Sliding-Mode Control for a DFIG-based Wind Turbine under Unbalanced Voltage

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Abstract: This paper reports a first-order sliding-mode control (1-SMC) design for controlling the doubly-fed induction generator (DFIG)-based wind turbine’s rotor-side power converter. The design is particularly focused on keeping the generator successfully in operation under unbalanced grid voltage conditions, as today’s grid codes require. Aside from controlling the stator-side active and reactive powers’ average value, the rotor-side converter is commanded so as to remove the fluctuations affecting the electromagnetic torque and the reactive power during unbalanced voltage. The paper aims to put forward the bases of the proposed design together with the described algorithm’s stability proof. Finally, the appropriateness of the sliding-mode control to deal with the aforementioned disturbed scenarios is supported by means of simulation results.

Keywords: Wind, power control, induction generators, variable-structure systems, unbalanced voltage.

1. INTRODUCTION

The most employed onshore wind turbine nowadays is the one based on the DFIG, covering up to 75% of the installed European wind turbines. Referring to Fig. 1, the general structure of the DFIG-based wind turbine can be found. The rotor-side power converter is connected to the machine’s rotor, while the stator is directly linked to the grid. Thanks to the coupling between the rotor and the stator, the latter’s current is governed by controlling the rotor current, and, consequently, so are the stator-side active and reactive powers exchanged with the grid.

Fig. 1. Structure of a DFIG-based wind turbine.

In the early versions of the DFIG, and always based upon the well-known vector control theory —see Vas (1998)—, rotor current control was mainly carried out by means of two proportional-integral (PI) controllers—one for each current component—and their respective decoupling control feedforward terms, which were in charge of improving the dynamic response of the system, as described in Tapia et al. (2003). Such a control scheme allowed for several years to independently control the stator-side active and reactive powers. However, this kind of strategies were not designed to deal with unbalanced scenarios. As a consequence, they cease to be sufficient from the moment in which the grid codes demand to keep the wind turbine successfully in operation during imbalances.

Under unbalanced grid voltage conditions—very common in weak grids—the current also turns out being unbalanced. Thus, according to the symmetrical components theory, a negative sequence current arises, which degrades the insulation of the generator windings. Moreover, if such a negative sequence current is not under control, not only the powers and the DC link voltage are affected by important oscillations of two times the grid frequency, but also the electromagnetic torque. Consequently, the shaft is subjected to an additional mechanical stress, leading to an audible mechanical noise, as tested in Kiani and Lee (2010).

The above-cited unwanted consequences can be solved by controlling the negative sequence of the current. Nevertheless, it is not possible to simultaneously solve all the cited issues, since there are not enough degrees of freedom. In this regard, the general strategy usually adopted, —as, for instance, in Xu (2008a) and Hu et al. (2009)—, as well as the one considered in this paper, is the following:

- The rotor-side converter removes the fluctuations affecting the electromagnetic torque and the stator reactive power.
The grid-side converter ensures both constant DC link voltage and steady active power output from the overall system.

As far as the control strategies adopted are concerned, the first solutions proposed revolve around PI controllers, like those in Xu (2008b) and Xu (2008a), in which the positive and the negative rotor current sequences are separately controlled. Moreover, in Hu et al. (2009) and Zhou et al. (2009) PI+Resonant (PI+R) controllers are proposed, avoiding the decomposition into sequences of the rotor current. This way, the slight errors originated by the decomposition procedure are eliminated. On the other hand, several authors have opted for direct power control (DPC) techniques, which are robust to parameter variations and simple to implement, as those in Santos-Martín et al. (2008) and Zhou et al. (2009).

Focusing on these subjects, it is desired to put forward how suitable the sliding-mode control is when facing unbalanced scenarios. First of all, because it allows to successfully achieve the positive and negative sequence de- 
ections, removing the afore-cited 
uctuations without the need of rotor current positive and negative sequences’ decom- position. Moreover, since in addition to being simple to implement, it is a fast-response solution and robust against DFIG parameter variations. And, finally, because it also turns out being an alternative that directly generates the signals to command the gates of the converter transistors, thus eluding modulating methods, such as PWM or SVM.

Within this framework, a DFIG active and reactive powers control algorithm is presented —based on the research work carried out by Utkin et al. (1999), and as a continuity to the work presented by Susperregui et al. (2010)—, specially conceived for facing either balanced or unbalanced grid voltage conditions, and whose results have been validated by carefully devised simulation tests.

2. SYSTEM MODELING AND ROTOR CURRENT SET-POINTS

The voltage equations of a DFIG denoted in space-vector form, considering the motor convention, and referred to the stator reference frame are given by Abad et al. (2010) as

\[
v_s = R_s i_s + \frac{d\phi_s}{dt}
\]

for its stator, and

\[
v_r = R_r i_r + \frac{d\phi_r}{dt} + \omega_r\Phi_r
\]

for its rotor, where

\[
\phi_s = L_s i_s + L_m i_d
\]

\[
\phi_r = L_r i_r + L_m i_q
\]

while \(\omega_r\) is the rotor electrical speed, \(R_s\) and \(R_r\) represent the stator and rotor resistances, \(L_s\), \(L_r\), and \(L_m\) are the stator, rotor, and magnetizing inductances, and \(v_s\), \(v_r\), \(i_s\), \(i_r\), \(\phi_s\) and \(\phi_r\) reflect the stator and rotor voltages, currents and fluxes, respectively. Additionally, subscript \(D\) and \(Q\) indicate if the variables belong either to the direct or quadrature component.

In section 3, the stator active and reactive powers are governed by two rotor current control loops—one for each current component. Thus, in order to fix the set-points for \(i_{rD}\) and \(i_{rQ}\), it is indispensable to determine the relations among the rotor current components and the stator active and reactive powers. Such relations are usually analyzed considering the DFIG model represented in the positive and negative sequences. In this paper, they are derived from the power expression provided in Wang and Xu (2007), and they have been particularized for the reference frames \(x_p = y_p\) and \(x_n = y_n\) depicted in Fig. 2. Note that the direct axes of these reference frames are collinear with the grid voltage positive and negative sequences’ space-vectors.

Thus, the average value of the stator active and reactive powers —\(P_{sa}\) and \(Q_{sa}\)— are governed by the rotor current positive sequence, as shown below:

\[
i_{rz}^s = -\frac{2L_s}{3L_m}\frac{|v_p|^2}{|v_s|^2}P_{sa}
\]

\[
i_{ry}^s = \frac{2L_s}{3L_m}\frac{|v_p|^2}{|v_s|^2}Q_{sa} - \frac{|v_p|^2}{L_m\omega_s}
\]

where, superscripts \(p\) and \(n\) indicate if the variables belong to the positive or negative sequence.

On the other hand, during unbalanced voltage conditions, it may be interesting to fulfill several control targets, as reflected in Xu (2008b). However, not all targets can be achieved simultaneously. Therefore, this paper is focused again —derived from Wang and Xu (2007)— to have been accomplished between the rotor current positive and negative sequences:

\[
i_{rz}^n = \frac{|v_p|}{|v_s|}i_{rz}^p; \quad i_{ry}^n = -\frac{|v_p|}{|v_s|}i_{ry}^p.
\]
However, the rotor current control carried out in section 3 is performed through the stationary reference frame—axes \(D-Q\). Thus, according to (5), the relations among the rotor current stationary components, \(i_{rD}\) and \(i_{rQ}\), and their corresponding \(i_{x}\) and \(i_{y}\) synchronous-frame components turn out finally being:

\[
i_{rD} = i_D^p + i_{rD}^n, \quad i_{rQ} = i_Q^p + i_{rQ}^n,
\]

where

\[
\begin{bmatrix}
i_D^p \\
i_Q^p \\
i_{rD}^n \\
i_{rQ}^n
\end{bmatrix} = \begin{bmatrix}
\cos \theta_p & -\sin \theta_p \\
\sin \theta_p & \cos \theta_p \\
\cos \theta_n & -\sin \theta_n \\
-\sin \theta_n & \cos \theta_n
\end{bmatrix} \begin{bmatrix}
i_x \\
i_y \\
i_{rD}^n \\
i_{rQ}^n
\end{bmatrix},
\]

(7)

while \(\theta_p\) and \(\theta_n\) are the phases of the stator voltage positive and negative sequences’ space-vectors, respectively —refer to Fig. 2.

3. SLIDING-MODE CONTROL

Fig. 3 shows the RSC scheme, in which \(s_{u1}, s_{u2}, s_{u3}, s_{u4}, s_{u5}\) and \(s_{u6}\) are the transistors’ on/off gating boolean signals, and \(U_{ra}, U_{rb}\) and \(U_{rc}\) represent the electric potential differences between the midpoints of the converter legs and \(N\). Therefore, the relationship between the latter and the gating signals is given by

\[
U_r = \begin{bmatrix} U_{ra} \\ U_{rb} \\ U_{rc} \end{bmatrix} = u_0 \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} s_{u1} \\ s_{u2} \\ s_{u3} \\ s_{u4} \\ s_{u5} \\ s_{u6} \end{bmatrix},
\]

(8)

where the backward transformation is:

\[
s_{u1} = 0.5(1 + U_{ra}/u_0); \quad s_{u4} = 1 - s_{u1} \\
s_{u2} = 0.5(1 + U_{rb}/u_0); \quad s_{u5} = 1 - s_{u2} \\
s_{u3} = 0.5(1 + U_{rc}/u_0); \quad s_{u6} = 1 - s_{u3}.
\]

(9)

Fig. 3. Scheme of DFIG RSC.

As the generator model handled is the one presented in (1) and (2), the rotor voltage signals considered during the design, \(v_{rD}\) and \(v_{rQ}\), are expressed in the stator reference frame. In this context, the relation between these voltages and \(U_{ra}, U_{rb}\) and \(U_{rc}\) turns out being

\[
\begin{bmatrix} v_{rD} \\ v_{rQ} \end{bmatrix} = \frac{2}{3} \begin{bmatrix}
\cos a & \cos b & \cos c \\
\sin a & \sin b & \sin c
\end{bmatrix} \begin{bmatrix} U_{ra} \\ U_{rb} \\ U_{rc} \end{bmatrix},
\]

(10)

while

\[
a = \theta_r; \quad b = \theta_r + \frac{2\pi}{3}; \quad c = \theta_r - \frac{2\pi}{3}.
\]

(11)

Aiming at governing each rotor current component, \(i_{rd}\) and \(i_{rq}\), the following switching variables are selected:

\[
s_D = e_D + e_D \int e_D dt, \quad (12)
\]

\[
s_Q = s_Q + e_Q \int e_Q dt,
\]

(13)

and the integrals of the errors in (12), weighted by \(e_D\) and \(e_Q\) constant positives, lead the steady-state errors to be zero.

Replacing (3) in (2), and solving for the time derivatives of the rotor current components from the latter, it turns out that

\[
\begin{aligned}
di_{rd} &= -\frac{L_r}{L_r} i_{rd} - \frac{L_m}{L_r} \frac{d\psi_D}{dt} - \omega_r (i_{rq} + \frac{L_m}{L_r} \phi_{sD}) \\
di_{rq} &= -\frac{L_r}{L_r} i_{rq} - \frac{L_m}{L_r} \frac{d\psi_Q}{dt} + \omega_r (i_{rd} + \frac{L_m}{L_r} \phi_{sD}),
\end{aligned}
\]

(14)

where \(i_{rd}\) and \(i_{rq}\) are the rotor current components, \(L_r\) is the inductance of the rotor winding, \(\omega_r\) is the rotor speed, \(\psi_D\) and \(\phi_{sD}\) are the flux linkage of the rotor and stator, respectively.

From (12) and (14), the dynamics of the switching variables can be expressed as:

\[
\begin{bmatrix} s_{DQ} \\ S_{DQ} \end{bmatrix} = \begin{bmatrix} F_D \\ F_Q \end{bmatrix} + \begin{bmatrix} \frac{a}{L_r} & 0 \\ 0 & \frac{a}{L_r} \end{bmatrix} \begin{bmatrix} v_{rD} \\ v_{rQ} \end{bmatrix},
\]

(15)

with

\[
F_D = f(i_{rd}^2, i_{rd}, i_{rq}, \omega_r, \phi_{sD}, \phi_{sD}), \quad F_Q = f(i_{rq}^2, i_{rd}, i_{rq}, \omega_r, \phi_{sD}, \phi_{sD}).
\]

(16)

According to (12) and (15), the relative degree of the system turns out being equal to one, since it is enough to take the time derivative of the switching variables just once to first appear control signals \(v_{rD}\) and \(v_{rQ}\). In this sense, taking into account that the order of the sliding-mode control to be designed must be at least equal to the relative degree of the system, a first-order sliding-mode control becomes enough to make the switching variables converge to zero, which has been performed by applying the following 1-SMC law:

\[
U_r = u_0 \text{sgn}(S_{abc}),
\]

(17)

with \(S_{abc} = [s_a s_b s_c]^T\).

Replacing (10) and (17) into (15), it turns out that

\[
\dot{S}_{DQ} = F_{DQ} - a M_1 \begin{bmatrix} u_0 \text{sgn}(s_a) \\ u_0 \text{sgn}(s_b) \\ u_0 \text{sgn}(s_c) \end{bmatrix}.
\]

(18)

Equation (18) establishes the direct relation between \(S_{DQ} = [s_D s_Q]^T\) and \(S_{abc} = [s_a s_b s_c]^T\), which should be designed so that, commanded by (17), \(s_D\) and \(s_Q\) vanish in finite time. Hence, it seems consistent to consider that a suitable candidate for \(S_{abc} = [s_a s_b s_c]^T\) might be:

\[
S_{abc} = (M_2)^+ S_{DQ},
\]

(19)

where \((M_2)^+\) denotes the Moore-Penrose pseudo-inverse of \(M_2\), being

\[
M_2^+ = \frac{3}{2a^2} M_2^T.
\]

(20)
Thus, the following is obtained for $S_{abc} = [s_a \ s_b \ s_c]^T$ from (19) and (20):
\[
\begin{align*}
  s_a &= \frac{1}{a} (s_D \cos a + s_Q \sin a) \\
  s_b &= \frac{1}{a} (s_D \cos b + s_Q \sin b) \\
  s_c &= \frac{1}{a} (s_D \cos c + s_Q \sin c). 
\end{align*}
\] (21)

Therefore, the discontinuous control signals to be applied can be deduced by replacing (21) into (17), hence yielding:
\[
\begin{align*}
  U_{ra} &= u_0 \text{sgn}(s_D \cos a + s_Q \sin a) \\
  U_{rb} &= u_0 \text{sgn}(s_D \cos b + s_Q \sin b) \\
  U_{rc} &= u_0 \text{sgn}(s_D \cos c + s_Q \sin c). 
\end{align*}
\] (22)

Note that the constant term $1/a$ in (21) has been omitted, since it has no effect in the sign of $S_{abc} = [s_a \ s_b \ s_c]^T$. Thus, the control signals are only dependent on the selected switching functions’ state, turning out being insensitive against parameters deviations.

Finally, by means of (22) and (9), transistors’ on-off gating boolean signals are derived.

3.1 Stability proof

In a stable system, the switching variables converge to zero. Thus, in order to prove the convergence, the following Lyapunov function is defined:
\[
V = \frac{1}{2} S_{DQ}^T S_{DQ} \geq 0, 
\] (23)
whose time derivative should be negative:
\[
\dot{V} = \frac{1}{2} (S_{DQ}^T \dot{S}_{DQ} + S_{DQ}^T \dot{S}_{DQ}) = S_{DQ}^T S_{DQ} \leq 0. 
\] (24)

Working out the term $S_{DQ}^T S_{DQ}$ in (24) by considering (18), the following is obtained:
\[
S_{DQ}^T F_{abc} S_{abc} = \frac{4a^2}{9} u_0 \begin{bmatrix}
  \text{sgn}(s_a) - 0.5 \text{sgn}(s_b) - 0.5 \text{sgn}(s_c) \\
  \text{sgn}(s_b) - 0.5 \text{sgn}(s_c) - 0.5 \text{sgn}(s_a) \\
  \text{sgn}(s_c) - 0.5 \text{sgn}(s_a) - 0.5 \text{sgn}(s_b)
\end{bmatrix} M_1^T M_2 U_r, 
\] (25)

where $F_{abc} = M_1^T F_{DQ} = [F_a \ F_b \ F_c]^T$.

Depending on the signs of $s_a$, $s_b$ and $s_c$, there are eight possible combinations of $\text{sgn}(s_a)$, $\text{sgn}(s_b)$ and $\text{sgn}(s_c)$ values. Note that, according to (21), they can never be all $+1$ or all $-1$ simultaneously. Thus, the remaining six combinations are presented in Table 1.

Table 1. All existing $V$ combinations

<table>
<thead>
<tr>
<th>$s_a$</th>
<th>$s_b$</th>
<th>$s_c$</th>
<th>$V$</th>
</tr>
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<tbody>
<tr>
<td>$&lt;0$</td>
<td>$&lt;0$</td>
<td>$&gt;0$</td>
<td>$V$</td>
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<td>$&lt;0$</td>
<td>$V$</td>
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<td>$&gt;0$</td>
<td>$V$</td>
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<td>$V$</td>
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<td>$V$</td>
</tr>
<tr>
<td>$&gt;0$</td>
<td>$&gt;0$</td>
<td>$&lt;0$</td>
<td>$V$</td>
</tr>
</tbody>
</table>

A general expression for $\dot{V}$, which reflects all possible combinations shown in Table 1, turns out being
\[
\dot{V} = \left( s_a F_a + s_b F_b + s_c F_c \right) = \frac{4a^2}{9} u_0 \left( |s_a| + |s_b| + |s_c| \right), 
\] (26)
where $\text{sgn}(s_i) = \text{sgn}(s_m)$, with $l \neq m \neq n$ and $l, m, n \in \{a, b, c\}$.

Subsequently, in order to ensure the convergence, $\dot{V} < 0$ has to be guaranteed for all possible $l$, $m$ and $n$, what means that $|g_l| > |f|$ must be fulfilled. Therefore, bearing in mind the most demanding case, the following condition should be accomplished:
\[
u_0 \geq \frac{9}{4a^2} \max(|F_a|, |F_b|, |F_c|). 
\] (27)

However, in practice, the whole system will usually remain stable with a lower value of $u_0$.

Thus, if condition in (27) is satisfied, even in presence of grid voltage disturbances, asymptotic convergence of $s_D$ and $s_Q$ to zero is guaranteed with the proposed control algorithm.

4. SIMULATION RESULTS

The above developed algorithm has been evaluated via simulation tests through Matlab/Simulink software. To this effect, a 660-kW DFIG mathematical model, whose electrical parameters are summarized in Table 2, is implemented by means of a $S$-function block programmed in C language. The rotor- and grid-side power converters are constructed using standard blocks from Simulink, while their control algorithms are also programmed in C.

Table 2. Electrical parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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</thead>
<tbody>
<tr>
<td>Nominal power</td>
<td>660 kW</td>
</tr>
<tr>
<td>Rated r.m.s. stator voltage</td>
<td>398/690 V</td>
</tr>
<tr>
<td>Rated peak rotor voltage</td>
<td>380 V</td>
</tr>
<tr>
<td>Rated peak rotor current</td>
<td>400 A</td>
</tr>
<tr>
<td>Stator resistance per phase, $R_s$</td>
<td>6.7 mΩ</td>
</tr>
<tr>
<td>Stator inductance per phase, $L_s$</td>
<td>7.5 mH</td>
</tr>
<tr>
<td>Magnetizing inductance, $L_m$</td>
<td>19.4 mH</td>
</tr>
<tr>
<td>Rotor resistance per phase, $R_r$</td>
<td>39.9 mΩ</td>
</tr>
<tr>
<td>Rotor inductance per phase, $L_r$</td>
<td>52 mH</td>
</tr>
<tr>
<td>General turns ratio, $n$</td>
<td>0.3806</td>
</tr>
<tr>
<td>Number of pole pairs, $P$</td>
<td>2</td>
</tr>
</tbody>
</table>

The grid voltage signal of the phase affected by the voltage drop is shown in Fig. 4(a). A 15% of the nominal grid voltage dip is simulated between seconds 2 and 4. –400 kW and 0.95 leading power factor are requested before the imbalance, while the amount of the demanded reactive power increases up to –400 kVAR from second 2 onward. On the other hand, the slip of the generator is assumed to be constant at 9%.

A zoom of the direct and quadrature rotor current components, $i_{rD}$ and $i_{rQ}$, together with their respective set-point signals, are displayed in Figs. 4(b) and 4(c), respectively. These figures allow to check how, indeed, the current successfully reaches its reference value. Hence, it can be concluded that the selected switching variables in (12) converge to zero when commanded by (22). Additionally, rotor current three-phase signal is shown in Fig. 4(d).
Fig. 4. Simulation results. (a) Stator ‘a’ phase voltage. (b) Zoom of the rotor current direct component. (c) Zoom of the rotor current quadrature component. (d) Three-phase rotor current. (e) Electromagnetic torque.

Fig. 5. Simulation results. (a) Stator-side reactive power. (b) Stator-side active power. (c) Active power of the overall system. (d) Reactive power of the overall system. (e) DC link voltage.
Given that the set-points are fixed in order to fulfill the relations in (5), oscillations affecting the electromagnetic torque and the reactive power are fully removed, as it is illustrated in Figs. 4(e) and 5(a), respectively. Besides, they contain a chatter around 6% of their rated value, percentage that is considered acceptable. Moreover, note that, in spite of the unbalanced grid voltage, the stator active and reactive powers’ average values follow their respective set-points —refer to Figs. 5(a) and 5(b). Furthermore, referring to Fig. 5(a), it can be checked that the control system is able to rapidly respond to abrupt changes in the set-points.

As far as the stator active power is concerned, Fig. 5(b) evidences there are not enough degrees of freedom to remove the oscillations arisen during the imbalance. However, the GSC is commanded in order to ensure steady active power output from the overall system. Thus, Fig. 5(c) shows that the total active power is free of fluctuations, similarly to the total reactive power reflected in Fig. 5(d). Moreover, as demonstrated in Zhou et al. (2009), the strategy adopted in regard to control targets distribution between converters allows also to remove the ripple in the DC link voltage, as shown in Fig. 5(e). In this context, although GSC control is not addressed in this paper, it should be remarked that the control algorithm employed for this converter is analogous to that described in section 3.

The suitability of sliding-mode control for handling unbalanced scenarios is therefore considered proven. However, the limited size of the power converters decreases the usual stator active and reactive powers’ controllable ranges, as analyzed in Hu and He (2011). Consequently, during adverse conditions —i.e., from a certain imbalance factor and with a slip close to its limit value—, it is not possible to generate the amount of reactive power that several countries’ grid codes demand nowadays. Therefore, some particular scenarios may need additional devices, such as FACTS, which can allow to generate exactly the amount of requested reactive power.

5. CONCLUSION

A 1-SMC has been presented, which is intended to control the RSC of a DFIG-based wind turbine both under balanced and unbalanced grid voltage conditions.

Fixing the set-points of the rotor current so that relations in (5) are fulfilled, the control algorithm proposed allows to completely remove the oscillations affecting the electromagnetic torque and the stator reactive power during imbalances.

Thanks to the sliding-mode control tracking ability, current control can be performed in the stationary reference frame. As a result, there is no need of rotor current sequences’ decomposition.

The commutation frequency of the IGBTs is given based on the state of the chosen switching variables. Therefore, despite the predominance of a given commutation frequency, it does not turn out being constant. However, on the other hand, the proposed solution does not require the use of SVM or PWM modulation techniques.

Finally, it aims to highlight the simple implementation of the control algorithm, as well as its fast response against disturbances and abrupt changes in the set-points.

REFERENCES


