

Article

Research on Fractional-Order Sliding Mode Control of Fractional-Order Permanent Magnet Direct-Drive Wind Power System

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Abstract

A large number of practical systems show pronounced fractional-order features. In comparison with integer-order calculus, fractional-order calculus has been demonstrated to possess enhanced precision in the description of the dynamic behavior of complex systems. The increase in control accuracy and flexibility results from this improvement. This study explores a direct-drive wind power generation system featuring permanent magnets, which incorporates fractional-order direct current bus (DC-bus) capacitor and fractional-order inductor–capacitor–inductor (FOLCL) grid-connected filter. For the machine-side rectifier, a fractional-order sliding mode (FOSM) speed outer-loop control and a fractional-order proportional–integral (FOPI) current inner-loop control were designed. A voltage outer-loop control using FOSM and a current inner-loop control using FOPI were developed for the grid-side inverter. Through simulation analyses under various wind speeds and grid fault conditions, it is demonstrated that compared to a control strategy using FOPI controllers in both inner and outer loops, the proposed control scheme which employs a FOSM outer-loop and reduces the overshoot of DC-bus voltage and grid-connected current by 21.51% and 32.49%, respectively, under sudden wind speed changes. Furthermore, during grid voltage sag faults, the maximum drop in DC-bus voltage and grid-connected active power are reduced by 65.38% and 33.38%, respectively. These results highlight the proposed method's superior dynamic and static performance, as well as enhanced resistance to disturbances.

Keywords: permanent magnet direct-drive wind power system; fractional-order calculus; fractional-order converter; fractional-order sliding mode



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1. Introduction

As the global economy develops rapidly, all countries have taken the development and utilization of clean energy as a key goal for sustainable development. As a type of environmentally friendly renewable energy, wind energy is playing an increasingly critical role in shaping the global energy structure [1]. Permanent magnet direct-drive wind power systems have been extensively utilized in the wind energy sector due to their straightforward design, elevated operational efficiency, and minimal maintenance requirements [2,3].

Conventional research on permanent magnet direct-drive wind power systems has predominantly relied on integer-order calculus for modeling and control. However, research has demonstrated that actual controlled plants and basic components, such as

capacitors and inductors, are inherently fractional-order. The application of fractional calculus has been demonstrated to enable more accurate modeling and improve control performance [4–7]. In recent years, scholars have conducted research on fractional-order modeling and fractional-order control for permanent magnet direct-drive wind power systems, yielding notable results. In permanent magnet direct-drive wind power systems, machine-side and grid-side converters, together with grid connection filters, are core components for energy conversion and grid connection. Their dynamic qualities directly affect the system's stability and efficiency. As reported in Reference [8], a fractional-order model was established for the single-phase pulse-width modulation (PWM) rectifier. The analysis based on the instantaneous reactive power theory and load variation tests was conducted to compare the integer-order and fractional-order models from two aspects. It was verified that the single-phase PWM rectifier fractional-order model is more in line with the real system. Reference [9] presents a fractional-order mathematical model of the three-phase grid-connected inverter, taking into account both high-frequency and low-frequency characteristics for photovoltaic grid-connected systems. The simulation results have shown that, in comparison to integer-order models, fractional-order photovoltaic three-phase inverters have better dynamic and static properties. For filters in grid-connected systems, Reference [10] integrated fractional-order inductor–capacitor–inductor (FOLCL) filters into the grid-side inverters of wind power systems and combined them with fractional-order proportional–integral (FOPI) control strategies to optimize the orders of fractional-order capacitors and inductors. This integration significantly suppressed the resonance of LCL filters. Furthermore, the study confirmed that fractional-order grid-connected inverter systems have greater advantages over integer-order systems in terms of grid-connected power factor and direct current bus (DC-bus) voltage stability. As demonstrated in the aforementioned literature, fractional-order models can more accurately describe the actual operational mechanisms of complex systems. It can be concluded that, for permanent magnet direct-drive wind power systems characterized by strong nonlinearity and multivariable coupling, the adoption of machine-side fractional-order converters, grid-side fractional-order converters, and fractional-order filters can more accurately reveal the systems' complex operating mechanisms.

It follows that, for permanent magnet direct-drive wind power systems with strong nonlinearity and multivariable coupling, the adoption of machine-side fractional-order converters, grid-side fractional-order converters, and fractional-order filters can more accurately characterize their complex operating mechanisms. Additionally, these systems are susceptible to external disturbances during operation, such as wind speed fluctuations and grid disturbances [11–13]. As a result, as the main control object of this system, the control of the machine-side and grid-side converters mainly focuses on the main part of this control. In recent years, the majority of studies have utilized conventional PI control [14], but its precision and disturbance rejection capabilities are still inadequate. Against this backdrop, the development of fractional calculus theory has provided new insights for the control of such complex systems. Common fractional-order control methods include fractional-order proportional–integral–derivative control (FOPID) [15–17], fractional-order sliding mode control (FOSMC) [18,19], fractional-order adaptive control [20], and fractional-order active disturbance rejection control (FOADRC) [21].

Reference [22] introduces a new FOPID controller to the control of permanent magnet direct-drive synchronous motors, which replaces the conventional PI regulator designed in accordance with traditional engineering principles, and thereby enhances the motor's speed regulation performance. Reference [23] proposes a combined method integrating an intelligent PI control algorithm and a sliding mode control algorithm. Although this method enhances the transient response and system robustness of permanent magnet

synchronous motors (PMSM), it fails to address the chattering problem inherent in sliding mode control.

To solve this issue, some scholars have incorporated fractional calculus operators into the method [24,25], confirming that FOSMC can reduce chattering to a certain extent and improve control performance. Reference [26] proposes integer-order PI controllers, fractional-order PI controllers, and fuzzy fractional-order controllers are designed to achieve Maximum Power Point Tracking (MPPT) control for permanent magnet synchronous wind power systems. The results show that whether under step wind speed or random wind speed, the fuzzy fractional-order controller exhibits more excellent MPPT performance and stronger robustness compared with the other two controllers. The above studies fully demonstrate that the adoption of fractional-order control can optimize the dynamic performance and anti-disturbance capability of systems under complex operating conditions.

As of right now, there are no published studies that examine the use of FOSMC and fractional calculus modeling in permanent magnet direct-drive wind power systems. Therefore, this study focuses on the fractional-order permanent magnet direct-drive wind power generation system that incorporates a DC-bus fractional-order capacitor and a FOLCL filter, and proceeds to design a corresponding FOSMC strategy. The subsequent chapters are arranged as follows: Section 1 introduces the fractional-order permanent magnet direct-drive wind power system model. Section 2 presents the design of the fractional-order control strategy. Section 3 conducts simulation analysis of the system. Section 4 draws conclusions.

2. Fractional-Order Permanent Magnet Direct-Drive Wind Power System Model

The main circuit of the fractional-order permanent magnet direct-drive wind power generation system investigated in this study is illustrated in Figure 1 [10]. The system is composed primarily of wind turbines, permanent magnet synchronous generators, fractional-order back-to-back converters (incorporating fractional-order DC-bus capacitors), FOLCL grid-connected filters, and other components. The key difference between fractional-order and traditional permanent magnet direct-drive wind power systems lies in the use of fractional-order elements in both the DC-bus capacitor and the inductors and capacitors of the LCL filter. As illustrated in Figure 1, v is the wind speed, i_{sa} , i_{sb} , and i_{sc} represent stator currents. i_{in} denotes the output current of the machine-side rectifier. i represents the input current of the grid-side inverter. e_{ga} , e_{gb} , and e_{gc} are defined as the voltages in a three-phase grid. The input currents of the FOLCL filter are denoted as i_{ga1} , i_{gb1} , and i_{gc1} . i_{ga2} , i_{gb2} , and i_{gc2} are represented as the output currents of the FOLCL filter. u_{dc} is represented as the DC-bus voltage. The fractional-order DC-bus capacitor and the capacitor in the FOLCL filter are represented by the symbols C_{dc} and C , respectively. α and β denote the orders of the fractional inductor and fractional capacitor, respectively, where $\alpha, \beta \in (0, 2)$. The filter inductors and resistors on the inverter-side and grid-side are labeled as L_1 , L_2 , R_1 , and R_2 , respectively. Wind turbines convert wind energy into mechanical energy, which is subsequently converted into electrical energy by permanent magnet direct-drive synchronous generators. Finally, the electrical energy is processed through fractional-order back-to-back converters and filters before being fed into the power grid.

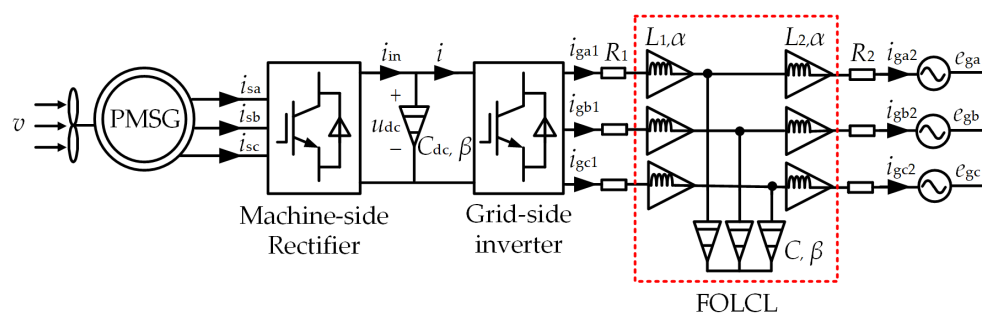


Figure 1. Primary circuit of a fractional-order permanent magnet direct-drive wind power generation system.

2.1. Wind Turbine Model

The mechanical power output (P_m) of the wind turbine is described by [27]:

$$P_m = \frac{1}{2} \rho \pi r^2 v^3 C_P(\lambda, \theta) \quad (1)$$

In this equation, ρ denotes the air density, r specifies the rotor radius, and C_P signifies the power coefficient, which is a function of the pitch angle θ and the tip-speed ratio λ .

In wind turbine control, when the wind speed exceeds the cut-in wind speed but below the rated wind speed, maximum power point tracking is achieved by setting the pitch angle $\theta = 0$ and managing the tip-speed ratio λ . For wind speeds that fall between the rated value and the cut-out threshold, regulation of the pitch angle θ ensures that the active power output remains stabilized at its rated level.

2.2. Permanent Magnet Direct-Drive Synchronous Generator Model

In the dq coordinate system, the stator voltage equation of a permanent magnet direct-drive synchronous generator is [28,29]:

$$\begin{cases} u_{sd} = R_s i_{sd} - \omega_e L_{sq} i_{sq} + L_{sd} D_t^1 i_{sd} \\ u_{sq} = R_s i_{sq} + \omega_e L_{sd} i_{sd} + \omega_e \psi_f + L_{sq} D_t^1 i_{sq} \end{cases} \quad (2)$$

In the equation, u_{sd} and u_{sq} represent the dq-axis components of the stator voltage, while i_{sd} and i_{sq} are the dq-axis components of the stator current. L_{sd} and L_{sq} represent the dq-axis components of the stator inductance, while R_s refers to the armature resistance of the generator. The electrical angular velocity of the generator is denoted as ω_e , while the mechanical angular velocity is represented by ω . These are related by the equation $\omega_e = p\omega = 2\pi p n$, where p is the number of pole pairs and n is the rotational speed. Additionally, ψ_f refers to the flux linkage of the permanent magnet, and D_t^1 represents the first-order differential operator from time 0 to time t .

The electromagnetic torque equation is expressed as follows:

$$T_e = \frac{3}{2} p [\psi_f i_{sq} + (L_{sd} - L_{sq}) i_{sd} i_{sq}] \quad (3)$$

For the generator, the following is the equation of motion:

$$T_e - T_L = J D_t^1 \omega + F \omega \quad (4)$$

In the equation, T_e and T_L represent the electromagnetic torque and load torque, respectively. F represents the friction coefficient, and J stands for the rotational inertia.

2.3. Fractional-Order Back-to-Back Converter Model

In the fractional-order permanent magnet direct-drive wind power system, the fractional-order back-to-back converter is a core component, as illustrated in Figure 2. The DC-bus capacitor exhibits fractional-order characteristics. S_1 – S_6 and S'_1 – S'_6 act as switching devices.

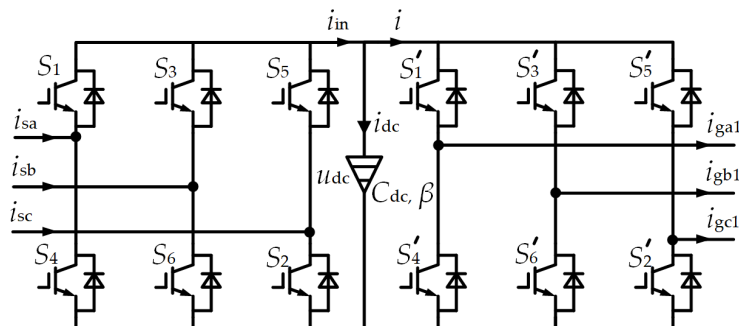


Figure 2. Converter topology.

The electrical equation of the fractional-order DC-bus capacitor is as follows:

$$i_{dc} = CD_t^\beta u_{dc} \quad (5)$$

In this equation, D_t^β denotes the β -order differential operator from time 0 to t .

2.4. Fractional-Order Grid-Connected Filter Model

In new energy grid-connected systems, the optimized design of filter topologies is crucial for suppressing resonance and improving power quality. Compared with conventional LCL filters, FOLCL filters exhibit greater flexibility in structural design and can effectively suppress resonance phenomena from the source [10]. The FOLCL filter topology is illustrated in Figure 3.

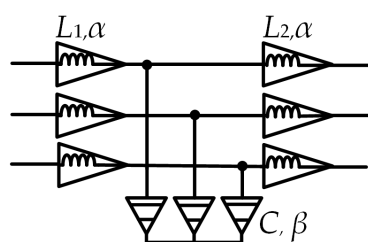


Figure 3. FOLCL topology diagram.

According to Kirchhoff's voltage and current laws, the state equations of the grid-side converter in the three-phase stationary coordinate system can be derived as follows:

$$\begin{cases} L_1 D_t^\alpha i_{gk1} = -R_1 i_{gk1} - u_{ck} + u_{gk} \\ L_2 D_t^\alpha i_{gk2} = -R_2 i_{gk2} + u_{ck} - e_{gk} \\ CD_t^\beta u_{ck} = i_{gk1} - i_{gk2} \end{cases} \quad (6)$$

In the aforementioned equation, k is equivalent to a, b, c , and u_{ck} represent the capacitor voltages of the FOLCL filter, while u_{gk} denotes the inverter output voltage.

To make control design and analysis easier, Equation (6) is transformed from the three-phase stationary coordinate system to the two-phase rotating coordinate system [30].

Subsequently, the dynamic equation of inductor L_1 in the dq coordinate system is deduced as follows:

$$\begin{cases} L_1 D_t^\alpha i_{gd1} = (L_1 A - R_1) i_{gd1} + L_1 B i_{gq1} + u_{gd} - u_{cd} \\ L_1 D_t^\alpha i_{gq1} = -L_1 B i_{gd1} + (L_1 A - R_1) i_{gq1} + u_{gq} - u_{cq} \end{cases} \quad (7)$$

Here, $A = \omega^\alpha \cos \frac{\alpha\pi}{2}$, $B = \omega^\alpha \sin \frac{\alpha\pi}{2}$.

The following is the dynamic equation of inductor L_2 on the dq-axis.

$$\begin{cases} L_2 D_t^\alpha i_{gd2} = (L_2 A - R_2) i_{gd2} + L_2 B i_{gq2} - e_{gd} + u_{cd} \\ L_2 D_t^\alpha i_{gq2} = -L_2 B i_{gd2} + (L_2 A - R_2) i_{gq2} - e_{gq} + u_{cq} \end{cases} \quad (8)$$

The dynamic equation of filter capacitor C on the dq-axis is as follows:

$$\begin{cases} C D_t^\beta u_{cd} = C A u_{cd} + C B u_{cq} + i_{gd1} - i_{gd2} \\ C D_t^\beta u_{cq} = -C B u_{cd} + C A u_{cq} + i_{gq1} - i_{gq2} \end{cases} \quad (9)$$

As can be seen from Equations (7) through (9), there is a cross-coupling effect between the d and q electrical quantities of the FOLCL filter.

3. Fractional-Order Control Design

3.1. Machine-Side Rectifier Fractional-Order Control

3.1.1. FOPI Current Inner-Loop Control with dq-Axis Decoupling

The goal of the machine-side converter's control is to manage the generator's electromagnetic torque to control its speed, allowing the wind turbine to function close to the optimal tip speed ratio. To improve the generator's efficiency and performance, the most commonly used control strategy is rotor field-oriented vector control with $i_{sd} = 0$, which simplifies Equation (3) to:

$$T_e = \frac{3}{2} p \psi_f i_{sq} \quad (10)$$

It can be demonstrated from the equation that the control of electromagnetic torque T_e can be achieved by regulating the current component i_{sq} . Since the stator current components i_{sd} and i_{sq} in Equation (2) are mutually coupled via mutual inductance, feed-forward control is introduced to achieve decoupling. To improve control performance, a current inner-loop FOPI controller is designed in this paper, with the corresponding control equation provided as follows:

$$\begin{cases} u_{sd} = R_s i_{sd} - \omega_e L_{sq} i_{sq} + k_p (i_{sd}^* - i_{sd}) + k_i s^{-\eta} (i_{sd}^* - i_{sd}) \\ u_{sq} = R_s i_{sq} + \omega_e L_{sd} i_{sd} + \omega_e \psi_f + k_p (i_{sq}^* - i_{sq}) + k_i s^{-\eta} (i_{sq}^* - i_{sq}) \end{cases} \quad (11)$$

In the formula, η denotes the fractional integration order, while k_p and k_i represent the proportional and integral coefficients, respectively.

3.1.2. FOSM Speed Outer-Loop Control

Reference [22] demonstrates that FOPI control in the speed outer-loop exhibits excellent speed regulation performance, with substantially improved dynamic performance and robustness compared to conventional PI control. However, in complex operating scenarios, the system requires enhanced anti-interference capability. To this end, this paper introduces FOSMC in the speed outer-loop design, by designing FOSM surfaces and FOSM reaching laws. This design optimizes the system's dynamic response and suppresses chattering by leveraging the fractional-order characteristics. By adopting rotor field-oriented vector control with $i_{sd} = 0$ and ignoring the damping coefficient, the mathematical model of the

permanent magnet direct-drive synchronous generator in the dq-axis coordinate system can be simplified to:

$$\begin{cases} L_{sq} D_t^1 i_{sq} = -R_s i_{sq} - p n \psi_f + u_{sq} \\ J D_t^1 n = \frac{3}{2} p \psi_f i_{sq} - T_L \end{cases} \quad (12)$$

Based on the mathematical model presented earlier in this paper, the following variables are chosen as the system state variables of the permanent magnet synchronous generator.

$$\begin{cases} x_1 = n^* - n \\ x_2 = D_t^1 x_1 = -D_t^1 n = -\frac{3}{2J} p \psi_f D_t^1 i_{sq} + \frac{1}{J} T_L \end{cases} \quad (13)$$

In the aforementioned equation, x_1 and x_2 represent the system state variables, while μ denotes the fractional derivative.

Subsequently, differentiating with respect to x_2 yields:

$$D_t^1 x_2 = -D_t^2 n = -\frac{3}{2J} p \psi_f D_t^2 i_{sq} \quad (14)$$

Let $H_1 = \frac{3p\psi_f}{2J}$. Then, the system's state-space equation can be expressed as:

$$D_t^1 \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} - D_t^2 \begin{bmatrix} 0 \\ H_1 \end{bmatrix} i_{sq} \quad (15)$$

The fractional-order sliding-mode surface is designed as follows:

$$f_1 = c_1 x_1 + D_t^\mu x_1 \quad (16)$$

In the formula, c_1 is a constant.

The power rate reaching law offers rapid convergence for systems distant from the sliding mode surface, while allowing parameter-based chattering reduction during approach. To further enhance these performance metrics, this paper develops a fractional-order fast power rate reaching law to improve both convergence speed and control performance, as shown in Equation (17):

$$D_t^\mu f_1 = -k_0 |f_1|^\sigma D_t^\mu \text{sgn}(f_1) - k_1 f_1 \quad (17)$$

In the equation, $\text{sgn}(f_1)$ denotes the sign function, while k_0 and k_1 represent constants greater than 0. These constants ensure that f_1 and $D_t^\mu f_1$ have opposite signs, thereby satisfying the stability conditions. This proves the stability of the system under fractional-order fast power-rate reaching law-based sliding mode control [31]. A fractional-order coefficient σ is introduced into the fractional-order power-rate model. The system's switching speed can be modulated by adjusting the coefficients σ , k_0 , and k_1 .

Upon differentiating Equation (16) and combining it with Equation (17), the FOSMC law for the speed outer-loop is obtained as follows:

$$i_{sq}^* = \frac{1}{H_1} D_t^{1-\mu} \left[c_1 x_1 + D_t^{-\mu} (k_0 |f_1|^\sigma D_t^\mu \text{sgn}(f_1) + k_1 f_1) \right] \quad (18)$$

The aforementioned design enables the establishment of a double closed-loop control structure for the machine-side rectifier, incorporating a "FOSM speed outer-loop + dq-axis decoupled FOPI current inner-loop," as depicted in Figure 4.

As illustrated in the figure, Y_1 and Y_2 represent the expressions of $|f_1|^\sigma$ and $\text{sgn}(f_1)$, respectively, in the complex frequency domain.

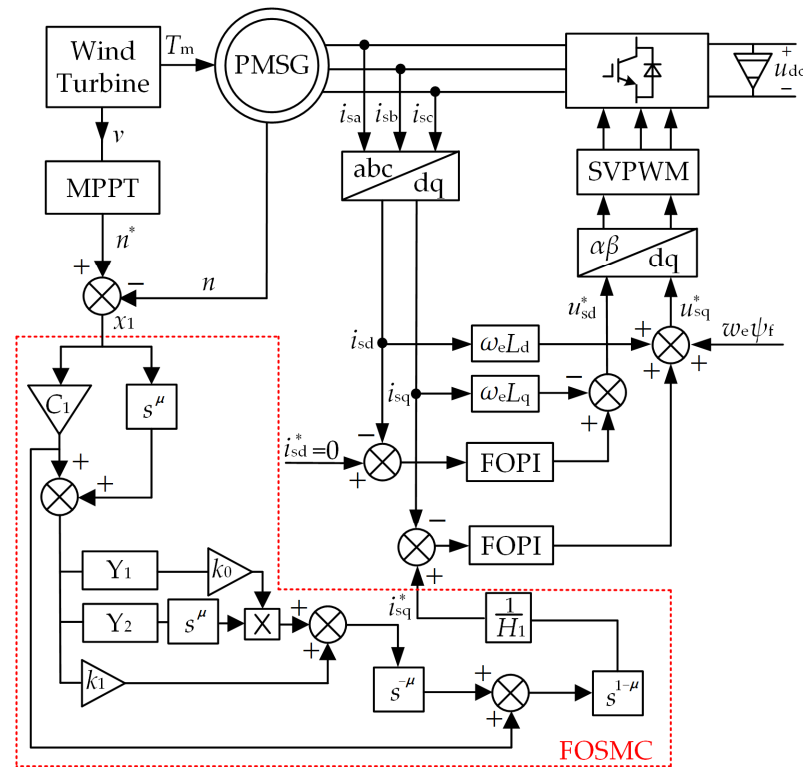


Figure 4. Control block diagram of the machine-side converter with FOSMC speed outer-loop and FOPI current inner-loop.

3.2. Grid-Side Inverter Fractional-Order Control

3.2.1. FOPI Current Inner-Loop Control with dq-Axis Decoupling

Given that grid-side converters generally involve the low-frequency dynamics of the power grid when controlling active and reactive power, and that capacitive reactance is significant at low frequencies, the capacitance C in the FOLCL filter can be neglected in control design. By equating the sum of L_1 and L_2 to a total fractional-order inductance L , the FOLCL filter can be simplified to a fractional-order inductor (FOL) filter. Equations (7) to (9) can thus be simplified as follows:

$$\begin{cases} LD_t^\alpha i_{gd} = (LA - R)i_{gd} + LBi_{gq} + u_{gd} - e_{gd} \\ LD_t^\alpha i_{gq} = -LBi_{gd} + (LA - R)i_{gq} + u_{gq} - e_{gq} \end{cases} \quad (19)$$

where $L = L_1 + L_2$ and $R = R_1 + R_2$.

The expression of Equation (19) in the complex frequency domain is as follows:

$$\begin{cases} (Ls^\alpha + R - LA)I_{gd}(s) = LBI_{gq}(s) + U_{gd}(s) - E_{gd}(s) \\ (Ls^\alpha + R - LA)I_{gq}(s) = -LBI_{gd}(s) + U_{gq}(s) - E_{gq}(s) \end{cases} \quad (20)$$

In the equation, I_{gd} , I_{gq} , U_{gd} , U_{gq} , E_{gd} , and E_{gq} are the expressions of i_{gd} , i_{gq} , u_{gd} , u_{gq} , e_{gd} , and e_{gq} , respectively, in the complex frequency domain. Given the cross-coupling between d-axis and q-axis currents in the rotating coordinate system, a FOPI current inner-loop control is proposed herein for dq-axis decoupling of the grid-side inverter. Based

on Equation (19), the control scheme is designed as Equation (21), and its control block diagram is presented in Figure 5.

$$\begin{cases} u_{gd}^* = e_{gd} - \omega^\alpha L \sin \frac{\alpha\pi}{2} i_{gq} + k_{p0} (i_{gd}^* - i_{gd}) \\ \quad + k_{i0} s^{-\eta_0} (i_{gd}^* - i_{gd}) \\ u_{gq}^* = e_{gd} + \omega^\alpha L \sin \frac{\alpha\pi}{2} i_{gd} + k_{p0} (i_{gq}^* - i_{gq}) \\ \quad + k_{i0} s^{-\eta_0} (i_{gq}^* - i_{gq}) \end{cases} \quad (21)$$

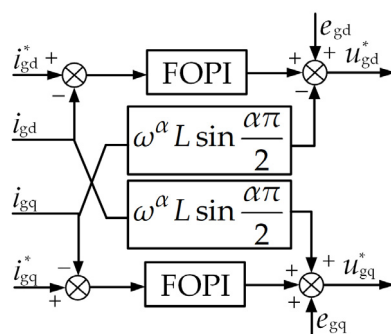


Figure 5. Fractional-order current inner-loop controller.

Through the above design, the decoupled d-axis current control diagram is shown in Figure 6.

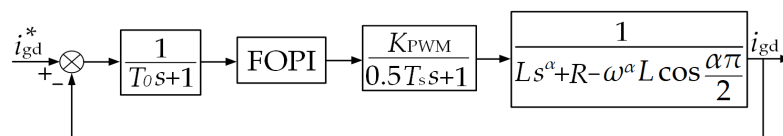


Figure 6. Block diagram of d-axis current control after decoupling.

As shown in the figure, K_{PWM} is the proportional gain, which characterizes the grid-side inverter's amplification characteristics under sufficiently high switching frequencies. The proportional and integral coefficients in FOPI are marked as k_{p0} and k_{i0} , while T_0 and T_s correspond to the time constants of the signal filtering delay link and modulation delay link, respectively.

3.2.2. Fractional-Order Sliding-Mode Voltage Outer-Loop Control

The operation principle of the voltage outer-loop is to send an error signal, derived from the discrepancy between the DC-bus voltage u_{dc} and its reference voltage, to the controller. This generates a reference current i_{gd} , which is then used as the input for the current inner loop. With the aim of enhancing control performance, this study proposes a fractional-order sliding-mode control strategy, and on this basis, develops a fractional-order sliding-mode voltage outer-loop control strategy specifically for the grid-side inverter.

In a three-phase inverter system, the active and reactive power of the system can be explicitly defined on the basis of power calculation equations:

$$\begin{cases} P = \frac{3}{2} (e_{gd} i_{gd} + e_{gq} i_{gq}) \\ Q = \frac{3}{2} (e_{gd} i_{gq} - e_{gq} i_{gd}) \end{cases} \quad (22)$$

In the equation, P and Q denote the active power and reactive power output from the grid-side inverter to the grid, respectively.

To simplify the control, the grid voltage directional vector is utilized, i.e., set $e_{gq} = 0$. Based on Equation (22), the power balance equation is derivable as follows:

$$\begin{cases} (i_{in} - i_{dc})u_{dc} = \frac{3}{2}e_{gd}i_{gd} \\ CD_t^\beta u_{dc} = i_{in} - \frac{3}{2}\frac{e_{gd}i_{gd}}{u_{dc}} \end{cases} \quad (23)$$

Based on the mathematical model of the grid-side inverter, the system's state variables are chosen as follows:

$$\begin{cases} x_3 = u_{dc}^* - u_{dc} \\ x_4 = D_t^\beta x_3 = -D_t^\beta u_{dc} = \frac{3e_{gd}}{2Cu_{dc}}i_{gd} - \frac{1}{C}i_{in} \end{cases} \quad (24)$$

In this equation, x_3 and x_4 denote the system's state variables.

Differentiating x_4 with respect to time, we obtain the following equation:

$$D_t^\beta x_4 = D_t^{2\beta} x_3 = -D_t^{2\beta} u_{dc} = \frac{3e_{gd}}{2Cu_{dc}} D_t^\beta i_{gd} \quad (25)$$

Define $H_2 = \frac{3e_{gd}}{2Cu_{dc}}$, then:

$$D_t^\beta \begin{bmatrix} x_3 \\ x_4 \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} x_3 \\ x_4 \end{bmatrix} + D_t^\beta \begin{bmatrix} 0 \\ H_2 \end{bmatrix} i_{gd} \quad (26)$$

The design for the fractional-order sliding-mode surface is given as follows:

$$f_2 = c_2 x_3 + D_t^\zeta x_3 \quad (27)$$

By differentiating Equation (27), the resulting equation is as follows:

$$D_t^\zeta f_2 = c_2 D_t^\zeta x_3 + D_t^{2\zeta} x_3 \quad (28)$$

The chosen fractional-order rapid power reaching law is given below:

$$D_t^\zeta f_2 = -k_2 |f_2|^\tau D_t^\zeta \text{sgn}(f_2) - k_3 f_2 \quad (29)$$

In the formula, ζ denotes the fractional differentiation order, while τ represents the fractional-order exponent.

The FOSM voltage outer-loop control law is derived by combining Equations (28) and (29) as follows:

$$i_{gd}^* = \frac{1}{H_2} D_t^{\beta-\zeta} \left[c_2 x_3 - D_t^{-\zeta} (k_2 |f_2|^\tau D_t^\zeta \text{sgn}(f_2) + k_3 f_2) \right] \quad (30)$$

In the equation, $\text{sgn}(f_2)$ represents the sign function, and k_2, k_3 are positive constants. These constants guarantee that f_2 and $D_t^\zeta f_2$ have opposite signs, thereby satisfying the stability conditions and proving the system stability under fractional-order fast power-law sliding mode control.

According to the design outlined above, the grid-side converter adopts a double closed-loop control structure of "FOSM voltage outer-loop + dq-axis decoupled FOPI current inner-loop," as shown in Figure 7.

As illustrated in the figure, Y_3 and Y_4 respectively represent the expressions of $|f_2|^\tau$ and $\text{sgn}(f_2)$, in the complex frequency domain.

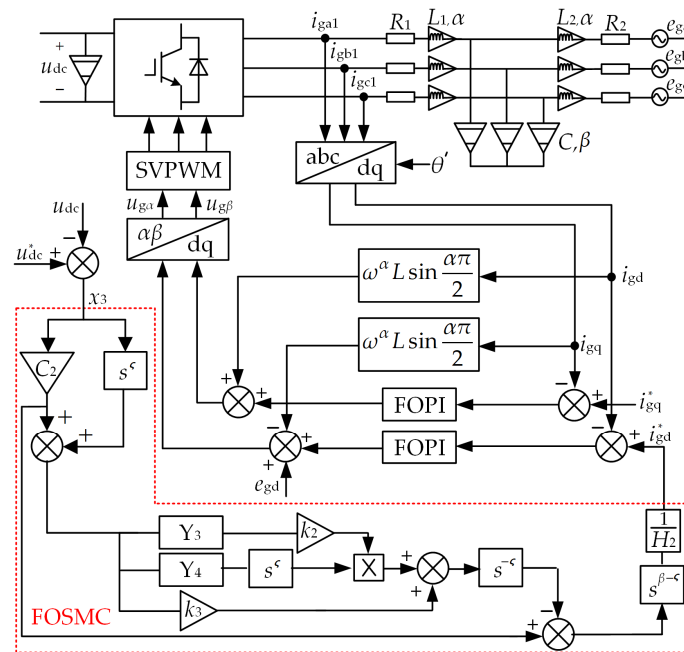


Figure 7. Control block diagram of the grid-side converter with FOSMC voltage outer-loop and FOPI current inner-loop.

4. Simulation Analysis

Following the above design, a digital simulation model for the fractional-order permanent magnet direct-drive wind power system was established in Matlab/Simulink, with the relevant system parameters presented in Table 1. The parameters of capacitance and inductance are adopted directly from reference [10], where $\alpha = 0.85$ and $\beta = 0.95$. The fractional-order values (α, β) are within $(0, 2)$. According to the fractional-order element theory, they cover the phase range of $(-\pi, \pi)$, and α and β make the element characteristics gradually change from positive resistance to integer-order and then to negative resistance. The current satisfies dynamic continuity, power conservation, and the total harmonic distortion (THD) of the power grid is $\leq 5\%$. The speed is constrained by the range of tip speed ratio, the overspeed upper limit (the rated wind speed is set to 12 m/s, and the cut-out wind speed is 25 m/s), and the electromechanical balance. The voltage meets the steady-state deviation of $\pm 5\%$, and the frequency meets the deviation $\leq \pm 0.2$ Hz. The sampling period is $20 \mu\text{s}$, and the toolbox is implemented based on the Caputo definition. The fractional-order sliding mode reaching law enhances robustness with fractional-order differential terms, and chattering suppression is achieved by adjusting the fractional order.

Table 1. Related system parameters.

Parameter	Value
u_{dc}	850 V
α	0.85
β	0.95
L_1	$0.72 \text{ mH/s}^{1-\alpha}$
L_2	$0.72 \text{ mH/s}^{1-\alpha}$
C_{dc}	$0.01 \text{ F/s}^{1-\beta}$
C	$0.001 \text{ F/s}^{1-\beta}$
p	10
J	$5 \text{ kg}\cdot\text{m}^2$
R_s	0.045Ω
ψ_f	1.832 W_b
L_{sd}, L_{sq}	$0.755 \text{ mH/s}^{1-\alpha}$

Next, simulations are carried out under varying wind speeds and asymmetric grid faults to compare the “FOSM outer-loop + FOPI inner-loop” control proposed in this paper with the “FOPI outer-loop + FOPI inner-loop” control proposed in [10], so as to verify the effectiveness of the fractional-order sliding-mode outer-loop control.

4.1. Control Parameter Design

The particle swarm optimization (PSO) method is employed in this paper for parameter tuning of the fractional-order controller. Similarly to other evolutionary algorithms, PSO is built around “swarm” and “evolution” concepts, relying on particle fitness values for operation. Particle velocity and position adjustments are shaped by both individual and collective experiences, with the objective of finding the optimal solution across the entire search domain [32]. The optimization process is depicted in Figure 8.

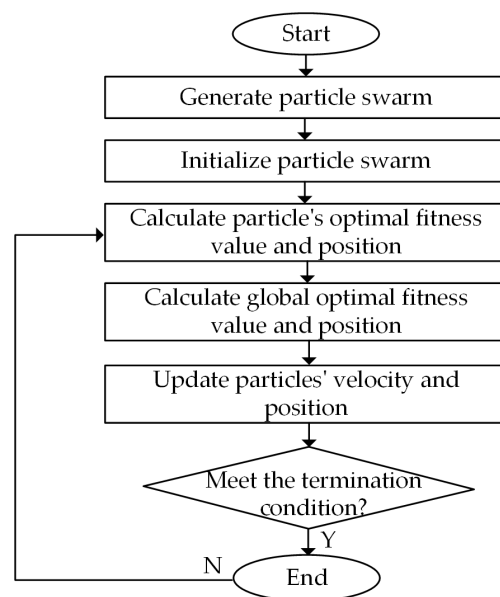


Figure 8. Particle swarm optimization algorithm design process.

The velocity and position of particles in the search space are determined via the following formula [33]:

$$\begin{cases} j_{t+1} = mj_t + l_1 o_1 (P_t - h_t) + l_2 o_2 (G_t - h_t) \\ h_{t+1} = h_t + j_{t+1} \end{cases} \quad (31)$$

Within the equation, h denotes the particle position, j denotes the particle velocity, and m serves as the inertia factor. Acceleration constants are represented by l_1 and l_2 , while o_1 and o_2 are random numbers in $[0, 1]$. P_t refers to the particle’s own historical optimal position, and G_t signifies the historical global optimal position of the entire particle swarm.

The fitness function is selected as follows:

$$Z = D_t^{-1}(t \cdot |e|) \quad (32)$$

Taking the voltage-speed outer loop of the machine-side FOSMC as an example, e is defined as the difference between the reference speed n^* and the actual speed n . Set the inertia factor $m = 0.6$, the acceleration constants $l_1 = l_2 = 2$, the dimension to 5, the particle swarm size to 50, the maximum number of iterations to 10, the minimum fitness value to 10^{-14} , the velocity range to $[-1, 1]$, the ranges of c_2 , k_0 , and k_1 all to $[0, 100]$, the range of μ

to $[0, 1]$, and the range of σ to $[0, 1]$. Other optimization conditions related to controllers have been placed in Appendix A.

After determining the spatial dimension of the fitness function and the population size, the PSO algorithm is executed to perform the optimization. Through continuous iteration, the particle with the minimum fitness function is identified, and its position in the search space represents the optimal parameters. Finally, the algorithm is checked to see whether it satisfies the constraint conditions: if it does, the optimization terminates and the algorithm outputs the optimal parameters; if it does not, the optimization continues. The variation curve of performance index Z obtained by PSO for the FOSM of the machine-side voltage speed outer-loop is shown in Figure 9.

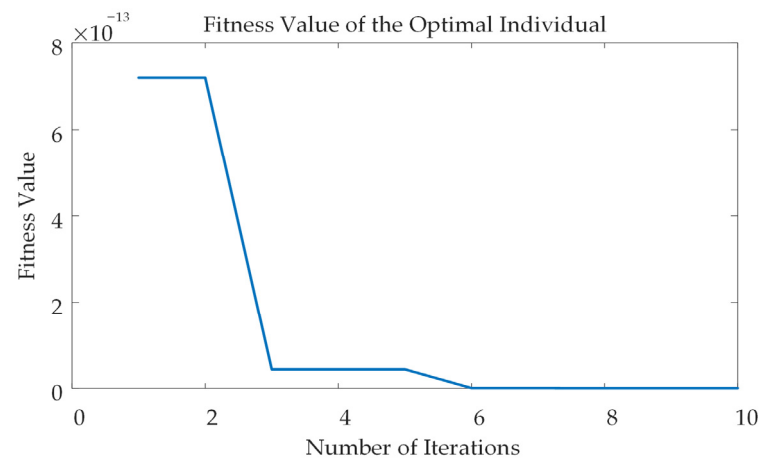


Figure 9. Response curve of performance index Z .

4.2. Simulation Analysis Under Changing Wind Speeds

The simulation experiment was performed based on the simulated wind speed presented in Figure 10. The wind speed is initially maintained at 7 m/s, then suddenly increases from 7 m/s to 12 m/s at 0.6 s. It is followed by a gradual rise from 12 m/s at 0.8 s, reaching 15 m/s at 1 s. After maintaining 15 m/s for 0.2 s, it gradually decreases to 12 m/s. After maintaining 15 m/s for 0.2 s, it gradually decreases to 12 m/s.

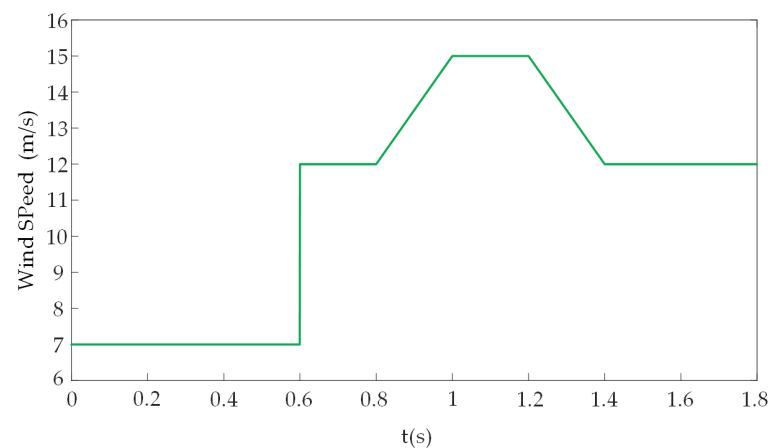


Figure 10. Wind speed curve.

The control parameters relevant to the machine-side converter are tabulated in Table 2.

Table 2. Machine-side configuration parameters.

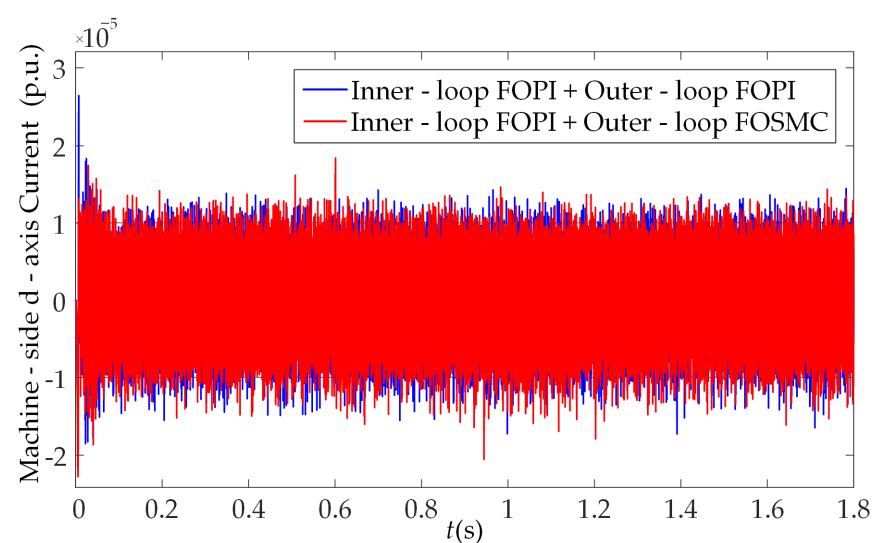
Loop	Parameter	Value
FOPI current loop	k_p	50
	k_i	1000
	η	0.8
FOPI speed loop	k_p'	20
	k_i'	3000
	η'	0.9
FOSMC speed loop	μ	0.976
	σ	0.142
	k_0	63
	k_1	59

Table 3 lists the relevant control parameters for grid-side converters.

Table 3. Grid-side configuration parameters.

Loop	Parameter	Value
FOPI current loop	k_{p0}	80
	k_{i0}	1000
	η_0	0.78
FOPI voltage loop	k_{p0}'	0.5
	k_{i0}'	60
	η_0'	0.95
FOSMC voltage loop	ς	0.1
	τ	0.869
	k_2	16
	k_3	100

With regard to the machine-side, the corresponding simulation results are shown in Figures 11 and 12.

**Figure 11.** Machine-side d-axis current comparison chart. Operating point: wind speed varies as shown in Figure 10, grid normal operation.

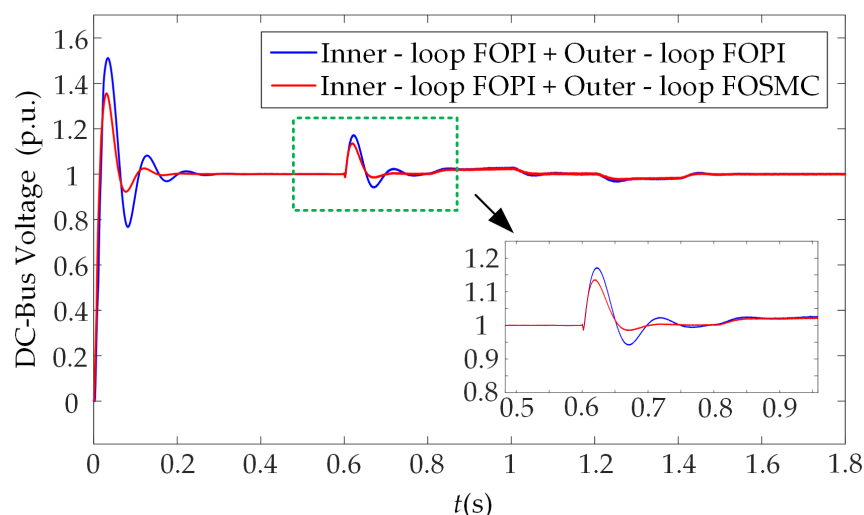


Figure 12. DC-bus voltage comparison chart. Operating point: wind speed varies as shown in Figure 10, grid normal operation.

Simulation analysis shows that both control approaches deliver good control performance. As illustrated in Figure 11, both control methods can effectively maintain the d-axis current i_{sd} near zero. However, during start-up and at 0.6 s when wind speed changes suddenly, the overshoot of FOSMC is smaller than that of FOPI control. As shown in Figure 12, FOSMC performs better in maintaining DC voltage stability. When wind speed changes suddenly, the DC-bus voltage has an overshoot of 0.135 p.u., while the outer-loop FOPI control has an overshoot of 0.172 p.u.

In contrast, FOSM speed outer-loop control not only enhances the system's disturbance rejection capability but also improves its stability and dynamic response. It exhibits stronger robustness when subjected to external disturbances (e.g., abrupt wind speed variations), thereby enabling faster attenuation of system fluctuations and consequently minimizing overshoot.

Grid-side simulation results are depicted in Figures 13–15, respectively.

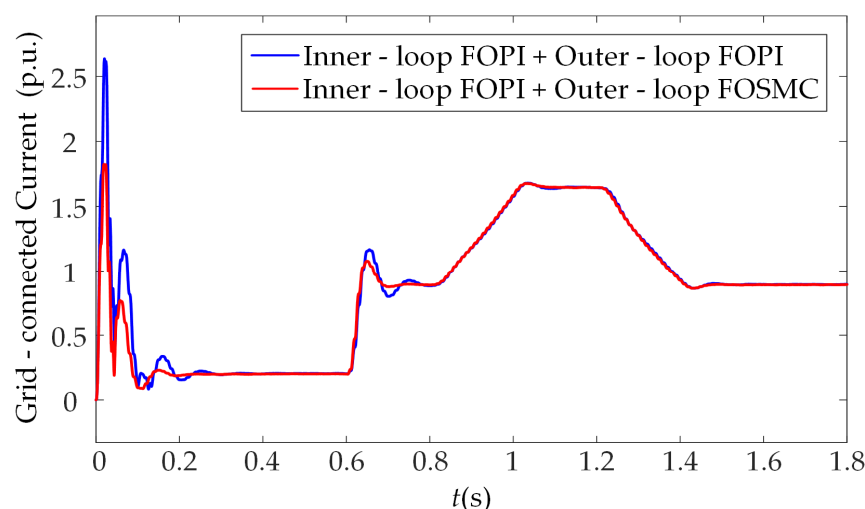


Figure 13. Grid-connected current comparison chart. Operating point: wind speed varies as shown in Figure 10, grid normal operation.

As demonstrated in Figures 13–15, adopting FOSMC in the voltage outer-loop significantly improves system performance in multiple aspects. During the start-up phase, the grid-connected current converges rapidly to a steady state. In the case of sudden wind

speed variations, FOSMC exhibits a maximum current overshoot of 0.187 p.u., which is smaller than FOPI control's 0.277 p.u., demonstrating superior fluctuation suppression and disturbance resistance. FOSMC also outperforms in power tracking, stabilizing within 0.2 s, and reduces power fluctuation amplitude by 31.74% relative to FOPI under abrupt wind speed changes—mitigating the grid-connection impact of wind energy fluctuations effectively. Moreover, it can quickly stabilize the power factor at 1, contributing to better power quality optimization.

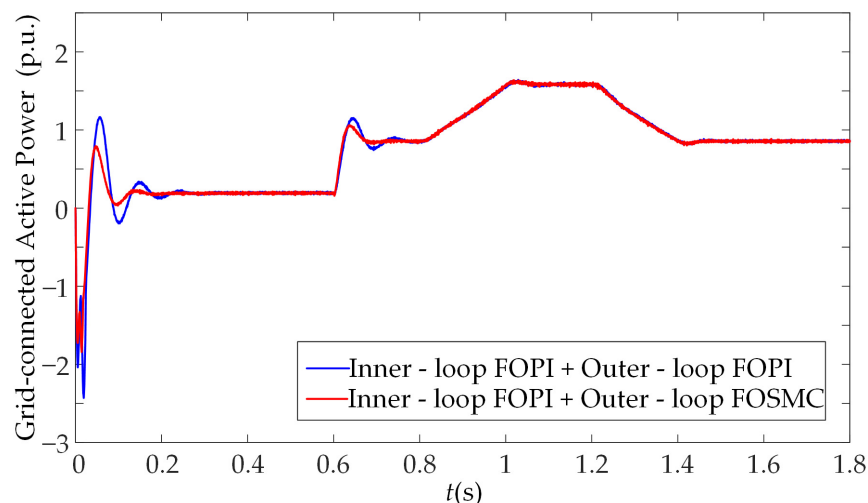


Figure 14. Grid-connected active power comparison chart. Operating point: wind speed varies as shown in Figure 10, grid normal operation.

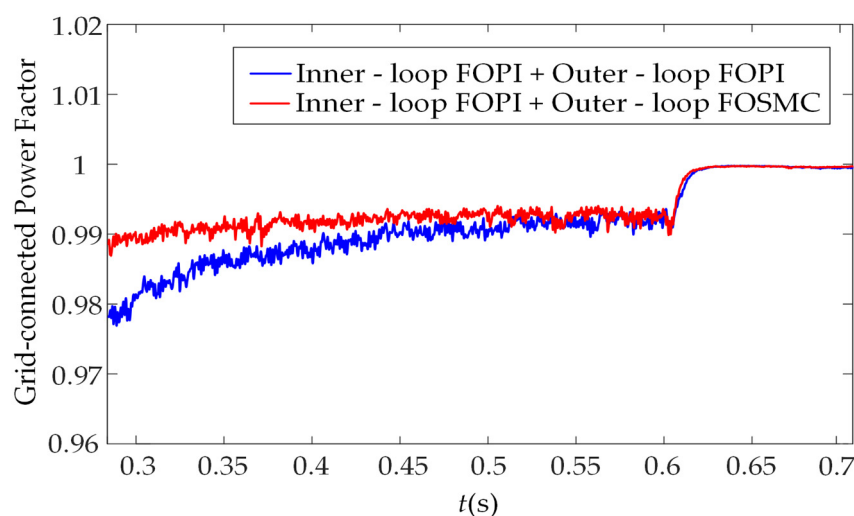


Figure 15. Grid-connected power factor comparison chart. Operating point: wind speed varies as shown in Figure 10, grid normal operation.

4.3. Simulation Analysis Under Asymmetric Grid Faults

In practical scenarios, the probability of asymmetric faults in the power grid is relatively high. Therefore, this study analyzes the simulation results of different control methods under the same wind speed conditions when a grid fault occurs. The simulation conditions are set as follows: at 0.5 s, the phase-C voltage sags by 50%; at 1 s, the fault is cleared, and the voltage is restored, as depicted in Figure 16.

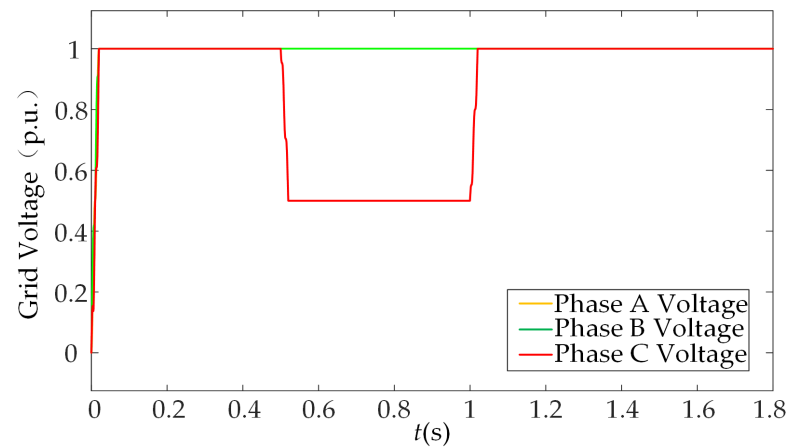


Figure 16. Grid voltage.

For both the converters, simulation experiments were conducted with two different control strategies to obtain the DC-bus voltage, grid-connected active power, and grid-connected current when asymmetric faults occur in the power grid. The results are presented in Figures 17–19.

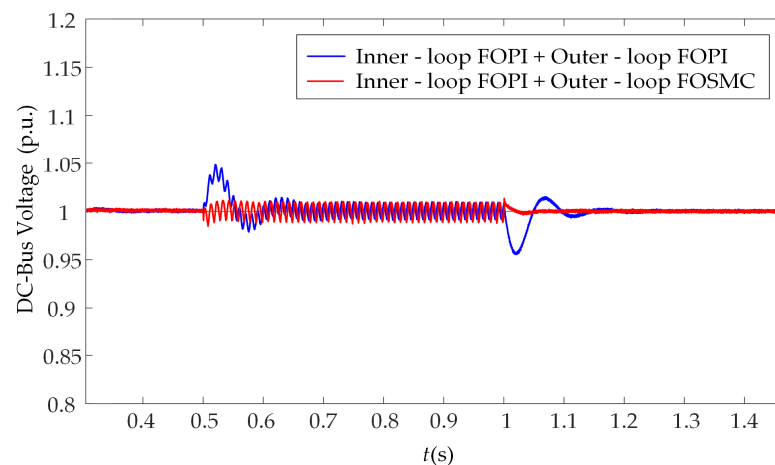


Figure 17. DC-bus voltage comparison chart. Operating point: Wind speed is 12 m/s, and phase C voltage drops by 50%.

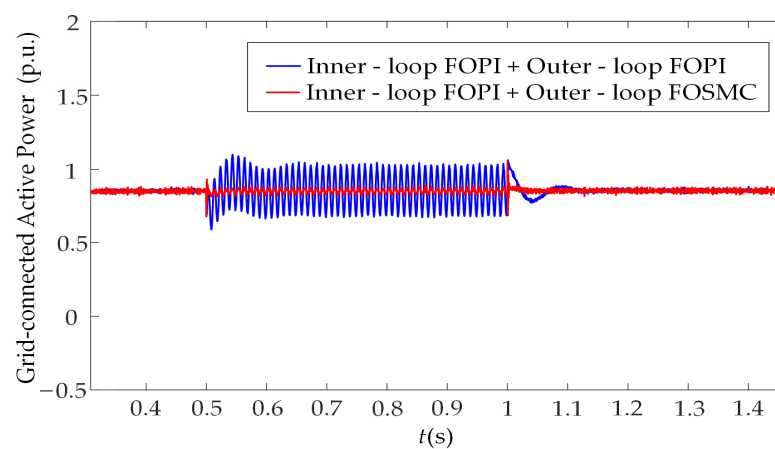


Figure 18. Grid-connected active power comparison chart. Operating point: Wind speed is 12 m/s, and phase C voltage drops by 50%.

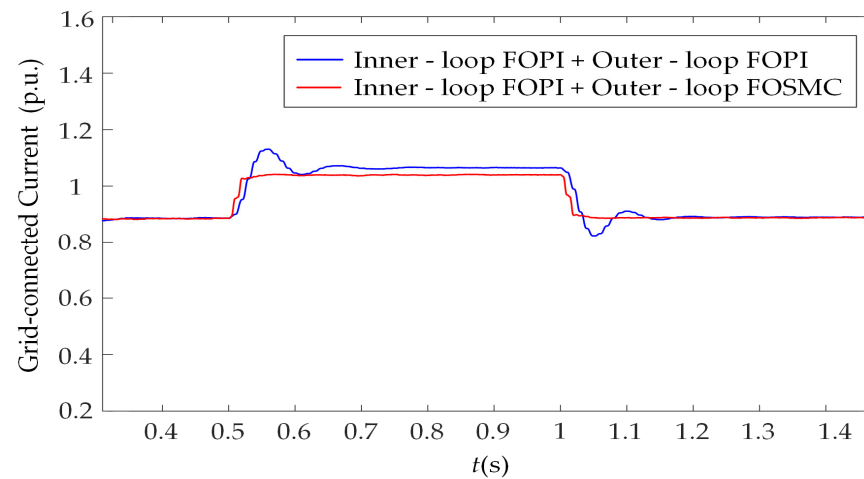


Figure 19. Grid-connected current comparison chart. Operating point: Wind speed is 12 m/s, and phase C voltage drops by 50%.

As demonstrated in Figure 17, during the period before and after the grid voltage fault, both control methods effectively stabilize the DC-bus voltage at approximately 850 V, demonstrating favorable control performance. However, during the grid voltage fault period (0.5 s–1 s), the DC-bus voltage exhibits significant fluctuations under both control methods. When both the machine-side and grid-side outer-loop controls employ FOSMC, the voltage drop is as low as 0.985 p.u., a 65.38% reduction compared to that with both outer-loop FOPI control. Additionally, during voltage recovery, employing FOSMC in the outer loops results in faster recovery speed and higher steady-state accuracy.

Meanwhile, there are obvious differences in the dynamic response of grid-connected active power under the two control methods during the fault, as illustrated in Figure 18. When the outer-loop uses FOPI control, the maximum drop in active power is 30.65%, accompanied by significant oscillations during the drop, indicating insufficient disturbance suppression capability of the control strategy. The overshoot during the recovery phase is 24.1%. In contrast, with FOSMC outer-loop control, the maximum drop in active power is 20.45%, representing a 33.38% reduction in drop magnitude compared to the FOPI control. Meanwhile, the fluctuation amplitude during the fault period is also significantly reduced—indicating that the strong robustness of sliding mode control enables effective disturbance suppression, facilitates faster recovery, and involves no obvious oscillations during the restoration process. Regarding the grid-connected current of the two control methods at fault occurrence and clearance, as shown in Figure 19, FOSMC exhibits a smaller overshoot than FOPI control.

In summary, FOSMC integrates fractional-order calculus with the robustness of sliding mode control, and is comparatively superior to FOPI control in terms of dynamic response, steady-state accuracy, and anti-interference capability.

4.4. Robustness Test of the FOSM Controller

Taking the grid-side voltage outer-loop FOSM controller as an example, this paper verifies the controller's robustness under wind speed variation and parameter uncertainty by changing system physical parameters and control parameters. Specifically, the machine-side inductance is perturbed by +10%, the grid-side inductance by −10%, while the core gain k_2 of the sliding mode switching term and the fractional-order parameters ζ and τ are varied within a $\pm 20\%$ range. Key indicators analyzed include the DC-bus voltage overshoot, settling time (with an error $\leq 0.5\%$), and the average of absolute errors in the stable phase at the later stage of simulation when wind speed suddenly changes at

0.6 s. The heat map of transient indicators for DC-bus voltage under different parameter combinations is shown in Figure 20.

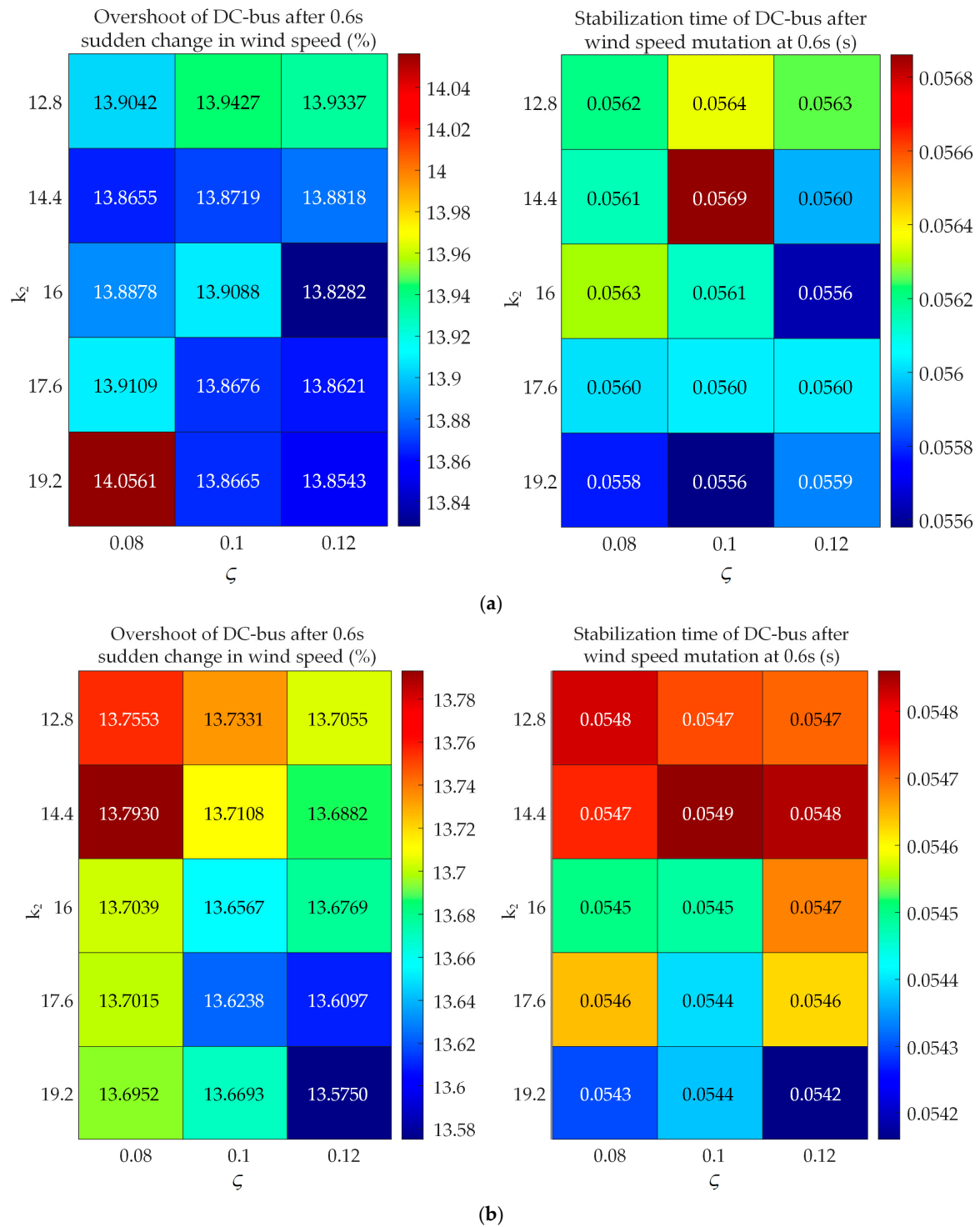


Figure 20. Cont.

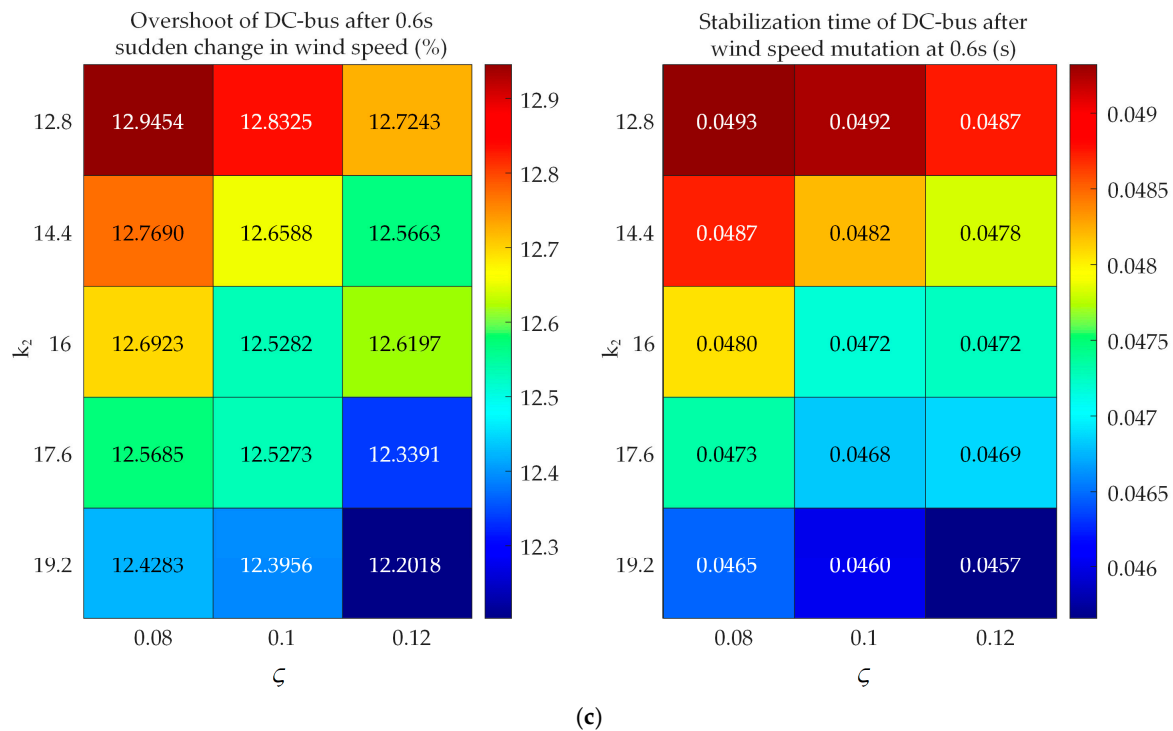


Figure 20. Heat map of DC-bus voltage transient indicators under different parameter combinations: (a) Heat map of transient indices when τ is 0.695; (b) Heat map of transient indices when τ is 0.869; (c) Heat map of transient indices when τ is 1.043.

It can be seen from Figure 20 that under parameter uncertainty, the DC-bus voltage overshoot fluctuates relatively smoothly, with no local sudden increase areas. The controller can effectively suppress the overshoot amplitude of the DC-bus voltage under parameter disturbances. Meanwhile, the distribution of the DC-bus voltage stabilization time is also relatively concentrated; even with parameter deviations, the controller can still quickly restore the system to a steady state.

The absolute value of the steady-state error of the DC-bus voltage under different parameter combinations is shown in Figure 21.

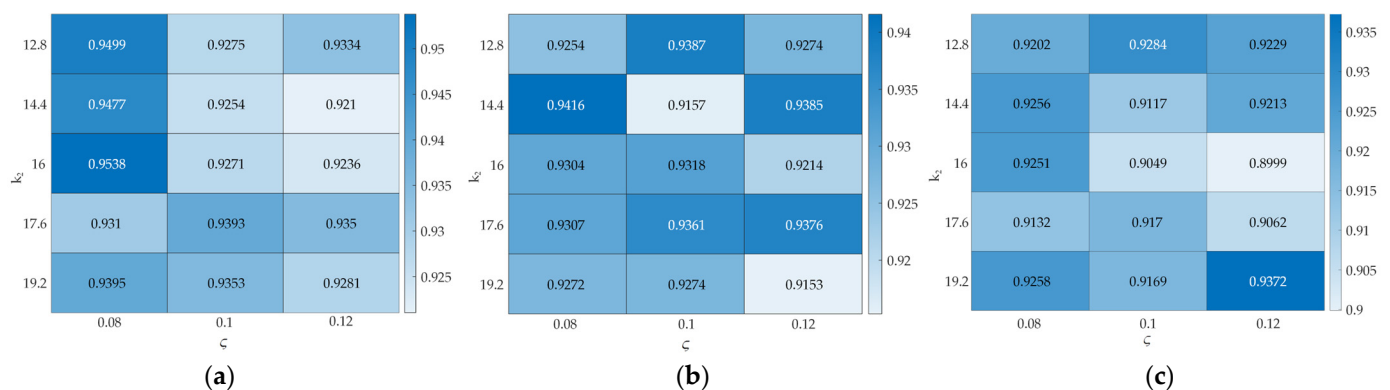


Figure 21. Heat map of the absolute value of the DC-bus voltage steady-state error under different parameter combinations: (a) Heat map when τ is 0.695; (b) Heat map when τ is 0.869; (c) Heat map when τ is 1.043.

It can be seen from Figure 21 that under different parameter combinations, the absolute value of the DC-bus voltage steady-state error is not only small in value, but also has a

small fluctuation amplitude when parameters change, fully demonstrating that the FOSM controller has strong adaptability under parameter disturbances.

Overall, the FOSM controller designed in this paper exhibits good robustness within the tested parameter range.

5. Conclusions

This study focuses on a fractional-order permanent magnet direct-drive wind power generation system, which integrates a fractional-order capacitor installed on the DC-bus and an FOLCL filter. Based on the system's mathematical model, a dual closed-loop control structure was designed: the machine-side rectifier employs an FOSM speed outer loop paired with a dq-axis decoupled FOPI current inner loop, while the grid-side converter adopts a FOSM voltage outer loop paired with a dq-axis decoupled FOPI current inner loop. The parameters of the FOSM controllers were tuned using the PSO method, and the validity of the theoretical research was subsequently verified via digital simulations. Key findings of this study are as follows:

- (1) Compared with the traditional FOPI outer-loop control, the FOSM outer-loop control demonstrates superior dynamic and static performance. In the presence of disturbances such as sudden wind speed fluctuations and grid voltage faults, FOSMC not only achieves lower overshoot and faster response speed but also exhibits notable superiority in terms of steady-state accuracy and anti-interference ability, thus providing a more dependable control support for the efficient grid integration of permanent magnet direct-drive wind power systems.
- (2) When both the speed outer-loop and the voltage outer-loop employ FOSMC, the system's performance improvement under different disturbance scenarios exhibits distinct focus areas. Applying FOSMC in the speed outer-loop can enhance the dynamic regulation capability for the d-axis current and speed; adopting FOSMC in the voltage outer-loop can significantly improve the stability of the DC-bus voltage, as well as the control accuracy of grid-connected current and power. Meanwhile, under complex operating conditions such as grid voltage faults, the robustness advantage of FOSMC can effectively suppress system fluctuations and facilitate faster system recovery.

However, this study still has some limitations:

- (1) The permanent magnet direct-drive generator studied in this paper is an integer-order model, and this research focuses on the basic effectiveness verification of the FOSMC strategy in the fractional-order permanent magnet direct-drive wind power grid-connected system. Therefore, in simulation modeling, only some delay links of the system (such as pulse width modulation delay and communication transmission delay) are taken into account, with the focus on verifying the control performance of the control strategy under wind speed variation and grid voltage faults. This study does not consider noise and quantization effects for the time being and does not extend to the influence of nonlinear disturbances in engineering scenarios.
- (2) This study focuses on the working conditions of wind speed changes and single-phase grid voltage sags, without considering more complex composite fault conditions, such as "wind speed change + grid fault". The control performance under multiple external disturbance conditions has not been verified, and there is a certain limitation in the coverage of working conditions.
- (3) The selection of relevant parameters for FOSMC is optimized according to the working conditions set in this study, and the adaptive matching between working condition parameters and control parameters has not been achieved. When the working con-

dition range is beyond the scope of this study, the parameters need to be re-tuned, resulting in a lack of cross-working condition universality.

Nonetheless, the applicability of the FOSMC method designed in this paper also has some limitations:

- (1) The proposed FOSMC can achieve smooth power output by quickly adjusting the electromagnetic torque above the rated wind speed. However, in the low-wind-speed range, due to the small output torque of the permanent magnet synchronous generator, the chattering suppression effect of FOSMC will weaken as the signal-to-noise ratio of the torque signal decreases, and there are certain limitations.
- (2) The algorithm proposed in this paper can suppress torque ripple by virtue of the fast convergence characteristic of FOSM when the single-phase voltage sag is 50%. But when the voltage sag is deep, the switching gain of the fractional-order sliding mode surface needs to be greatly increased to maintain control stability. At this time, the chattering phenomenon of the algorithm will be significantly aggravated, and its applicability in deep sag scenarios is limited.

Future research of this study can be carried out in the following aspects (but not limited to):

- (1) For the permanent magnet synchronous generator in the fractional-order permanent magnet direct-drive wind power system investigated in this paper—currently modeled using an integer-order mathematical framework—it is necessary to conduct in-depth studies on its fractional-order mathematical model.
- (2) This paper employs FOPI control for the inner-loop control of back-to-back converters. In subsequent work, FOSMC can be applied to the inner-loop control for comparative analysis. Additionally, based on (1), FOSMC can be used for full-system control and then compared and analyzed in detail with other fractional-order control strategies.
- (3) Noise and quantization effects play a crucial role in practical system performance. In future work, we will introduce noise to analyze its impact on the control accuracy of FOSMC; explore the coupling mechanism between quantization effects and the dynamic characteristics of fractional-order capacitors; and verify the anti-disturbance capability of the proposed control strategy when such disturbances are considered through comparative simulations, thereby further enhancing the engineering applicability of the method.

Author Contributions: Conceptualization, J.X.; data curation, Y.L. (Yue Lan), C.W., B.L., Y.L. (Yingheng Li) and Y.X.; formal analysis, Y.L. (Yue Lan); investigation, Y.L. (Yue Lan); methodology, J.X.; project administration, J.X.; resources, J.X., C.W., B.L., Y.L. (Yingheng Li) and Y.X.; software, Y.L. (Yue Lan); supervision, J.X.; validation, Y.L. (Yue Lan); visualization, Y.L. (Yue Lan); writing—original draft, Y.L. (Yue Lan); writing—review and editing, J.X. and Y.L. (Yue Lan) All authors have read and agreed to the published version of the manuscript.

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Conflicts of Interest: The authors declare no conflicts of interest.

Appendix A

Information related to the particle swarm optimization algorithm, such as the state objective function, parameter boundaries, population size, number of iterations, and stopping criteria, is presented in Tables A1 and A2. Among them, the objective function is integral of

time-weighted absolute error (ITAE), the number of iterations is 10 and the population size is 50.

Table A1. Machine-side parameter.

Loop	Parameter	Parameter Boundaries	Stopping Criteria
FOPI current loop	k_p	[0, 100]	10^{-6}
	k_i	[800, 1200]	
	η	[0, 1]	
FOPI speed loop	k_p'	[0, 100]	10^{-6}
	k_i'	[2800, 3500]	
	η'	[0, 1]	
FOSMC speed loop	μ	[0, 1]	10^{-14}
	σ	[0, 1]	
	k_0	[0, 100]	
	k_1	[0, 100]	

Table A2. Grid-side parameter.

Loop	Parameter	Parameter Boundaries	Stopping Criteria
FOPI current loop	k_{p0}	[0, 100]	10^{-6}
	k_{i0}	[800, 1200]	
	η_0	[0, 1]	
FOPI voltage loop	k_{p0}'	[0, 10]	10^{-6}
	k_{i0}'	[0, 100]	
	η_0'	[0, 1]	
FOSMC voltage loop	ς	[0, 1]	10^{-14}
	τ	[0, 1]	
	k_2	[0, 50]	
	k_3	[50, 150]	

The relevant convergence graph is shown in Figure A1.

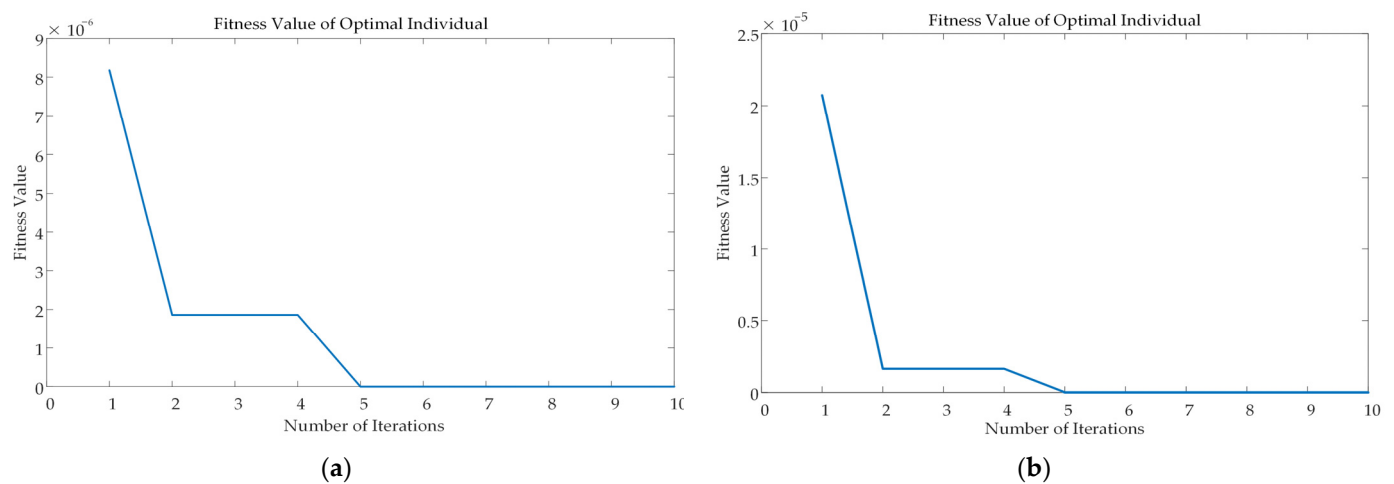


Figure A1. Cont.

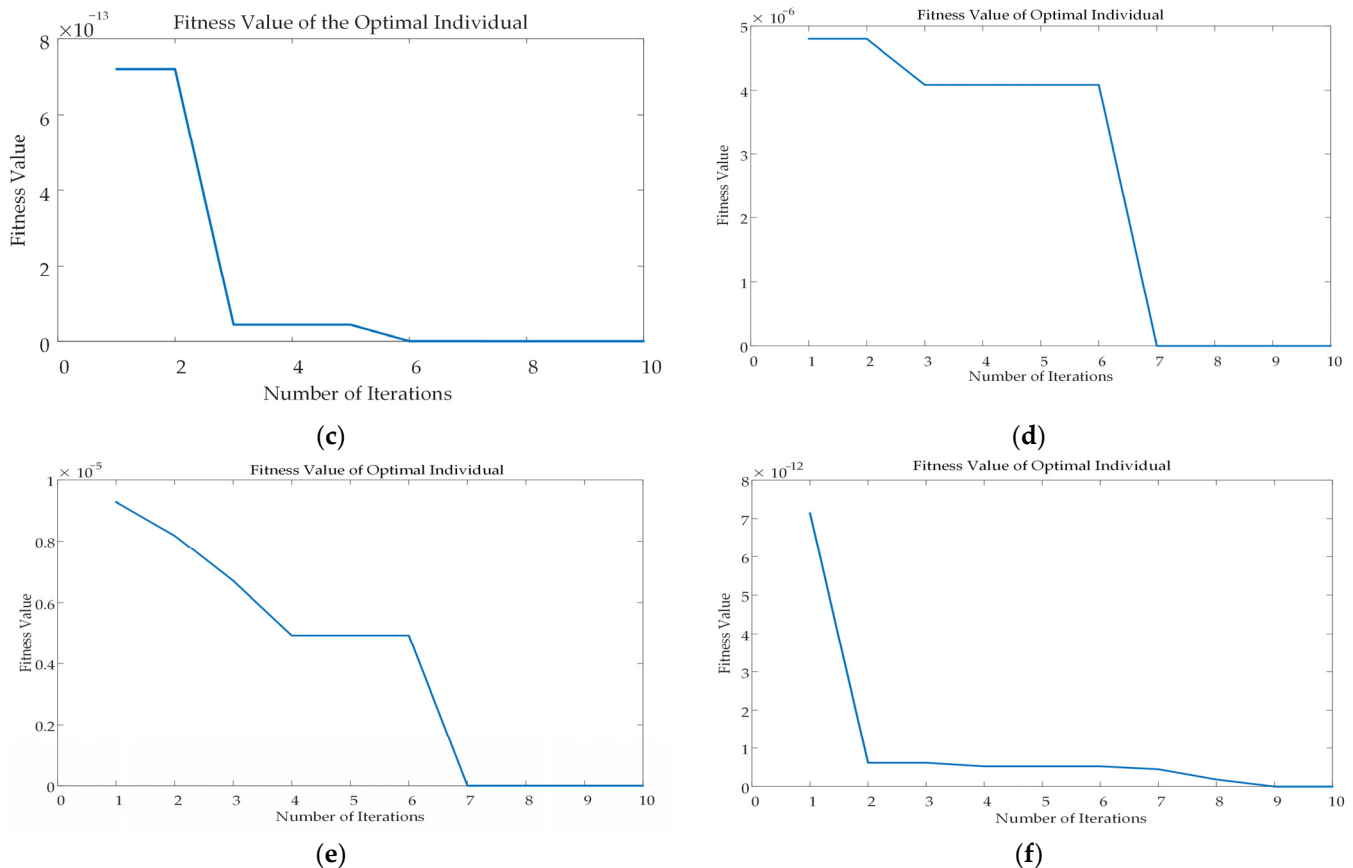


Figure A1. Convergence graph of the optimization algorithm: (a) Convergence graph of the machine-side FOPI current loop; (b) Convergence graph of the machine-side FOPI speed loop; (c) Convergence graph of the machine-side FOSMC speed loop; (d) Convergence graph of the grid-side FOPI current loop; (e) Convergence graph of the grid-side FOPI voltage loop; (f) Convergence graph of the grid-side FOSMC voltage loop.

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