed method).

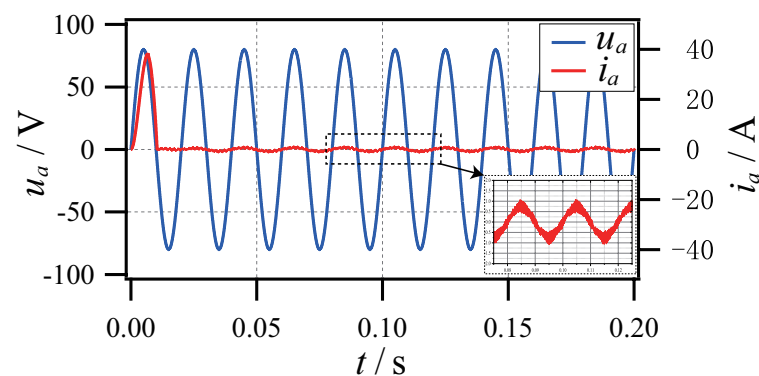


Figure 7. A-phase grid voltage and current waveforms in steady state (comparison method).

Figures 8 and 9 represent the instantaneous power waveforms of grids, and it is thus clear that the system stabilizes the reactive power above and below 0 after steady-state, basically realizing unit power operation.

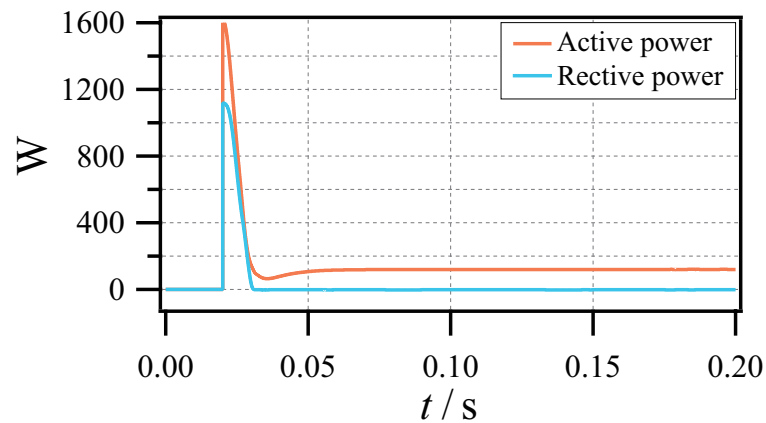


Figure 8. Grid-instantaneous power waveforms in steady state (proposed method).

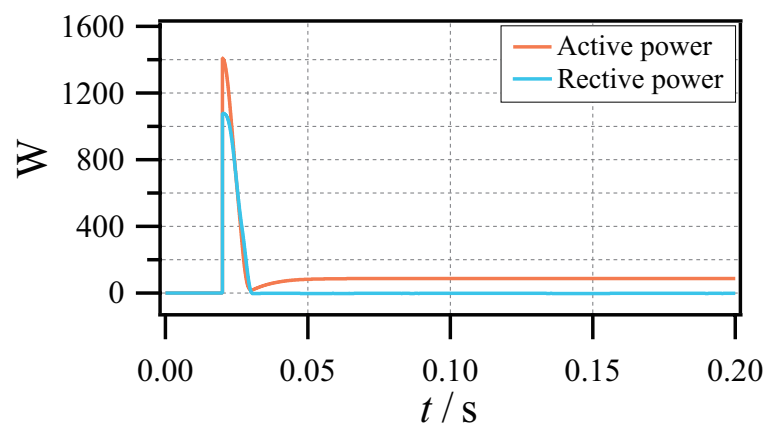


Figure 9. Grid-instantaneous power waveforms in steady state (comparison method).

Figures 10 and 11 represent the three-phase stator current waveforms and electromagnetic torque diagrams of the motor. We vary the torque from 0 to 2 N·m at 0.1 s to better show the stator current waveform graphs for both methods. As can be seen from Figure 10, the proposed method has a good effect on the control of the stator current and can achieve stability in a relatively fast time. After the torque changes, the stator current of the proposed method can achieve stability in one cycle. Figure 11 shows that the method proposed in this paper has a more prominent effect on the response rate for the torque.

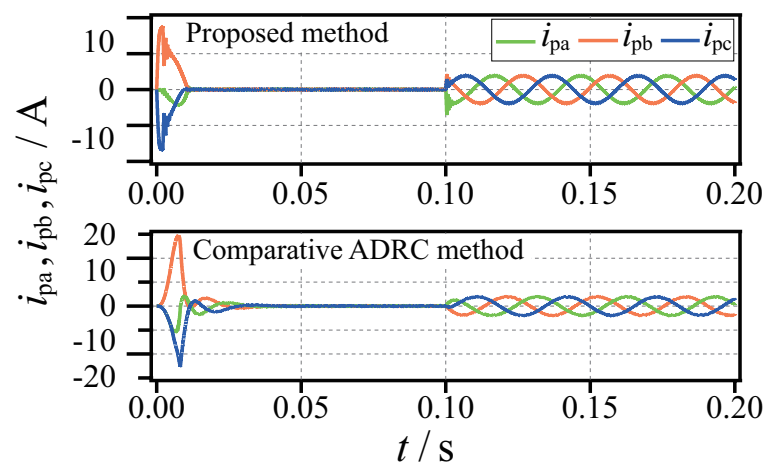


Figure 10. Motor three-phase stator current in steady state.

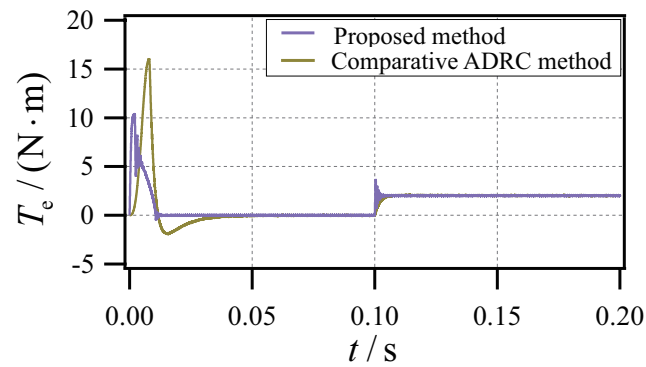


Figure 11. Electromagnetic torque waveform in steady state.

4.2. Rotation Speed Sudden Change Experiment

We design a speed change from 500 r/min to 1000 r/min at 0.1 s to simulate a sudden speed change in a real system.

Figure 12 represents the rotational speed waveform after the mutation. As is clear from the figure, the proposed method also achieves the new steady-state equilibrium quickly and with less jitter under the new steady-state target after the mutation, while the comparison method appears to have a sudden spike under the new parameter.

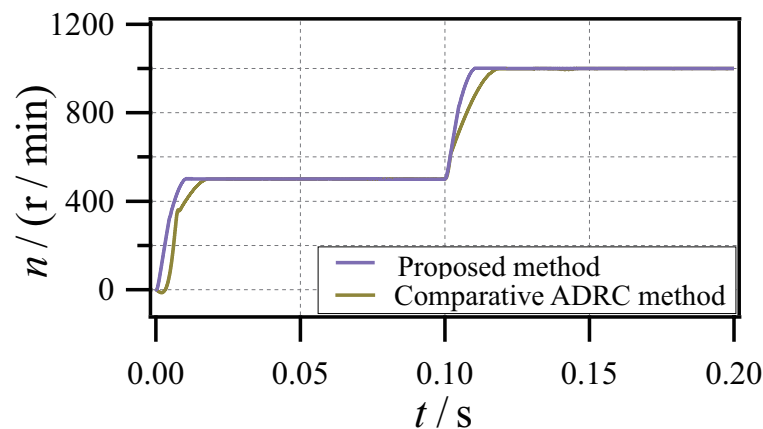


Figure 12. Rotation speed sudden change waveforms.

Figures 13 and 14 represent a-phase stator voltage–current waveforms after the mutation, and the proposed method achieves amplitude stabilization after one voltage cycle, while the comparison method requires two cycles.

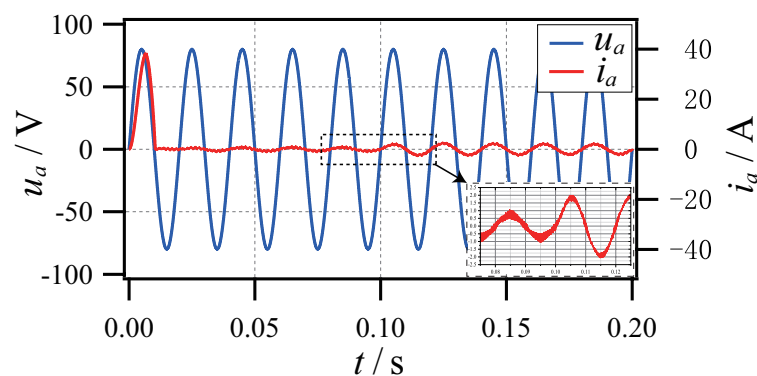


Figure 13. A-phase grid voltage and current waveforms in rotation speed sudden (proposed method).

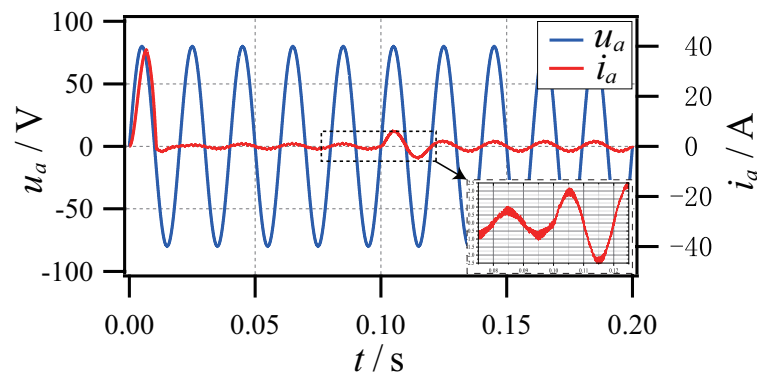


Figure 14. A-phase grid voltage and current waveforms in rotation speed sudden (comparison method).

Figures 15 and 16 represent the instantaneous power waveforms of the grid after the abrupt change. As can be seen from the figure, the method proposed in this paper can stabilize the power of the system in a short time after the system mutation, while the comparison method takes a longer time to achieve stability.

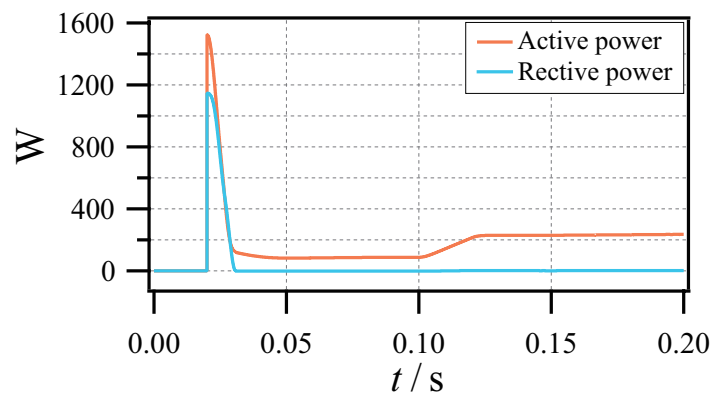


Figure 15. Grid-instantaneous power waveforms in rotation speed sudden (proposed method).

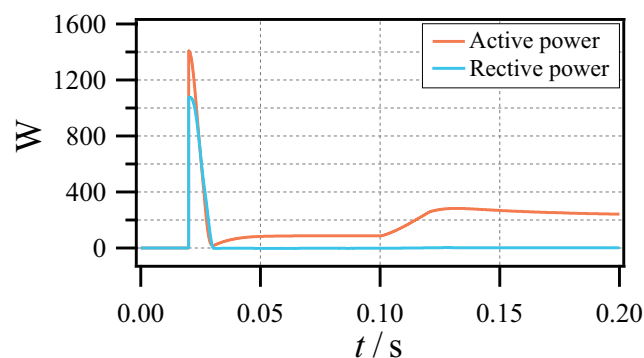


Figure 16. Grid-instantaneous power waveforms in rotation speed sudden (comparison method).

Figures 17 and 18 show the three-phase stator current waveform and electromagnetic torque of the motor. As can be seen from Figure 17, when the speed changes abruptly, the stator current change of the proposed method is more gentle than that of the comparison method, and the time to reach the new steady state is shorter, which has certain advantages. Figure 18 shows that under the condition of a sudden change in speed, the electromagnetic torque of the proposed method takes a short time to reach the new steady state.

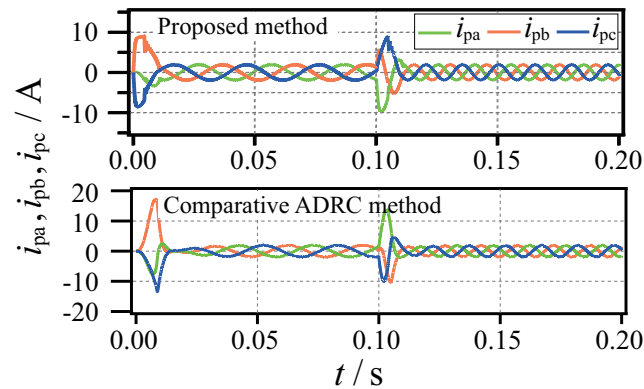


Figure 17. Motor three-phase stator current in rotation speed sudden.

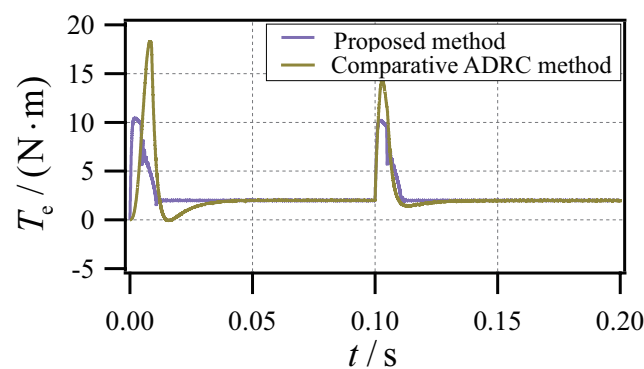


Figure 18. Electromagnetic torque waveform in rotation speed sudden.

4.3. DC Voltage Sudden Change Experiment

We design the DC voltage to change from 200 V to 250 V at 0.1 s to simulate a sudden voltage change in a real situation.

Figure 19 shows the voltage waveforms during sudden voltage changes. It is clear from the figure that the proposed method has a fast response time and little voltage jitter in the new steady state.

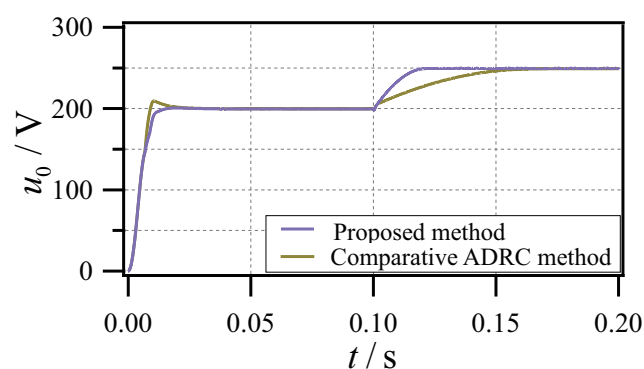


Figure 19. Voltage waveforms.

Figures 20 and 21 show the voltage–current waveforms of one phase of the grid during a sudden voltage change. It is clear from the figures that the proposed method achieves stabilization of the current amplitude after one voltage cycle, whereas the comparison method requires three voltage cycles to reach the new steady state.

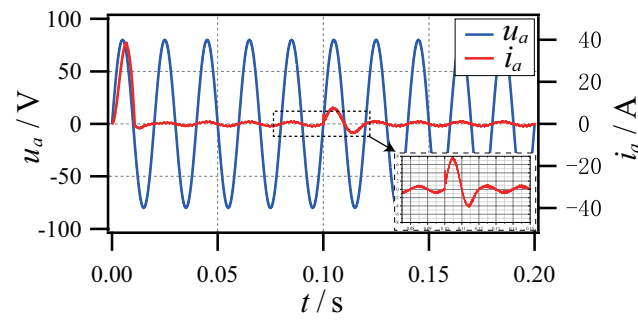


Figure 20. A-phase grid voltage and current waveforms in voltage sudden (proposed method).

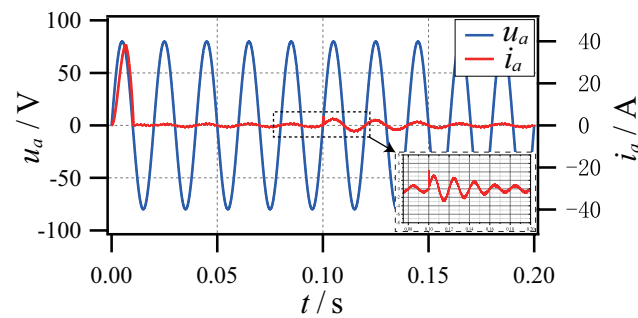


Figure 21. A-phase grid voltage and current waveforms in voltage sudden (comparison method).

Figures 22 and 23 show the instantaneous power waveforms of the grid, and it can be seen that the power fluctuates after a sudden change in voltage, but the proposed method can bring the power to a steady state in a short period of time.

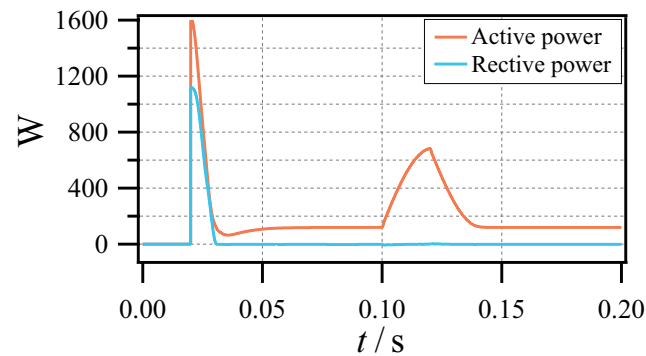


Figure 22. Grid-instantaneous power waveforms in voltage sudden (proposed method).

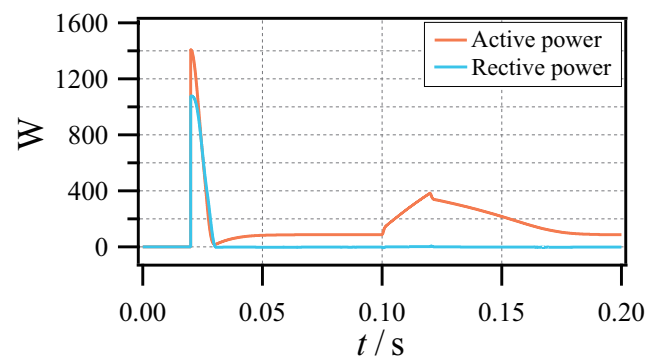


Figure 23. Grid-instantaneous power waveforms in voltage sudden (comparison method).

Figures 24 and 25 show that the sudden change in voltage has no significant effect on the stator current and electromagnetic torque on the motor side.

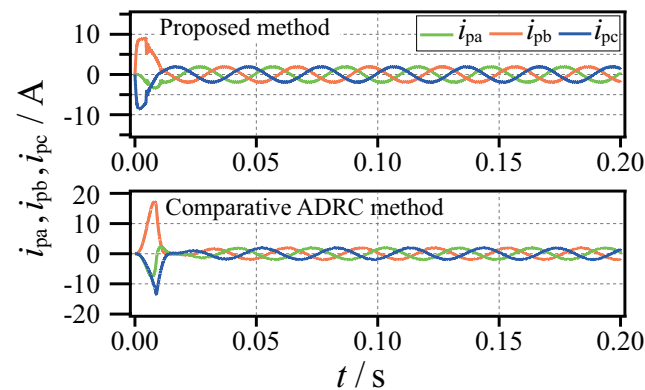


Figure 24. Motor three-phase stator current in voltage sudden.

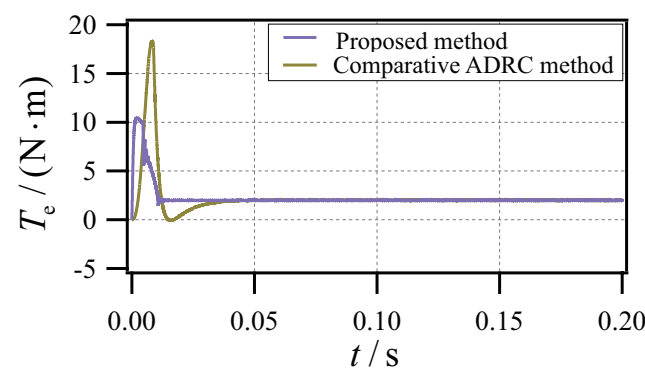


Figure 25. Electromagnetic torque waveform in voltage sudden.

5. Conclusions

In this paper, an improved control method is proposed to solve the disturbance immunity and improve the power quality of a dual-PWM converter system. For the grid side, a terminal fuzzy sliding mode control method based on EID error estimation is proposed to solve the current jitter problem caused by the traditional sliding mode, and EID error estimation is utilized to reduce the voltage response rate problem of ADRC control. For the motor side, we improve the disturbance capability and power quality of the system by introducing a self-resistant control with EID error estimation into the traditional SVM-DTC control method, and the final experiment verifies that the proposed method has good steady-state performance, and it also has better performance in the case of external disturbances. The control method proposed in this paper effectively improves the stability of the PWM converter when it is subjected to common external perturbations and increases the efficiency of energy flow and conversion. In the future, we will continue to study the immunity of PWM converter to disturbances and generalize these strategies to other systems.

In the future, we will study the control methods using PLL phase-locked loops in PWM rectifiers to adapt to more application scenarios. Meanwhile, this paper reduces the current jitter during the control of the current, which further improves the control performance of the system under the premise of ensuring the normal operation of the system.

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