Research Article

Robust Switching Gain-Based Fractional-Order Sliding Mode Control for Wind-Powered Microgrids

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This study proposes a novel fractional-order sliding-mode control strategy with robust switching gain to achieve reliable and high quality of wind-powered microgrid systems. Three fractional-order sliding mode controllers are designed to generate continuous control signals and regulate the outer DC-link voltage loop and inner current loop in the grid-side inverters. High robustness and stability of the grid-side inverter can be guaranteed even in the presence of parameter variations and external disturbances. Owing to the fractional-order sliding manifold and fractional-order integral control law, the chattering is attenuated. The fractional-order robust adaptive switching gain is designed to avoid overestimating the upper bound of matched/unmatched uncertainties, save the control energy, and guarantee the rapidity and robustness of the convergence. Simulations validate the proposed method.

1. Introduction

Distributed generation (DG) has paid more and more attention because of its advantages, such as low investment, clean environment protection, high reliability of power supply, and flexible power generation [1]. Beside the aforementioned advantages, DG unit may bring some problems to the grid, such as voltage fluctuations, voltage deviations, and bidirectional power flow. To give full play to the efficiency of the distributed generation systems, the most effective way is to combine the distributed generation and loads to form a microgrid connecting to the main grid. The active power and reactive power between the microgrid and the main grid can be flexibly exchanged. If the main power grid fails, the microgrid will be instantly separated from the main grid and run in the isolated island mode to guarantee the power supply of important loads [2].

In the wind-powered microgrid system, the voltage source inverters (VSIs) are widely used as grid-side inverters. However, this type of converter has a characteristic that the high-frequency harmonics exist in the voltage waveform. Therefore, filters must be used to filter the harmonics out. LCL filters, as commonly used filters, have superior

performance in harmonic attenuation and lower current ripple if compared with the L filter, another commonly used filter. Due to the structures and resonances of LCL filters, the LCL filters are rather complicated, which means the control schemes for LCL filters are difficult to implement. The traditional control methods cannot have good performance when applied in the control of LCL filters [3, 4], which may have poor reference tracking and stability problems. In [5], a proportional resonant (PR) controller with a PI controller is designed to get a better dynamic response in terms of disturbance compensation. However, the high gain was introduced by PR, which can cause a larger steady-state error [6].

There are many uncertainties existed in the wind powered microgrid system, such as parameter variations of LCL filters, grid voltage sag, and frequency fluctuation. Sliding mode control (SMC) is extensively used to deal with aforementioned problems due to its high robustness, fast dynamics responses, and simplicity in the control algorithm. It is suitable to use SMC for the control of the grid-side inverters (GSI), on account of its attractive advantages [7–12]. In [9], an integral-type terminal sliding mode control is proposed to get a high dynamics response in the wind

energy conversion system with LCL filters. In [13], a sliding mode controller with Kalman filter (KF) is proposed in a three-phase unity power factor rectifier. An SMC method by modifying reaching law for the variable-speed direct-driven wind energy conversion systems is proposed in [14]. In [15, 16], PI-type sliding surfaces are presented to remove the steady-state error. In [17], a super-twisting sliding mode (STSM) for a gearless wind turbine is developed, so that the system has distinguished robustness against disturbances. In [18], a novel direct active and reactive power control of grid-connected DFIG-based wind turbine systems is presented, which employed a novel SMC to calculate the required rotor control voltage. In [19], an advanced SMC scheme is designed for wind energy conversion systems with a non-linear disturbance observer.

The aforementioned SMC strategies are all based on integer order calculus which restricts the development of the control theory [20]. With the development of fractional-order calculus, the fractional-order SMC (FOSMC) has been a topic of intense research. Compared with traditional integral order calculus, fractional calculus increases the variability of differential and integral degrees of freedom, which brings new flexibility to the design of control systems [21, 22], so that it has both advantages of fractional calculus and the SMC. The memorial and genetic traits are the typical characteristics of FOSMC, which can predict the control behavior of the system to a certain extent, reduce the switching probability of the control behavior in the reaching process, and thus improve the continuity of the control [23, 24]. A new time-delay estimation-based fractional-order NTSM controller and an intelligent robust fractional-order LSM controller are proposed in [25, 26], but the boundary layer leads to steady-state errors. An adaptive fractional-order TSM is developed in [27]; however, the singularity exists.

To achieve higher performances of wind-powered microgrid systems, this study proposes a fractional-order SMC method with robust adaptive switching gain for GSI. Fractional-order sliding mode controllers are designed for the outer DC-link voltage loop and inner current loop in GSI. The DC-link voltage and current errors can be controlled to converge to zero and chattering can be avoided. High performances of GSI can be guaranteed even in the presence of external disturbances and parameter variations. The remainder of this study is organized as follows: Section 2 briefly introduces the fractional calculus and the wind-powered microgrid systems with a squirrel cage induction generator (SCIG) and the model of GSI with LCL. Section 3 presents the fractional-order SMC design for GSI. Section 4 shows the simulation results. Finally, Section 5 gives a conclusion.

2. Preliminaries

Composed of a wind turbine, an SCIG, a converter, and an LCL-type filter, a SCIG wind turbine system is shown in Figure 1. Figure 2 shows the detailed circuit of the wind-powered microgrid control system, and the definition of symbols in Figure 2 is given in Table 1. The PWM converter

consists of a machine-side rectifier (MSR), a DC-link capacitor C with its voltage u_{DC} , and a grid-side inverter. The wind turbine delivers the energy P_{Ge} to MSR through the SCIG. The MSR and GSI are controlled separately using fractional-order SMC strategies proposed in the study.

Due to the existence of the uncertainties in the GSI system, which may be caused by aging, temperature, saturation effects, and some parameter variations have to be considered as follows when modeling the system:

$$\begin{cases} L_{lj} = L'_{lj} + \Delta L_{lj}, \\ R_{lj} = R'_{lj} + \Delta R_{lj}, \\ L_{Gj} = L'_{Gj} + \Delta L_{Gj}, \\ R_{Gj} = R'_{Gj} + \Delta R_{Gj}, \\ C_j = C_j + \Delta C_j, \\ R_{Cj} = R_{Cj} + \Delta R_{Cj}, \end{cases} \quad (1)$$

$j = a, b, c,$

where Δ stands for the parameter variations.

With the harmonic of grid voltages and the loss of switches ignored, the model of the GSI systems can be expressed as follows:

$$\begin{cases} C\dot{u}_{DC} = i_L - i_I, \\ \dot{i}_j = -\frac{R_{lj}}{L_{lj}}i_j - \frac{1}{L_{lj}}u_{Cj} + \frac{u_{DC}}{L_{lj}}S_j, \\ \dot{i}_{Gj} = -\frac{R_{Gj}}{L_{Gj}}i_{Gj} + \frac{1}{L_{Gj}}u_{Ck} - \frac{1}{L_{Gj}}e_j, \\ \dot{u}_{Cj} = \frac{1}{C_{lj}}i_j - \frac{1}{C_{lj}}i_{Gj} + R_{Cj}(i_j - i_{Gj}), \end{cases} \quad (2)$$

$j = a, b, c,$

where S_j , $j = a, b, c$ are the inputs of the PWM modulator, and $i_I = S_a i_a + S_b i_b + S_c i_c$.

In view of the uncertainties of parameters, the model of the GSI systems (2) can be rewritten in the vector form as

$$\begin{cases} C\dot{u}_{DC} = i_L - i_I, \\ \dot{\mathbf{i}}_G = -\mathbf{L}_G^{-1}\mathbf{R}_G\mathbf{i}_G - \mathbf{L}_G^{-1}\mathbf{e} + \mathbf{L}_G^{-1}\mathbf{u}_C + \boldsymbol{\sigma}_G\mathbf{u}_C + \Delta\mathbf{d}_G, \\ \dot{\mathbf{u}}_C = (\mathbf{C}^{-1} - \mathbf{L}_I^{-1}\mathbf{R}_C\mathbf{R}_I)\mathbf{i} - (\mathbf{C}^{-1} - \mathbf{L}_G^{-1}\mathbf{R}_C\mathbf{R}_G)\mathbf{i}_G + \mathbf{L}_G^{-1}\mathbf{R}_C\mathbf{e} \\ \quad - (\mathbf{L}_I^{-1} + \mathbf{L}_G^{-1})\mathbf{R}_C\mathbf{u}_C + \mathbf{L}_G^{-1}\mathbf{R}_C u_{DC}\mathbf{S} + \boldsymbol{\sigma}_C u_{DC}\mathbf{S} + \boldsymbol{\rho}_C, \end{cases} \quad (3)$$

where $\mathbf{S} = [S_a, S_b, S_c]^T$, $\mathbf{i}_G = [i_{Ga}, i_{Gb}, i_{Gc}]^T$, $\mathbf{u}_C = [u_{Ca}, u_{Cb}, u_{Cc}]^T$, $\mathbf{i} = [i_a, i_b, i_c]^T$, and $\mathbf{e} = [e_a, e_b, e_c]^T$, and the parameters $\mathbf{L}_I = [L_{Ia}, L_{Ib}, L_{Ic}]^T$, $\mathbf{R}_k = \text{diag}(R'_{ka}, R'_{kb}, R'_{kc})$, $k = I, G$, C ; $\mathbf{L}_l = \text{diag}(L'_{la}, L'_{lb}, L'_{lc})$, $l = I, G$; and $\mathbf{C} = \text{diag}(C'_a, C'_b, C'_c)$. It can be assumed that the uncertainties $\boldsymbol{\sigma}_G$, $\Delta\mathbf{d}_G$, $\boldsymbol{\sigma}_C$, and $\boldsymbol{\rho}_C$ are upper-bounded.

FIGURE 1: An SCIG wind turbine system.

FIGURE 2: Wind powered microgrid control system.

TABLE 1: Definition of the symbols in Figure 2 ($j = a, b, c$).

Symbol	Definition
e'_j, i'_j	Generator voltages and currents
L', R	Inductors/resistances of the L -type filter of the generator
i_L	Output current of the MSR
i_I	Input current of GSI
i_{DC}	Current through the capacitor
e_j	Grid voltages
i_j, i_{Gj}	Currents flowing through the LCL filters L_{Ij}, L_{Gj}
u_{Cj}	Voltages in LCL filter capacitors C_j

For the controller design, the model of the GSI system (3) can be transformed to the form in the $\alpha\beta$ -stationary

coordinate by the Clarke transformation; then, equation (3) can be rearranged as follows:

$$\begin{cases} C\dot{u}_{DC} = i_L - i_I, \\ i_{G2} = -\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_G \mathbf{T}_2 \mathbf{i}_{G2} - \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{e}_2 + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \sigma_G \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \Delta \mathbf{d}_G, \\ \dot{\mathbf{u}}_{C2} = \mathbf{T}_1 (\mathbf{C}^{-1} - \mathbf{L}_I^{-1} \mathbf{R}_C \mathbf{R}_I) \mathbf{T}_2 \mathbf{i}_2 - \mathbf{T}_1 (\mathbf{C}^{-1} - \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{R}_G) \mathbf{T}_2 \mathbf{i}_{G2} + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{T}_2 \mathbf{e}_2 \\ - \mathbf{T}_1 (\mathbf{L}_I^{-1} + \mathbf{L}_G^{-1}) \mathbf{R}_C \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \sigma_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \rho_C, \end{cases} \quad (4)$$

where $\mathbf{S}_2 = [S_\alpha, S_\beta]^T$, $\mathbf{i}_{G2} = [i_{G\alpha}, i_{G\beta}]^T$, $\mathbf{i}_2 = [i_\alpha, i_\beta]^T$, $\mathbf{u}_{C2} = [u_{C\alpha}, u_{C\beta}]^T$, and $\mathbf{e}_2 = [e_\alpha, e_\beta]^T$.

3. FOSM Controller Design for GSI

The GSI control system is presented in Figure 3. There are three feedback control loops in the system: the DC-link voltage loop, the current loop, and the LCL filter capacitor voltage loop, where * stands for the reference. To depress the effect of unbalanced grid-voltages and deal with the parameter variations, three FOSM controllers with robust switching gains are designed for the three loops. The fractional-order calculus used in three controllers is defined by α^{th} -order Caputo.

To ensure the stability of the controller, the following are the reasonable assumptions:

$$\left| {}^C D_t^{\alpha-1} i_L \right| \leq U_{DC}, \quad (5a)$$

$$\left\| {}^C D_t^\alpha (\mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2}^* + \mathbf{T}_1 \Delta \mathbf{d}_G) \right\| \leq U_G, \quad (5b)$$

$$\left\| {}^C D_t^\alpha (\mathbf{T}_1 \boldsymbol{\sigma}_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \boldsymbol{\rho}_C) \right\| \leq U_C, \quad (5c)$$

where U_{DC} , U_G , and U_C are the unknown upper bounds of the uncertainty vectors.

Figure 4 shows the flow chart of the whole algorithms of the fractional-order controllers for the grid-side converters. These three controllers are designed to guarantee that the errors of the DC voltage, current voltage, and the LCL capacitor voltage can converge to zero.

Lemma 1 (see [28]). *The nonautonomous fractional-order system ${}^C D_t^\alpha (\mathbf{x}(t)) = f(t, \mathbf{x}(t))$ is Mittag-Leffler stable at the equilibrium point $\mathbf{x}(t) = 0$, if a continuous and differentiable Lyapunov function $V(t, \mathbf{x}(t)): \mathbb{R}^+ \rightarrow \mathbb{R}$ and class-K function ϕ_i ($i = 1, 2, 3$) satisfies*

$$\begin{aligned} \phi_1(\|\mathbf{x}(t)\|) &\leq V(t, \mathbf{x}(t)) \leq \phi_2(\|\mathbf{x}(t)\|), \\ {}^C D_t^\alpha V(t, \mathbf{x}(t)) &\leq -\phi_3(\|\mathbf{x}(t)\|). \end{aligned} \quad (6)$$

3.1. DC-Link Voltage Controller. Define the DC-link voltage error as $\bar{u}_{DC} = u_{DC} - u_{DC}^*$. Then, the error dynamics of the DC-link voltage can be obtained from equation (4) as follows:

$$\dot{\bar{u}}_{DC} = \frac{(i_L - i_I)}{C}. \quad (7)$$

A fractional-order sliding manifold for the DC-link voltage error dynamic (7) is chosen as follows:

$$\mathbf{s}_{DC} = {}^C D_t^\alpha \bar{u}_{DC} + \lambda_{DC} \bar{u}_{DC}, \quad (8)$$

where $0 < \alpha < 1$ is the fractional order, and $\lambda_{DC} > 0$ is a constant.

Theorem 1. *Along the fractional-order sliding manifold (8), the DC-link voltage error dynamic (7) can be guaranteed to converge to its equilibrium point, if a fractional-order integral*

virtual control law with its robust adaptive switching gain is designed as follows:

$$i_I^* = i_{\text{leq}}^* + i_{\text{In}}^*, \quad (9a)$$

$$i_{\text{leq}}^* = C \lambda_{DC} {}^C D_t^{1-\alpha} \bar{u}_{DC}, \quad (9b)$$

$$i_{\text{In}}^* = {}^C I_t^\alpha (k_{DC} \text{sgn}(s_{DC})), \quad (9c)$$

$$k_{DC} = \mu {}^C I_t^\beta |s_{DC}|, \quad (9d)$$

where $k_{DC} > 0$ is the robust adaptive switching gain, $0 < \beta < 1$ is the fractional order of the fractional integrator, and μ is a positive constant.

Proof. Substituting the error dynamics of the DC voltage (7) into the fractional-order sliding manifold (8) yields

$$s_{DC} = \frac{{}^C D_t^{\alpha-1} (i_L - i_I^*)}{C} + \lambda_{DC} \bar{u}_{DC}. \quad (10)$$

Considering the fractional-order integral virtual control laws (9a) and (9b) gives

$$s_{DC} = \frac{{}^C D_t^{\alpha-1} (i_L - i_{\text{In}}^*)}{C}. \quad (11)$$

Consider a Lyapunov function $V_{DC} = 0.5 s_{DC}^2$. Differentiating V_{DC} gives

$$\dot{V}_{DC} = s_{DC} \dot{s}_{DC} = \frac{s_{DC} ({}^C D_t^\alpha i_L - {}^C D_t^\alpha i_{\text{In}}^*)}{C}. \quad (12)$$

Substituting the fractional-order integral virtual control law (9c) with its robust adaptive switching gain (9d) into the above yields

$$\begin{aligned} \dot{V}_{DC} &= \frac{s_{DC} ({}^C D_t^\alpha i_L - k \text{sgn}(s))}{C} \\ &\leq -\frac{|s_{DC}| \left(k - \left| {}^C D_t^\alpha i_L \right| \right)}{C} < 0, \end{aligned} \quad (13)$$

which indicates that the DC-link voltage error dynamic (7) will reach to the equilibrium point. On $s_{DC} = 0$, from (8), it can be seen that the system will behave in the identical fashion as

$${}^C D_t^\alpha \bar{u}_{DC} + \lambda_{DC} \bar{u}_{DC} = 0. \quad (14)$$

Based on Lemma 1, it can be obtained that when $s_{DC} = 0$, the DC-link voltage error dynamic (7) will converge to its equilibrium points along the fractional-order sliding manifold. This completes the proof. And the block diagram of the control algorithm of DC-link voltage is shown in Figure 5.

Based on equation (4), the following power equation set can be obtained:

$$\dot{\bar{i}}_{G2} = -\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_G \mathbf{T}_2 \mathbf{i}_{G2} - \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{e}_2 + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \Delta \mathbf{d}_G - \dot{i}_{G2}^* \quad (17)$$

A fractional-order sliding manifold for current error dynamic (17) is chosen as follows:

$$\mathbf{s}_{G2} = {}^C D_t^{\alpha} \bar{i}_{G2} + \boldsymbol{\lambda}_{G2} \bar{i}_{G2}, \quad (18)$$

where $0 < \alpha < 1$ is the fractional order, and $\boldsymbol{\lambda}_{G2} = \text{diag}(\lambda_{G2\alpha}, \lambda_{G2\beta}, \lambda_{G2\gamma}) > 0$ are all constants.

Theorem 2. Along the fractional-order sliding manifold (18), the current error dynamic (17) can be guaranteed to converge to its equilibrium point, if a fractional-order integral virtual control law with its robust adaptive switching gain is designed as follows:

$$\mathbf{u}_{C2}^* = \mathbf{u}_{C2eq}^* + \mathbf{u}_{C2n}^*, \quad (19a)$$

$$\mathbf{u}_{C2eq}^* = (\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2)^{-1} ({}^C D_t^{1-\alpha} (\boldsymbol{\lambda}_{G2} \bar{i}_{G2}) + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_G \mathbf{T}_2 \mathbf{i}_{G2} + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{e}_2 + \dot{i}_{G2}^*), \quad (19b)$$

$$\mathbf{u}_{C2n}^* = -(\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2)^{-1} {}^C I_t^\alpha \left(\frac{k_{C2} \mathbf{s}_{G2}}{\|\mathbf{s}_{G2}\|} \right), \quad (19c)$$

$$k_{C2n} = \mu {}^C I_t^\beta \|\mathbf{s}_{G2}\|, \quad (19d)$$

where $k_{C2n} > 0$ is the robust adaptive switching gain, $0 < \beta < 1$ is the fractional order of the fractional integrator, and $\mu > 0$ is a constant.

Proof. Substituting the current error dynamic (17) into the fractional-order sliding manifold (18) yields

$$\mathbf{s}_{G2} = {}^C D_t^{\alpha-1} \left(-\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_G \mathbf{T}_2 \mathbf{i}_{G2} - \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{e}_2 + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \Delta \mathbf{d}_G - \dot{i}_{G2}^* \right) + \boldsymbol{\lambda}_{G2} \bar{i}_{G2}. \quad (20)$$

Considering the fractional-order integral virtual control laws (19a) and (19b) gives

$$\mathbf{s}_{G2} = {}^C D_t^{\alpha-1} (\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{u}_{C2n}^* + \mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2}^* + \mathbf{T}_1 \Delta \mathbf{d}_G). \quad (21)$$

Differentiating the above with respect to time t gives

$$\dot{\mathbf{s}}_{G2} = {}^C D_t^\alpha (\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{u}_{C2n}^* + \mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2}^* + \mathbf{T}_1 \Delta \mathbf{d}_G). \quad (22)$$

Consider a Lyapunov function $V_{G2} = 0.5 \mathbf{s}_{G2}^T \mathbf{s}_{G2}$. Differentiating V_{G2} gives

$$\dot{V}_{G2} = \mathbf{s}_{G2}^T \dot{\mathbf{s}}_{G2} = \mathbf{s}_{G2}^T {}^C D_t^\alpha (\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{T}_2 \mathbf{u}_{C2n}^* + \mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2}^* + \mathbf{T}_1 \Delta \mathbf{d}_G). \quad (23)$$

Substituting the fractional-order integral virtual control law (19c) with its robust adaptive switching gain (19d) into the above yields

$$\begin{aligned} \dot{V}_{G2} &\leq \mathbf{s}_{G2}^T \left(-k_{G2} \frac{\mathbf{s}_{G2}}{\|\mathbf{s}_{G2}\|} + {}^C D_t^\alpha (\mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2}^* + \mathbf{T}_1 \Delta \mathbf{d}_G) \right) \\ &\leq -\|\mathbf{s}_{G2}\| \left(k_{G2} - \left\| {}^C D_t^\alpha (\mathbf{T}_1 \boldsymbol{\sigma}_G \mathbf{T}_2 \mathbf{u}_{C2}^* + \mathbf{T}_1 \Delta \mathbf{d}_G) \right\| \right) < 0, \end{aligned} \quad (24)$$

which means that the error dynamics of the current (17) will reach to zero. On $\mathbf{s}_{G2} = 0$, it can be seen from (18) that the system will behave according to the following equation:

$${}^C D_t^{\alpha} \bar{i}_{G2} + \boldsymbol{\lambda}_{G2} \bar{i}_{G2} = 0. \quad (25)$$

Based on Lemma 1, it can be obtained that when $\mathbf{s}_{G2} = 0$, the current error dynamic (26) will converge to its equilibrium points along the fractional-order sliding manifold. This completes the proof.

3.3. Filter Capacitor Voltage Controller. The switching control signal $\mathbf{S}_2$ in the system (4) is designed to force the actual voltage to track its virtual control reference. Define the current error as $\bar{\mathbf{u}}_{C2} = \mathbf{u}_{C2} - \mathbf{u}_{C2}^*$. Then, filter capacitor voltage error dynamics are as follows:

$$\begin{aligned} \dot{\bar{\mathbf{u}}}_{C2} &= \mathbf{T}_1 (\mathbf{C}^{-1} - \mathbf{L}_I^{-1} \mathbf{R}_C \mathbf{R}_I) \mathbf{T}_2 \mathbf{i}_2 - \mathbf{T}_1 (\mathbf{C}^{-1} - \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{R}_G) \mathbf{T}_2 \mathbf{i}_{G2} \\ &\quad + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{T}_2 \mathbf{e}_2 - \mathbf{T}_1 (\mathbf{L}_I^{-1} + \mathbf{L}_G^{-1}) \mathbf{R}_C \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{u}_{DC} \mathbf{T}_2 \mathbf{S}_2 \\ &\quad + \mathbf{T}_1 \boldsymbol{\sigma}_C \mathbf{u}_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \boldsymbol{\rho}_C - \dot{\mathbf{u}}_{C2}^*. \end{aligned} \quad (26)$$

A fractional-order sliding manifold for filter capacitor voltage error dynamic (17) is chosen as

$$\mathbf{s}_{C2} = {}^C D_t^\alpha \bar{\mathbf{u}}_{C2} + \boldsymbol{\lambda}_{C2} \bar{\mathbf{u}}_{C2}, \quad (27)$$

where $0 < \alpha < 1$ is the fractional order, and $\boldsymbol{\lambda}_{C2} = \text{diag}(\lambda_{C2\alpha}, \lambda_{C2\beta})$, $\lambda_{C2\alpha}, \lambda_{C2\beta} > 0$ are all constants.

Theorem 3. Along the fractional-order sliding manifold (27), the LCL capacitor voltage error dynamics (26) can be guaranteed to converge to its equilibrium point, if a fractional-order integral actual control law with its robust adaptive switching gain is designed as follows:

$$\mathbf{S}_2 = \mathbf{S}_{2eq} + \mathbf{S}_{2n}, \quad (28a)$$

$$\begin{aligned} \mathbf{S}_{2eq} = & \left(u_{DC} \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{T}_2 \right)^{-1} \left[-\mathbf{T}_1 \left(\mathbf{C}^{-1} - \mathbf{L}_I^{-1} \mathbf{R}_C \mathbf{R}_I \right) \mathbf{T}_2 \mathbf{i}_2 - \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{T}_2 \mathbf{e}_2 + \dot{\mathbf{u}}_{C2}^* \right. \\ & \left. + \mathbf{T}_1 \left(\mathbf{C}^{-1} - \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{R}_G \right) \mathbf{T}_2 \mathbf{i}_{G2} + \mathbf{T}_1 \left(\mathbf{L}_I^{-1} + \mathbf{L}_G^{-1} \right) \mathbf{R}_C \mathbf{T}_2 \mathbf{u}_{C2} - {}^C D_t^{1-\alpha} \boldsymbol{\lambda}_{C2} \bar{\mathbf{u}}_{C2} \right], \end{aligned} \quad (28b)$$

$$\mathbf{S}_{2n} = - \left(u_{DC} \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{T}_2 \right)^{-1} {}^C I_t^\alpha \left(\frac{k_2 \mathbf{s}_{C2}}{\|\mathbf{s}_{C2}\|} \right), \quad (28c)$$

$$k_2 = \mu_{2n} {}^C I_t^\beta \|\mathbf{s}_{C2}\|. \quad (28d)$$

Proof. Substituting the LCL capacitor voltage error dynamic (26) into the fractional-order sliding manifold (27) yields

$$\mathbf{s}_{C2} = {}^C D_t^{\alpha-1} \left(\begin{aligned} & \mathbf{T}_1 \left(\mathbf{C}^{-1} - \mathbf{L}_I^{-1} \mathbf{R}_C \mathbf{R}_I \right) \mathbf{T}_2 \mathbf{i}_2 - \mathbf{T}_1 \left(\mathbf{C}^{-1} - \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{R}_G \right) \mathbf{T}_2 \mathbf{i}_{G2} \\ & + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C \mathbf{T}_2 \mathbf{e}_2 - \mathbf{T}_1 \left(\mathbf{L}_I^{-1} + \mathbf{L}_G^{-1} \right) \mathbf{R}_C \mathbf{T}_2 \mathbf{u}_{C2} + \mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 \boldsymbol{\lambda}_{C2} \bar{\mathbf{u}}_{C2} \\ & + \mathbf{T}_1 \sigma_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \rho_C - \dot{\mathbf{u}}_{C2}^* + {}^C D_t^{1-\alpha} \end{aligned} \right). \quad (29)$$

Considering the fractional-order integral virtual control laws (28a) and (28b) gives

$$\mathbf{s}_{C2} = {}^C D_t^{\alpha-1} \left(\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C u_{DC} \mathbf{T}_2 \mathbf{S}_{2n} + \mathbf{T}_1 \sigma_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \rho_C \right). \quad (30)$$

Differentiating the above with respect to time t gives

$$\dot{\mathbf{s}}_{C2} = {}^C D_t^\alpha \left(\mathbf{T}_1 \mathbf{L}_G^{-1} \mathbf{R}_C u_{DC} \mathbf{T}_2 \mathbf{S}_{2n} + \mathbf{T}_1 \sigma_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \rho_C \right). \quad (31)$$

Consider Lyapunov function $V_{C2} = 0.5 \mathbf{s}_{C2}^T \mathbf{s}_{C2}$. According to the fractional-order integral virtual control law (24c) with its robust adaptive switching gain (24d) into the above yields

$$\dot{V}_{C2} = \mathbf{s}_{C2} \dot{\mathbf{s}}_{C2} = -\|\mathbf{s}_{C2}\| \left(k_2 - \left\| {}^C D_t^\alpha \left(\mathbf{T}_1 \sigma_C u_{DC} \mathbf{T}_2 \mathbf{S}_2 + \mathbf{T}_1 \rho_C \right) \right\| \right) < 0, \quad (32)$$

which indicates that the error dynamics of filter capacitor (26) will reach to zero. On $\mathbf{s}_{C2} = 0$, it can be seen from (27) that the system will behave in the following identical fashion:

$${}^C D_t^\alpha \bar{\mathbf{u}}_{C2} + \boldsymbol{\lambda}_{C2} \bar{\mathbf{u}}_{C2} = 0. \quad (33)$$

Based on Lemma 1, it can be obtained that when $\mathbf{s}_{C2} = 0$, the current error dynamics (25) will converge to its equilibrium points along the fractional-order sliding manifold. This completes the proof.

4. Simulations

In order to verify the effectiveness of the proposed method, some simulations are carried out using Matlab. Table 2 provides the parameters of GSI. The parameters of the controllers are as follows: $k = 60$, $k_d = 100$, $\eta = 100$, $\mathbf{C} = \text{diag}(80, 80)$, $\mu = \text{diag}(3/5, 3/5)$, and $\mathbf{T} = \text{diag}(100, 100)$.

The simulations of PI and FOSMC are shown in Figure 6. It can be seen that the DC-link voltage is quickly converged to 700 V and stays around at 700 V under both control strategies. Compared with PI, the proposed FOSMC method can effectively improve the dynamic response of DC-link voltage and reduce the peak value and settling time of DC-link voltage, which enhance the disturbance rejection. From Figures 6(b) and 6(c), the instantaneous active power and reactive power are controlled to converge to their reference value. Therefore, it can be concluded that the control objective has been achieved.

The simulations of the current controller are shown in Figure 7. In Figure 7(a), it can be seen that the current can track the reference quickly and precisely. In Figure 7(b), u_{ca}^* and u_{cb}^* have a high degree of sinusoidal, which means the harmonics barely exist in the voltage. Meanwhile, total harmonic distortion (THD) of the output currents of GSI are shown in Figure 7(c) which is a measurement of the harmonic distortion present. It can be known from Figure 7(c) that THD is extremely closed to zero, which means that the

TABLE 2: Parameters of GSI.

Parameter	Value (kW)
Rated power	2
DC-link voltage	700
DC-link capacitor	6800
Filter inductor	1.5
Phase resistor	0.4
Phase voltage of the grid	311

(a)

(b)

(c)

FIGURE 6: The simulation results of the grid-side converter. (a) u_{DC} (SMC)(V)/ u_{DC} (PI)(V), (b) P (SMC)(KW)/ P (PI)(KW), and (c) Q (SMC)(KW)/ Q (PI)(KW).

(a)

FIGURE 7: Continued.

FIGURE 7: The simulation results of current controller. (a) Current (A), (b) controller, and (c) THD (i_{abc}).

FIGURE 8: The simulation results for the case where the three filter inductances change their values precisely and quickly. Compared with the traditional SMC, the proposed FOSMC method can eliminate chattering, and the adaptive gain can avoid overestimating the upper boundary of the parameter variations and external disturbance to obtain a better control performance. The simulation results validate the proposed method. (a) P (KW)/ Q (KVar), (b) i_{abc} (A), and (c) $e_a(A)/i_{abc}$ (A).

FIGURE 9: The simulation results for the case where the three resistors change their values. (a) P (KW)/ Q (KVar), (b) i_{abc} (A), and (c) $e_a(A)/i_{abc}$ (A).

FIGURE 10: Continued.

FIGURE 10: The simulation results when resistors and inductance are all different. (a) $P(KW)/Q(KVar)$, (b) i_{abc} (A), and (c) $e_a(A)/i_{abc}$ (A).

harmonic in the current is eliminated and the quality of grid energy can be guaranteed.

Figures 8–10 show the simulation results when the filter inductances and resistors are changed. Figure 8 depicts the case that values of filter inductances are changed and the resistors keep unchanged. The filter inductances are changed from 1.5 mH to 2 mH. Figure 9 depicts the case that values of the resistors are changed from 0.4 Ω to 0.6 Ω , and three filter inductances keep unchanged. Figure 10 depicts the case for varying resistors and varying filter inductances. The values of resistors and the filter inductances are changed to 1.5 Ω and 2 mH. It can be seen that the changes are not obvious under FOSMC in the view of parameter change, which means that the proposed method has a strong robustness against the parameter variation. From the simulation results in Figures 8(c)–10(c), it can be concluded that the voltage and current have the same phase angle, which means that the unit power factor can be guaranteed. The basics of fractional calculus are shown in appendix.

5. Conclusions

In this study, a novel fractional-order sliding mode control strategy with robust switching gain is proposed to achieve reliable and high quality of wind-powered microgrid systems. Fractional-order sliding mode controllers generate continuous control signals and regulate the outer DC-link voltage loop and current loop in GSI. Considering the parameters variation and external disturbance, the outer DC-link voltage and inner current can track their references.

Appendix

A. Basics of Fractional Calculus

The theory of fractional-order calculus is closely related to the traditional integral order calculus. Grünwald–Letnikov (GL), Riemann–Liouville (RL), and Caputo definitions, shown as follows, are three most definitions of the fractional-order differentiation used in literature and practical applications.

Definition 1 (see [29]). The α^{th} -order GL fractional derivative for a function $f(t): R^+ \rightarrow R$ is defined as

$${}_{t_0}^{GL}D_t^\alpha f(t) = \lim_{h \rightarrow 0} h^{-\alpha} \sum_{m=0}^{[(t-t_0)/h]} (-1)^m \binom{\alpha}{m} f(t-mh), \quad (\text{A.1})$$

where $\alpha \in R$ is the fractional order, t_0 and t are the limits of the operation, $[\cdot]$ denotes the integer value function, and h is the time increment, and

$$\binom{\alpha}{m} = \frac{\Gamma(\alpha+1)}{\Gamma(m+1)\Gamma(\alpha-m+1)} = \frac{\alpha!}{m!(\alpha-m)!}, \quad (\text{A.2})$$

where $\Gamma(\cdot)$ is the well-known Euler gamma function.

Definition 2 (see [29]). The α^{th} -order RL fractional derivative for a function $f(t): R^+ \rightarrow R$ is defined as

$${}_{t_0}^{RL}D_t^{-\alpha} f(t) = \frac{1}{\Gamma(u-\alpha)} \frac{d^u}{dt^u} \int_{t_0}^t \frac{f(\tau)}{(t-\tau)^{\alpha-u+1}} d\tau, \quad (\text{A.3})$$

where $u-1 < \alpha \leq u$, $u \in N$.

Definition 3 (see [29]). The α^{th} -order Caputo fractional derivative for a continuous function $f(t): R^+ \rightarrow R$ is defined as

$${}_t^CD_t^\alpha f(t) = \frac{1}{\Gamma(u-\alpha)} \int_{t_0}^t \frac{f^{(u)}(\tau)}{(t-\tau)^{\alpha-u+1}} d\tau, \quad (\text{A.4})$$

where u is defined in (2). Meanwhile, the α^{th} -order Caputo fractional integration for a function $f(t): R^+ \rightarrow R$ is defined as

$${}_t^CI_t^\alpha f(t) = {}_t^CD_t^{-\alpha} f(t) = \frac{1}{\Gamma(\alpha)} \int_{t_0}^t \frac{f(\tau)}{(t-\tau)^{1-\alpha}} d\tau. \quad (\text{A.5})$$

Data Availability

The data used in the simulations are carried out using Matlab-Simulink and are available from the corresponding author upon request.

Conflicts of Interest

The authors declare that they have no conflicts of interest.

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