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Control of Doubly-fed Induction Generator with Extended State Observer under Unbalanced Grid Conditions

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Abstract: Under unbalanced grid condition, in a doubly-fed induction generator (DFIG), voltage, current, and flux of the stator become asymmetric. Therefore, active-reactive power and torque will be oscillating. In DFIG controlling rotor side converter (RSC) aims to eliminate power and torque oscillations. However, simultaneous elimination of the power and torque oscillations is not possible. Also, grid side converter (GSC) aims to regulate DC-Link voltage. In this paper, in order to regulate DC-Link voltage, an extended state observer (ESO) based on a generalized proportional-integral (GPI) controller, is employed. In this controlling method, DC-Link voltage is controlled without measuring the GSC current, and due to using the GPI controller, the improved dynamic response is resistant against voltage changes, and the settling time is reduced. A versatile rotor position computation algorithm (RPCA) is utilized to measure the rotor speed. This algorithm is simple, yet it is accurate and is resistant to changes in the resistance of the rotor and stator. The simulations are implemented by MATLAB software in the synchronous positive and negative sequence reference (d-q).

Keywords: DFIG, RSC, GSC, ESO, Unbalanced grid.

1.1. Introduction:

Dynamic Modeling and Control of doubly-fed induction generator (DFIG)-based wind generation system operating during unbalanced grid conditions were studied in [1-2]. In a DFIG under unbalanced grid condition, voltage, current, and flux of the stator become asymmetric. If voltage unbalance is not taken into account by the control system, the stator current could be highly unbalanced even with a small unbalanced stator voltage [3]. Under unbalance grid conditions, the negative sequence component of stator voltage leads to the oscillation of power (active-reactive) and
electromagnetic torque with twice the grid frequency. Power fluctuations lead to increased temperature in the stator winding, and torque fluctuations lead to mechanical tension on the rotor [4]. Control methods of DFIG are done based on the voltage orientation control (VOC) or flux orientation control (FOC). In the VOC method, the current signal controls the active power and along with the direct axis (Id) and reactive power along with the quadrature axis (Iq), and in the FOC method is conversely [5-6-7]. Control of DFIG for eliminating power and torque oscillation under unbalanced grid voltage was studied in [8]. In DFIG controlling rotor side converter (RSC) aims to eliminate power and torque oscillations. However, simultaneous elimination of the stator active/reactive power and torque oscillations is not possible [9]. For the purpose of controlling the power and torque by RSC, multiple controller methods have been used. In [10], a sliding mode controller with fractional order sliding surface (SMC-F) based on robust control designed for controlling the performance of DFIG's RSC is used. In this control method to compensate uncertainties and incoming disturbances to the system, a sliding mode controller has been used. In order to increase the degree of freedom and further robustness of the controller, the sliding surface is selected as a fractional order form. In addition to increasing the performance of the controller, using a robust sliding mode controller cause a reduction to the chattering phenomenon in the input control signal. In [11] for the performance controlling of DFIG's RSC nonlinear robust control approach, an extended state observer based on the backstepping controller (ESO-BS) has been used. In this controller, the approach has a simple design and small computational complexity, which can eliminate nonlinear factors. As a result, computation complexity is reduced. This paper aims to make the steady-state error value zero and improve the accuracy of controlling the RSC, proportional-integral (PI) controller is used. Finally, this method control is compared with SMC-F and ESO-BS. In recent years for calculating the rotor speed of the DFIG, several methods of sensor-less have been proposed [12-13]. In [14-15], to calculate the rotor speed of the DFIG, Model reference adaptive system (MRAS) is used. The weakness of MRAS observers is that it runs in the stator reference frame, thus designing the control parameter can be difficult, and results may be incorrect or be unstable. In this paper, to measure the rotor speed of DFIG, a versatile rotor position computation algorithm (RPCA) is used. In the algorithm, without the need for estimating the stator flux, the rotor position is estimated by means of measured stator as well
as rotor voltages and current. This algorithm is simple and effective and does not require a direct estimation of the stator flux. One of the most important advantages of this scheme is that the calculations do not use any approximation. This algorithm is robust against variations of stator and rotor resistances and is only dependent on the estimation of the DFIG mutual inductance [16-18].

Under unbalanced grid conditions, common DC-link voltage is controlled by a grid side converter (GSC). In order to control the DC-link voltage, multiple control methods have been used. In [19] to control the DC-Link voltage, the FUZZY-PI controller has been used. Complexity in performance and analysis is one of the most important disadvantages of this controller. In [20] to control DC-Link voltage, a nonlinear observer controller based on sliding mode control has been used. This controller has a suitable performance in controlling the DC-Link voltage, but it is slightly complicated in terms of analysis and regulation. To solve this problem, in [21] a nonlinear observer controller based on high order sliding mode control has been used. This controller has a high sensitivity to a variety of disturbances and uncertainties. In [22-23] to control the DC-Link voltage, one cycle controller has been used. In this controller, the dynamic response is improved, and it is resistant to voltage changes, but it is complex in terms of the optimization of responses. In this paper, for controlling the DC-Link voltage, an expanded state observer (ESO) based on a generalized proportional-integral (GPI) controller is introduced. In this controller, due to using the GPI controller, the improved dynamic response is resistant against voltage changes, and settling time is reduced [24-29]. Finally, this control method is compared with SMC-F, ESO-BS, PI, and proportional (P) controller.

1.2. Materials and Methods:

1.2.1. Dynamic model of DFIG:

The existing DFIG models are primarily developed on the basis of balanced grid conditions. Under unbalanced grid conditions, both positive and negative sequence components of voltage and current need to be considered in order to accurately describe the system behavior. With the purpose of obtaining a decoupled control between torque/active power and reactive power, the d-axis in the synchronous reference frame is generally oriented along the stator flux vector or the stator voltage vector.
The equivalent circuit of a DFIG can be expressed in different reference frames such as the stationary frame, the rotor frame, or the synchronous frame fixed to either the stator voltage or the stator flux.

A general expression of a DFIG model in an arbitrary (d-q) reference frame rotating at the angular speed of $\omega_s$ is shown in figure (1). The stator and rotor flux are given, respectively, by:

$$\Psi_s = L_s i_s + L_m i_r$$,  $$\Psi_r = L_r i_r + L_m i_s$$ \hspace{1cm} \text{Equation (1)}$$

From figure (1), the stator and rotor voltages in the arbitrary reference frame can be expressed, respectively, as:

$$V_s = R_s i_s + \frac{d}{dt} \psi_s + j \omega_s \psi_s$$ \hspace{1cm} \text{Equation (2)}$$

$$V_r = R_r i_r + \frac{d}{dt} \psi_r + j (\omega_s - \omega_r) \psi_r$$ \hspace{1cm} \text{Equation (3)}$$

According to Equation (1), the rotor flux and stator current can be expressed, respectively, as:

$$\psi_r = \frac{L_m}{L_s} \psi_s + \sigma L_r i_r$$,  $$i_s = \frac{1}{L_s} (\psi_s - L_m i_r)$$ \hspace{1cm} \text{Equation (4)}$$

Substituting Equation (4) into Equation (3) yields the rotor voltage in the arbitrary rotating reference frame as:

$$V_r = R_r i_r + \sigma L_r \frac{d}{dt} i_r + \frac{L_m}{L_s} \frac{d}{dt} \psi_s + j (\omega_{\text{slip}}) (\sigma L_r i_r + \frac{L_m}{L_s} \psi_s)$$ \hspace{1cm} \text{Equation (5)}$$

Where $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ is the leakage factor. In equations Equation (1) to Equation (5) $V_s$, $V_r$ are terminal voltages of stator and rotor, $i_s$, $i_r$ are output currents of stator and rotor, $\psi_s$, $\psi_r$ are leakage flux of stator and rotor, $R_s$, $R_r$ are the resistance of stator and rotor, $L_s$, $L_r$ are inductance of stator and rotor and $L_m$ is magnetizing inductance, $\omega_s$, $\omega_r$, $\omega_{\text{slip}}$ are synchronous speed of stator, rotor speed, and slip frequency, respectively[1-4].

1.2.1.1. Balanced grid conditions:

In numerous articles, a detailed model of the DFIG system under a balanced network supply has been studied. Thus, only a brief description is given here. Figure (2) shows the phasor diagram of the variable F, which represents voltage, current, or flux, in the stator flux oriented (d-q), stator ($\alpha-\beta$), and
rotor ($\alpha_r - \beta_r$) reference frames. The transformation between (dq), ($\alpha$-$\beta$) and ($\alpha_r - \beta_r$) reference frames are given by:

$$
\text{F}_{\text{dq}} = \text{F}_{\alpha\beta} e^{-j(\omega_s t)} \quad \text{F}_{\alpha\beta} = \text{F}_{\alpha\beta r} e^{j(\omega_r t)}
$$

Equation (6)

In balanced grid conditions, stator voltage is usually constant, which results in constant stator flux relationship Equation (5) can be expressed as:

$$
V_r = R_{s} i_r + \sigma L_{s} i_r + j \omega_{s} i_r
$$

Equation (7)

Equation (7) in (d-q) reference frame components yields:

$$
\frac{d}{dt} \begin{bmatrix} I_{rd} \\ I_{rq} \end{bmatrix} = \begin{bmatrix} -\frac{R_{r}}{\sigma L_{r}} & \omega_{\text{slip}} \\ -\omega_{\text{slip}} & -\frac{R_{r}}{\sigma L_{r}} \end{bmatrix} \begin{bmatrix} I_{rd} \\ I_{rq} \end{bmatrix} + \frac{1}{\sigma L_{r}} \begin{bmatrix} V_{rd} \\ V_{rq} \end{bmatrix}
$$

Equation (8)

The stator's output active power ($p_s$) and reactive power ($q_s$) can be calculated as:

$$
p_s + jq_s = \frac{3}{2} V_{sdq} \chi_{dq} = \frac{3}{2} j \omega_s \psi_s \chi (\psi_s - L_m I_r)
$$

$$
= \frac{3}{2} \omega_s (\psi_s - L_m I_r) [j \dot{\psi}_s (\psi_s - L_m I_r)]
$$

Equation (9)

Thus, the stator active and reactive powers are given by

$$
p_s = \frac{3}{2} \omega_s \psi_s L_m I_r, \quad q_s = -\frac{3}{2} \omega_s \psi_s \chi (\psi_s - L_m I_r)
$$

Equation (10)

1.2.1.2. Unbalanced grid conditions:

Under unbalanced conditions, variable F in ($\alpha$-$\beta$) reference frame can be described as follows:

$$
\text{F}_{\alpha\beta} = \text{F}_{\alpha\beta+} + \text{F}_{\alpha\beta-} = \text{F}_{\alpha\beta+} e^{j(\omega_{s} t + \phi_s)} + \text{F}_{\alpha\beta-} e^{-j(\omega_{s} t + \phi_s)}
$$

Equation (11)

Where $\phi_s$ and $\phi_c$ are phase shifts in positive and negative sequences [1].

As shown in figure (2), the conversions between the reference frame positive sequence (dq+) and negative (dq−) are given as follows:

$$
\text{F}_{dq+} = \text{F}_{dq} e^{-2j\omega_s t}, \quad \text{F}_{dq-} = \text{F}_{dq} e^{2j\omega_s t}
$$

Equation (12)
According to Equation (11), variable F is the sum of the positive and negative sequences. Therefore, in the reference frame \( (dq^+) \), variable F in Equation (12) can be written as follows:

\[
F_{dq}^+ = F_{dq_+}^+ + F_{dq_+}^+ e^{-2j\omega_s t} \quad \text{Equation (13)}
\]

According to equation (13), the positive component is a constant value and the negative component oscillates with frequency \( (2\omega_s) \) [5-7]. Equation (13) can be expressed as a constant value and an oscillatory value:

\[
F_{dq}^+ = F_{dq_+}^+ + F_{dq_+}^+ \cos(2\omega_s t) - F_{dq_+}^+ \sin(2\omega_s t) \quad \text{Equation (14)}
\]

The stator's output active power and reactive power in the positive sequence reference frame is:

\[
p_{s+} = \frac{3}{2} V_{sdq_+} I_{sdq_+} - \frac{3}{2} (V_{sq_-}^+ I_{sdq_+}^+ + I_{sdq_-}^+ V_{sq_-}^+)
\]

\[
\text{Equation (15)}
\]

Where

\[
V_{sdq_+}^+ = V_{sdq_+}^+ + \frac{d}{dt} V_{sdq_+}^+ = -\omega_s (\psi_{sq_+}^+ + \psi_{sdq_+}^+ \cos(2\omega_s t) - \psi_{sq_-}^+ \sin(2\omega_s t)) + \frac{d}{dt} (\psi_{sdq_+}^+ - \psi_{sq_+}^+ \cos(2\omega_s t) + \psi_{sq_-}^+ \sin(2\omega_s t))
\]

\[
= -\omega_s \psi_{sq_+}^+ + \omega_s \psi_{sdq_+}^+ \cos(2\omega_s t) - \omega_s \psi_{sq_-}^+ \sin(2\omega_s t) \quad \text{Equation (16)}
\]

\[
I_{sdq_+}^+ = \frac{1}{L_s} \left[ \left( \psi_{sdq_+}^+ - L_m I_{rdq_+}^+ \right) - \frac{1}{L_s} \left( \left( \psi_{sdq_+}^+ - L_m I_{rdq_+}^+ \right) - \psi_{sq_+}^+ \cos(2\omega_s t) - \psi_{sq_-}^+ \sin(2\omega_s t) \right) \right]
\]

\[
\text{Equation (17)}
\]

\[
V_{sq_-}^+ = \psi_{sq_-}^+ - \frac{d}{dt} \psi_{sq_-}^+ = \frac{d}{dt} (\psi_{sdq_-}^+ - \psi_{sq_-}^+ \cos(2\omega_s t) + \psi_{sq_-}^+ \sin(2\omega_s t)) + \frac{d}{dt} (\psi_{sq_-}^+ - \psi_{sdq_-}^+ \cos(2\omega_s t) - \psi_{sq_-}^+ \sin(2\omega_s t))
\]

\[
= \psi_{sq_-}^+ - \psi_{sq_-}^+ \cos(2\omega_s t) - \psi_{sq_-}^+ \sin(2\omega_s t) \quad \text{Equation (18)}
\]

\[
I_{sq_-}^+ = \frac{1}{L_s} \left[ \left( \psi_{sq_-}^+ - L_m I_{rq_+}^+ \right) - \frac{1}{L_s} \left( \left( \psi_{sq_-}^+ - L_m I_{rq_+}^+ \right) - \psi_{sq_-}^+ \cos(2\omega_s t) + \psi_{sq_-}^+ \sin(2\omega_s t) \right) \right]
\]

\[
\text{Equation (19)}
\]

In DFIG, we assume that the stator resistance is negligible. Therefore, in reference \( (dq^+) \):
\[ V_{sdq}^+ = \frac{d}{dt} \Psi_{sdq}^+ + j\omega_s \Psi_{sdq}^+ \]
\[ = \frac{d}{dt} (\Psi_{sdq}^+ + j\Psi_{sdq}^+ e^{-2j\omega_s t} + j\omega_s (\Psi_{sdq}^+ + j\Psi_{sdq}^+ e^{-2j\omega_s t})) \]
\[ = j\omega_s \Psi_{sdq}^+ + j\omega_s \Psi_{sdq}^+ e^{-2j\omega_s t} = V_{sdq}^- + V_{sdq}^- e^{-2j\omega_s t} \]

Therefore:

\[ \begin{cases} V_{sdq}^+ = j\omega_s \Psi_{sdq}^+ & \Rightarrow (V_{sq}^+ + jV_{sd}^+) = j\omega_s (\Psi_{sq}^+ + j\Psi_{sd}^+) \\ V_{sdq}^- = j\omega_s \Psi_{sdq}^- & \Rightarrow (V_{sq}^- + jV_{sd}^-) = -j\omega_s (\Psi_{sq}^- + j\Psi_{sd}^-) \end{cases} \]

And:

\[ \Psi_{sd}^+ = \frac{1}{\omega_s} V_{sq}^+ + \Psi_{sdq}^+ = \frac{1}{\omega_s} V_{sd}^+ + \Psi_{sdq}^+ = \frac{1}{\omega_s} V_{sq}^- - \Psi_{sq}^- = \frac{1}{\omega_s} V_{sd}^- - \Psi_{sd}^- \]

According to equation (14), although unbalanced, the stator voltage can still be regarded as constant (\( F_{d+}^+ \) and \( F_{d+}^- \)). Consequently:

\[ \frac{d}{dt} (\Psi_{sd}^+) = 0, \frac{d}{dt} (\Psi_{sd}^-) = 0, \frac{d}{dt} (\Psi_{sq}^+) = 0, \frac{d}{dt} (\Psi_{sq}^-) = 0 \]

By replacing relations equation (22) into equations (16, 17, 18 and 19) and considering Relation (23), active and reactive power can be expressed as:

\[ P_s = P_{s-av} + P_{s-sin^2(2\omega_s t)} + P_{s-cos^2(2\omega_s t)}, Q_s = Q_{s-av} + Q_{s-sin^2(2\omega_s t)} + Q_{s-cos^2(2\omega_s t)} \]

Considering that in VOC control \( V_{sq}^+ = 0 \), the active and reactive matrix form can be described as [4]:

\[ \begin{bmatrix} P_{s-av} \\ P_{s-sin^2} \\ P_{s-cos^2} \end{bmatrix} = \sum_{m=1}^{M} \begin{bmatrix} 0 \\ -2V_{sd}^+ V_{sd}^- \\ 2V_{sq}^- V_{sd}^+ \end{bmatrix} + \begin{bmatrix} V_{sd}^+ \\ V_{sq}^- \\ V_{sd}^- \end{bmatrix} \]

\[ = \begin{bmatrix} 3L \frac{m}{s} & \frac{m}{s} & \frac{m}{s} \\ \frac{m}{s} & \frac{m}{s} & \frac{m}{s} \end{bmatrix} \begin{bmatrix} V_{sd}^+ \\ V_{sq}^- \\ V_{sd}^- \end{bmatrix} \]

\[ \begin{bmatrix} 0 \\ -V_{sd}^- V_{sd}^+ \\ V_{sq}^+ V_{sd}^- \end{bmatrix} \]

\[ \begin{bmatrix} I_{rd}^+ \\ I_{rq}^+ \\ I_{rd}^- \end{bmatrix} \]

\[ \begin{bmatrix} I_{rd}^- \\ I_{rq}^- \end{bmatrix} \]

\[ \text{Equation (25)} \]
\[
\begin{bmatrix}
Q_{s-av} \\
Q_{s-sin2} \\
Q_{s-cos2}
\end{bmatrix}
= \frac{3}{2\omega L_s} \begin{bmatrix}
-(V_{sd+})^2 + (V_{sd-})^2 + (V_{sq-})^2 \\
0 \\
0
\end{bmatrix}
\begin{bmatrix}
I_{rd+} \\
I_{rq+} \\
I_{rd-} \\
I_{rq-}
\end{bmatrix}
\]  
Equation (26)

The electromagnetic torque \( (T_e) \) of a DFIG can be described as:

\[
T_e = \frac{3}{2} \text{Plm}[\psi_s(t) i_s^*(t)] = \frac{3}{2} \text{Plm} \left[ \psi_s(t) \psi_s(t) - L_{mR}(t) \right] = \frac{3\text{Plm}}{2L_s} \text{Im}[\psi_s(t) i_s^*(t)]
\]  
Equation (27)

\( (P \text{ is pole pair}) \), where the matrix form of electromagnetic torque is:

\[
\begin{bmatrix}
T_{e-av} \\
T_{e-sin2} \\
T_{e-cos2}
\end{bmatrix}
= \frac{3L_m}{2L_s} \frac{\begin{bmatrix}
V_{sd+} & 0 & -V_{sd-} \\
-\psi_{sq-} & V_{sd-} & 0 \\
-V_{sq-} & \psi_{sq-} & V_{sd+}
\end{bmatrix}}{\begin{bmatrix}
I_{rd+} \\
I_{rq+} \\
I_{rd-} \\
I_{rq-}
\end{bmatrix}}
\]  
Equation (28)

1.2.2. Control of rotor side converter (RSC)
1.2.2.1. Balanced conditions:

Equation (10) indicates active power controlled by regulating \( (I_{rq}) \) and reactive power can be controlled by regulating \( (I_{rd}) \). As a result, the active and reactive power can be controlled independently using rotor current \( (I_{rq}) \) and \( (I_{rd}) \), respectively. According to (8), the required rotor control voltages in the \( (d-q) \) reference frame are given by [8-9]:

\[
\begin{bmatrix}
V_{rd} \\
V_{rq}
\end{bmatrix}
= \sigma L_r \frac{d}{dt} \begin{bmatrix}
I_{rd} \\
I_{rq}
\end{bmatrix}
- \begin{bmatrix}
-R_r \\
\sigma L_r
\end{bmatrix}
\begin{bmatrix}
\omega_{\text{slip}} \\
\omega_{\text{slip}}
\end{bmatrix}
I_{rd}
- \frac{L_s \omega_{\text{slip}}}{L_m}
\begin{bmatrix}
\psi_{sq} \\
\psi_{sq}
\end{bmatrix}
\]
Equation (29)

Where:

\[
\frac{d}{dt} \begin{bmatrix}
I_{rd} \\
I_{rq}
\end{bmatrix}
= \frac{d}{dt} (I_{rdq} - I_{rdq}) (K_{pd} + K_{id}) (\omega_{\text{rdq}} - \omega_{\text{rdq}})
\]
Equation (30)

Where \( K_{pd} \) and \( K_{id} \) are the proportional and integral gains of the current controllers.

1.2.2.2. Unbalanced Conditions:
Under unbalance grid conditions, the negative sequence component of stator voltage leads to the oscillation of power (active-reactive) and electromagnetic torque with frequency \(2\omega_s\). Power fluctuations lead to increased temperature in the stator winding, and torque fluctuations lead to mechanical tension on the rotor. In order to make the steady-state error value zero and improve the accuracy of controlling, proportional-integral (PI) controller is used. As the simultaneous elimination of power and torque oscillations is not possible, the controller is set to Target 1 (Eliminating fluctuation of stator active power) and Target 2 (Eliminating fluctuation of electromagnetic torque).

### 1.2.2.2.1. Target 1 (Eliminating fluctuation of stator active power):

In this case, the stator winding temperature is increased. According to equation (25), the required negative sequence rotor current leads to \(P_{s-sin2} = 0\), \(P_{s-cos2} = 0\) is given as:

\[
P_{s-sin2} = 0 \Rightarrow I^{\text{(ref)}}_{rq} = -\frac{1}{V_{sd+}^+} \left( V_{sd-}^+ I_{rd+}^- - V_{sq-}^- I_{rd+}^+ \right) + \frac{2V_{sq-}^-}{\omega L s ms} \]

Equation (31)

\[
P_{s-cos2} = 0 \Rightarrow I^{\text{(ref)}}_{rd-} = -\frac{1}{V_{sd+}^+} \left( V_{sd-}^+ I_{rd+}^- + V_{sq-}^- I_{rq+}^- \right) - \frac{2V_{sq-}^-}{\omega L s ms} \]

Equation (32)

Accordingly, power (active-reactive) and the electromagnetic torque are given by:

\[
P_{s_{-av}} = \frac{3L_m}{2L_s} \left( V_{sd+}^+ \right)^2 \left( V_{sq-}^- \right)^2 I_{rd+}^+ \]

Equation (33)

\[
Q_{s_{-av}} = -\frac{3}{2L_s} \left( V_{sd-}^- \right)^2 \left( V_{sq-}^- \right)^2 \left( V_{sd+}^+ \right)^2 \left( V_{sq+}^+ \right) \left( V_{sd+}^+ \right)^2 \left( V_{sq-}^- \right)^2 \left( V_{sq+}^+ \right) \left( V_{sq+}^+ \right) \]

Equation (34)

\[
T_{e_{-av}} = \frac{3PL_m}{2\omega s L_s} \left( V_{sd+}^+ \right)^2 \left( V_{sq-}^- \right)^2 I_{rd+}^+ \]

Equation (35)

### 1.2.2.2.2. Target 2 (Eliminating fluctuation of electromagnetic torque):
In this case, mechanical tension on the rotor decreases. According to equation (28), the required negative sequence rotor current leads to \((T_\text{e-sin}^2 = 0, T_\text{e-cos}^2 = 0)\) is given as:

\[
T_\text{e-sin}^2 = 0 \Rightarrow I_{\text{rq}}^- (\text{ref}) = \frac{1}{\sqrt{V_{sd}^+}} \left( V_{sq}^- I_{rd}^+ V_{sd}^- I_{rq}^+ \right)
\]

Equation (36)

\[
T_\text{e-cos}^2 = 0 \Rightarrow I_{\text{rd}}^- (\text{ref}) = \frac{1}{\sqrt{V_{sd}^+}} \left( V_{sq}^- I_{rd}^+ + V_{sq}^- I_{rq}^+ \right)
\]

Equation (37)

Therefore, power (active-reactive) and the electromagnetic torque are given by:

\[
P_{\text{e-av}} = \frac{3L}{2m} \left( \frac{(V_{sd}^+)^2 + (V_{sq}^-)^2 + (V_{sq}^-)^2}{V_{sd}^+} \right) I_{rd}^+
\]

Equation (38)

\[
Q_{\text{s-av}} = -\frac{3}{2L} \left( \frac{(V_{sq}^-)^2 + (V_{sq}^-)^2 + (V_{sq}^-)^2}{V_{sd}^+} \right) \left\{ V_{sd}^+ + L \frac{\omega}{m} I_{rq}^+ \right\}
\]

Equation (39)

\[
T_{\text{e-av}} = \frac{3PL}{2\omega_s L_s} \left( \frac{(V_{sd}^+)^2 + (V_{sq}^-)^2 + (V_{sq}^-)^2}{V_{sd}^+} \right) I_{rd}^+
\]

Equation (40)

Equations (33, 34, 35) and (38, 39, 40) express active power and electromagnetic torque controlled by regulating \((I_{rd}^+)^2\) and reactive power can be controlled by regulating \((I_{rq}^+)^2\). In the mean while controlling \((I_{rd}^-)^2\) and \((I_{rq}^-)^2\) can eliminate the oscillations of either stator active power based on equation (31, 32) or electromagnetic torque based on equation (36, 37). As a result, the average power and electromagnetic torque can be controlled independently using the rotor current positive sequence component \((I_{rd}^+, I_{rq}^+)\) and the oscillation can be eliminated by using the rotor current negative sequence component \((I_{rd}^-, I_{rq}^-)\). Figure (3) shows the block diagram phase-locked loop (PLL). The stator’s voltage frequency \((\omega_s)\), the phase angle of the stator voltage \((\theta_s)\), negative, and positive sequence components of stator voltage and the negative and positive sequence components of the
rotor current can be measured by PLL. Under unbalance grid conditions, to reduce the effects of the negative sequence component, the band stop filter (B.S.F) with twice the grid frequency is used. Figure (4) shows the block diagram of a voltage orientation control (VOC) for RSC control under unbalanced grid. The purpose of control RSC is to eliminate power fluctuations (Target 1) and torque fluctuations (Target 2). To achieve Target 1, control components' positive and negative sequences can be calculated by equation (33, 34, 35) and (31, 32), respectively. To achieve Target 2, components' positive and negative sequences can be calculated by equation (38, 39, 40) and (36, 37), respectively.

Where:

\[ V_{rd} = \frac{d}{dt}(\Psi_{rd}) - (\omega_{slip})\Psi_{rq} + R_{r}i_{rd} \]  \hspace{1cm} \text{Equation (41)}

\[ V_{rq} = \frac{d}{dt}(\Psi_{rq}) - (\omega_{slip})\Psi_{rd} + R_{r}i_{rq} \]  \hspace{1cm} \text{Equation (42)}

\[ \Psi_{rd} = L_{r}i_{rd} + L_{m}\frac{i_{sd}}{L_{s}} + L_{m}\left(\Psi_{sd} - L_{m}i_{rd}\right) = \sigma L_{r}i_{rd} + \frac{L_{m}}{L_{s}}\Psi_{sd} \]  \hspace{1cm} \text{Equation (43)}

\[ \Psi_{rq} = L_{r}i_{rq} + L_{m}\frac{i_{sq}}{L_{s}} + L_{m}\left(\Psi_{sq} - L_{m}i_{rd}\right) = \sigma L_{r}i_{rq} + \frac{L_{m}}{L_{s}}\Psi_{sq} \]  \hspace{1cm} \text{Equation (44)}

With substituting equations (41) and (42) into equations (43) and (44) control voltages are given by:

\[ V_{rd+}^{+}(\text{ref}) = \sigma L_{r}i_{rd} + \frac{\omega_{slip}}{L_{s}}L_{m}V_{sd+} \]  \hspace{1cm} \text{Equation (45)}

\[ V_{rq+}^{+}(\text{ref}) = \sigma L_{r}i_{rq} + \frac{\omega_{slip}}{L_{s}}L_{m}V_{sq+} \]  \hspace{1cm} \text{Equation (46)}

\[ V_{rd-}^{-}(\text{ref}) = \sigma L_{r}i_{rd} - \frac{\omega_{slip}}{L_{s}}L_{m}V_{sd-} \]  \hspace{1cm} \text{Equation (47)}
\[ V_{\text{rq}}\text{-(ref)}=\alpha L_r \frac{d}{dt} i_{\text{rq}}+V_{\text{rq}}^+ V_{\text{rq}}^- = (\omega_{\text{slip}}) \alpha L_r \frac{d}{dt} i_{\text{rd}} \frac{\omega_{\text{slip}} L_m}{s} V_{\text{sq}}^- \]  
Equation (48)

\[ \frac{d}{dt} (I_{\text{rdq}})= (I_{\text{rdq}} \text{-ref})L_{\text{rdq}}(k_{pd}+k_{id}) \]  
Equation (49)

Where \( k_{pd} \) and \( k_{id} \) are the proportional and integral gains of the current controllers.

### 1.2.3. Measure the rotor speed:

To measure the rotor speed of DFIG, a RPCA is used. In this algorithm, without the need for estimating the stator flux, the rotor position is measured by means of measured stator as well as rotor voltages and current. This algorithm is simple and effective and does not require a direct estimation of the stator flux. One of the most important advantages of this scheme is that the calculations do not use any approximation. This algorithm is robust against variations of stator and rotor resistances and is only dependent on the estimation of the DFIG mutual inductance [16].

The stator flux in the stationary reference frame (\( \alpha - \beta \)) axis can be expressed as:

\[ \Psi_{s\alpha}=L_s i_{s\alpha}+L_m i_{s\beta} \Psi_{s\beta}=L_s i_{s\beta}+L_m i_{s\alpha} \]  
Equation (50)

The stator voltage components in the stationary reference frame can be expressed as:

\[ V_{s\alpha}=R_s i_{s\alpha}+\frac{d}{dt} \psi_{s\beta} V_{s\beta}=R_s i_{s\beta}+\frac{d}{dt} \psi_{s\alpha} \]  
Equation (51)

Therefore:

\[ i_{r\beta} = \int \frac{(v_{s\alpha}-R_s i_{s\alpha})dt-L_s i_{s\beta}}{L_m} \]  
and \[ i_{r\alpha} = \int \frac{(v_{s\beta}-R_s i_{s\beta})dt-L_s i_{s\alpha}}{L_m} \]  
Equation (52)

In figure (5), the rotor current space vector makes the angle \( \theta_1 \) with respect to the \( \alpha \)-axis of the stator reference frame and angle \( \theta_2 \) with respect to the \( \alpha \)-axis of the rotor reference frame. The difference between \( \theta_1 \) and \( \theta_2 \) is \( \epsilon \) and can be expressed as:

\[ \theta_1 = \text{atan2}(\frac{i_{r\beta}}{i_{r\alpha}}), \theta_2 = \text{atan2}(\frac{i_{r\beta}}{i_{r\alpha}}), \epsilon = \theta_1 - \theta_2 = \text{atan2}(\frac{i_{r\beta}i_{r\alpha}-i_{r\alpha}i_{r\beta}}{i_{r\alpha}i_{r\alpha}+i_{r\beta}i_{r\beta}}) \]  
Equation (53)

Where:
\[
\begin{align*}
i_{r_a} = & \frac{2}{3} (i_{ra} - \frac{i_{rb}}{2} - \frac{i_{rc}}{2}), \quad i_{r_b} = \frac{1}{\sqrt{3}} (i_{rb} + \frac{i_{ra}}{2} - \frac{i_{rc}}{2}) \\
i_{r_c} = & \frac{1}{\sqrt{3}} (i_{rc} - \frac{i_{ra}}{2} + \frac{i_{rb}}{2})
\end{align*}
\]  
Equation (54)

Figure (6) shows the block diagram of RPCA algorithm which is used to measure the rotor speed of DFIG.

1.2.4. Control of grid side converter (GSC):

To regulate DC-Link voltage, an ESO, based on a GPI controller, is employed. Figure (7) shows the block diagram of the ESO controller. In this controlling method, the voltage of the DC-Link, is controlled without requiring to measure the current of GSC converter, and because of GPI, the dynamic response is improved, voltage-settling time is reduced, and it is resistant to changes in voltage parameters. Power transfer relationship in GSC can be described as:

\[
\frac{d}{dt} (0.5CV_{DC}^2) = \frac{V_{DC}^2}{R_{loos}} P_r(t) + P_g(t) \Rightarrow V_{DC}^2(S) = \frac{2R_{loos}s}{CR_{loos}s + 2} P_g(s) - \frac{2R_{loos}s}{CR_{loos}s + 2} P_r(s)
\]

Equation (55)

\( R_{loos} \) is the loss of switching, \( V_{DC} \) is voltage DC-Link, \( C \) is the capacity, \( P_r \) is output active power of the rotor, and \( P_g \geq P_{g(ref)} \) where \( P_{g(ref)} \) and \( P_g \) are reference and output power of GSC respectively.

1.2.5. Designing of an extended state observer (ESO)

Figure (8) shows the block diagram of voltage control of the DC-Link voltage by GSC. This diagram consists of an internal loop for controlling converter current and an external loop for controlling and regulating the DC-Link voltage [25-26]. In order to prevent dynamic interference between internal and external loops, the bandwidth of the internal loop is considered much larger than that of the external loop. Equation (55) can be rewritten as:

\[
\frac{d}{dt} (V_{DC}^2) = \frac{2}{C} P_g(t) - \frac{V_{DC}^2}{R_{loos}} P_r(t) \Rightarrow \frac{2}{C} P_g(t) + f_{total}
\]

Equation (56)

Where \( f_{total} = \frac{2}{C} \left( \frac{V_{DC}^2}{R_{loos}} + P_r(t) \right) \) is a total disturbance that consists of the external disturbance \( -\frac{2}{C} (P_r(t)) \) and internally dynamic variation \(-\frac{2}{C} \left( \frac{V_{DC}^2}{R_{loos}} \right) \). In (56), \( V_{DC}^2 \) is considered as a state
variable and expressed $x_1 = \sqrt{\frac{2}{C}} \cdot f_{\text{total}}$ is considered as based on augmented input estimation (AIE) of state variable $x_2$ and expressed\cite{27-28} as: $x_2 = f_{\text{total}} = -\frac{2}{C} \left( \frac{V_{\text{DC}}^2}{R_{\text{loos}}} + P(t) \right)$. $P(t)$ is the system input and indicated as: $u = P_g(t) \cdot b_0 = \frac{2}{C}$ is the coefficient of input. Moreover, the derivative time of $x_2$ is signified $h$, and the expression is shown as: $\frac{dx_2}{dt} = h$. Therefore, from the aforementioned analysis, the state-space model is derived as:

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} b_0 \\ 0 \end{bmatrix} u + \begin{bmatrix} 0 \\ 1 \end{bmatrix} h$$

Equation (57)

Based on Equation (57), the ESO is constructed as:

$$\begin{bmatrix} \dot{z}_1 \\ \dot{z}_2 \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} z_1 \\ z_2 \end{bmatrix} + \begin{bmatrix} b_0 \\ 0 \end{bmatrix} u + \begin{bmatrix} \beta_1 \\ \beta_2 \end{bmatrix} \begin{bmatrix} x_1 - z_1 \\ x_2 - z_2 \end{bmatrix}$$

Equation (58)

Where $z_1$, $z_2$ are the estimated values of $x_1$, $x_2$, and $\begin{bmatrix} \beta_1 \\ \beta_2 \end{bmatrix}$ is the gain vector of ESO. The state error equation (e_1 and e_2) can be subtracting equation (58) from equation (57), as follows:

$$\begin{bmatrix} \dot{e}_1 \\ \dot{e}_2 \end{bmatrix} = \begin{bmatrix} -\beta_1 & 1 \\ -\beta_2 & 0 \end{bmatrix} \begin{bmatrix} e_1 \\ e_2 \end{bmatrix} + \begin{bmatrix} b_0 \\ 0 \end{bmatrix} u + \begin{bmatrix} 0 \\ 1 \end{bmatrix} h \text{ and } \begin{bmatrix} -\beta_1 & 1 \\ -\beta_2 & 0 \end{bmatrix} = H_e$$

Equation (59)

From (59), it is known that the matrix ($H_e$) will be Routh-Hurwitz stable if all the roots of the characteristic polynomial of $H_e$, i.e.:

$$\lambda(s) = s^2 + \beta_1 s + \beta_2$$

Equation (60)

Are in the left half-plane. Suppose all poles of ESO are designed to stay on bandwidth ($-\omega_0$), which is indicated below\cite{29}:

$$\lambda(s) = s^2 + \beta_1 s + \beta_2 = (s + \omega_0)^2 \rightarrow \beta_1 = 2\omega_0, \quad \beta_2 = \omega_0^2$$

Equation (61)

From equation (61), it is shown that the design of ESO is simplified to tune the ($\omega_0$) of the ESO, which greatly simplifies the design process. Choosing ($\omega_0$) is an essential factor that affects the system’s function. Observer’s bandwidth is usually chosen 5 to 15 times the bandwidth of the controller of the DC-Link voltage. Therefore, the dynamics of the estimated state has fast tracking
performance. Because high bandwidth reduces safety versus system noise, the bandwidth of ESO cannot be very high. In order to achieve a fast dynamic response, the bandwidth of the current loop is chosen 300 rad/s, and the voltage controller of the DC-Link is chosen 20 rad/s. As a result, all three loops are separated, and ESO has a high dynamic response for estimating the real state.

By substituting $\beta_1=2\omega_0$ and $\beta_2=\omega_0^2$ in equation (58), the ESO is described as:

$$
\begin{bmatrix}
  z_1 \\
  z_2
\end{bmatrix} =
\begin{bmatrix}
  -2\omega_0 & 1 \\
  -\omega_0^2 & 0
\end{bmatrix}
\begin{bmatrix}
  z_1 \\
  z_2
\end{bmatrix}
+ \begin{bmatrix}
  b_0 \\
  0
\end{bmatrix}
\begin{bmatrix}
  2\omega_0 \\
  \omega_0^2
\end{bmatrix}
\begin{bmatrix}
  u \\
  x_1
\end{bmatrix}
$$

Equation (62)

Equation (62) is corresponding in the Laplace domain to (figure 9):

$$
\hat{G}_{f-u(s)} = \frac{f_{total}(s)}{u(s)} = [0 \quad 1][SI-A_z]^{-1} \begin{bmatrix}
  b_0 \\
  0
\end{bmatrix} = \frac{b_0\omega_0^2}{(s+\omega_0)^2}
$$

Equation (63)

$$
\hat{G}_{f-V_{DC}^2} = \frac{f_{total}(s)}{V_{DC}^2} = [0 \quad 1][SI-A_z]^{-1} \begin{bmatrix}
  2\omega_0 \\
  \omega_0^2
\end{bmatrix} = \frac{s\omega_0^2}{(s+\omega_0)^2}
$$

Equation (64)

By combining equation (63) and equation (64), the transfer function of ESO would be as follows:

$$
\hat{f}_{total}(s) = \frac{b_0\omega_0^2}{(s+\omega_0)^2} u(s) + \frac{s\omega_0^2}{(s+\omega_0)^2} V_{DC}^2
$$

Equation (65)

1.3. Results and Discussion:

1.3.1. Simulation results:

Simulations of the proposed control strategy, for a DFIG-based generation system, were carried out using Matlab/Simulink. Figure 10, shows the schematic diagram of the implemented system. The DFIG was rated at (2 MW) and its parameters are given in Table I. The nominal converter DC-link voltage was set at (1200 V), the switching frequencies for both converters were (2 kHz), the rotor speed is (1.1pu) and wind speed is (16.2m/s). As shown in figure (10), to absorb the switching harmonics generated by the two converters to the stator side a high-frequency AC filter is connected. The main objective of the RSC is to control active-reactive power and torque, in this paper in order to achieve zero state error it was controlled using a proportional-integral (PI) controller. In [10] for controlling the performance of the RSC sliding mode controller with fractional order (SMC.F) based on robust control was used. In this controller to increase the degree of freedom and further robustness of the controller, the sliding surface is selected as a fractional order form. Using a robust sliding mode
controller in addition to increasing the performance of the controller leads to reduce the chattering phenomenon in the input control signal. Using the decoupling of the current loops approach as a very powerful design tool for nonlinear systems makes the designed controller more robust against incoming disturbances to the system. In [11] for controlling performance of RSC nonlinear robust control approach an ESO-BS has been used. This controller is designed by combining the advantages of ESO with those of the backstepping theory. This controller has the advantages of rapid response and insensitivity to disturbances. Then, ESO is constructed to compensate for model error and unknown disturbances, thereby reducing the calculation burden. This controller has a simple design and small computational complexity and it has a very fast transient response that effectively eliminates the overcurrent in the RSC and eliminates electromagnetic torque in DFIG fluctuations. In this paper to prove the results of the simulation, control of RSC with a PI controller is compared with the methods mentioned in references [10] and [11]. In the simulation, unbalanced grid conditions as a short circuit single line to ground fault in time duration at [3s-3.2s] was applied. Under unbalanced grid conditions for the conventional control scheme, it can be seen from figure (11) voltage, current of stator and rotor current becomes unbalanced. Figure (12) shows the positive component of stator voltage (Vspd) and the negative sequence component of stator voltage (Vsnd, Vsnq) conventional control. Also, figure (13) shows the positive component of rotor current (Irpd, Irpq) and negative sequence component of rotor current (Irnd, Irnq) conventional control. Under unbalance grid conditions, the negative sequence component of stator voltage and rotor current lead to the oscillation of power (active-reactive) and electromagnetic torque with frequency \(2\omega_s=100\text{HZ}\). Figure (14) shows the power (active-reactive) and electromagnetic torque conventional control. Power fluctuations lead to increased temperature in the stator winding and torque fluctuations lead to mechanical tension on the rotor. To reduce the effects caused by the negative sequence component, the band stop filter is employed with twice the grid frequency. Figure (15) shows the positive and negative sequence components of stator voltage proposed control and figure (16) shows the positive and negative sequence components of the rotor current after passing from the band stop filter. As the simultaneous elimination of power and torque oscillations by RSC is not possible, the controller of
RSC is set to Target 1 (Eliminating fluctuation of stator active power) during [3s-3.1s] and switched to Target 2 (Eliminating fluctuation of electromagnetic torque) during [3.1s-3.2s]. As can be seen in figure (17) oscillations of the active power and torque have been reduced significantly during [3s-3.1s] and [3.1s-3.2s] respectively. Figures (18-23) show control steps of RSC control with SMC-F and figures (24-29) show control steps of RSC control with ESO-BS. From the comparison of the results control of RSC by PI, SMC-F, and ESO-BS, it was found that under unbalanced grid conditions when ESO-BS was used the amplitude of negative sequence component of stator voltage, rotor current and fluctuations range of power and torque was less than the state, in which the SMC-F and (PI) controller were used. Thus control of DFIG’s RSC with ESO.BS controller due to avoids complex derivation of adaptive backstepping had better behavior than when PI and SMC-F controller were used. The main objective of the GSC is to control the DC-link voltage, and it was controlled using an ESO based on a GPI controller. To prove the results of the simulation, control of GSC with ESO controller is compared with the methods PI, SMC-F, ESO-BS and proportional (P) controller. Figure (30) shows the DC-link voltage and figure (31) shows the variations of DC-link voltage during an unbalanced grid. As can be seen in the ESO proposed to control for GSC control, the maximum overshoot of DC-Link voltage in this controller is less than the maximum overshoot of DC-Link voltage when ESO-BS, SMC-F, PI, P controller was employed. Also in ESO controller, due to using the GPI controller, settling time was reduced, and voltage changes in duration [3s-3.2s] was less than when the PI, SMC.F, and P controller was employed. Therefore, the simulation results confirmed that the ESO controller for GSC control had a better behavior than in the other controller. In order to calculate the rotor speed, a versatile RPCA was employed. Figure (32) shows the mechanical speed of the rotor and figure (33) shows rotor speed changes duration at [3s-3.2s]. As can be seen, this algorithm is resistant to changes in stator voltage and was stable under grid conditions.

1.4. Conclusion:

Under unbalance grid conditions, the negative sequence component of stator voltage leads to the oscillation of power (active-reactive) and electromagnetic torque with twice the grid frequency. The controlling RSC aims to eliminate power and torque oscillations. In order to regulate DC-Link
voltage, an ESO is employed. In this controlling method, DC-Link voltage was controlled without measuring the GSC current and, due to using a GPI controller, the improved dynamic response was resistant against voltage changes, and settling time was reduced. To measure the rotor speed, a versatile RPCA was employed. This algorithm was resistant to changes in stator voltage and was stable under grid conditions. Simulation results confirmed that controlling DFIG’s RSC control with an ESO-BS controller had a better behavior than when PI controller or Sliding mode controller with fractional order sliding surface (SMC-F) controller were used. However, the ESO proposed controller for GSC control had a better behavior than ESO-BS, SMC-F, PI, and the proportional (P) controller, which were compared.

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Abbreviations:

B.S.F: band stop filter

DFIG: doubly-fed induction generator

ESO: expanded state observer

ESO-BS: extended state observer based on the backstepping controller

FOC: flux orientation control

GPI: generalized proportional-integral

GSC: grid side converter

PI: proportional-integral

RPCA: rotor position computation algorithm

RSC: rotor side converter

SMC.F: sliding mode controller with fractional order

SMC-F: sliding surface

VOC: voltage orientation control

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**Figures:**

Figure 1: DFIG (doubly-fed induction generator) model in (d-q) reference frame

Figure 2: Diagram of reference frame, positive (dq⁺) and negative (dq⁻)

Figure 3: Block diagram of the phase-locked loop (PLL)

Figure 4: Block diagram of RSC control

Figure 5: Rotor current space vector

Figure 6: Block diagram of RPCA (rotor position computation algorithm) algorithm

Figure 7: Block diagram of the ESO (expanded state observer)

Figure 8: Block diagram of control voltage DC-Link

Figure 9: Block diagram equivalent transfer function of ESO

Figure 10: Schematic diagram of the simulated system

Figure 11: Voltage, current of stator and rotor current

Figure 12: Stator voltage sequence conventional control with PI controller

Figure 13: Rotor current sequence conventional control with PI controller

Figure 14: (Ps - Qs) and Te conventional control with PI controller

Figure 15: Stator voltage sequence proposed control with PI controller

Figure 16: Rotor current sequence proposed control with PI controller

Figure 17: (Ps - Qs) and Te proposed control with PI controller

Figure 18: Stator voltage sequence conventional control with SMC-F (sliding mode controller with fractional order sliding surface) controller

Figure 19: Rotor current sequence conventional control with SMC-F (sliding mode controller with fractional order sliding surface) controller

Figure 20: (Ps - Qs) and Te conventional control with SMC-F (sliding mode controller with fractional order sliding surface) controller

Figure 21: Stator voltage sequence proposed control with SMC-F (sliding mode controller with fractional order sliding surface) controller

Figure 22: Rotor current sequence proposed control with SMC-F (sliding mode controller with fractional order sliding surface) controller

Figure 23: (Ps - Qs) and Te proposed control with SMC-F (sliding mode controller with fractional order sliding surface) controller
Figure 24: Stator voltage sequence conventional control with ESO-BS (extended state observer based on the backstepping controller)

Figure 25: Rotor current sequence conventional control with ESO-BS (extended state observer based on the backstepping controller)

Figure 26: (P_s - Q_s) and T_e conventional control with ESO-BS (extended state observer based on the backstepping controller)

Figure 27: Stator voltage sequence proposed control with ESO-BS (extended state observer based on the backstepping controller)

Figure 28: Rotor current sequence proposed control with ESO-BS (extended state observer based on the backstepping controller)

Figure 29: (P_s - Q_s) and T_e proposed control with ESO-BS (extended state observer based on the backstepping controller)

Figure 30: Voltage DC-link

Figure 31: Variation voltage DC-link during unbalanced grid

Figure 32: Rotor speed by RPCA (rotor position computation algorithm) algorithm

Figure 33: Variation rotor speed during unbalanced grid

**Tables:**

Table 1. DFIG (doubly-fed induction generator) parameters
Figures and Tables:

Figure 1: DFIG (doubly-fed induction generator) model in (d-q) reference frame

\[ V_s \rightarrow l_s \rightarrow d \frac{d}{dt} \psi_s \rightarrow L_{ls} \rightarrow L_{tr} \rightarrow R_f \rightarrow V_r \]

\[ j\omega_s \psi_s \rightarrow j(\omega_s - \omega_r) \psi_s \]

Figure 2: Diagram of reference frame, positive (dq+) and negative (dq-)

\[ q^+ \rightarrow \beta \rightarrow q^- \rightarrow F \rightarrow d^+ \rightarrow \omega_s \rightarrow d^- \]

\[ V_s \rightarrow \alpha \rightarrow \beta \rightarrow \alpha \rightarrow \beta \rightarrow \sin(\theta) \rightarrow \cos(\theta) \rightarrow \sin(\theta) \rightarrow \cos(\theta) \]

\[ V_s = \theta_r + s \theta = \theta - \theta_s \]

Figure 3: Block diagram of the phase-locked loop (PLL)
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Figure 28: Rotor current sequence proposed control with ESO-BS (extended state observer based on the backstepping controller)
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Figure 30: Voltage DC-link

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Figure 33: Variation rotor speed during unbalanced grid

Tables:

Table 1. DFIG (doubly-fed induction generator) parameters

| DFIG                      | Rated power          | 2MW                        |
|---------------------------|----------------------|----------------------------|
| voltage /Frequency       | 690V /5Hz            |                            |
| \(R_s/R_r\)               | 0.0108pu /0.12pu     |                            |
| \(L_m\)                   | 3.362pu              |                            |
| \(L_{\sigma s}/L_{\sigma r}\)| 0.102pu /0.11pu      |                            |
| Inertia constant         | 1.5s                 |                            |
| Dc-link Capacitor        | \(C_{dc}\)           | 0.0001F                    |
| Choke                    | \(L_g\)              | 0.25mH                     |
| Filter                   | \(R_f/C_f\)          | 0.06Ω /1000 µF             |