

Article

Sensorless Control of Doubly Fed Induction Machines Using Only Rotor-Side Variables

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Abstract: In this study, a sensorless vector control method was proposed for a doubly fed induction machine (DFIM), where the stator is directly connected to the grid. The DFIM is a three-phase symmetric system without saliency, and when the stator side is directly connected to the grid, the magnitude and frequency of the stator flux are almost fixed and determined by the grid voltage. Due to its three-phase symmetric configuration, this structure can be modeled in a manner similar to that of a symmetric permanent-magnet synchronous motor (PMSM). It enables the application of back-EMF-based sensorless control methods commonly used for symmetric PMSMs. In PMSMs, sensorless estimators typically estimate the back-EMF using only stator voltage and current measurements. By extending this modeling concept to DFIMs, a similar estimator can be designed that utilizes only rotor-side voltage and current for sensorless back-EMF estimation. This paper proposes a back-EMF estimator using only rotor-side voltages and currents, which were implemented on a stator flux reference frame. The proposed algorithm also estimates the stator-side variables, including the magnitudes of stator voltage, current, and stator power factor. These variables can be used to detect grid faults. The feasibility of the proposed method was validated via experiments using a 2.4 kW DFIM. It was confirmed that the sensorless operation functioned properly even during speed acceleration/deceleration and step load conditions. Additionally, the system maintained stable operation and achieved an accurate estimation of stator voltage and current, even under a 30% voltage sag in the stator grid voltage.



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Keywords: doubly fed induction machine; sensorless control

1. Introduction

A doubly fed induction machine (DFIM) offers several advantages in high-power applications such as wind turbine generators. Nevertheless, as the share of conventional synchronous generators declines, the overall system rotational inertia is also reduced [1], posing new challenges for the stability and control of DFIM-based systems. When the stator windings are directly connected to the grid (Figure 1), the DFIM converts mechanical power to electrical power at a constant frequency using the rotor-side converter with a reduced power rating [1–15]. Compared with conventional singly fed electric machines, the DFIM requires additional sensors to be installed on rotor and stator sides; this makes its system configuration more complex and costlier. Therefore, sensorless vector control is highly desirable for DFIM drive systems, contrary to singly fed electric machines.

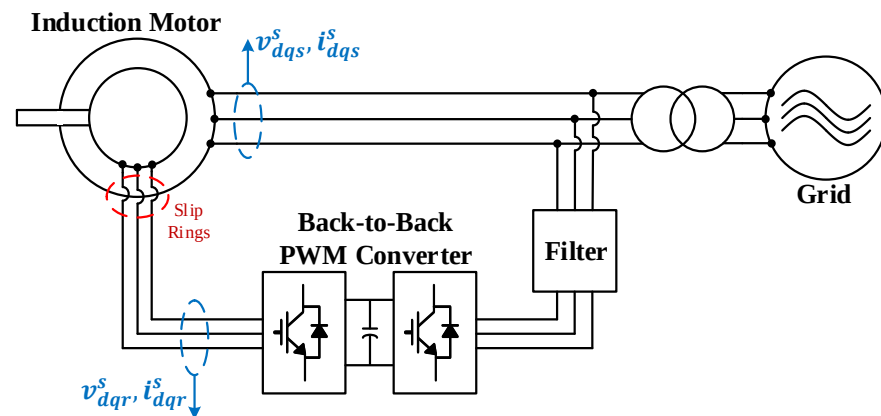


Figure 1. Configuration of a grid-connected DFIM drive system.

Motor position sensors generally present challenges related to installation and maintenance; therefore, sensorless control methods are being extensively studied. Back-EMF-based methods are commonly used for the sensorless control of permanent-magnet synchronous motors (PMSMs) and induction motors used in singly fed systems [16–25]. Typically, in single-inverter-fed systems, sensorless algorithms are realized based solely on the stator-side voltage and current measurements. In contrast, for doubly fed systems such as DFIMs, sensors are typically required on both the stator and rotor sides. Therefore, research has focused not only on eliminating the rotor position sensor, but also on reducing the number of voltage or current sensors [4–9,15].

Sensorless control methods have been extensively studied for use in DFIMs to reduce the number of sensors and simplify system configuration [4–9]. For instance, stator flux estimation techniques [4,5] based on inverse-plant estimators calculate the stator flux using voltage and measured current. Although these methods are effective, their performance relies heavily on the accuracy of motor parameters [7]. Model reference adaptive systems (MRASs) were proposed to address this issue [6–9], which rely on the feedback from measured signals for parameter adjustment. These observers are robust to parameter errors, and their performance can be enhanced via gain adaptations. However, stator voltage and current have to be measured for the MRAS observer to estimate parameters [5–8].

In this study, the sensorless vector control of DFIMs [15]—where rotational transducers or stator-side measurements are not used for estimating rotor speed and position—was revised. Also, the estimator for the back-EMF was extended to various types of estimators and their mathematical expressions were derived. The observer estimated the voltage induced by the rotating stator flux. As the amplitude and frequency of the stator flux were almost fixed and determined by the grid voltage, these values were assumed to be similar to those of PMSM drive systems. Notably, a symmetric PMSM without impedance saliency exhibits a voltage–current relationship that is nearly identical to that of a DFIM, which is a perfectly symmetric three-phase system. Therefore, the back-EMF estimator used in sensorless control schemes in PMSMs was introduced in grid-connected DFIMs. This observer used only rotor voltages and currents to estimate the voltage induced by the stator flux. Based on this estimation, the slip angle for orientation with respect to the stator flux reference frame was determined. The rotational speed for speed control was determined from the estimated slip frequency and grid frequency. The estimated slip angle provided information about the stator flux for estimating stator-side variables. Thus, this method could detect grid failures such as voltage sag from the estimated stator voltage along with the power factor of stator windings. With the proposed sensorless method, the DFIG can operate without measurement of stator-side signals, while still enabling power factor control (suppressing the reactive power), speed control, and fault-tolerant operation

under grid disturbance conditions. The effectiveness and feasibility of the proposed control scheme were validated by using it in a 2.4 kW DFIM.

The organization of the content after Section 1 is as follows: Section 2 reviews the conventional stator-flux-oriented vector control of the DFIM. In Section 3, a mathematical model is presented that interprets the DFIM as a PMSM. Section 4 proposes a back-EMF estimator using only rotor-side variables, and Section 5 provides experimental results to validate the proposed method. Finally, Section 6 concludes the paper with a summary.

2. Stator-Flux-Oriented Vector Control of DFIMs

The stator and rotor voltage and flux equations and stator-flux-oriented vector control of DFIMs are reviewed and summarized herein.

$$v_{dqs}^e = R_s i_{dqs}^e + \frac{d}{dt} \lambda_{dqs}^e + j\omega_e \lambda_{dqs}^e, \quad (1)$$

$$v_{dqr}^e = R_r i_{dqr}^e + \frac{d}{dt} \lambda_{dqr}^e + j(\omega_e - \omega_r) \lambda_{dqr}^e, \quad (2)$$

$$\lambda_{dqs}^e = L_s i_{dqs}^e + L_m i_{dqr}^e, \quad (3)$$

$$\lambda_{dqr}^e = L_m i_{dqs}^e + L_r i_{dqr}^e, \quad (4)$$

where e denotes the synchronous reference frame; s and r denote the stator and rotor variables, respectively; ω_e is the synchronous frequency; and ω_r is the angular frequency of the rotor speed. R_s and R_r are the resistances of stator and rotor windings; L_s , L_r , and L_m are the stator, rotor, and magnetizing inductances; v_{dqs}^e and v_{dqr}^e are the stator and rotor voltages; i_{dqs}^e and i_{dqr}^e are the stator and rotor currents; and λ_{dqs}^e and λ_{dqr}^e are the stator and rotor fluxes, respectively.

The torque, active power, and reactive power of the DFIM on the synchronous reference frame can be expressed using the d-q voltage, current, and flux as follows.

$$T_e = \frac{3}{2} p p \frac{L_m}{L_s} (i_{dr}^e \lambda_{qs}^e - i_{qr}^e \lambda_{ds}^e), \quad (5)$$

$$P_s = \frac{3}{2} (v_{ds}^e i_{ds}^e + v_{qs}^e i_{qs}^e), \quad (6)$$

$$Q_s = \frac{3}{2} (v_{qs}^e i_{ds}^e - v_{ds}^e i_{qs}^e), \quad (7)$$

In Equation (5), pp denotes the pole pairs and P_s and Q_s denote the stator active power and reactive power, respectively.

By aligning the stator flux vector to the d-axis, Equations (5)–(7) can be simplified. The stator-flux-oriented vector control is usually adopted for the DFIM, in which the stator windings are directly connected to the grid. Assuming that the amplitude and frequency of the grid voltage are fixed, the stator-side voltages of the DFIM can be derived using Equation (1) by applying the stationary reference frame “s”.

$$v_{dqs}^s = R_s i_{dqs}^s + \frac{d}{dt} \lambda_{dqs}^s \approx \frac{d}{dt} \lambda_{dqs}^s, \quad (8)$$

By neglecting the resistance voltage drop in Equation (8), the stator flux frequency ω_e is almost the same as the grid frequency ω_g [10]. If the stator flux reference frame is applied, the d-q-axis stator fluxes can be determined as follows.

$$\lambda_{ds}^e \approx \frac{E_g}{\omega_e} \approx \frac{E_g}{\omega_g}, \frac{d}{dt} \lambda_{ds}^e \approx 0, \quad (9)$$

$$\lambda_{qs}^e = 0, \frac{d}{dt}\lambda_{qs}^e = 0, \quad (10)$$

where E_g is the grid voltage amplitude and ω_g is the grid frequency. The d-axis stator flux in the stator flux reference frame denotes the function of rotor current. However, if the resistance voltage drop is ignored within the range of rated rotor current, the magnitude of the d-axis stator flux can be determined by the ratio between the magnitude and frequency of grid voltage (Equation (9)) regardless of the stator-side current.

Grid-connected DFIMs require torque control in motor drives and power quality control at the stator side. By applying Equations (9) and (10) on the stator-flux-oriented reference frame, the equations of torque and stator-side active and reactive powers can be rewritten as follows.

$$T_e = -\frac{3}{2}pp \frac{L_m}{L_s} (i_{qr}^e \lambda_{ds}^e) \approx -\left(\frac{3}{2}pp \frac{L_m}{L_s} \frac{E_g}{\omega_e}\right) i_{qr}^e = -K_T i_{qr}^e, \quad (11)$$

$$P_s \approx -\frac{3}{2} \frac{L_m}{L_s} \omega_e \lambda_{ds}^e i_{qr}^e = -\left(\frac{3}{2} \frac{L_m}{L_s} \omega_e \lambda_{ds}^e\right) i_{qr}^e = -K_P i_{qr}^e, \quad (12)$$

$$Q_s = \frac{3}{2} \frac{\omega_e \lambda_{ds}^e}{L_s} (\lambda_{ds}^e - L_m i_{dr}^e) = \left(\frac{3}{2} \frac{L_m}{L_s} \omega_e \lambda_{ds}^e\right) \left(\frac{\lambda_{ds}^e}{L_m} - i_{dr}^e\right) \approx K_Q \left(\frac{E_g}{\omega_e L_m} - i_{dr}^e\right), \quad (13)$$

By assuming that the d-axis stator flux in the stator flux reference frame is fixed in Equation (9), torque and stator-side active power control can be obtained using the q-axis rotor current. Here, K_T , K_P , and K_Q can be adopted as the torque, stator-side active power and reactive power constants, respectively. Similarly, the stator-side reactive power can be regulated by the d-axis rotor current. For the unity power factor, the d-axis current can be determined as $\frac{\lambda_{ds}^e}{L_m}$, regardless of the reactive power constant K_Q .

$$i_{dr}^e = \frac{\lambda_{ds}^e}{L_m} \approx \frac{E_g}{\omega_e L_m}, \quad (14)$$

Thus, active power control via the q-axis current and reactive power control via the d-axis current are decoupled, and the power factor on the stator side can be determined by regulating the d-axis current in Equation (14).

3. Grid-Connected DFIM as Synchronous Machine

In squirrel-cage induction motors, the inverter-controlled stator current generates flux. Therefore, errors in the sensorless control angle can lead to an unstable rotor flux, necessitating a more complex sensorless algorithm for flux and rotor angle estimations [24,25]. The stator side in the DFIM is directly connected to the grid, and the speed and magnitude of stator flux are almost constant (Equations (9) and (10)). Similarly, PMSMs have inherent and constant rotor fluxes due to permanent magnets. Therefore, the sensorless control algorithm in PMSMs can be applied to grid-connected DFIMs. At zero speed, the induced voltage (back-EMF) in a PMSM is zero; however, the induced voltage is zero in the DFIM at synchronous rotor speed, which corresponds to zero slip speed at the rotor side. Similar to the sensorless control in PMSMs, which has a dead zone at zero rotor speed due to the back-EMF, the proposed DFIM sensorless control includes the dead zone at the synchronous rotor speed, where the slip speed is zero. This similarity is further corroborated by comparing the voltage model equations for the PMSM and DFIM.

For simplicity, voltage equations in the synchronous reference frame for the PMSM are presented first.

$$\begin{aligned} v_{ds}^e &= R_s i_{ds}^e + L_s \frac{d}{dt} i_{ds}^e - \omega_r L_s i_{qs}^e \\ v_{qs}^e &= R_s i_{qs}^e + L_s \frac{d}{dt} i_{qs}^e + \omega_r L_s i_{ds}^e + \omega_r \lambda_f' \end{aligned} \quad (15)$$

where L_s is the stator d–q inductance and λ_f is the flux linkage of the permanent magnet in the rotor. For the grid-connected DFIM, the rotor voltage equations can be rewritten by substituting the stator fluxes in Equations (9) and (10) to Equations (1) and (2) as follows.

$$\begin{aligned} v_{dr}^e &= R_r i_{dr}^e + \sigma L_r \frac{d}{dt} i_{dr}^e - (\omega_e - \omega_r) \sigma L_r i_{qr}^e + \frac{L_m}{L_s} \frac{d}{dt} \lambda_{ds}^e \\ v_{qr}^e &= R_r i_{qr}^e + \sigma L_r \frac{d}{dt} i_{qr}^e + (\omega_e - \omega_r) \sigma L_r i_{dr}^e + \frac{L_m}{L_s} (\omega_e - \omega_r) \lambda_{ds}^e \end{aligned} \quad (16)$$

A comparison of Equations (15) and (16) shows that the stator inductance L_s corresponds to σL_r and the synchronous speed ω_r of the PMSM corresponds to the slip speed $(\omega_e - \omega_r)$ of the DFIM. The flux linkage λ_f induced by the rotor permanent magnets in the PMSM corresponds to the stator flux induced by the grid voltage in the DFIM. To employ the back-EMF-based sensorless algorithm in the PMSM, Equation (16) can be rewritten in matrix form as follows.

$$\begin{bmatrix} v_{dr}^e \\ v_{qr}^e \end{bmatrix} = \begin{bmatrix} R_r + \sigma L_r \frac{d}{dt} & -(\omega_e - \omega_r) \sigma L_r \\ (\omega_e - \omega_r) \sigma L_r & R_r + \sigma L_r \frac{d}{dt} \end{bmatrix} \begin{bmatrix} i_{dr}^e \\ i_{qr}^e \end{bmatrix} + \begin{bmatrix} E_d^e \\ E_q^e \end{bmatrix}, \quad (17)$$

$$E_d^e = \frac{L_m}{L_s} \frac{d}{dt} \lambda_{ds}^e \approx 0, E_q^e = \frac{L_m}{L_s} (\omega_e - \omega_r) \lambda_{ds}^e, \quad (18)$$

where E_d^e and E_q^e are the d–q-induced voltages from the stator side to the rotor side and can be analyzed using the back-EMF that is proportional to the slip speed. If the grid voltage has constant frequency and amplitude and the sampling frequency in the digital control system is sufficiently high to neglect the variations in the d-axis stator flux, the approximation of Equation (18) can be applied; this has also been expressed in Equation (9). Therefore, the sensorless control approach for PMSMs can be applied to the grid-connected DFIM. The back-EMF-based sensorless control of PMSMs uses only stator voltage and current to estimate the rotor flux angle. Similarly, the slip angle and frequency between the stator flux and rotor position in the DFIM can be determined using only the rotor voltage and current.

4. Sensorless Control of DFIM Using Only Rotor-Side Variables

In a PMSM, the back-EMF-based sensorless algorithm estimates the rotor position based on the back-EMF components. To this end, voltage equations in the estimated coordinate plane must be analyzed. An observer must also be designed for estimating the back-EMF and its parameter dependence must be analyzed.

Herein, the rotor voltage equations in the estimated stator flux reference frame for the DFIM were analyzed. Based on the results, an observer was proposed to estimate the back-EMF in the DFIM on the synchronous reference frame. A method for estimating stator-side variables using the observer was proposed, which was also used for measuring stator-side voltage and current variations without requiring sensors on the stator side.

4.1. Rotor Voltage Equations in the Estimated Stator Flux Reference Frame for the DFIM

To simplify equation development, the following variables were defined and used.

$$\omega_{slip} = \omega_e - \omega_r, \quad (19)$$

$$\theta_{slip} = \theta_e - \theta_r, \quad (20)$$

$$\tilde{\theta}_{slip} = \theta_{slip} - \hat{\theta}_{slip}, \quad (21)$$

$$R(\theta) = \begin{bmatrix} \cos\theta & \sin\theta \\ -\sin\theta & \cos\theta \end{bmatrix}, \quad (22)$$

where ω_{slip} and θ_{slip} are the slip frequency and angle, respectively, and $\hat{\cdot}$ and $\tilde{\cdot}$ denote the value estimated by the sensorless control and the error between the estimated and actual values, respectively. Figure 2 shows these parameters in a space vector diagram.

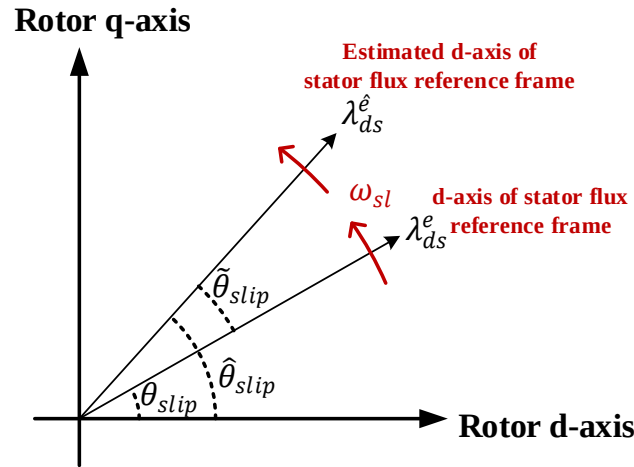


Figure 2. Space vector diagram for the grid-connected DFIM.

$R(\theta)$ corresponds to the Park transform. The actual stator flux reference frame cannot be applied for the sensorless control of the DFIM; therefore, the estimated stator flux reference frame was used. To derive the rotor voltage equation in the estimated stator flux reference frame, $R(-\tilde{\theta}_{slip})$ was applied to the rotor voltage (Equation (17)). The resulting equations are expressed below.

$$\begin{bmatrix} v_{dr}^{\hat{e}} \\ v_{qr}^{\hat{e}} \end{bmatrix} = \begin{bmatrix} R_r + \sigma L_r \frac{d}{dt} & -\omega_{slip} \sigma L_r \\ \omega_{slip} \sigma L_r & R_r + \sigma L_r \frac{d}{dt} \end{bmatrix} \begin{bmatrix} i_{dr}^{\hat{e}} \\ i_{qr}^{\hat{e}} \end{bmatrix} + \begin{bmatrix} E_d^{\hat{e}} \\ E_q^{\hat{e}} \end{bmatrix} + \sigma L_r \tilde{\omega}_{slip} \begin{bmatrix} i_{qr}^{\hat{e}} \\ -i_{dr}^{\hat{e}} \end{bmatrix} \quad (23)$$

where $\begin{bmatrix} E_d^{\hat{e}} \\ E_q^{\hat{e}} \end{bmatrix} = E \begin{bmatrix} -\sin(\tilde{\theta}_{slip}) \\ \cos(\tilde{\theta}_{slip}) \end{bmatrix}$, $E = \frac{L_m}{L_s} (\omega_e - \omega_r) \lambda_{ds}^e$,

where $\hat{e}$ denotes the estimated stator flux reference frame. Under the steady-state condition, the third term on the right side of Equation (23) can be neglected assuming that the bandwidth of the speed observer (Section 4.2) is sufficiently higher than the variations in the mechanical speed. Then, the rotor voltage equations in the estimated stator flux reference frame can be simply rewritten as follows.

$$\begin{bmatrix} v_{dr}^{\hat{e}} \\ v_{qr}^{\hat{e}} \end{bmatrix} = \begin{bmatrix} R_r + \sigma L_r \frac{d}{dt} & -\omega_{slip} \sigma L_r \\ \omega_{slip} \sigma L_r & R_r + \sigma L_r \frac{d}{dt} \end{bmatrix} \begin{bmatrix} i_{dr}^{\hat{e}} \\ i_{qr}^{\hat{e}} \end{bmatrix} + \begin{bmatrix} E_d^{\hat{e}} \\ E_q^{\hat{e}} \end{bmatrix} \quad (24)$$

In Equation (24), the error of the estimated slip angle can be obtained from the back-EMF terms, $E_d^{\hat{e}}$ and $E_q^{\hat{e}}$. The arctangent function is usually used to extract only the angle error term as follows.

$$\tilde{\theta}_{slip} = \text{atan}\left(\frac{-E_d^{\hat{e}}}{E_q^{\hat{e}}}\right). \quad (25)$$

These back-EMF terms can be directly calculated using Equation (24). As the derivative of the current was included, a low-pass filter (LPF) was applied to the final output for reducing the influence of various noise components on it (Figure 3). Here, the superscript ‘*’ means the reference value. Morimoto et al. presented the back-EMF estimator in PMSM sensorless control [16].

$$v_{dqr}^{\wedge} - \hat{R}_r i_{dqr}^{\wedge} - \sigma \hat{L}_r \frac{d}{dt} i_{dqr}^{\wedge} - j \hat{\omega}_{sl} \sigma \hat{L}_r i_{dqr}^{\wedge} \longrightarrow \frac{\omega_{LPF}}{s + \omega_{LPF}} \longrightarrow \begin{bmatrix} \hat{E}_d^{\wedge} \\ \hat{E}_q^{\wedge} \end{bmatrix}$$

Figure 3. Back-EMF estimation method using rotor voltage equation and low-pass filter (LPF).

As the parameters used for back-EMF estimation may contain errors, $\wedge$ was applied to all parameters. The voltage error was assumed to be sufficiently small to be neglected based on inverter nonlinearity compensation. The performance of this estimator was validated based on the accuracy of motor parameters and the cutoff frequency of the LPF. In cases wherein the motor parameters do not contain errors, the transfer function between the actual and estimated back-EMF terms can be expressed as follows. The cutoff frequency of this LPF is generally set to approximately 50% of the bandwidth of the current controller.

$$\frac{\hat{E}_d^{\wedge}}{E_d^{\wedge}} = \frac{\hat{E}_q^{\wedge}}{E_q^{\wedge}} = \frac{\omega_{LPF}}{s + \omega_{LPF}}. \quad (26)$$

4.2. Back-EMF Estimator in DFIM Using PI-Type State Filter

A PI-type state filter was used in the back-EMF estimator configuration (Figure 3), thereby avoiding the use of the differential term (Figure 4), where the back-EMF estimator is based on the estimation of the rotor current. The output of the PI-type state filter provides an estimated back-EMF term. The current is estimated based on the voltage information and impedance drop. The estimated current is then employed to correct and estimate the back-EMF term using a PI controller based on the difference between the estimated and actual current components. Based on the resistance and leakage inductance (σL_r) at the rotor side, the gain of the PI controller can be set as follows.

$$k_p = \sigma \hat{L}_r \omega_{LPF}, \quad k_i = \hat{R}_r \omega_{LPF}. \quad (27)$$

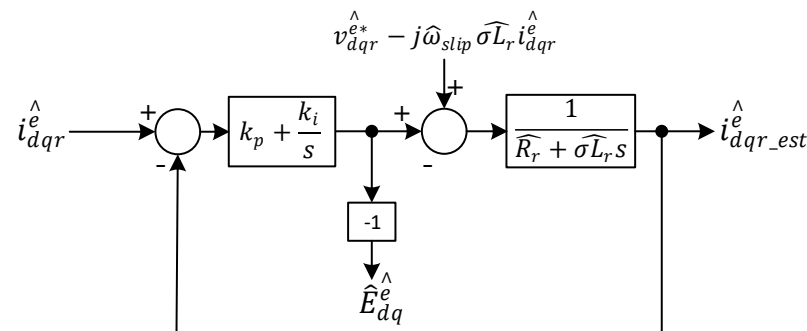


Figure 4. Back-EMF estimation method using PI-type state filter and estimated current.

If the parameters used in the estimator do not contain errors, the transfer function between the actual and estimated back-EMF terms can be expressed as

$$\frac{\hat{E}_d^{\wedge}}{E_d^{\wedge}} = \frac{\hat{E}_q^{\wedge}}{E_q^{\wedge}} = \frac{\omega_{LPF}}{s + \omega_{LPF}}. \quad (28)$$

4.3. Back-EMF Estimator in DFIM Using Reduced-Order Observer

As the back-EMF was estimated over a short sampling period, the derivatives of the back-EMF terms can be assumed as negligible. Under this assumption, the state equation

that uses rotor currents and back-EMF terms as the state variables can be derived in the reference frame of the estimated stator flux.

$$\frac{d}{dt} \begin{bmatrix} \hat{i}_{dr}^e \\ \hat{i}_{qr}^e \\ \hat{E}_d^e \\ \hat{E}_q^e \end{bmatrix} = \begin{bmatrix} -\frac{R_r}{\sigma L_r} & \omega_{slip} & -\frac{1}{\sigma L_r} & 0 \\ -\omega_{slip} & -\frac{R_r}{\sigma L_r} & 0 & -\frac{1}{\sigma L_r} \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} \hat{i}_{dr}^e \\ \hat{i}_{qr}^e \\ \hat{E}_d^e \\ \hat{E}_q^e \end{bmatrix} + \begin{bmatrix} \frac{\hat{v}_{dr}^{e*}}{\sigma L_r} \\ \frac{\hat{v}_{qr}^{e*}}{\sigma L_r} \\ 0 \\ 0 \end{bmatrix}. \quad (29)$$

The measurable rotor current can be chosen as the output variables.

$$Y = \begin{bmatrix} \hat{i}_{dr}^e \\ \hat{i}_{qr}^e \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \begin{bmatrix} \hat{i}_{dr}^e \\ \hat{i}_{qr}^e \\ \hat{E}_d^e \\ \hat{E}_q^e \end{bmatrix}. \quad (30)$$

To reduce the number of state variables, new output variables can be defined using the measurable rotor current as follows.

$$-\frac{1}{\sigma L_r} \hat{E}_d^e = \frac{d}{dt} \hat{i}_{dr}^e + \frac{R_r}{\sigma L_r} \hat{i}_{dr}^e - \omega_{slip} \hat{i}_{qr}^e - \frac{\hat{v}_{dr}^{e*}}{\sigma L_r} \quad (31)$$

$$-\frac{1}{\sigma L_r} \hat{E}_q^e = \frac{d}{dt} \hat{i}_{qr}^e + \frac{R_r}{\sigma L_r} \hat{i}_{qr}^e + \omega_{slip} \hat{i}_{dr}^e - \frac{\hat{v}_{qr}^{e*}}{\sigma L_r} \quad (32)$$

The new output variables from Equations (31) and (32) can be applied to the state equation to obtain a reduced-order observer with only two state variables.

$$\frac{d}{dt} \hat{E}_d^e = k_r \left(-\frac{1}{\sigma \hat{L}_r} \hat{E}_d^e - \left(-\frac{1}{\sigma \hat{L}_r} \hat{E}_d^e \right) \right) = \frac{k_r}{\sigma \hat{L}_r} \hat{E}_d^e + k_r \left(\frac{d}{dt} \hat{i}_{dr}^e + \frac{\hat{R}_r}{\sigma \hat{L}_r} \hat{i}_{dr}^e - \omega_{slip} \hat{i}_{qr}^e - \frac{\hat{v}_{dr}^{e*}}{\sigma \hat{L}_r} \right), \quad (33)$$

$$\frac{d}{dt} \hat{E}_q^e = k_r \left(-\frac{1}{\sigma \hat{L}_r} \hat{E}_q^e - \left(-\frac{1}{\sigma \hat{L}_r} \hat{E}_q^e \right) \right) = \frac{k_r}{\sigma \hat{L}_r} \hat{E}_q^e + k_r \left(\frac{d}{dt} \hat{i}_{qr}^e + \frac{\hat{R}_r}{\sigma \hat{L}_r} \hat{i}_{qr}^e + \omega_{slip} \hat{i}_{dr}^e - \frac{\hat{v}_{qr}^{e*}}{\sigma \hat{L}_r} \right), \quad (34)$$

where k_r is the feedback gain of the observer. To eliminate the derivatives of the rotor currents, the state variables $\hat{\eta}_1$ and $\hat{\eta}_2$ can be defined as follows.

$$\hat{\eta}_1 = \hat{E}_d^e - k_r \hat{i}_{dr}^e, \quad \hat{\eta}_2 = \hat{E}_q^e - k_r \hat{i}_{qr}^e. \quad (35)$$

By applying $\hat{\eta}_1$ and $\hat{\eta}_2$ as the state variables, the reduced-order observer can be rewritten as follows.

$$\frac{d}{dt} \hat{\eta}_d = \frac{k_r}{\sigma \hat{L}_r} \hat{\eta}_d + \frac{k_r}{\sigma \hat{L}_r} \left((k_r + \hat{R}_r) \hat{i}_{dr}^e - \omega_{slip} \sigma \hat{L}_r \hat{i}_{qr}^e - \hat{v}_{dr}^{e*} \right), \quad (36)$$

$$\frac{d}{dt} \hat{\eta}_q = \frac{k_r}{\sigma \hat{L}_r} \hat{\eta}_q + \frac{k_r}{\sigma \hat{L}_r} \left((k_r + \hat{R}_r) \hat{i}_{qr}^e + \omega_{slip} \sigma \hat{L}_r \hat{i}_{dr}^e - \hat{v}_{qr}^{e*} \right), \quad (37)$$

Figure 5 shows the reduced-order observer used for estimating the dq-axis back-EMF using Equations (35)–(37). The feedback gain, k_r , can be determined by selecting the bandwidth of the observer as ω_{LPF} .

$$k_r = -\sigma \hat{L}_r \omega_{LPF}. \quad (38)$$

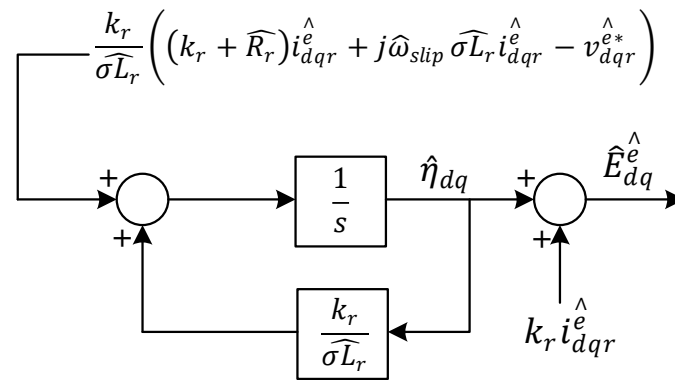


Figure 5. Back-EMF estimation method using reduced-order observer.

If the parameters used in the estimator does not contain errors, the transfer function between the actual and estimated back-EMF terms can be expressed as

$$\frac{\hat{E}_d^e}{E_d^e} = \frac{\hat{E}_q^e}{E_q^e} = \frac{\omega_{LPF}}{s + \omega_{LPF}}. \quad (39)$$

4.4. Estimation of Slip Angle and Speed

The three estimators (Equations (26), (28), and (39)) exhibited identical performance in back-EMF estimation. The estimator employing the voltage model and LPF (Figure 3), that using a PI-type state filter (Figure 4), and that using the reduced-order observer (Figure 5) were used for estimating the back-EMF of a PMSM; these estimators are mathematically equivalent [22]. Since the observer stability is structured identically to that of conventional back-EMF-based sensorless estimators for PMSMs, a detailed stability analysis is omitted in this paper. The three proposed estimators based on the synchronous motor voltage equation for the sensorless control of the DFIM were also used for back-EMF estimation, and their performance was verified to be identical.

Based on the estimated back-EMF, a method for generating the slip angle and slip speed used for vector control was proposed. To this end, angle and speed estimators were constructed based on the structure of the PI-type state filter. Figure 6 shows the block diagram of slip angle and speed generation for vector control from the output of the back-EMF estimator.

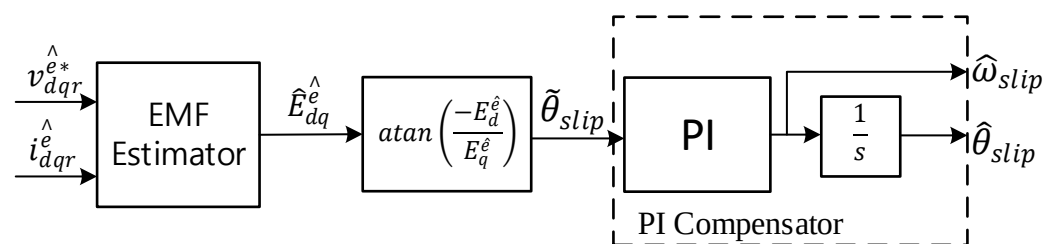


Figure 6. Block diagram of slip angle and speed estimation.

The gain value of the PI compensator was set based on the closed-loop transfer function of the slip angle as follows.

$$\frac{\hat{\theta}_{slip}}{\theta_{slip}} = \frac{k_{pc}s + k_{ic}}{s^2 + k_{pc}s + k_{ic}}. \quad (40)$$

As the closed-loop transfer function is a second-order system (Equation (40)), the gain values k_{pc} and k_{ic} , where k_{pc} is a proportional gain and k_{ic} is an integral gain, can be set based on the natural frequency ω_n and the damping ratio ζ of the second-order system.

$$k_{pc} = 2\zeta\omega_n, \quad k_{ic} = \omega_n^2. \quad (41)$$

where ζ was set to about 0.707 based on the Butterworth filter; however, Equation (40) shows that the zero-frequency of the system can contribute to an overshoot. The damping ratio is therefore set to a value larger than 0.707. When the damping ratio is 1.5~3, the overshoot is clearly reduced (Figure 7). Although increasing the damping ratio theoretically reduces overshoot, setting considerably higher damping ratios in practical systems can result in a high k_{pc} -gain. This can introduce high-frequency noise components in final angle measurements because the estimated back-EMF with noise is directly reflected in the measured values. As a result, the damping ratio needs to be determined between 1 and 2. The natural frequency can be determined based on a given damping ratio (ζ) in order to set the desired bandwidth. It should be set significantly lower than the bandwidth of the previously designed back-EMF estimator—typically less than one-tenth—to ensure stable operation.

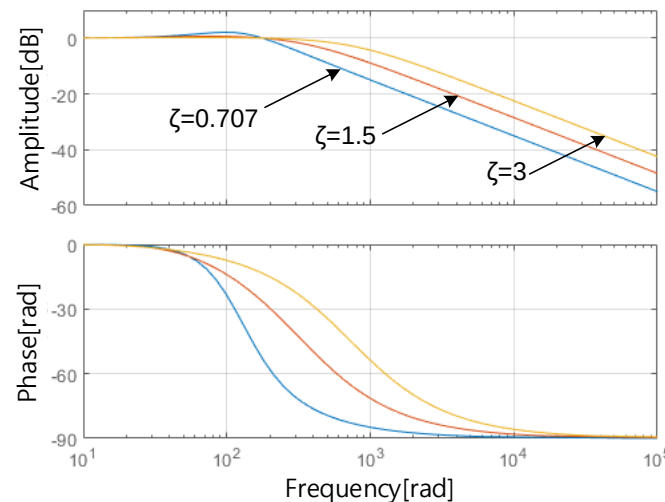


Figure 7. Bode plot of closed-loop transfer function of slip angle estimator ($\omega_n = 2 \times 20 \times \pi$).

4.5. Estimation of Stator-Side Voltage and Current

As expressed in Equations (15)–(17), the d-axis stator flux is required for calculating torque constant and grid power. For protecting the DFIM against grid accidents such as voltage sag, information about the stator voltage and current is required. As stator-side measurements are not used in the proposed sensorless control, the stator-side voltage and current must be estimated.

The d-axis stator flux can be estimated from the reduced-order observer and PI compensator as follows.

$$\hat{\lambda}_{ds}^e = \frac{\hat{L}_s}{\hat{L}_m} \frac{\hat{e}_q^e}{\hat{\omega}_{slip}}. \quad (42)$$

Using the d-axis stator flux, the torque and stator power can be regulated without any measurements of the stator-side voltage and current. The magnitude of the d–q stator current and voltage in the estimated stator flux reference frame can be derived as follows.

$$\hat{i}_{ds}^e = \frac{1}{\hat{L}_s} \left(\hat{\lambda}_{ds}^e - \hat{L}_m \hat{i}_{dr}^e \right), \quad \hat{i}_{qs}^e = \frac{1}{\hat{L}_s} \left(\hat{L}_m \hat{i}_{qr}^e \right), \quad (43)$$

$$\hat{v}_{ds}^e \approx \hat{R}_s \hat{i}_{ds}^e, \hat{v}_{qs}^e \approx \hat{R}_s \hat{i}_{qs}^e + \omega_e \hat{\lambda}_{ds}^e. \quad (44)$$

From Equations (43) and (44), the angle of stator voltage and current vectors can be calculated.

$$\hat{\theta}_V = \text{atan2} \left(\frac{\hat{v}_{qs}^e}{\hat{v}_{ds}^e} \right), \hat{\theta}_i = \text{atan2} \left(\frac{\hat{i}_{qs}^e}{\hat{i}_{ds}^e} \right). \quad (45)$$

These estimated stator variables can be used to detect grid failures and employ grid fault handling operations [26,27]. The stator-side active and reactive powers can also be calculated by substituting Equations (43) and (44) with Equations (6) and (7), respectively, and regulated without any stator-side sensors. Figure 8 shows the block diagram of the proposed sensorless vector control with active power and reactive power controllers. Here, the superscript ‘r’ means the rotor reference frame, and this coordinate should be applied to the rotor-side inverter. Instead of the active power controller, the speed controller can also be adopted based on Equation (11).

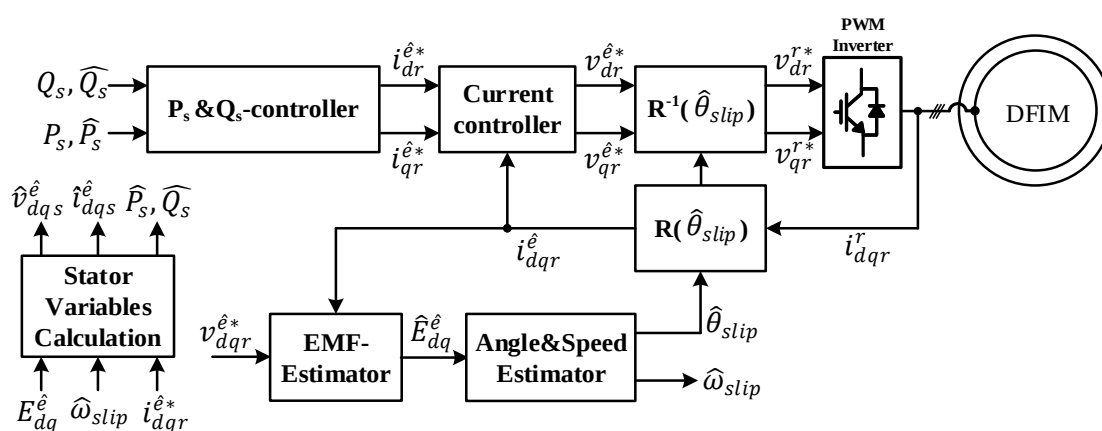


Figure 8. Block diagram of proposed sensorless control of DFIM.

5. Experimental Verification

Experiments were conducted using a 2.4 kW wound rotor induction machine, with its stator windings directly connected to the grid. A load machine was coupled with the test machine to simulate the load conditions. The machine parameters and experimental conditions are summarized in Table 1.

Table 1. Machine parameters and grid conditions.

Parameters	Values	Parameters	Values
Rated power	2.4 kW	Rated speed	1720 r/min
Stator resistance (R_s)	0.6 Ω	Rated torque	12 Nm
Rotor resistance (R_r)	0.7 Ω	Rated current	10 Arms
Stator inductance (L_s)	54 mH	Number of pole pairs	2
Rotor inductance (L_r)	56 mH	Grid voltage	220 Vrms
Mutual inductance (L_m)	49 mH	Grid frequency	60 Hz

5.1. State Observer for Stator Flux Estimation

The performance of the sensorless control was verified by comparing the estimated flux with the actual stator flux. As applying a flux measurement equipment is generally difficult, a state observer was designed and implemented to estimate the stator flux based on the measured stator voltage and rotor position. The detailed structure of the state observer has been explained below. Equations (1)–(4) can be rewritten in the stationary reference frame (superscript ‘s’) as follows.

$$\frac{d}{dt} i_{dr}^s = -\frac{1}{\sigma L_r} \left(R_r + R_s \frac{L_m^2}{L_s^2} \right) i_{dr}^s - \omega_r i_{qr}^s + \frac{R_s L_m}{\sigma L_s^2 L_r} \lambda_{ds}^s - \omega_r \frac{L_m}{\sigma L_s L_r} \lambda_{qs}^s + \frac{1}{\sigma L_r} v_{dr}^s - \frac{L_m}{\sigma L_s L_r} v_{ds}^s, \quad (46)$$

$$\frac{d}{dt} i_{qr}^s = -\frac{1}{\sigma L_r} \left(R_r + R_s \frac{L_m^2}{L_s^2} \right) i_{qr}^s + \omega_r i_{dr}^s + \omega_r \frac{L_m}{\sigma L_s L_r} \lambda_{ds}^s + \frac{R_s L_m}{\sigma L_s^2 L_r} \lambda_{qs}^s + \frac{1}{\sigma L_r} v_{dr}^s - \frac{L_m}{\sigma L_s L_r} v_{ds}^s, \quad (47)$$

$$\frac{d}{dt} \lambda_{ds}^s = R_s \frac{L_m}{L_s} i_{dr}^s - \frac{R_s}{L_s} \lambda_{ds}^s + v_{ds}^s, \quad (48)$$

$$\frac{d}{dt} \lambda_{qs}^s = R_s \frac{L_m}{L_s} i_{qr}^s - \frac{R_s}{L_s} \lambda_{qs}^s + v_{qs}^s. \quad (49)$$

The state observer was designed using Equations (46)–(49), which includes the d- and q-axis rotor currents and stator fluxes as four-state variables; the stator and rotor voltages were used as inputs. The difference between the estimated and measured currents was used as the feedback signal. An encoder equipped with a rotor shaft was used for transforming the rotor reference frame to a stationary reference frame.

$$\frac{d}{dt} \begin{bmatrix} \hat{i}_{dr}^s \\ \hat{i}_{qr}^s \\ \hat{\lambda}_{ds}^s \\ \hat{\lambda}_{qs}^s \end{bmatrix} = A \begin{bmatrix} \hat{i}_{dr}^s \\ \hat{i}_{qr}^s \\ \hat{\lambda}_{ds}^s \\ \hat{\lambda}_{qs}^s \end{bmatrix} + B \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \\ v_{dr}^{s*} \\ v_{qr}^{s*} \end{bmatrix} + G \begin{bmatrix} \hat{i}_{dr}^s - i_{dr}^s \\ \hat{i}_{qr}^s - i_{qr}^s \end{bmatrix}, \quad (50)$$

$$\sigma = 1 - \frac{L_m^2}{L_s L_r}, \tau_s = \frac{L_s}{R_s}, c = \frac{\sigma L_s L_r}{L_m}, \quad (51)$$

$$A = \begin{bmatrix} -\left\{ \frac{R_r}{\sigma L_r} + \frac{1-\sigma}{\sigma \tau_s} \right\} & -\omega_r & \frac{1}{\tau_s c} & -\frac{\omega_r}{c} \\ \omega_r & -\left\{ \frac{R_r}{\sigma L_r} + \frac{1-\sigma}{\sigma \tau_s} \right\} & \frac{\omega_r}{c} & \frac{1}{\tau_s c} \\ \frac{L_m}{\tau_s} & 0 & -\frac{1}{\tau_s} & 0 \\ 0 & \frac{L_m}{\tau_s} & 0 & -\frac{1}{\tau_s} \end{bmatrix} \quad (52)$$

$$B = \begin{bmatrix} -\frac{1}{c} & 0 & \frac{1}{\sigma L_r} & 0 \\ 0 & -\frac{1}{c} & 0 & \frac{1}{\sigma L_r} \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}, G = \begin{bmatrix} g_1 & -g_2 \\ g_2 & g_1 \\ g_3 & -g_4 \\ g_4 & g_3 \end{bmatrix}. \quad (53)$$

The feedback gain, G, was calculated using the pole placement method or by selecting the poles of the state observer to be k-times the number of machine poles, similar to the squirrel-cage induction motor [25].

5.2. Experimental Results

Figure 9 shows the estimated d-axis stator fluxes under various conditions: acceleration and deceleration between 0 and 1200 rpm, and $\pm$ rated step load torques at 1200 rpm. In the steady state, the estimation error remains nearly constant and close to zero. Although rotor current variations impact the d-axis stator flux, the impact is minimal (blue line in Figure 9). The variation in the estimated flux under the rated q-axis current is less than $\pm 5\%$ of the stator flux at zero rotor current.

Figure 10 shows the dynamic performance of speed control with rated load torque, reversing at a slip speed of 0.05 (1710 rpm). The slip angle error remains within ± 0.125 rad but the load torque undergoes a step-wise variation.

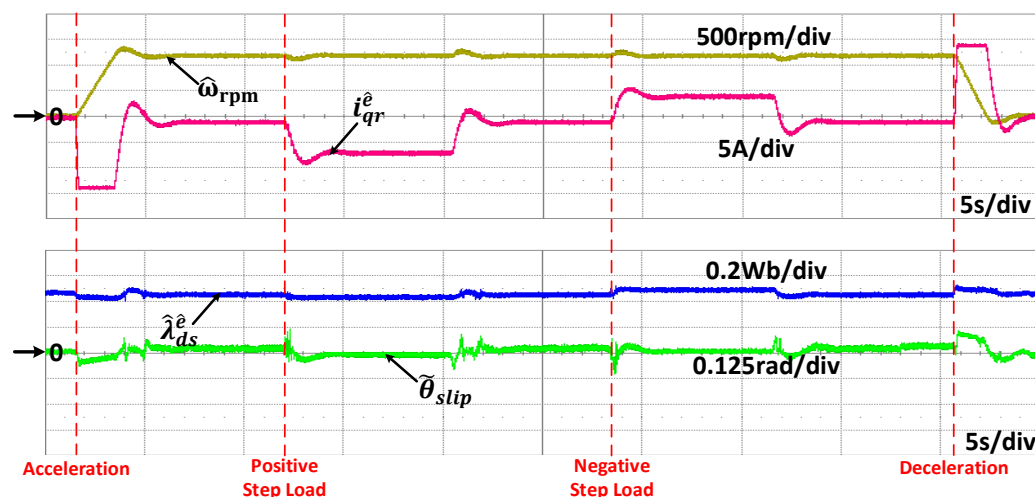


Figure 9. Experimental results: The speed control performance where the speed is accelerated to 1200 rpm and the rated step load is applied; (yellow) rotating speed, (red) q-axis rotor current, (blue) estimated d-axis stator flux, and (green) error in the estimated slip angle.

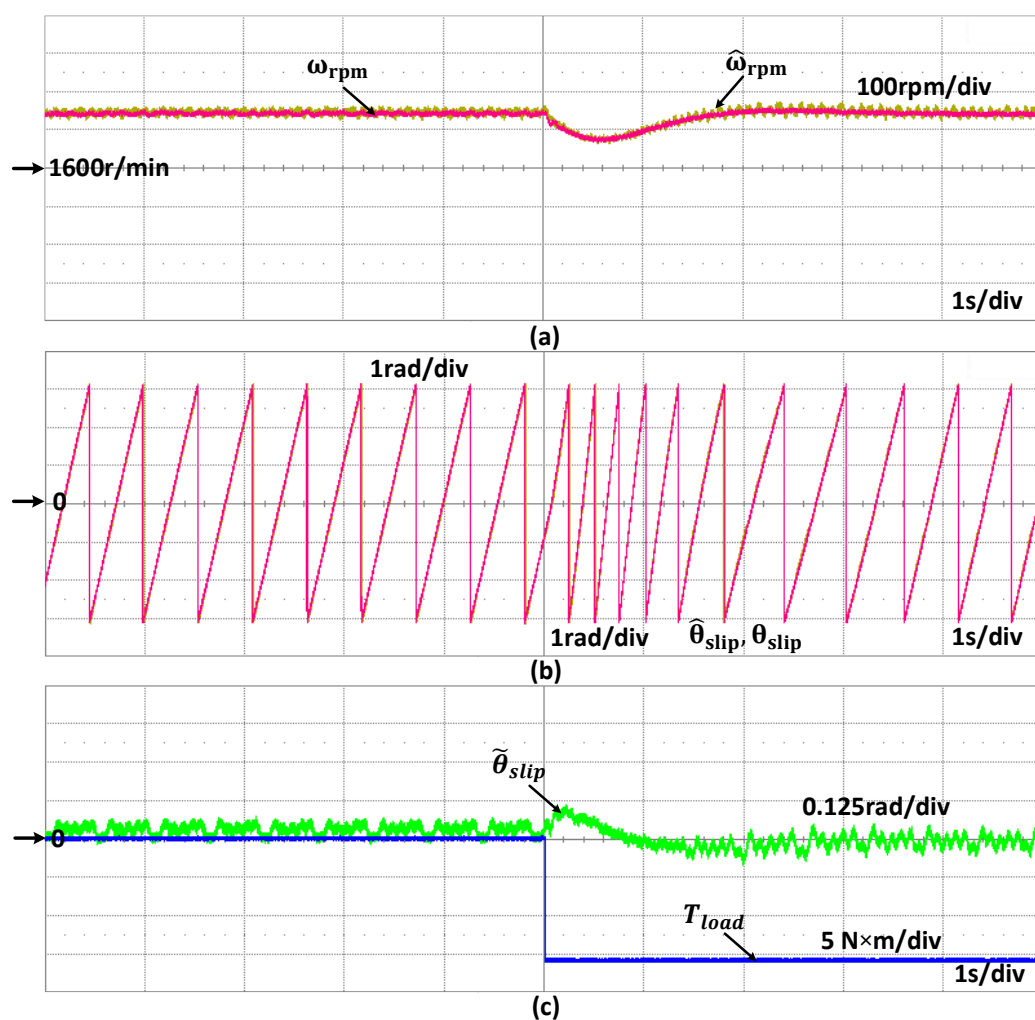


Figure 10. Experimental results: The low-slip operation performance where the operating speed is set to 1710 rpm (0.05 slip) and the rated step load is applied; (a) actual and estimated rotating speeds; (b) estimated and actual slip angles; and (c) slip angle and load torque errors.

Figure 11 shows the performance of the power factor control. The d-axis rotor current in the estimated stator flux reference frame is regulated by eliminating the reactive power component in the stator-side current; this ensures that the stator-side power factor approaches unity, as shown in the first waveforms. Moreover, power factor control is achieved without affecting the torque and active power control, as evidenced by no variations in the q-axis rotor current.

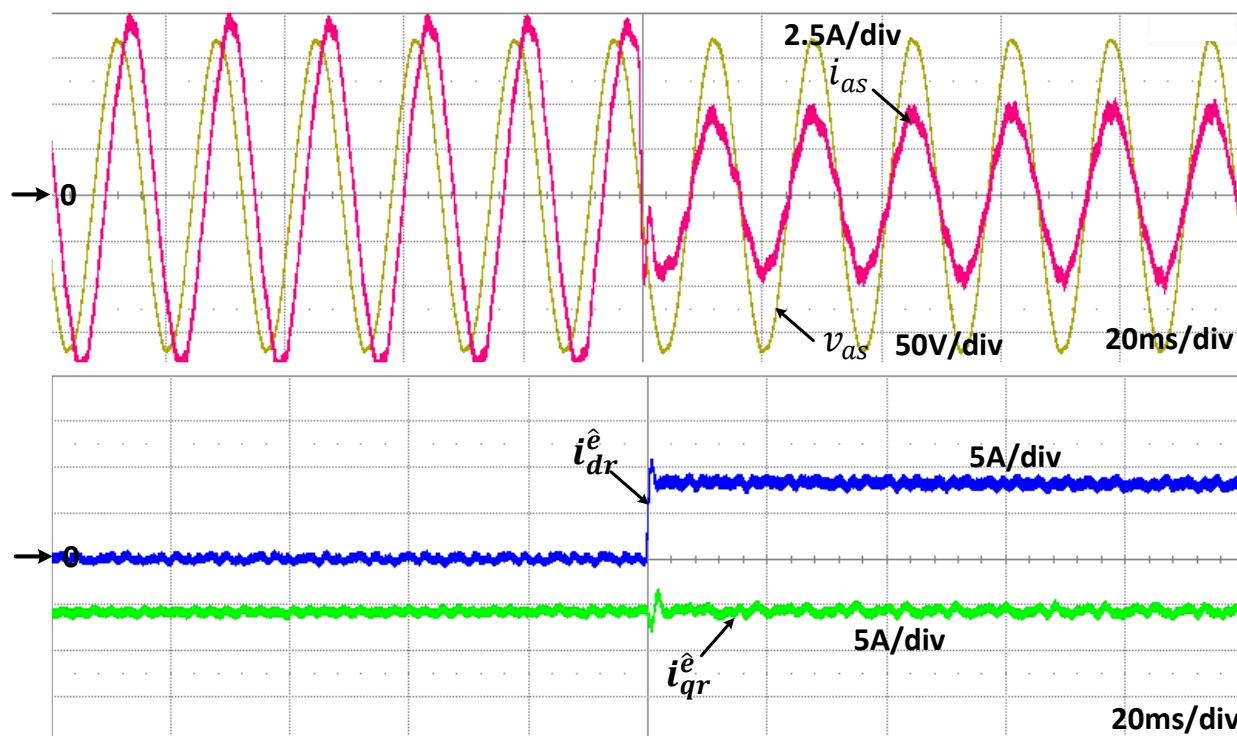


Figure 11. Experimental results: Stator-side power factor control where the d-axis rotor current is utilized to suppress the stator reactive power; (yellow) a-phase stator voltage, (red) a-phase stator current, (blue) d-axis rotor current, and (green) q-axis rotor current.

Figure 12 shows the estimated stator variables during a grid fault with a 70% voltage dip. The stator voltage, current, and d-axis stator flux are estimated using the proposed observer (Figure 12a–d). The power factor angle ($\theta_v - \theta_i$) on the stator side is calculated based on Equation (45), as shown in Figure 12c. As the grid voltage decreases, the stator voltage, current, d-axis stator flux, and estimated values also decrease. These findings confirm that the proposed observer successfully estimates stator variables and continues to operate during a sudden voltage drop.

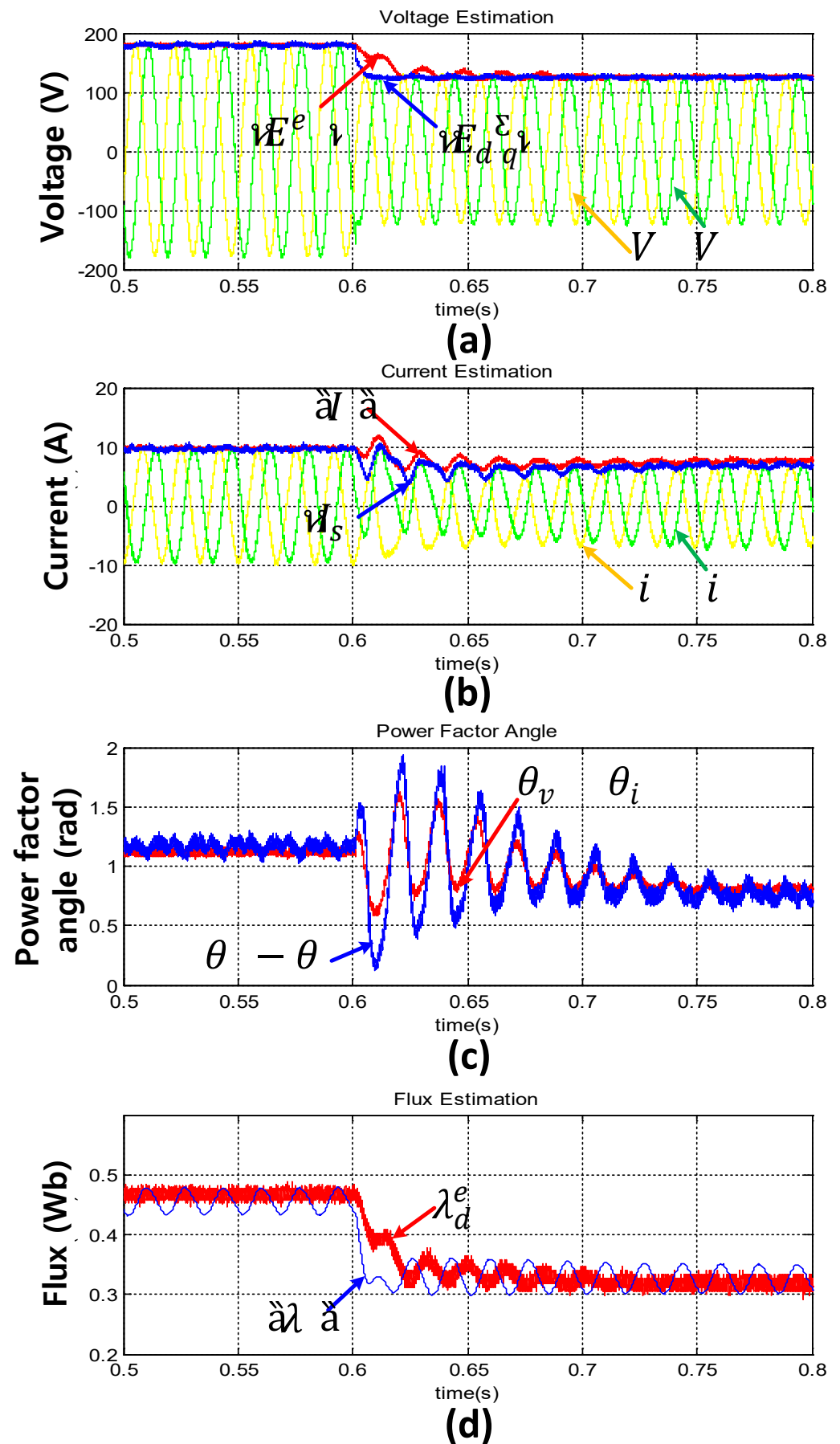


Figure 12. Experimental results: Estimation performance of stator-side variables. (a) A-phase stator voltage; (b) a-phase stator current; (c) d-axis rotor current; and (d) q-axis rotor current.

6. Conclusions

In this study, a sensorless vector control was proposed for the DFIM that does not require position/speed, stator voltage, or current sensors. The similarities between the DFIM and PMSM were analyzed, based on which a reduced-order observer was designed to estimate the induced voltage. Using only the rotor voltage and current, the slip angle and speed were estimated by the reduced-order observer. These estimated values were used for achieving speed and power factor controls. The estimated induced voltage can be used to detect and mitigate variations or faults in grid voltages. Thus, previously proposed grid fault strategies [26,27] can be employed during voltage dips or unbalanced grid voltage to address issues related to stator current ripple, electric torque, and reactive power.

The proposed sensorless operation method is not only intended to reduce the number of sensors but can also be applied as a fault-tolerant strategy for DFIM operation in the event of sensor failures. In general, DFIMs are employed in high-power applications such as wind power generation systems, where the cost contribution of rotor position or voltage sensors is relatively small. Therefore, in practical systems, the proposed method can be utilized to detect and respond to faults in stator-side sensors or rotor position sensors in conventional DFIM systems. To ensure the robust operation of the proposed method, an analysis of the effects of parameter uncertainties and sensor errors is necessary. However, since the proposed method shares the same mathematical structure as the conventional back-EMF-based sensorless control of PMSMs [22], it is relatively straightforward to analyze these effects based on existing studies. Furthermore, recent AI-based sensorless control approaches of PMSMs [28] could be extended to DFIMs, and incorporating such techniques is considered a promising direction for future research.

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Conflicts of Interest: The author declares no conflicts of interest.

Nomenclature

$\omega_e, \omega_r, \omega_{slip}$	Synchronous, rotor, and slip frequency
$\theta_e, \theta_r, \theta_{slip}$	Synchronous, rotor, and slip angle
pp	Number of pole pairs
R_s, R_r	Stator and rotor resistance
L_s, L_r, L_m	Stator, rotor, and mutual inductance
σ	Leakage factor ($= 1 - L_m^2 / (L_s L_r)$)
v_{dqs}^e, v_{dqr}^e	Stator and rotor dq-axis voltages on the stator flux reference frame
i_{dqs}^e, i_{dqr}^e	Stator and rotor dq-axis currents on the stator flux reference frame
$\lambda_{dqs}^e, \lambda_{dqr}^e$	Stator and rotor dq-axis fluxes on the stator flux reference frame
E_g, ω_g	Amplitude and frequency of grid phase voltage
ω_g	Frequency of grid voltage
P_s, Q_s	Stator-side active and reactive power
E_d^e, E_q^e	d–q-induced voltages from the stator side to the rotor side
T_e	Electrical torque
$\hat{}$	Estimated value
$\sim$	Error between actual and estimated values
$R(\theta)$	Clockwise rotation matrix as much as θ

Abbreviations

The following abbreviations are used in this manuscript:

DFIM	Doubly fed induction machine
MRASs	Model reference adaptive systems
PMSM	Permanent-magnet synchronous motor

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