A novel direct field-oriented control strategy for fault-tolerant control of induction machine drives based on EKF

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Abstract
This article presents a novel Direct Field-Oriented Control (DFOC) strategy for Fault-Tolerant Control (FTC) of wye-connected 3-Phase Induction Machine (3-PIM) drives under the stator winding open-phase failure. In the proposed control strategy, instead of flux measurement, an Extended Kalman Filter (EKF) for flux estimation is used. The introduced controller with minor modifications can be used during normal and stator winding open-phase failure conditions. With the proposed DFOC system, the speed and torque pulsations that normally happen during the open-phase failure can be reduced. The performances of the proposed EKF-based DFOC strategy and the conventional control strategy for a faulted machine using simulations and experiments under different operation conditions are compared. Simulation and experimental results demonstrated an important improvement in speed and torque pulsations through this type of fault. Results also confirmed the superiority of the proposed EKF-based DFOC scheme over the conventional control scheme to balance the faulted machine phase currents.

1 | INTRODUCTION

3-Phase Induction Machines (3-PIMs) are broadly used in a number of industrial processes because of the low cost, high efficiency and reliability, rugged structure, high robustness, and simple construction. They contribute to around 60% of the whole industrial electricity consumption. 3-PIMs are used in various industrial applications such as electric vehicles, wind turbines, electrical drives [1–3].

Electrical drives play significant roles in many commercial and industrial applications. Vector-based control methods such as Direct Torque Control (DTC) and Field-Oriented Control (FOC) methods for electrical machine drives can provide a good dynamic response and increase the drive system efficiency. In comparison with the FOC methods, control techniques based on DTC have some drawbacks such as the lack of control on machine currents, variable switching frequency and torque oscillations particularly at low-speed operations [3–5]. Generally, FOC strategy can be classified as Indirect FOC (IFOC) and Direct FOC (DFOC). The approach of the rotor/stator flux position calculation specifies the form of FOC strategies [6]. In the IFOC strategy, the flux position is calculated based on the model of the electrical machine. Owing to using pure integrator difficulties for the flux position calculation, IFOC strategy suffers from drift, offset, and saturation problems. In DFOC strategy, the flux position is measured using the flux sensing coils or Hall-effect sensors. However, in some industries, installing the flux sensor is difficult due to the physical limitations. Furthermore, the use of flux sensors increases the cost and hardware complexity of the drive system. In this case, to improve the drive system performance, flux estimation is more preferable [7–11].

Electrical drives are exposed to various types of electrical and/or mechanical faults such as: inverter faults [12, 13], machine faults [14, 15] and sensor faults [16, 17]. In some critical industrial applications such as transportation, medical equipment, and aircrafts, electrical drive operation cannot be disturbed by faulty conditions. Accordingly, for these applications, a Fault-Tolerant Control (FTC) system is very important [18, 19]. The main purpose of the FTC systems is to continue the performance quality of the faulted drive system as close as possible to the performance quality of the healthy drive system. Generally, the FTC techniques for electrical drive systems can be classified as passive FTC [20, 21] and active FTC [22–24].
the passive FTC system, using a robust controller and without requiring fault detection mechanism, healthy and faulty drive systems can be controlled. Despite the simple structure, robustness, and very fast operation to any fault situation, the performance of this system under the normal situation is not optimized. In the active FTC system, a control strategy for the electrical drive system under the normal situation and a control strategy for the electrical drive system under the faulty situation are designed. In general, active FTC systems include the following subsystems: fault detection mechanism, a reconfiguration system, and a reconfigurable control strategy. According to the provided information from fault detection unit, the reconfiguration system selects an appropriate control strategy for the new situation. Unlike passive FTC systems, active FTC systems have the ability to deal with various classes of faults and have good performances during both normal and faulty situations. However, active FTC systems are sensitive to the results provided from the fault detection unit. Moreover, these systems have complex control structure owing to the use of different control systems [25, 26].

One of the popular types of faults in the electrical drives is the open-phase fault in the stator windings of the electrical machine [27–43]. This fault occurs because of the mechanical shaking of the machine, opening of coils and so forth. For FTC of wye-connected 3-PIM drives in the case of open-phase fault, a new topology for the inverter should be designed to minimize the influence of the fault. Different inverter topologies have been used to supply the wye-connected 3-phase machine drives under open-phase failure; for example, the redundant leg of the inverter [15, 28] or the cascaded inverter [29]. However, these topologies increase the cost of drive system. Reference [30] showed that a wye-connected 3-PIM drive under the 2-phase mode operation results in significant torque oscillations. In this situation, the currents in the healthy phases are dependent on each other and cannot be controlled independently. For this, the 3-PIM Neutral Point (NP) should be available and connected to the DC Bus Midpoint (DCBM) of the drive system (see Figure 1). This article uses the inverter topology proposed in [30] but it investigates the use of the DFOC strategy to control a wye-connected 3-PIM under open-phase failure.

Investigations on the open-phase faults mostly focus on the multi-phase (more than 3 phases) induction machines and multi-phase permanent magnet synchronous machines [31–34]. Different studies have also been presented in literature in the area of open-phase failure for 3-PIM drive systems which can be summarized as follows: In [35], a technique for open-loop scalar control of 3-PIM drives under open-phase failure using injecting odd harmonic voltages was proposed. Open-loop scalar control strategy is commonly cheap, simple, and well-implementable. In spite of these positive features, this control technique cannot provide an accurate speed-torque control, such as in the FOC systems. Besides, injecting the odd harmonic voltage increases the stator rms currents. In [36], a scalar control method for a Δ-connected motor, in [30], FOC of a wye-connected motor and in [37], FOC of a Δ-connected motor during the single-phase open fault based on calculation of the Magnetic Motive Force (MMF) have been presented. The purpose of these control techniques is to produce the stator rotating MMF for a 2-PIM similar to a 3-PIM. The presented strategy in [30] exploited a current controller strategy. Owing to the lack of control on machine currents in the current controller strategy, this method is not an appropriate method during load conditions. Moreover, the used control scheme in [37] needs extra PI controllers during the fault mode. In [38], FOC of a 3-PIM with wye-connected under open-circuit fault using different converter configurations based on a unified feedforward compensation technique has been proposed. In [27, 39], and [40], IFOC of a wye-connected 3-PIM under one-phase failure condition based on the voltage controller strategy have been proposed. In [27, 39, 40], it is shown that in order to gain a non-pulsating electromagnetic torque during the fault, it is essential to use the Unbalanced Rotational Transformation Matrices (URTMs) for stator variables. However, these methods are very complex, depend on simultaneous tuning of the PI controller coefficients and suffer from pure integrator problems. In [41], FOC of a wye-connected 3-PIM under one-phase failure without speed measurement based on using two URTMs has been proposed. The proposed method in [41] cannot be used for sensored 3-PIM drives. Moreover, the used URTMs in this research are very complex and change with the variations of motor parameters. As a result, the presented control method in [41] requires heavy real time computation and depends on 3-PIM parameters. In [42], FOC of a wye-connected 3-PIM under one-phase failure condition based on the Balanced Rotational Transformation Matrix (BRTM), has been proposed. It was shown in [42] using the conventional BRTM, the FOC equations of a wye-connected 3-PIM during open-phase failure can be written as forward and backward terms. The structure of these two terms is similar to the healthy 3-PIM equations. Therefore, using two FOC algorithms, control of a faulted 3-PIM is possible. In [42], the rotor flux has been estimated using two modified Extended Kalman Filter (EKF) algorithms. The proposed control strategy in [42] has a low dynamic response and this method is sensitive to the PI controller coefficients because of using two FOC systems. Moreover, this strategy suffers from computational complexity because of using two EKF algorithms. The performance of the proposed method in [42] is only verified by simulations.

An active FTC system based on the DFOC strategy for the wye-connected 3-PIM drives using URTMs in order to

**FIGURE 1** Inverter fed a wye-connected 3-PIM under open-phase failure
improve the performance and reliability of the drive systems under the stator winding open-phase failure is presented. In the presented FTC system, a flux estimator based on an EKF algorithm is used. Many flux estimators have been presented in the past years, such as the Model Reference Adaptive System (MRAS), Sliding Mode Observer (SMO), reduced-order and full-order observers, Luenberger observer [17, 44–48]. There are also EKF applications in the literature, to estimate the motor flux. The EKF algorithm is a stochastic estimator, which can estimate the state variables considering process and measurement noises and disturbances. The EKF algorithm has a good disturbance rejection and is particularly suitable under conditions where the parameter uncertainties and noises are unavoidable [49–51].

The proposed EKF-based DFOC system in this paper with some changes in the control parameters and the Transformation Matrix (TM) can be developed for wye-connected 3-PIM drives during normal and stator winding open-phase failure conditions. In other words, unlike the conventional active FTC systems (e.g. [22, 23]), the introduced FTC system in this article does not require different control systems during normal and fault operations. Moreover, the proposed DFOC strategy in comparison to the previous published papers (e.g. [27, 40]) has good performances over a wide range of speed including low- and zero-speed operation. Additionally, different from [42], the presented control system in this article has a simple structure and a low computational complexity as a result of using a modified DFOC algorithm. The main contributions of this research can be summarized as follows:

- A technique for improving reliability of the wye-connected 3-PIM drives under open-phase failure based on the DFOC is proposed.
- A method for rotor flux estimation of healthy and faulted machines based on an EKF is proposed.
- A modified control technique can be shared for 3-PIM drives during normal and open-phase fault conditions.

In this work, simulation and experimental results over a wide speed range including operation at low- and zero-speed approve the effectiveness of the introduced EKF-based DFOC strategy in decreasing the torque and speed oscillations for the wye-connected 3-PIM drive during the 2-phase mode operation. The results also approve the introduced controller ability to balance the faulted machine phase currents. Furthermore, the comparisons made demonstrate the superiority of the introduced DFOC in this research over the previous studies.

2 | 3-PIM MODEL DURING THE STATOR WINDING OPEN-PHASE FAULT (2-PIM MODEL)

Under open-phase failure, the d-q axes, the stator magnetic axes, and the rotor magnetic axes of a wye-connected 3-PIM under open-phase failure can be illustrated as Figure 2 (it is assumed that the stator windings midpoint is accessible and single-phase open fault happens in phase ‘c’ of stator windings):

From Figure 2, the normalized TMs for stator and rotor components are obtained as [43]:

$$[T_s] = \sqrt{2} \begin{bmatrix} 0.5 & -0.5 \\ 0.5 & 0.5 \end{bmatrix}$$

(1a)

$$[T_r] = \frac{1}{\sqrt{3}} \begin{bmatrix} \cos(\alpha_0 - \alpha_e) & \cos(\alpha_0 - \alpha_e + \frac{2\pi}{3}) & \cos(\alpha_0 - \alpha_e + \frac{4\pi}{3}) \\ \sin(\alpha_0 - \alpha_e) & \sin(\alpha_0 - \alpha_e + \frac{2\pi}{3}) & \sin(\alpha_0 - \alpha_e + \frac{4\pi}{3}) \end{bmatrix}$$

(1b)

Using Equations (1a) and (1b), the d-q model of a wye-connected 3-PIM during an open-phase failure can be expressed by Equations (2a–2g) [43]:

$$\psi'_{ds} = r_s l_{ds} + p(L_d l_{ds} + L_q l_{dq})$$

(2a)

$$\psi'_{qr} = r_s l_{qr} + p(L_q l_{qr} + L_d l_{dq})$$

(2b)

$$\psi'_{dr} = 0 = r_s l_{dr} + p(L_d l_{dr} + L_q l_{dq}) + \omega_r (L_r l_{dr} + L_q l_{qr})$$

(2c)

$$\psi'_{qr} = 0 = r_s l_{qr} + p(L_q l_{qr} + L_d l_{dq}) - \omega_r (L_r l_{dr} + L_q l_{qr})$$

(2d)

$$\phi'_{ds} = L_d l_{ds} + L_d l_{dr}$$

(2e)

$$\phi'_{qr} = L_q l_{qr} + L_q l_{qr}$$

(2f)

$$\phi'_{dr} = L_d l_{dr} + L_r l_{dr}$$

(2g)

$$\phi'_{qr} = L_q l_{qr} + L_r l_{qr}$$

(2h)

$$\tau_e = \frac{\text{pole}}{2} (L_q l_{qr} l_{dr} - L_d l_{ds} l_{qr})$$

(2i)

$$\frac{\text{pole}}{2} (\tau_e - \tau_i) = J\omega_r + B\alpha_r$$

(2j)
where, superscript ‘s’ indicates that the variables are in the Stationary Reference (SR) frame. Furthermore [43]:

\[ L_{ds} = L_{ds} + 1.5L_{ms}, \quad L_d = 1.5L_{ms} \]
\[ L_{qs} = L_{ds} + 0.5L_{ms}, \quad L_q = \left(\sqrt{3}/2\right)L_{ms} \quad (2k) \]

It can be noted that Equations (2a–2j) can also represent the d-q model of a wye-connected 3-PIM during an open-phase failure when the fault happens in phase ‘a’ or ‘b’ of stator windings. Moreover, Equations (2a–2j) represent the d-q model of a healthy 3-PIM with the following parameters [52]:

\[ L_{ds} = L_{qs} = L_{ds} + 1.5L_{ms}, \quad L_d = L_q = 1.5L_{ms} \quad (2L) \]

In Equations (2a–2j), \( L_{ds}, L_{qs} \) and \( L_r \) are the stator and rotor d-q axes self-inductances. \( L_{ds} \) and \( L_q \) are the stator d-q axes mutual inductances. \( L_{ms} \) and \( L_{ls} \) are the magnetizing and stator leakage inductances, \( f \) and \( B \) are the moment of inertia and the viscous friction coefficient. \( \tau_l \) is the electromagnetic torque. \( \tau_f \) is the load torque. \( \phi'_{ds}, \phi'_{qs}, \phi'_{ds} \) and \( \phi'_{qs} \) are the rotor and stator d-q axes voltages. \( I_{ds}, I_{qs}, I_{dd} \) and \( I_{dq} \) are the rotor and stator d-q axes currents. \( \phi_{ds}, \phi_{qs}, \phi_{dd} \) and \( \phi_{dq} \) are the rotor and stator d-q axes fluxes. \( r_r \) and \( r_s \) indicate the rotor and stator resistances. \( p = d/dt \). \( \omega_r \) is the motor speed.

### 3 | EKF FOR THE 2-PIM ROTOR FLUX ESTIMATION

Considering the process and measurement noises, the extended 2-PIM state-space model can be expressed in the following overall form:

\[
\dot{X} = f(X, U, W) \\
Y = h(X, V) 
\]

where \( f(X, U, W) = AX + BU + W \) and \( h(X, V) = CX + V \). In addition, \( X, Y \) and \( U \) are state, output, and input matrices, respectively. Furthermore, \( W \) and \( V \) are the process and measurement noises. Using Equations (3) and (4), the 2-PIM model can be developed to estimate the rotor flux as given in Equations (5) and (6):

\[
\begin{bmatrix}
Y \\
P_{ds} \\
P_{qs}
\end{bmatrix}
= 
\begin{bmatrix}
C & 0 & 0 \\
1 & 0 & 0 \\
0 & 1 & 0
\end{bmatrix}
\begin{bmatrix}
P_{ds} \\
P_{qs} \\
\phi_{ds} \\
\phi_{qs}
\end{bmatrix}
+ 
V(t)
\quad (6)
\]

Notice that in this article DFOC of 2-PIM drives based on the rotor flux is presented. The stochastic continuous time system in Equations (3) and (4) must be written in the discrete time form using the Taylor series around a desired reference point \( \hat{X}_T \) in order to be acceptable with the EKF structure. The non-linear dynamic model of the 2-PIM can be written in the following discrete time overall form:

\[
X_{T+1} = f(X_T, U_T, W_T) \\
Y_T = h(X_T, V_T)
\quad (7)
\]

Considering the process and measurement noises as white noises, the recursive EKF algorithm including prediction and filtering stages can be written as Equations (8a–8e) [52]:

\[
\dot{X}_{T+1} = f\left(\hat{X}_{T}/T, U_T, 0\right) \\
P_{T+1/T} = F_T P_{T/T} F_T^T + Q \\
K_{T+1} = P_{T+1/T} H_T^T (H_T P_{T+1/T} H_T^T + R)^{-1} \\
\hat{X}_{T+1/T+1} = \hat{X}_{T+1/T} + K_{T+1} \left(Y_T - h\left(\hat{X}_{T+1/T}, 0\right)\right) \\
P_{T+1/T+1} = (I - K_{T+1} H_T) P_{T+1/T}
\quad (8d) \quad (8e)
\]

where \( R \) and \( Q \) are the covariance matrices of the measurement noise and the process noise, \( P \) is the state estimation error covariance matrix. Consequently, the rotor flux amplitude \( |\phi_r| \) and the rotor flux angle \( \alpha_r \) are obtained based on the estimated values of state variables as Equations (9) and (10):

\[
|\phi_r| = \sqrt{\left(\phi'_{dr}\right)^2 + \left(\phi'_{qr}\right)^2} \\
\alpha_r = \tan^{-1}\left(\phi'_{qr}/\phi'_{dr}\right)
\quad (9) \quad (10)
\]

To confirm the performances of the introduced EKF for a 2-PIM, some simulations are conducted using the MATLAB/ M-File software. The parameters of the studied 3-PIM in this paper are listed in Table 1.

### Table 1 Parameters of the studied 3-PIM

| Parameter                      | Value          |
|--------------------------------|----------------|
| Nominal Voltage (\(V_o\))     | 400 V          |
| Frequency (\(f\))             | 50 Hz          |
| Nominal power (\(P_{nom}\))   | 0.75 kW        |
| Pole                           | 4              |
| \(r_r\)                        | 14.64 \(\Omega\)|
| \(r_s\)                        | 10.44 \(\Omega\)|
| \(L_{ms}\)                     | 0.398 H        |
| \(L_s\)                        | 0.0167 H       |
| \(J\)                          | 0.016 kg.m\(^2\) |
In simulations, the fourth-order Runge-Kutta algorithm has been utilized to solve equations of healthy and faulted machines. During simulations, the machine is directly connected to an AC power source. In simulations, the healthy 3-PIM machine is simulated based on Equations (2a–2i). Moreover, the faulty 3-PIM machine is simulated using 2-PIM equations given by Equations (2a–2k).

Figure 3 shows real and estimated stator currents and rotor fluxes for the 2-PIM using the conventional EKF and the proposed EKF during the transient period (in Figure 3, 76 rad/s < \( \omega_r \) < 152 rad/s). In Figure 3a, the machine d-q axes currents and fluxes during the fault condition are estimated using the conventional EKF (the conventional EKF has been discussed in [52]). In Figure 3b, the machine d-q...
axes currents and fluxes during the fault condition are estimated using the introduced EKF (based on Equations (2b), (5) and (6)).

It is observed that the conventional EKF and the proposed EKF can estimate the stator currents. However, using the introduced EKF algorithm a lower error compared to the conventional EKF algorithm can be seen. Moreover, based on simulation results of Figure 3, the conventional EKF algorithm cannot estimate the rotor fluxes during the single-phase open fault condition. Nevertheless, using the introduced EKF system, good tracking compared to the conventional EKF system can be observed.

Figure 4 shows real and estimated stator currents and rotor fluxes at a nominal speed during both normal and single-phase open fault situations using the conventional EKF and the introduced EKF. In Figure 4a, the stator currents and the rotor fluxes during both normal and fault conditions are estimated using the conventional EKF. In Figure 4b, the stator currents and the rotor fluxes during both normal and fault conditions are estimated using the introduced EKF. In Figure 4a, at \( t = 1 \) s, single-phase open fault is introduced. In other words, at \( t < 1 \) s, the machine is simulated using 3-PIM equations given by Equations (2a–2l). Moreover, at \( t > 1 \) s, the machine is simulated using 2-PIM equations given by Equations (2a–2k).

From Figure 4, it is observed that the conventional EKF and the proposed EKF can estimate the stator currents and the rotor fluxes during both the normal and fault situations. However, during the fault condition, using the proposed EKF algorithm, a lower error can be seen compared to the conventional EKF algorithm.

4 | URTMs FOR 2-PIM CONTROL

Owing to the asymmetrical structure of 2-PIM equations, the use of flux-orientation principles requires a special attention. This asymmetry in the structure of the 2-PIM equations is the result of the different \( d \) and \( q \) inductances \( (L_{ds} \neq L_{qs} \text{ and } L_{d} \neq L_{q}) \) and causes an oscillating term in the 2-PIM electromagnetic torque (see Appendix). It is possible to eliminate the machine electromagnetic torque oscillating term using an appropriate URTM for stator current variables. The aim of using URTM is to obtain a balanced structure for the faulty 3-PIM electromagnetic torque equation.

4.1 | URTM for stator current variables

For this type of fault investigated in this article, the BRTM as shown in Equation (11) can be used for rotor variables [52].

![Figure 4](image-url)
\[ [T_r^e] = \begin{bmatrix} \cos \alpha_e & \sin \alpha_e \\ -\sin \alpha_e & \cos \alpha_e \end{bmatrix} \]  
(11)

For stator current variables, URTM is assumed as:

\[ [T_{\text{Rs}}] = \begin{bmatrix} f_1(\alpha_e) & f_2(\alpha_e) \\ f_3(\alpha_e) & f_4(\alpha_e) \end{bmatrix} \]  
(12)

where \( f_i(\alpha_e) \) is a function in terms of \( \alpha_e \). The 2-PIM electromagnetic torque equation can be expressed by:

\[ \tau_e = \frac{\text{pole}}{2} \left( L_d I_d' I_{d'q} - L_d I_{d'q}' I_{d'} \right) \]

\[ = \frac{\text{pole}}{2} \begin{bmatrix} I_d' & I_{d'q}' \end{bmatrix} \begin{bmatrix} 0 & L_d \\ -L_d & 0 \end{bmatrix} \begin{bmatrix} I_d' \\ I_{d'q}' \end{bmatrix} \]

\[ = \frac{\text{pole}}{2} \begin{bmatrix} I_d' & I_{d'q}' \end{bmatrix} \begin{bmatrix} f_1(\alpha_e) & f_2(\alpha_e) \\ f_3(\alpha_e) & f_4(\alpha_e) \end{bmatrix} \begin{bmatrix} I_d' \\ I_{d'q}' \end{bmatrix} \]  
(13)

then by substituting Equations (11) and (12) into Equation (13), we have:

\[ \tau_e = \left( f_1(\alpha_e)f_4(\alpha_e) - f_3(\alpha_e)f_2(\alpha_e) \right) \frac{\text{pole}}{2} \begin{bmatrix} I_d' & I_{d'q}' \end{bmatrix} \]

\[ \cdot \begin{bmatrix} L_d f_3(\alpha_e) & L_d f_1(\alpha_e) \\ -L_d f_4(\alpha_e) & L_d f_2(\alpha_e) \end{bmatrix} \begin{bmatrix} I_d' \\ I_{d'q}' \end{bmatrix} \]  
(14)

The 3-PIM torque equation in the healthy mode (superscript ‘b’) based on Equations (2) and (2) can be written as:

\[ \tau^b = \frac{\text{pole}}{2} \left( L_d I_q' I_{d'q} - L_d I_{d'q}' I_q' \right) \]

\[ = \frac{\text{pole}}{2} \begin{bmatrix} I_d' & I_{d'q}' \end{bmatrix} \begin{bmatrix} 0 & L_d \\ -L_d & 0 \end{bmatrix} \begin{bmatrix} I_d' \\ I_{d'q}' \end{bmatrix} \]

\[ = \frac{\text{pole}}{2} \begin{bmatrix} I_d' & I_{d'q}' \end{bmatrix} \begin{bmatrix} f_1(\alpha_e) & f_2(\alpha_e) \\ f_3(\alpha_e) & f_4(\alpha_e) \end{bmatrix} \begin{bmatrix} I_d' \\ I_{d'q}' \end{bmatrix} \]  
(15)

Based on Equations (14) and (15), it can be shown that the structure of the 2-PIM electromagnetic torque equation can be gained similar to that of the 3-PIM electromagnetic torque equation using \( f_1(\alpha_e) = (L_d/L_q)\cos \alpha_e, \ f_2(\alpha_e) = \sin \alpha_e, \ f_3(\alpha_e) = -L_d/L_q\sin \alpha_e, \ f_4(\alpha_e) = \cos \alpha_e \). Consequently, URTM for stator current variables can be expressed as Equation (16):

\[ [T_{\text{Rs}}] = \begin{bmatrix} (L_d/L_q)\cos \alpha_e & \sin \alpha_e \\ -L_d/L_q & \cos \alpha_e \end{bmatrix} \]  
(16)

### 4.2 URTM for stator voltage variables

As can be seen from Equations (2a–2i), the faulty 3-PIM model in the SR frame is similar to the model of a Single-phase Induction Machine (SPIM) with \( N_d \) and \( N_q \) turn numbers. Consequently, a 3-PIM with wye-connected under open-circuit fault can be considered equivalent to an asymmetrical SPIM with different turn numbers as shown in Figure 5.

The MMF of a healthy 3-PIM in the Rotating Reference (RR) frame (superscript ‘e’) and the MMF of the faulty 3-PIM in the SR frame can be written as Equations (17a) and (17b), respectively:

\[ \begin{bmatrix} F^e_{ds} \\ F^e_{d'q} \end{bmatrix} = \begin{bmatrix} \cos \alpha_e & \sin \alpha_e \\ -\sin \alpha_e & \cos \alpha_e \end{bmatrix} \begin{bmatrix} F^b_{ds} \\ F^b_{d'q} \end{bmatrix} \]  
(17a)

\[ \begin{bmatrix} F^{f}_{ds} \\ F^{f}_{d'q} \end{bmatrix} = \begin{bmatrix} N_d & 0 \\ 0 & N_q \end{bmatrix} \begin{bmatrix} I_d' \\ I_{d'q}' \end{bmatrix} \]  
(17b)

where \( F^b_{ds}, F^b_{d'q}, F^f_{ds}, F^f_{d'q} \) are the stator d-q MMFs of a healthy 3-PIM in the RR frame, the stator d-q MMFs of a healthy 3-PIM in the SR frame, and the stator d-q MMFs of the faulty 3-PIM in the SR frame. In order to control the faulty 3-PIM, the stator d-q MMFs of a healthy 3-PIM is considered equal to the stator d-q MMFs of the faulty 3-PIM. In other words, \( \begin{bmatrix} F^b_{ds} \\ F^b_{d'q} \end{bmatrix} = \begin{bmatrix} F^f_{ds} \\ F^f_{d'q} \end{bmatrix} \). As a result, based on Equations

![FIGURE 5 An asymmetrical SPIM](image-url)
(16), (17a) and (17b) with assuming \( N_d/N_q \cong L_d/L_q \) can be written as Equation (18):

\[
\begin{bmatrix}
F_{ds}^e \\
F_{qs}^e \\
F_{dq}^e \\
F_{qp}^e
\end{bmatrix} = \begin{bmatrix}
\cos \alpha_e & \sin \alpha_e \\
-\sin \alpha_e & \cos \alpha_e \\
N_d & 0 \\
0 & N_q
\end{bmatrix} \begin{bmatrix}
I_{ds} \\
I_{qs} \\
I_{dq} \\
I_{qp}
\end{bmatrix} = \begin{bmatrix}
N_d \cos \alpha_e & N_q \sin \alpha_e \\
-N_d \sin \alpha_e & N_q \cos \alpha_e \\
(N_d/N_q) \cos \alpha_e & (N_q/N_d) \sin \alpha_e \\
-(N_q/N_d) \sin \alpha_e & (N_d/N_q) \cos \alpha_e
\end{bmatrix}^{-1} \begin{bmatrix}
I_{ds} \\
I_{qs} \\
I_{dq} \\
I_{qp}
\end{bmatrix}
\]

(18)

As can be seen from Equation (18), using the presented URTM for stator current variables (Equation 16), the stator d-q MMFs of the faulty 3-PIM in the RR frame can be written as Equation (19):

\[
\begin{bmatrix}
F_{ds}^e \\
F_{qs}^e \\
F_{dq}^e \\
F_{qp}^e
\end{bmatrix} = \begin{bmatrix}
N_q & 0 \\
0 & N_q
\end{bmatrix} \begin{bmatrix}
I_{ds} \\
I_{qs}
\end{bmatrix}
\]

(19)

From Equation (19), it is seen that the faulty 3-PIM model in the RR frame can be considered equivalent to a symmetrical SPIM as shown in Figure 6.

According to Figures 5 and 6, following equations can be written:

\[
\begin{align*}
\frac{z_{ds}^e}{z_{ds}^e} &\cong \left(\frac{N_q}{N_d}\right)^2, \frac{z_{qs}^e}{z_{qs}^e} \cong \left(\frac{N_q}{N_d}\right)^2 = 1 \\
\frac{z_{ds}^e}{z_{qs}^e} &\cong \left(\frac{N_d}{N_q}\right)^2, \psi_3 = z_{ds}^e I_{ds}^e, \psi_3^e = z_{qs}^e I_{qs}^e
\end{align*}
\]

(20a)

As a result, the stator voltage equations of the faulty 3-PIM in the RR frame can be described by Equation (20b):

\[
\begin{bmatrix}
\psi_{d3}^e \\
\psi_{q3}^e
\end{bmatrix} = \begin{bmatrix}
0 \\
0
\end{bmatrix}
\]

(20b)

Consequently, the URTM for stator voltage variables can be expressed as:

\[
[T_{cs}^e] = \begin{bmatrix}
(L_q/L_d) \cos \alpha_e & \sin \alpha_e \\
-(L_q/L_d) \sin \alpha_e & \cos \alpha_e
\end{bmatrix}
\]

(21)

5 | FOC OF 2-PIM DRIVES

Employing URTMs given by Equations (16) and (21) and the BRTM for rotor components given by Equation (11), a new mathematical model can be developed to describe the vector control scheme. From Equations (11) and (16), the 2-PIM torque equation (Equation (13)) can be written as:

\[
\tau_e = \frac{\text{pole}}{2} L_q (I_{qs}^e I_{dr}^e - I_{ds}^e I_{qr}^e)
\]

(22)

The 2-PIM rotor voltage equations can be written as:

\[
[T_r^e] \begin{bmatrix}
\psi_{d3}^e \\
\psi_{q3}^e
\end{bmatrix} = [T_{r3}] \begin{bmatrix}
0 \\
0
\end{bmatrix}
\]

\[
= [T_{r3}] \begin{bmatrix}
L_d p & \omega_r L_q \\
-\omega_r L_d & L_q p
\end{bmatrix} [T_{r3}]^{-1} [T_{r3}] \begin{bmatrix}
I_{ds}^e \\
I_{qs}^e
\end{bmatrix} + [T_{r3}] \begin{bmatrix}
r + L_p & \omega_r L_r \\
r - \omega_r L_p & r + L_p
\end{bmatrix} [T_{r3}]^{-1} [T_{r3}] \begin{bmatrix}
I_{dr}^e \\
I_{qr}^e
\end{bmatrix}
\]

(23)
which gives,

$$
\begin{align*}
0 &= \left[ T_r^e \right] \begin{bmatrix} 0 & \omega_r L_q & 0 \\ 0 & 0 & -\omega_r L_d \\ -\omega_r L_d & 0 & 0 \end{bmatrix} \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} \\
+ [T_r^e] \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \left( p \left( \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} \right) + \left[ T_h^e \right]^{-1} \begin{bmatrix} p I_{ds}^r \\ p I_{qs}^r \\ p I_{qp}^r \end{bmatrix} \right) \\
+ [T_r^e] \begin{bmatrix} r_r & 0 & 0 \\ 0 & \omega_r L_r & 0 \\ 0 & 0 & \omega_r L_r \end{bmatrix} \left( \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} \right) \\
+ [T_r^e] \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \left( \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} \right)
\end{align*}
$$

(24)

Substituting Equations (11) and (16) into Equation (24), the 2-PIM rotor voltage equations can be simplified as Equation (25):

$$
\begin{align*}
0 &= \begin{bmatrix} L_d p & -(\omega_e - \omega_r) L_q \\ (\omega_e - \omega_r) L_q & L_d p \\ r_r + L_r p & -(\omega_e - \omega_r) L_r \\ (\omega_e - \omega_r) L_r & r_r + L_r p \end{bmatrix} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{dr}^r \\ I_{qr}^r \end{bmatrix} \\
+ \begin{bmatrix} L_d p & -(\omega_e - \omega_r) L_q \\ (\omega_e - \omega_r) L_q & L_d p \\ r_r + L_r p & -(\omega_e - \omega_r) L_r \\ (\omega_e - \omega_r) L_r & r_r + L_r p \end{bmatrix} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{dr}^r \\ I_{qr}^r \end{bmatrix}
\end{align*}
$$

(25)

where \( \omega_e \) is the angular speed of the rotor flux oriented reference frame. By applying Equations (11) and (16) to the 2-PIM rotor flux equations, we have:

$$
\begin{align*}
\left[ T_r^e \right] \begin{bmatrix} \phi_{dr}^e \\ \phi_{qr}^e \end{bmatrix} &= \left[ T_r^e \right] \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} \\
+ [T_r^e] \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix}
\end{align*}
$$

(26)

Substituting Equations (11) and (16) into Equation (26), the 2-PIM rotor flux equations can be simplified as Equation (27):

$$
\begin{align*}
\phi_{dr}^e &= \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} + \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} \\
\phi_{qr}^e &= \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix} + \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{qp}^r \end{bmatrix}
\end{align*}
$$

(27)

The 2-PIM stator voltage equations (Equation 18) can be written as Equation (28):

$$
\begin{align*}
\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix} &= \begin{bmatrix} T_{vs}^e \\ T_{qs}^e \end{bmatrix} \begin{bmatrix} r_s & 0 & 0 \\ 0 & r_s & 0 \\ 0 & 0 & r_s \end{bmatrix} \begin{bmatrix} T_h^e \end{bmatrix}^{-1} \begin{bmatrix} T_h^e \\ T_h^e \\ T_h^e \end{bmatrix} \begin{bmatrix} I_{ds}^s \\ I_{qs}^s \\ I_{qp}^s \end{bmatrix} \\
+ \begin{bmatrix} T_{vs}^e \\ T_{qs}^e \end{bmatrix} \begin{bmatrix} L_{ds} & 0 & 0 \\ 0 & L_{qs} & 0 \\ 0 & 0 & L_{qp} \end{bmatrix} \left( p \left( \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{ds}^s \\ I_{qs}^s \\ I_{qp}^s \end{bmatrix} \right) + \left[ T_h^e \right]^{-1} \begin{bmatrix} p I_{ds}^s \\ p I_{qs}^s \\ p I_{qp}^s \end{bmatrix} \right) \\
+ \begin{bmatrix} T_{vs}^e \\ T_{qs}^e \end{bmatrix} \begin{bmatrix} L_d & 0 & 0 \\ 0 & L_q & 0 \\ 0 & 0 & L_r \end{bmatrix} \left( p \left( \left[ T_h^e \right]^{-1} \begin{bmatrix} I_{dr}^s \\ I_{qr}^s \end{bmatrix} \right) + \left[ T_h^e \right]^{-1} \begin{bmatrix} p I_{dr}^s \\ p I_{qr}^s \end{bmatrix} \right)
\end{align*}
$$

(28)

Substituting Equations (11), (16) and (21) into Equation (28), the 2-PIM stator voltage equations considering \((L_q/L_d)^2 = L_{qp}/L_{ds}\) can be expressed as Equation (29):

$$
\begin{align*}
\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix} &= \begin{bmatrix} r_s + L_q p & -\omega_e L_q \\ \omega_e L_q & r_s + L_q p \end{bmatrix} \begin{bmatrix} I_{ds}^s \\ I_{qs}^s \end{bmatrix} \\
+ \begin{bmatrix} L_d p & -(\omega_e - \omega_r) L_q \\ (\omega_e - \omega_r) L_q & L_d p \\ r_r + L_r p & -(\omega_e - \omega_r) L_r \\ (\omega_e - \omega_r) L_r & r_r + L_r p \end{bmatrix} \begin{bmatrix} I_{ds}^r \\ I_{qs}^r \\ I_{dr}^r \\ I_{qr}^r \end{bmatrix}
\end{align*}
$$

(29)

where

$$
\begin{align*}
\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix} &= \begin{bmatrix} L_q^2 & -r_s \\ -r_s & -0.5 \sin 2\alpha_e & \sin 2\alpha_e \end{bmatrix} \begin{bmatrix} I_{ds}^e \\ I_{qs}^e \end{bmatrix}
\end{align*}
$$

(30)

As expected Equations (22), (25), (27) and (29) are similar to that of the 3-PIM equations. Therefore, it is promising to apply the DFOC principles to control 2-PIM drives. From Equations (22), (25), (27) and (29), the 2-PIM equations with assumption \( \phi_{dr}^e = |\phi_r| \) and \( \phi_{qr}^e = 0 \) can be written as Equation (31):

$$
\begin{align*}
\tau_e &= \frac{\text{pole}}{2} \frac{L_q}{L_r} |\phi_r| I_{qs}^e, \quad |\phi_r| + \frac{L_r}{r_r} p |\phi_r| = L_q I_{ds}^e \\
v_{ds}^e &= v_{ds}^d + v_{ds}^* + \psi_{dr}^e, \quad v_{qs}^e = v_{qs}^d + v_{qs}^* + \psi_{qr}^e \\
\alpha_e &= \omega_r + \frac{r_r L_q I_{qs}^e}{L_r |\phi_r|}
\end{align*}
$$

(31)
\[
\begin{align*}
v_{ds}^d &= \frac{r_s L_s}{L_r} \left( L_q I_{ds}^* - \frac{\phi_r}{L_r} \right) + \omega_r I_{qs} \left( -L_{qs} + \frac{L_s^2}{L_r} \right), \\
v_{qs}^d &= \frac{\omega_r L_s}{L_r} \phi_r - \omega_r I_{ds}^* \left( -L_{qs} + \frac{L_s^2}{L_r} \right), \\
v_{ds}^* &= r_s I_{ds}^* + \left( L_{qs} - \frac{L_s^2}{L_r} \right) \alpha^*_{ds}, \\
v_{qs}^* &= r_s I_{qs}^* + \left( L_{qs} - \frac{L_s^2}{L_r} \right) \alpha^*_{qs}, \\
\end{align*}
\]

Based on Equations (31) and (32), Figure 7 can be introduced for DFOC of the wye-connected induction machine drives. It can be noted that the introduced controller of Figure 7 can be exploited for both 2-PIM and 3-PIM drives. In other words, unlike the conventional EKF-based DFOC, the introduced controller in this article can be used for a wye-connected 3-PIM during normal and stator winding open-phase failure conditions. In Figure 7, the control parameters and the TM to control 2-PIM and 3-PIM drives are given in Tables 2 and 3, respectively.

In Figure 7, \( |\phi_r^*|, r_s^*, \) and \( \omega_r^* \) denote the reference rotor flux amplitude, the reference electromagnetic torque, and the reference electromagnetic torque, respectively.

**TABLE 2** Control parameters and TM to control the 2-PIM drives

| Control Parameters | TM |
|--------------------|----|
| \( L_d = L_d + 1.5L_{ms} \) | \([T_e] = \sqrt{2} \begin{bmatrix} 0.5 & -0.5 \\ 0.5 & 0.5 \end{bmatrix} \) |
| \( L_q = L_q + 0.5L_{ms} \) |
| \( L_d = 1.5L_{ms}, L_q = (\sqrt{3}/2)L_{ms} \) |

**TABLE 3** Control parameters and TM to control the 3-PIM drives

| Control Parameters | TM |
|--------------------|----|
| \( L_d = L_d + 1.5L_{ms} \) | \([T_e] = \sqrt{2} \begin{bmatrix} 1 & -0.5 & -0.5 \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \) |
| \( L_q = L_q + 0.5L_{ms} \) |
| \( L_d = 1.5L_{ms}, L_q = (\sqrt{3}/2)L_{ms} \) |

**FIGURE 7** Block diagram of the introduced EKF-based DFOC.
reference speed, respectively. From Equation (31), $I_{sq}^r$ controls the electromagnetic torque and $I_{ds}^r$ controls the rotor flux, respectively. Block $[T_{qs}^r]$ is the URTM to transform the stator current components from the SR frame to the RR frame. Block $[T_{es}^r]$ is the URTM to transform the stator voltage components from the RR frame to the SR frame. Block $[T]$ is the TM to transform the stator components from a-b frame to d-q SR frame. $I_{ds}^e$ and $I_{qs}^e$ represent reference stator currents and are determined from the flux controller and the torque controller, respectively. As shown in Figure 7 the machine voltages ($v_{ds}^e, v_{qs}^e$) are obtained from Decoupling Circuit block, current controller blocks, and $e_{ds}^r, e_{qs}^r$.

For comparison of the proposed scheme with the conventional DFOC strategy, the block diagram of the conventional DFOC based on EKF for 3-PIM drives has been shown in Figure 8 (the conventional EKF-based DFOC of 3-PIM drives has been discussed in [52]). In Figure 8:

$$[T] = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -0.5 & -0.5 \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix}, \quad [T^e_r] = \begin{bmatrix} \cos \alpha_r & \sin \alpha_r \\ -\sin \alpha_r & \cos \alpha_r \end{bmatrix}$$

$L_s = L_{ts} + 1.5L_{ms}, \quad L = 1.5L_{ms}$

To show the performance of the introduced EKF-based DFOC scheme, some simulations using MATLAB/M-File software have been performed. In simulations, single-phase open fault happens in phase ‘c’ of the stator windings and

**FIGURE 8** Block diagram of the conventional EKF-based DFOC

**FIGURE 9** Simulation results of the estimated rotor flux, the stator q-axis current in the RR frame and the speed of the 3-PIM drive during normal and fault conditions using the proposed EKF-based DFOC system. (a) 2-PIM vector control using the proposed method, (b) 3-PIM vector control using the proposed method.
the machine reference flux is set to 1 wb. The parameters of the simulated 3-PIM are listed in Table 1.

The values of the covariance matrices are very effective in the EKF performance. The values of the covariance matrices should be determined by considering the stochastic properties of the corresponding noises. Nevertheless, since these are typically unknown, these matrices are used as weighting factors [53]. In simulations, to avoid the computational complexity, the process noise covariance $Q$ which includes model uncertainty, machine parameter inaccuracy, rounding and truncation error, system disturbances [49] and the measurement noise covariance $R$ which includes current and voltage measurement errors and A/D quantization errors [49] are chosen in the diagonal form. In this paper, $Q$ and $R$ have been considered as:

$$Q = \text{diag}[4.1 \times 10^{-6}, 1.3 \times 10^{-6}, 5.3 \times 10^{-9}, 1.7 \times 10^{-9}]$$
$$R = \text{diag}[1 \times 10^{-2}, 1 \times 10^{-2}]$$

To show the performances of the introduced EKF-based DFOC scheme for vector control of 3-PIM during normal and fault conditions, two simulations have been performed. Figure 9a shows the simulation results of the 2-PIM based on Figure 7 and Table 2. Furthermore, Figure 9b shows the simulation results of the 3-PIM based on Figure 7 and Table 3. In Figure 9, the reference speed is changed from 100 rad/s to 60 rad/s at $t = 22.2$ s. Figure 9 shows the estimated rotor flux amplitude, the stator $q$-axis current in the RR frame and the speed of the machine.

As it is seen from Figure 9, during both the normal and fault conditions, the estimated rotor flux amplitude and the motor speed can follow their reference values properly. Additionally, in both simulations, the stator $q$-axis current magnitude has reasonable oscillations and its average value except during the transient period is zero.

To check the performances of the proposed strategy at low- and zero-speed, three different simulations have been conducted. Figure 10a shows the simulation results of the 2-PIM based on [40], Figure 10b shows the simulation results of the 2-PIM based on [27], and Figure 10c shows the simulation results of the 2-PIM based on Figure 7 and Table 2. In Figure 10, the machine reference speed is changed from 0 rad/s to 3 rad/s at $t = 26.7$ s. Figure 10 shows the comparison of the estimated rotor flux amplitude, the stator $q$-axis current in the RR frame and the speed of the 2-PIM. As can be observed in all simulations of Figure 10, the estimated rotor flux amplitude and the speed of the 2-PIM can follow their reference values. Nevertheless, as shown in

![Figure 10](image-url)  
**Figure 10** Simulation results of the estimated rotor flux, the stator $q$-axis current in the RR frame and the speed of the 2-PIM drive based on [27,40] and the proposed EKF-based DFOC system. (a) 2-PIM vector control based on [40], (b) 2-PIM vector control based on [27], (c) 2-PIM vector control using the proposed method.
Figure 10, using the proposed EKF-based DFOC scheme in this paper, low flux and speed oscillations compared to the presented strategies in [27, 40] can be achieved. It is also seen that in all simulations of Figure 10, the average value of the stator q-axis current in the RR frame has zero value. Nevertheless, using the proposed control system, low stator q-axis current oscillations compared to other control strategies can be seen.

6 | EXPERIMENTAL RESULTS

To verify the effectiveness of the proposed EKF-based DFOC scheme, a laboratory prototype is made and tested. The main elements which are used in the laboratory tests are shown in Figure 11.

The experiments of the wye-connected 3-PIM were implemented by DSP/TMS320F28335. In tests, the NP of the wye-connected 3-PIM is accessible, and the 3-PIM is coupled to a DC generator as the load. In the proposed method and during the fault condition, the NP of the 3-PIM is connected to the DCBM. In this condition, the currents of the 2-PIM are not dependent on each other and can be controlled separately [30]. The sampling time is considered 100 μs. It can be pointed out that, when the fault happens, high frequency current in the faulted phase reduces to zero in few milliseconds, permitting fast fault detection [28]. The parameters of the experimented 3-PIM are listed in Table 1.

6.1 | Evaluation of the conventional and the proposed EKF-based DFOC strategies during normal and fault conditions

To evaluate the performances of conventional and introduced EKF-based DFOC schemes for vector control of 3-PIM during normal and fault conditions, three tests have been conducted. Figures 12a and 13a show the experimental results of the 2-PIM using the conventional EKF-based DFOC (based on Figure 8). Moreover, Figures 12b and 13b show the experimental results of the 2-PIM using the introduced EKF-based DFOC (based on Figure 7 and Table 2). In addition, Figure 12c shows the experimental results of the healthy 3-PIM using the introduced EKF-based DFOC (based on Figure 7 and Table 3). In tests, single-phase open fault happens in phase ‘c’ of stator windings.

In Figures 12a, 13a, and 12c, the NP of the machine is not connected to the DCBM and in Figures 12b and 13b the NP of the machine is connected to the DCBM. In these tests, the reference speed is changed from 100 rad/s to 60 rad/s at t = 22.2 s. Besides, the machine reference flux is set to 1 wb.

Figure 12a,b displays the estimated rotor flux amplitude, the stator q-axis current in the RR frame and the speed of the 2-PIM drive. As it is observed in both tests the estimated rotor flux amplitude and the motor speed can track their reference values. Nevertheless, as exposed in Figure 12a,b, using the proposed DFOC approach, low rotor flux amplitude and speed oscillations compared to the conventional DFOC system can be seen. As shown in Figure 12a,b, using the introduced scheme, fewer stator q-axis current oscillations compared to the conventional scheme can be obtained. As in the FOC strategy based on the rotor flux, the stator q-axis current in the RR frame is proportional to the machine electromagnetic torque \( \tau_e = \frac{\text{pole} \cdot L_s}{L_r} |\phi_s| |I_{q1}| \), it can be concluded that using the introduced EKF-based DFOC, the electromagnetic torque oscillations of the 2-PIM can be significantly decreased. As it is seen from Figure 12c, during the normal condition, the estimated rotor flux amplitude and the motor speed can follow their reference values with reasonable oscillations. It can be mentioned that the performance of the proposed strategy during the normal mode is the same as the performance of the conventional EKF-based DFOC strategy during the normal mode.

In summary, the peak-to-peak ripples of the estimated rotor flux amplitude, the stator q-axis current in the RR frame and the speed using conventional and proposed controllers during normal and fault conditions when 24 s < t < 26 s are compared in Table 4.
Figure 13a,b shows the stator phase currents and the estimated rotor flux angle of the 2-PIM using conventional and proposed EKF-based DFOC approaches for the scenario shown in Figure 12.

It can be seen that the proposed technique has performed well under the open-phase fault operation. From Figure 13a, it is also seen that, as the NP of the 2-PIM is not connected to the DCBM, the healthy currents of the machine using the conventional controller depend on each other \( I_{\text{d}} = -I_{\text{q}} \) and cannot be controlled independently. As a result, using the conventional controller, the stator d-q axes currents cannot be decoupled into two components and therefore an independent control of the torque and flux is not possible. From Figure 13b, it is seen that as the NP of 2-PIM is connected to the DCBM, the currents of healthy phases can be controlled separately. In this case, the independent control of the torque and flux is possible. The waveforms exposed in Figure 13 evidence the ability of the proposed EKF-based DFOC system to balance the faulted machine currents.

6.2 Evaluation of the FTC of 3-PIM drive using conventional and proposed EKF-based DFOC strategies

To check the FTC performance of 3-PIM drive using the conventional EKF-based DFOC strategy and the introduced EKF-based DFOC strategy, two experiments have been conducted. In Figures 14a and 15a, the 3-PIM and the 2-PIM vector control using the conventional method, (b) 2-PIM vector control using the proposed method, (c) 3-PIM vector control using the proposed method.
The peak-to-peak ripples of the estimated rotor flux amplitude, the stator q-axis current in the RR frame and the speed using conventional and proposed controllers during normal and fault conditions.

|                          | 2-PIM vector control using the conventional method | 2-PIM vector control using the proposed method | 3-PIM vector control using the proposed method |
|--------------------------|--------------------------------------------------|-----------------------------------------------|-----------------------------------------------|
| The peak-to-peak ripples of the estimated rotor flux amplitude | 0.1 wb                                            | 0.07 wb                                       | 0.05 wb                                       |
| The peak-to-peak ripples of the stator q-axis current in the RR frame | 0.95 A                                           | 0.4 A                                         | 0.3 A                                         |
| The peak-to-peak ripples of the speed                            | 11.3 rad/s                                       | 3.2 rad/s                                     | 2.1 rad/s                                     |

PIM are run using the conventional EKF-based DFOC (based on Figure 8). Additionally, in Figures 14b and 15b, the 3-PIM and the 2-PIM are run using the introduced EKF-based DFOC (based on Figure 7, Tables 2 and 3). In Figures 14a and 15a, the fault happens at $t = 24.1$ s and in Figures 14b and 15b, the fault happens at $t = 23.7$ s. In both tests, the NP of the machine is connected to the DCBM. Moreover, the machine reference speed and machine reference flux are set to 100 rad/s and 1 wb, respectively. In tests, the open-phase fault happens in phase 'c' of the stator windings.

Figure 14a, b illustrates the estimated rotor flux amplitude, the stator q-axis current in the RR frame and the speed of the 3-PIM and the 2-PIM. As it is shown in Figure 14, both conventional and proposed EKF-based DFOC systems during normal and fault operation conditions regulate the estimated rotor flux amplitude and the rotor speed at their reference values. However, the proposed DFOC method during fault produces smaller oscillations compared to the conventional DFOC method. Based on experimental results of the stator q-axis current in the RR frame, it can be concluded that the proposed scheme can decrease the torque oscillations of the 2-PIM.

Figure 15a, b displays the stator phase currents and the estimated rotor flux angle of the 2-PIM using conventional and proposed EKF-based DFOC systems. (a) Conventional method, (b) Proposed method.
forward and backward components appear in the equations of the faulty motor. The backward components cause oscillations in the machine torque and the unbalance in the motor currents. To solve this problem, in this article, two URTMs have been used in order to eliminate the backward components.

6.3 | Evaluation of conventional and proposed EKF-based DFOC strategies during the load condition

To show the performances of the conventional EKF-based DFOC and the introduced EKF-based DFOC during the load condition, two experiments have been performed. Figure 16a displays the results of the 2-PIM using the conventional EKF-based DFOC technique (based on Figure 8). Furthermore, Figure 16b displays the results of the 2-PIM using the introduced EKF-based DFOC technique (based on Figure 7 and Table 2). In both tests, the NP of the machine is connected to the DCBM. Furthermore, the reference speed, the load torque and the reference rotor flux are set to 120 rad/s, 1.5 N.m and 1 wb, respectively. It should be noted that the load torque limitation for a wye-connected 3-PIM during single-phase open fault is about 38% of the nominal torque [27]. In tests, fault happens in phase ‘b’ of stator windings.

Figure 16 illustrates the estimated rotor flux amplitude, the stator q-axis current in the RR frame, the motor speed and the stator phase currents of the 2-PIM. As it is shown in Figure 16, the introduced DFOC system during fault produces smaller ripples in the rotor flux, the stator q-axis current and the motor speed responses compared to the conventional DFOC system. It can be also observed that the stator phase currents of the proposed DFOC strategy in comparison with the classic controller are more sinusoidal.

6.4 | Evaluation of the proposed EKF-based DFOC strategy during the load condition and different speeds

To check the performances of the proposed EKF-based DFOC strategy during the load condition and different speeds, an experiment based on Figure 7 and Table 2 has been conducted. Figure 17 shows the estimated rotor flux amplitude, the stator q-axis current in the RR frame, the motor speed and the stator phase currents of the 2-PIM during no-load and load conditions. In this figure, a step load torque of 1 N.m is applied at \( t = 11.1 \) s. In Figure 17, the machine reference speed is changed from 150 rad/s to 40 rad/s at \( t = 21.7 \)s. Besides, the machine reference flux is set to 1wb. In this test, the NP of the machine is connected to the DCBM and single-phase open fault happens in phase ‘c’ of the stator windings.

It can be seen that during the load condition, the speed and the estimated rotor flux of the 2-PIM closely follow their reference values. The average value of the stator q-axis current magnitude except during the transient period in both no-load and load conditions, is constant. Moreover, except during the transient period, the average value of the stator q-axis current magnitude is proportional to the applied load. As shown in Figure 17, the 2-PIM currents can be controlled independently, and they are perfectly sinusoidal.

6.5 | Evaluation of the proposed EKF-based DFOC strategy at low- and zero-speed

Figure 18 demonstrates the experimental results of the proposed EKF-based DFOC system based on Figure 7 and Table 2 at low- and zero-speed. Figure 18 shows the estimated rotor flux amplitude, the stator q-axis current in the RR frame, and the motor speed of the 2-PIM. In Figure 18, the machine reference speed is changed from 0 rad/s to 3 rad/s at \( t = 26.7 \)s. Moreover, the machine reference flux is set to 1wb. In this test,
the NP of the 2-PIM is connected to the DCBM and the single-phase open fault happens in phase ‘c’ of the stator windings.

It is easily observed in Figure 18 that the flux and speed tracking is satisfactory. Furthermore, the stator q-axis current magnitude has reasonable oscillations and its average value except during the transient period is zero. In addition, the results of Figure 18 agree with those of simulation in Figure 10c.

7 | CONCLUSION

This article presents a novel DFOC strategy for open-phase FTC of wye-connected 3-PIM drives. In the proposed control system, instead of the flux measurement, a flux estimator based on an EKF is utilized. The proposed EKF-based DFOC strategy with some modifications in the control parameters and the TM can be used for normal and stator winding open-circuit failure operations of 3-PIM drives. Simulation and experimental results in different operating conditions confirm the good performance of the proposed EKF-based DFOC system. The introduced controller exhibits a significant improvement in the reduction of the flux, torque, and speed oscillations compared to the conventional control scheme. The proposed strategy is suitable for critical industrial applications of the 3-PIM drives, it will retain the system running even during the single-phase open fault.

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APPENDIX

The 2-PIM electromagnetic torque equation using the BRTM can be expressed by Equations (A1) and (A2):

\[
\tau_e = \frac{\text{pole}}{2} \left( L_d I_q p_{d} - L_d I_d p_{q} \right) = \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & -L_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & L_q \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right]^T \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right] \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left( L_d \right) \frac{\text{pole}}{2} \left( L_d \right) \tau_e
\]

\[
\tau_e = \frac{\text{pole}}{2} \left( L_d I_q p_{d} - L_d I_d p_{q} \right) = \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & -L_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & L_q \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right]^T \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right] \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left( L_d \right) \frac{\text{pole}}{2} \left( L_d \right) \tau_e
\]

\[
\tau_e = \frac{\text{pole}}{2} \left( L_d I_q p_{d} - L_d I_d p_{q} \right) = \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & -L_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & L_q \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right]^T \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right] \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left( L_d \right) \frac{\text{pole}}{2} \left( L_d \right) \tau_e
\]

\[
\tau_e = \frac{\text{pole}}{2} \left( L_d I_q p_{d} - L_d I_d p_{q} \right) = \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & -L_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & L_q \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right]^T \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right] \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left( L_d \right) \frac{\text{pole}}{2} \left( L_d \right) \tau_e
\]

\[
\tau_e = \frac{\text{pole}}{2} \left( L_d I_q p_{d} - L_d I_d p_{q} \right) = \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & -L_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ I_d \quad I_q \right] \begin{bmatrix} 0 & L_q \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right]^T \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left[ \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \right] \begin{bmatrix} \begin{bmatrix} I_d \end{bmatrix} \end{bmatrix}
\]

\[
= \frac{\text{pole}}{2} \left( L_d \right) \frac{\text{pole}}{2} \left( L_d \right) \tau_e
\]