Super-Twisting Sliding Mode Direct Power Control of Brushless Doubly Fed Induction Generator

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Abstract- This paper proposes and implements a supertwisting sliding mode direct power control (SSM-DPC) strategy for brushless doubly fed induction generator (BDFIG). Direct power control has fast and robust response under transient conditions, however, suffers from active and reactive power ripples and current distortions, which degrades the quality of output power. In contrast, vector control has good steady-state current harmonic spectra, however, not robust to machine parameters variations thus needs phase locked loop for synchronous coordinate transformations. The SSM-DPC strategy controls active and reactive power directly without the need of phase locked loop. Moreover, its transient performance is similar to DPC and its steadystate performance is the same as vector control. The proposed controller is robust to uncertainties toward parameter variations and achieves constant converter switching frequency, by using space vector modulation. Simulation and experimental results of 2 MW and 3 kW laboratory scale BDFIG are provided and compared with those of integral-sliding mode and direct power control to validate the effectiveness, correctness and the robustness of the proposed strategy.

Index Terms— Brushless machines, mathematical model, power control, super-twisting sliding mode control, wind power generation.

NOMENCLATURE

 $\vec{v}, \vec{i}, \vec{\lambda}$ Space-vector of voltage, current and flux

- \tilde{U} , **U** Phasor and voltage vector
- $\omega_r, \omega_c, \omega_e$ Electrical angular speed of rotor, controlwinding and grid

R, *L* Resistance and inductance

- L_{mp} , L_{mc} Magnet. inductance of power-winding and control-winding
- *P*, *Q* Active and reactive power
- S_{b} , V_{b} , I_{b} Base value of power, voltage and current

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S	Sliding surface
sgn(S)	Switch function
W	Candidate Lyapunov function
A, B	Positive gains of the switching control laws
Η	System disturbance
σ	Leakage coefficient, $\sigma = 1 - L_m^2 / (L_p' L_c')$
<i>x</i> ₁ , <i>x</i> ₂ , <i>u</i> , <i>y</i>	Position, velocity, control force and output
f	Disturbance force (i.e., dry and viscous
	friction)
р	Differential operator (d/dt)

- *p*, *r*, *c* Subscripts denoting power-winding, rotor and control-winding quantities
- *a*, *b*, *c* Subscripts denoting phase *a*,*b*,*c* quantities
- *s, r, c* Superscript denoting stationary, rotor and control-winding reference frame
 - Superscript denoting reference value

I. INTRODUCTION

B rushless doubly-fed induction generators (BDFIGs) have shown promising prospect as an alternative to the conventional doubly-fed induction generator (DFIG) in commercial wind turbine applications especially for offshore wind farms, since it does not require brushes and slip-rings (as the name implies) which results in higher reliability and lower maintenance cost operations [1-3]. Another member of the doubly-fed machine family, BDFIG inherits all the advantages of the conventional DFIG and has been regarded as a viable replacement [3-6]. The BDFIG with nested-loop/wound rotor consists of power-stator, rotor, and control-stator windings, as depicted in Fig. 1. The power winding (PW) is directly connected to the grid, while control winding (CW) is supplied via a fractionally-rated dual-bridge converter in 'back-to-back' configuration for bi-directional power flow [4]. Controller designs for the BDFIG are somewhat difficult to implement which could be down to the model complexities yielding to complex structure caused by the inner rotor loop (presence of the rotor resistance), heavy parameter dependence, large number of degrees of freedom and multiple-input/multipleoutput [5], thus appropriate BDFIG control strategies are very demanding but highly required.

Vector control (VC) is a common and practical scheme pertinent to induction machines is also applied to the BDFIG [6, 7]. Sensitivity to parameter variation, detuning effect of PI controllers, deficiency during grid disturbances and necessity of position sensors are accounted as the main drawbacks of

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Fig. 1. A generic conceptual diagram of the BDFIG system for adjustable speed constant frequency grid-connected applications.

VC. Efforts towards prominent brushless doubly fed reluctance generator (BDFRG) have been carried out to conquer these drawbacks [8, 9], which could potentially be applicable universally to the BDFIG with minor adjustments. As an alternative to VC, the direct torque control (DTC) [10] and direct power control (DPC) [11], are proposed for the BDFIG. Such strategies have fast dynamic response, simple implementation and robustness. The DTC/DPC provides direct regulation of the machines torque/power by selecting proper voltage vectors from the lookup-tables. However, converterswitching frequency varies with operating conditions, which results in large torque/power ripples and current distortions. To improve such shortcomings, while keeping the advantages of DTC/DPC over the VC, the sliding-mode control (SMC) is proposed for induction machines [12] and DFIGs [13]. The SMC is a robust control method for nonlinear systems with large perturbations and parameter variations [14]. The SMC has been proposed to control the DFIG under non-ideal grid voltage conditions [15, 16].

The application of SMC for BDFIG in [17], illustrates the robustness of this method against parameter variations. Advantages of the SMC are the simple implementation, disturbance rejection, strong robustness, and fast dynamic responses. However, its stabilization time is not necessarily finite and undesired chattering appears on the controlled states. To overcome these drawbacks, forms of internal-SMC are proposed for DFIG, which delivers smooth active power to an unbalanced grid with minimized torque/power ripple [18]. However, this method requires a discrete control action, which requires high switching frequency for applying the control output via the inverter. Alternatively, the second-order sliding mode does not have a discrete output and mitigates the chattering [19, 20], in the presence of disturbances and model inaccuracies. Thus, the second-order SMC suffers from complex mathematical calculations and its implementation is difficult when the state variables are increased [14]. The integral-SMC for BDFIG is improved by means of boundary layer and feed-forward terms but make the controller sensitive to parameter variations, while the control output is not discrete and provides fast dynamic response [21].

Super-twisting is a recently developed theory in the SMC design, which is proven to be efficient for electromechanical systems [22]. In the light of the latest development in the field, this paper proposes a super-twisting sliding-mode direct power control (SSM-DPC) for controlling active and reactive power of BDFIG, without using inner current loop regulator and

phase-lock loop (PLL). The main advantage of super-twisting SMC is that it only requires a sliding surface functions (S) and not its derivative (dS/dt) [23]. This controller mitigates the chattering that exists in most SMC strategies, while keeping SMC excellent static/dynamic performances and robustness, in case of uncertainties and parameter variations. The transient performance of the proposed strategy is similar to DPC and its steady-state performance is comparable to VC.

Rest of the paper is organized as follows. Dynamic modelling of the BDFIG and configuration is presented in Section II. Super-twisting SMC technique is introduced, whilst control objectives of sliding variables, and controller design of the proposed SSM-DPC are constructed in Section III. Comparative qualitative and quantitative results are presented in Section IV, which confirms the feasibility and efficiency of the proposed controller. Simulation of 2 MW and experimental verification of a 3 kW BDFIG are presented and outlined in Section V. Finally, Section VI draws conclusions.

II. BDFIG DYNAMIC MODEL AND CONFIGURATION

The BDFIG consists of a complicated mathematical model and structural arrangement due to existing of voltage sources and resistances in its rotor loop. In order to design a controller for the BDFIG, the model complexity needs simplifying. In [10], a dynamic model of BDFIG is presented in rotor reference frame, which resembles close alignments to DFIG equivalent circuit depicted in Fig. 2.



Fig. 2. Representation of the BDFIG presented model.

The dynamic model equations of this BDFIG model are expressed in space-vector form as follows [10]:

$$\vec{v}_c^r = r_c \, \vec{i}_c^r + \, \mathcal{P} \vec{\lambda}_c^r + j \, (\omega_r - \, \omega_c) \vec{\lambda}_c^r \tag{1}$$

$$\vec{v}_p^r = r_p \, \vec{i}_p^r + p \, \vec{\lambda}_p^r + j \, \omega_r \, \vec{\lambda}_p^r \tag{2}$$

$$\vec{\lambda}_c^r = \left(L_{lc} + \frac{L_{mc}}{L_r}L_{lr}\right)\vec{i}_c^r + \frac{L_{mc}L_{mp}}{L_r}\vec{i}_p^r + \frac{L_{mc}r_r}{L_r}\int\vec{i}_c^r dt$$
(3)

$$\vec{\lambda}_p^r = \left(L_{lp} + \frac{L_{mp}}{L_r}L_{lr}\right)\vec{i}_p^r + \frac{L_{mc}L_{mp}}{L_r}\vec{i}_c^r + \frac{L_{mp}r_r}{L_r}\int\vec{i}_p^r\,dt \tag{4}$$

Due to the effect of integrators in the flux equations in (3) and (4), cannot be transformed to other reference frames as per model representation in [10], therefore, this model does not simplify the controller design. Since, the variation range of rotor-speed/CW-frequency in BDFIG is under $\pm 30\%$ of the synchronous speed ($\omega_e \approx 100\pi$), the variation range of the rotor frequency is $20\pi < \omega_r < 50\pi$ [10]. Therefore,

$$r_r L_{mp}/(L_{mc} + L_{mp}) \ll (\omega_e - \omega_r)L'_{lp},$$

$$r_r L_{mc}/(L_{mc} + L_{mp}) \ll (\omega_e - \omega_r)L'_{lc}$$

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Consequently, resistances $r_r L_{mp} / (L_{mc} + L_{mp})$ and $r_r L_{mc} / (L_{mc} + L_{mp})$ are neglected. Neglecting those resistances makes the BDFIG model analogous to DFIG, since the proposed controller is based on SMC, it is robust to the error due to this simplification. The model is further modified by transferring to the CW reference frame depicted in Fig. 3, which its dynamic equations are as follows:

$$\vec{v}_c = r_c \, \vec{i}_c^c + \, \mathscr{p} \vec{\lambda}_c^c \tag{5}$$

$$\vec{v}_p = r_p \, \vec{i}_p^c + \mathcal{p} \, \vec{\lambda}_p^c + j \, \omega_c \, \vec{\lambda}_p^c \tag{6}$$

$$\vec{\lambda}_c = L'_c \, \vec{i}_c + L_m \, \vec{i}_p \tag{7}$$

$$\hat{\lambda}_p = L'_p \hat{i}_p + L_m \, \hat{i}_c \tag{8}$$

where, $L'_{lc} \triangleq L_{lc} + L_{mc}L_{lr}/L_r$, $L'_{lp} \triangleq L_{lp} + L_{mp}L_{lr}/L_r$, $L_m \triangleq L_{mc}L_{mp}/L_r$, $L'_p = L'_{lp} + L_m$ and $L'_c = L'_{lc} + L_m$.



Fig. 3. Reduced dq-model of the BDFIG in the CW reference frame.

According to (7) and (8), the relation between PW and CW fluxes is presented as

$$\vec{\lambda}_p = \frac{L_m}{L'_c} \, \vec{\lambda}_c + \sigma L'_p \, \vec{i}_p \tag{9}$$

The model (Fig. 3) is appropriate for implementing control strategies like vector control, direct torque/power control and sliding mode control for the BDFIG.

III. SUPER-TWISTING SMC AND PROPOSED SSM-DPC STRATEGY

A. Super-Twisting Sliding Mode Control Mechanism

The main drawbacks of the traditional SMC are chattering effect and discontinuous high-frequency switching control which is impractical. To overcome these problems, supertwisting controller is used. A single-dimensional motion of a unit mass system (10) is employed [24]:

$$\begin{cases} \dot{x}_1 = x_2 & x_1(0) = x_{10} \\ \dot{x}_2 = u + f(x_1, x_2, t) & x_2(0) = x_{20} \\ y = x_1 \end{cases}$$
(10)

It is desired to reduce the order of this system to one, by defining the output tracking error: $e = y_c(t) - y(t)$, where y_c is desired output. A sliding surface is selected as:

$$S = \dot{e} + ke \qquad k > 0 \tag{11}$$

When the sliding variable *S* approaches zero, *y*(*t*) reaches $y_c(t)$. If the disturbance magnitude has an upper-boundary, i.e, $|f| \le L$, *u* is designed as follows to drive $S \rightarrow 0$ in finite time and keep it at zero.

$$\begin{cases} u = w_1 + w_2 \\ w_1 = b \int sgn(S) dt \\ w_2 = c|S|^{1/2} sgn(S) \end{cases}$$
(12)

where, $c = 1.5\sqrt{L}$ and b = 1.1 L. The super-twisting control (12) is continuous, since both terms are continuous [24]. The discontinues high-frequency switching term sgn(s) is 'hidden' under the integral. Figure 4 shows the block diagram of super-twisting SMC, where, w_1 compensates the disturbance f in finite time, and w_2 forces S to become zero. This means that both e and \dot{e} become zero, and the system trajectory stays on the surface thereafter. In other words, the control u drives e to zero, i.e. y approaches to y_c , in the presence of the bounded disturbance f.



Fig. 4. Schematic control diagram of the super-twisting SMC.

B. Proposed Super-twisting Sliding Mode DPC Strategy

The target of the proposed super-twisting SMC-DPC is to control PW active-power (P_p) and reactive-power (Q_p) independently and directly. In the proposed strategy, the CW voltage components, u_{dc} and u_{qc} , are the inputs to the system, whilst the power components P_p and Q_p are the outputs of the system, which are calculated as follows:

$$\begin{cases} P_p = -\frac{3}{2} Re\{\vec{u}_p^c \vec{i}_p^{c^*}\} \\ Q_p = -\frac{3}{2} Im\{\vec{u}_p^c \vec{i}_p^{c^*}\} \end{cases}$$
(13)

where, \vec{u} is space-vector of voltage and quantity P_p is positive if BDFIG is operating in the generator mode.

1) Sliding Surface

In order to minimize the steady-state error and make the transient response faster, instantaneous power errors are chosen as sliding surface:

$$\mathbf{S} = \begin{bmatrix} S_1 & S_2 \end{bmatrix}^T \tag{14}$$

where, $S_1 = P_{ref} - P_p$ and $S_2 = Q_{ref} - Q_p$.

The surface S = 0 represent the accurate power tracking. When the system states reach and stay on the sliding surface, S = dS/dt = 0.

2) SMC Law

To derive conditions on the control law that will drive the system-states to the sliding surfaces, a candidate Lyapunov function is introduced:

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$$\boldsymbol{W} = \frac{1}{2}\boldsymbol{S}^{T}\boldsymbol{S} > 0 \tag{15}$$

Time derivative of the above Lyapunov function is calculated

$$\frac{dW}{dt} = \frac{1}{2} \left(\boldsymbol{S}^T \frac{d\boldsymbol{S}}{dt} + \boldsymbol{S} \frac{d\boldsymbol{S}^T}{dt} \right) = \boldsymbol{S}^T \frac{d\boldsymbol{S}}{dt}$$
(16)

According to (14), time derivative of S is given by:

$$\frac{dS}{dt} = \frac{d}{dt} \begin{bmatrix} S_1 \\ S_2 \end{bmatrix} = -\frac{d}{dt} \begin{bmatrix} P_p \\ Q_p \end{bmatrix}$$
(17)

To calculate time derivative of P_p and Q_p , it is only required to calculate time derivative of $\vec{u}_p^c \vec{i}_p^{c^*}$ in (13):

$$\frac{d(\vec{u}_{p}^{c}\vec{i}_{p}^{c^{*}})}{dt} = \frac{d\vec{u}_{p}^{c}}{dt}\vec{i}_{p}^{c^{*}} + \vec{u}_{p}^{c}\frac{d\vec{i}_{p}^{c^{*}}}{dt}$$
(18)

The PW voltage magnitude is considered constant, since it is connected to the grid, therefore

$$\vec{u}_p^c = \tilde{U}_p e^{j(\omega_e - \omega_c)t} \tag{19}$$

The time derivative of PW voltage is:

$$\frac{d\vec{u}_p^c}{dt} = j(\omega_e - \omega_c)\tilde{U}_p e^{j(\omega_e - \omega_c)t} = j(\omega_e - \omega_c)\vec{u}_p^c$$
(20)

Time derivative of the PW current is obtained from (5), (6) and (7),

$$\frac{d\vec{l}_p^c}{dt} = -\frac{L_m}{\sigma L_p' L_c'} \vec{u}_c^c + \left(\frac{r_c L_m}{\sigma L_p' L_c'} - j \frac{\omega_c L_m}{\sigma L_p'}\right) \vec{i}_c^c - \left(\frac{r_p}{\sigma L_p'} + j \frac{\omega_c}{\sigma}\right) \vec{i}_p^c + \frac{\vec{u}_p^c}{\sigma L_p'} \tag{21}$$

Equations (20) and (21) are substituted into (18), thus,

$$\frac{d\left(\vec{u}_{p}^{c}\vec{i}_{p}^{c^{*}}\right)}{dt} = \left[-\frac{r_{p}}{\sigma L_{p}'} + j\left(\omega_{e} - \omega_{c} + \frac{\omega_{c}}{\sigma}\right)\right]\vec{u}_{p}^{c}\vec{i}_{p}^{c^{*}} - \frac{L_{m}}{\sigma L_{p}'L_{c}'}\vec{u}_{p}^{c}\vec{u}_{c}^{c^{*}} + \left(\frac{r_{c}L_{m}}{\sigma L_{p}'L_{c}'} + j\frac{\omega_{c}L_{m}}{\sigma L_{p}'}\right)\vec{u}_{p}^{c}\vec{i}_{p}^{c^{*}} + \frac{|u_{p}^{c}|^{2}}{\sigma L_{p}'}$$

$$(22)$$

Equation (22) is decomposed to dq by using (13),

$$\frac{d\left(\vec{u}_{p}^{c}\vec{l}_{p}^{c}\right)^{*}}{dt} = -\frac{2}{3} \left[-\frac{r_{p}}{\sigma L_{p}'} + j\left(\omega_{e} - \omega_{c} + \frac{\omega_{c}}{\sigma}\right) \right] P_{p} \\
+ \frac{2}{3} \left(\omega_{e} - \omega_{c} + \frac{\omega_{c}}{\sigma} + j\frac{r_{p}}{\sigma L_{p}'}\right) Q_{p} \\
- \frac{L_{m}}{\sigma L_{p}' L_{c}'} \left(u_{pq}^{c} u_{cd}^{c} + u_{pq}^{c} u_{cq}^{c}\right) \\
- j\frac{L_{m}}{\sigma L_{p}' L_{c}'} \left(u_{pq}^{c} u_{cd}^{c} - u_{pd}^{c} u_{cq}^{c}\right) \\
+ \left(\frac{r_{c} L_{m}}{\sigma L_{p}' L_{c}'} + j\frac{\omega_{c} L_{m}}{\sigma L_{p}'}\right) \left(u_{pq}^{c} i_{cd}^{c} + u_{pq}^{c} i_{cq}^{c}\right) \\
+ \left(-\frac{\omega_{c} L_{m}}{\sigma L_{p}'} + j\frac{r_{c} L_{m}}{\sigma L_{p}' L_{c}'}\right) \left(u_{pq}^{c} i_{cd}^{c} - u_{pd}^{c} i_{cq}^{c}\right) \\
+ \frac{u_{pd}^{c}^{2} + u_{pq}^{2}^{2}}{\sigma L_{p}'} \tag{23}$$

According to (13), (17) and (23), time-derivative of sliding surface is obtained as:

$$\frac{dS}{dt} = F + DU_{cdq}^c \tag{24}$$

where,

$$\mathbf{F} = \frac{3}{2} \frac{L_m}{\sigma L'_p} \begin{bmatrix} \frac{r_c}{L'_c} u_{pd}^c - \omega_c u_{pq}^c & \frac{r_c}{L'_c} u_{pq}^c + \omega_c u_{pd}^c \\ \frac{r_c}{L'_c} u_{pq}^c + \omega_c u_{pd}^c & -\frac{r_c}{L'_c} u_{pd}^c + \omega_c u_{pq}^c \end{bmatrix} \begin{bmatrix} i_{cd}^c \\ i_{cq}^c \end{bmatrix} + \\ \frac{3}{2} \frac{1}{\sigma L'_p} \begin{bmatrix} u_{pd}^c ^2 + u_{pq}^c ^2 \\ 0 \end{bmatrix} + \\ \begin{bmatrix} \frac{r_p}{\sigma L'_p} & \omega_e - \left(1 - \frac{1}{\sigma}\right) \omega_c \\ -\omega_e + \left(1 - \frac{1}{\sigma}\right) \omega_c & \frac{r_p}{\sigma L'_p} \end{bmatrix} \begin{bmatrix} P_p \\ Q_p \end{bmatrix} \\ \mathbf{D} = -\frac{3}{2} \frac{L_m}{\sigma L'_p L'_c} \begin{bmatrix} u_{pd}^c & u_{pq}^c \\ u_{pq}^c & -u_{pd}^c \end{bmatrix} \quad \mathbf{U}_{cdq}^c = \begin{bmatrix} u_{cd}^c \\ u_{cq}^c \end{bmatrix}$$

The switch control law is chosen based on super-twisting SMC [24] to make dW / dt < 0 for $S \neq 0$. Thus, the following control law can be designed as:

$$U_{cdq}^{c} = -D^{-1}[F + U_{c}^{*}]$$
(25)

where,

$$\begin{aligned} \boldsymbol{U}_{c}^{*} &= \int \boldsymbol{A} sgn(\boldsymbol{S}) + \boldsymbol{B} |\boldsymbol{S}|^{0.5} sgn(\boldsymbol{S}) \\ \boldsymbol{A} &= \begin{bmatrix} A_{1} & 0 \\ 0 & A_{2} \end{bmatrix}, \ \boldsymbol{B} &= \begin{bmatrix} B_{1} & 0 \\ 0 & B_{2} \end{bmatrix}, \ sgn(\boldsymbol{S}) = \begin{bmatrix} sgn(S_{1}) \\ sgn(S_{2}) \end{bmatrix}^{T} \end{aligned}$$

The required CW voltage vector \boldsymbol{U}_{cdq}^{c} is generated by the SVM module.

3) Proof of the Stability

For stability in the sliding surfaces, dW/dt < 0 must be satisfied based on (16), (24) and (25):

$$\frac{d\boldsymbol{W}}{dt} = -\boldsymbol{S}(\int \boldsymbol{A} sgn(\boldsymbol{S}) dt + \boldsymbol{B} |\boldsymbol{S}|^{0.5} sgn(\boldsymbol{S}))$$
(26)

where, $S \cdot sgn(S) > 0$. Setting appropriate positive control gains yields dW/dt < 0. Since *W* is a positive-definite function and its time derivative (dW/dt) is a negative-definite function, S_1 and S_2 approaches zero asymptotically and the proposed controller becomes asymptotically stable.

4) Proof of the Robustness

In practical application conditions, the performance of the control system is impressed by system disturbances such as parameter variations, measurement noises, analogue-digital sample errors, thus, (24) can be rewritten as:

$$\frac{dS}{dt} = F + DU_{cdq}^{c} + H$$
(27)
where, $H = [H_1 H_2]^{\text{T}}$. Thus, (26) is rewritten as:

$$\frac{dW}{dt} = \mathbf{S}^T \frac{dS}{dt} = -\mathbf{S}(\int \mathbf{A} sgn(\mathbf{S}) \, dt + \mathbf{B} |\mathbf{S}|^{0.5} sgn(\mathbf{S}) - \mathbf{H}) < 0$$
(28)

If the positive control gains matrices A and B are set large enough to fulfill (28), dW/dt is still definitely negative. According to Lyapunov stability theorem, the proposed controller features strong robustness, if the control gains are selected properly.

$$\left|\int Asgn(\mathbf{S}) \, dt + \mathbf{B}|\mathbf{S}|^{0.5}sgn(\mathbf{S})\right| > |\mathbf{H}| \tag{29}$$

Figure 5 shows the block diagram of the proposed SSM-DPC. First, the measured PW voltages and currents, U_{pabc} and $I_{p \ abc}$, are transformed into the dq stationary reference frame:



Fig. 5. A structural block diagram of the proposed SSM-DPC strategy control system for the inverter-fed BDFIG.

 U_{pdq}^{s} and I_{pdq}^{s} . Then, the PW active and reactive power can be calculated as follows:

$$\begin{cases} P_p = -\frac{3}{2} (u_{dp}^s i_{dp}^s + u_{qp}^s i_{qp}^s) \\ Q_p = -\frac{3}{2} (u_{qp}^s i_{dp}^s - u_{dp}^s i_{qp}^s) \end{cases}$$
(30)

Consequently, PW voltage U_{pdq}^{s} is transformed into the CW reference frame U_{pdq}^{c} and matrices *F* and *D* are obtained according to (24). In addition, errors of instantaneous active and reactive power of PW are used as the input of the supertwisting SMC-based power controller. The CW voltage reference can be directly deduced in the CW reference frame, while, other presented SMC strategies use stationary reference frame and need reference frame transformation [15-17, 19-20]. The required CW voltage vector U_{cdq}^{c} is generated by space vector modulation (SVM) unit and it is worth noting that the proposed controller is simple and needs no synchronous coordinate transformations, PLL block and tuning PI parameters which makes such method favorable for the target application.

IV. SIMULATION RESULTS OF 2 MW BDFIG

In order to verify the proposed SSM-DPC strategy, for large machines, a 2 MW BDFIG is simulated in Matlab/Simulink[®] platform and the parameters of the generator are tabulated in

Table I. During simulation, the sampling time and the simulation time-step were 100 μs and 5 μs respectively. The nominal speed of the investigated generator is 600 rpm which is considered as 1 *pu*. Since wind turbine inertia is large, the rotor speed variations are small and negligible [25]. Therefore, the simulation set-up evaluated considers the rotor speed constant, whereby $\omega_m = 0.8 pu$.

TABLE I					
3 kW and 2 MW BDFIG Parameters and Ratings					
BDFIG	BDFIG 3 kW-Experimental 2 MW-Simulati				
ρ _{p,} ρ _c	3, 2	3, 2			
f_p, f_c (Hz)	50, 50	50,50			
V _{pn} , V _{cn} (V _{rms})	380, 380	690, 690			
$r_{\rm p}, r_{\rm c}, r_{\rm r}(\Omega)$	2.025, 0.96, 0.282	0.408, 1.186, 1.531			
L _{lp} , L _{lp} , L _{lr} (mH)	11.9, 7.1, 19	0.014, 0.012, 0.026			
L _{mp} , L _{mc} (mH)	260.7, 149.7	0.626, 0.373			
S _b (kW)	3.9	2103.5			
V _b (V _{rms})	380/√3	690/√3			

The PW (Fig. 1) is connected to three-phase, 690 V_{rms} , and 50 Hz, where the nominal dc-link voltage of the machine side converter (MSC) is set to 1200 V and the dc capacitor is 16000 μF . The control parameters of the proposed SSM-DPC are tabulated in Table II. To evaluate performance and the effectiveness of the proposed SSM-DPC, comparative simulations involving DPC [11] and integral-SMC [13] strategies are investigated under the same conditions. Integral-SMC in [13] is presented for DFIG, and is adopted for



Fig. 6. Simulation results under step change conditions for active and reactive power. (a) Proposed SSM-DPC (b) Integral-SMC and (c) DPC.

BDFIG. Switching frequency of the SSM-DPC and integral-SMC strategies is set at 5 kHz.

TABLE II				
Control Parameters of the Proposed SSM-DPC Strategy				
BDFIG A_1 A_2 B_1 E				
2 MW-Simulation	100	35	7	6.5
3 kW-Experimental	4	20	2.2	8

Fig. 6(a)-(c) compares the BDFIG response during different active and reactive power steps for the proposed SSM-DPC, integral-SMC and DPC, respectively. The waveforms shown in Fig. 6 include PW active power, PW reactive power, PW phase current and CW phase current. Initially, PW active and reactive power references are set to zero. The PW active power is stepped from zero to 1 pu at t = 0 s and then backed to zero at t = 1.5 s, while the reactive power is stepped from zero to -1 pu at t = 0.5 s and backed to zero at t = 1 s. The transient response of active and reactive power for three mentioned methods are within few milliseconds, which their accurate values are tabulated in Table III.

TABLE III QUANTITATIVE COMPARISON OF SIMULATION RESULTS

QUANTITATIVE COMITATISON OF CIMOLATION RESOLTS						
Control method	Transitory response (<i>ms</i>)		Power Ripple (%)		THD (%)	
	P_{p}	Q_p	Pp	Q_p	i _p	i _c
SSM-DPC	1.2	1.3	4	3	0.84	4.22
Integral-SMC	1.32	4.3	18	9	3.61	6.11
DPC	1	1	19	11	4.20	6.23

It is evident that time-response of reactive-power of the proposed SSM-DPC is up to three times faster than the integral-SMC. As shown in Fig. 6(a) for the proposed SSM-DPC, the step change of one control variable, i.e., PW active or reactive-power, does not affect the other. Moreover, there is no over-shoot in the PW current. Conversely, Fig 6(c) shows that, the DPC method has the highest ripples in power and current compared to the other methods. According to Fig. 6 and Table III, the proposed SSM-DPC has lower ripples in both active and reactive power ripple than those of DPC and integral-SMC strategies. The aforementioned comparative advancements show the accuracy and effectiveness of the proposed SSM-DPC during transient and steady-state conditions accordingly. Additional, simulation scenarios have been deduced and carried out to further evaluate the performance of the proposed SSM-DPC against rotor-speed variations, whereby the PW active and reactive powers are set to 1 pu and zero, respectively. The rotor speed is changed from 0.9 pu to 1.1 pu. Figure 7 shows that, during the speed variation, the PW active and reactive powers are controlled, whilst the CW current frequency initially decreases proportional to the BDFIG slip changes, reaching zero at the synchronous speed of 600 r/min, and increases after passing 600 r/min. Therefore, the proposed SSM-DPC has proven to be robust to rotor speed variations.

V. EXPERIMENTAL VERIFICATION OF THE 3 KW BDFIG

The performance of the proposed SSM-DPC is deduced, designed, simulated and experimentally verified and heavily scrutinized taking into consideration the active and reactive power transient step changes and also variable rotor speed variations typical for wind energy conversion system



Fig. 7. Simulation results under rotor speed variation.

applications. Consequently, comparative advancements against the system response of DPC [11] and integral-SMC [13] under the same operating conditions have been benchmarked and their trade-offs have been evaluated accordingly. The model adopted in BDFIGs with wound or nested-loop rotor by most researchers is similar to the model adopted in the cascaded BDFIGs [6, 10, 11, 20]. Hence, experimental tests are run on a 3 kW CBDFIG setup instead of BDFIG with wound rotor [6, 10]. The CBDFIG consists of two DFIGs namely power and control machines which are connected in both mechanical and electrical manner as depicted in Fig. 8.

The stator windings of power and control machines play the role of the PW and CW in BDFIG, respectively. The parameters of a 3 kW CBDFIG are tabulated in Table I. Rotor impendences (i.e. rotor resistances and leakage inductances) of power and control machine are added together which is consider as equivalent rotor impedance of BDFIG. The 3 kW CBDFIG is driven by a controlled induction motor which emulates the wind-turbine and is mechanically coupled with the shaft of CBDFIG. The PW is directly connected to threephase grid voltage 380 V_{rms} , and 50 Hz. The CW is fed by a voltage source converter (VSC) which is controlled by TI TMS320F28335 DSP, and the sampling frequency is 10 kHz. The control parameters of the proposed SSM-DPC, employed for experimental test purposes are listed in Table II. The control of GSC is not the focus of this paper, while the dc-link voltage is considered constant and set at 545 V, where dc-link capacitor is 560 μ F. The measurement of speed is done via an incremental optical encoder with 3,600 pulses per revolution for shaft rotational speed. LEM LV-25p and LA55/p transducers are used to measure the voltage and current of PW and CW. In order to demonstrate the PW active and reactive



Fig. 8. The BDFIG test setup used for experimental validation purposes.

power, information from PW voltage and current sensors are gathered by a Data Acquisition Card (DAQ) made by AdvantechTM.

The experimental results of active/reactive power and PW/CW currents are shown in Fig. 9(a)-(c) for the proposed SSM-DPC, integral-SMC and DPC strategies, respectively. The selected PW active power variation is a step of 1 pu, i.e. the nominal PW active power, and the selected PW reactive power variation is a step of -0.5 pu, i.e. half of the nominal PW reactive power, with the constant rotor speed of 0.8 pu. The PW active power reference steps from zero to 1 pu at t = 2

s while the PW reactive power reference is stepped from zero to -0.5 pu at t = 2.5 s and then is stepped to -1 pu at t = 3 s.

The transitory responses during PW active power steps for the proposed SSM-DPC and DPC are similar, both within a few milliseconds, Fig. 9(a) and (c). In contrast, the transitory response of integral-SMC during PW reactive power is twice time slower, Fig. 9(b). Moreover, the step change of one control variable, i.e., PW active/reactive power, in the proposed SSM-DPC does not significantly affect the other, and there is no over-shoot of the PW/CW currents, Fig. 9(a). Moreover, the active and reactive power ripples of the proposed SSM-DPC are smaller than other comparative strategies. In case of the PW active power, ripples of the proposed SSM-DPC are about four times smaller than other strategies. In case of the PW reactive power, ripples of the proposed SSM-DPC are about three times smaller than DPC. Accurate comparison of the transient time-response, powerripples and currents THD for three considered and evaluated methods are summarized in Table IV.

Table IV	
Quantitative Comparison of the Experimenta	I Results

Control method	Transitory response (<i>ms</i>)		Power Ripple (%)		THD (%)	
	Pp	Q_{p}	P_{p}	Q_p	i _p	i _c
SSM-DPC	1.2	1.1	5.4	5.1	1.24	4.62
Integral-SMC	1.2	2.64	18.2	6.7	11.97	10.5
DPC	1	1	22.5	18.2	8.18	7.7

Figure 10, shows the harmonic spectra of PW and CW currents during the steady-state conditions of $P_p = 1 pu$ and $Q_p = -0.5 pu$. Furthermore, Fig 10(c) shows that, the currents harmonics of DPC are spread over a wide frequency range. Whereas Fig. 10(a) and (b) indicate that, the proposed SSM-DPC and integral-SMC produces deterministic harmonics with dominant harmonics around the switching frequency of 5 kHz and ~ 4 kHz. Based on Table III, the proposed SSM-DPC have lower PW and CW current harmonic distortion than other methods. The THDs of the PW and CW currents of the



Fig. 9. Experimental results, stepped changes in both active and reactive power. (a) Proposed SSM-DPC (b) Integral-SMC and (c) DPC.



Fig. 10. PW and CW current harmonic spectra of Fig. 8: Po=1 pu, Qo= -0.5 pu. (a) Proposed SSM-DPC (b) Integral-SMC and (c) DPC.

proposed SSM-DPC strategy are 1.24% and 4.62%, integral-SMC (11.97% and 10.46%) and DPC (8.18% and 7.7%). It is concluded that the proposed SSM-DPC has better overall performance than DPC and integral-SMC. The proposed SSM-DPC provides enhanced transient performance similar to the DPC, and meanwhile provides the excellent steady-state performance, compared to DPC and integral-SMC.

To evaluate the performance of the proposed SSM-DPC against rotor-speed variations, the rotor speed is changed from 0.9 pu to 1.1 pu. In this test, the PW active and reactive powers are constant at 1 pu and zero, respectively. Figure 11 shows that, during the speed variation, the CW current frequency initially decreases proportional to the BDFIG slip changes, reaching zero at the synchronous speed of 600 r/min, and increases after passing 600 r/min. The PW active and reactive powers are controlled and the performance of the proposed controller is not affected by rotor speed variations. The proposed SSM-DPC is robust to rotor speed variations and has appreciate performance.

Further tests are carry out to evaluate the robustness of the proposed SSM-DPC against machine parameters variations, which is a crucial point of any controller performance, since they used to calculate the terms F and D and may be affected the performance of the proposed SSM-DPC strategy. The impact of the variation of mutual inductances and resistances on system performance is studied, since parameters variations can occur due to possible machine saturation, temperature variation, skin effect etc. The stator and rotor leakage inductances are considered constant since their variations during machine operation are negligible [26]. The test results are compared in Fig. 12 with values of mutual-inductances and resistances used in the control system varied by $\pm 50\%$, which are used to calculate the terms F and D. The step changes in both PW active and reactive power are applied: active power reference is changed from 1 to 0.5 pu at t = 2 sand then backed to 1 pu at t = 3.5 s and, while the reactive power is stepped from -1 to -0.5 pu at t = 2.5 s and backed to -1 pu at t = 3 s. Comparing Fig. 11(a)–(d), these parameters variations have insignificant influence on system dynamic and steady-state performances and the system maintains superb performance, even with such large errors in the mutual inductances and resistances. As a result, the proposed SSM-



Fig. 11. Experimental results under rotor speed variation.

DPC is robust to generator parameters' variations with excellent dynamic and static performances.

VI. CONCLUSION

This paper proposes an improved SSM-DPC strategy for the machine side converter of the BDFIG which controls both the active and reactive power independently. The proposed controller directly determines the required CW voltage based on the CW current, PW voltage, rotor speed, and values of active and reactive powers. The proposed method obtains the CW voltage in the CW-reference frame, outperforming the integral-SMC [13] which is very sensitive to the rotor-position error for transformation from PW to CW reference-frame.

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Fig. 12. Experimental results. (a) without parameters variations, (b) with -50% variation in mutual inductances and, rotor and stator resistances (c) with +50% variation in mutual inductances and -50% variation in rotor and stator resistances (d) with +50% variation in mutual inductances and, rotor and stator resistances.

However, the proposed method needs the rotor position and speed to calculate the control matrices F and D. Since the proposed SSM-DPC strategy is based on SMC method, it can be made robust to the effect of rotor position error in F and D by selecting appropriate gains for super-twisting mechanism.

Simulation results on a 2 MW range and experimental results on a 3 kW BDFIG test rig have been provided and equated with those of DPC [11] and integral-SMC [13]. The main features of the proposed control system can be summarized as:

- lower power ripple compared to integral-SMC and DPC,

- less THD than integral-SMC and DPC strategies, due to its continuous controller output,

- ability to control both active and reactive power directly without overshoot and with constant switching frequency,

- provides excellent steady-state performance while having high transient response in contrast to DPC,

- robustness against the machine parameters variation.

Simulations and experimental tests have validated the effectiveness, robustness and feasibility of the proposed SSM-DPC during numerous operating conditions (power transient changes and rotor speed variations) and machine parameter variations.

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