Passive Fault Tolerant Control of Induction Motors Using Nonlinear Block Control

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Abstract: This paper studies passive fault tolerant control (PFTC) of three phase induction motors (IMs). First, a nonlinear block control (NBC) transformation is applied to handle nonlinearities of motor faulty model. Afterwards, a fault tolerant speed controller is developed based on sliding mode control (SMC) technique with linear sliding surface. However, the utilized NBC transformation needs multiple derivatives calculations which make it computationally burden. In addition, this NBC transformation has unmatched fault rejection problem which SMC controller is not able to remove the fault. To avoid the mentioned drawbacks, a novel and simple NBC transformation is proposed for IM. Also, a new nonlinear sliding surface is presented in order to enhance the SMC controller performance. Comparative simulations demonstrated performance of the proposed controllers for matched and unmatched faults and load torque uncertainties.

Keywords: Induction motor, Fault tolerant control, Nonlinear block control, Sliding mode control, Matched and unmatched faults.

1. INTRODUCTION

When a fault occurs in a system components including sensors, actuators, and plant, it can cause performance degradation and even instability of the system. Therefore, there is a crucial need to design fault tolerant controllers (FTCs) to compensate faults effects and guarantee system reliability and stability with an acceptable performance. Generally speaking, the FTCs can be classified into two types (Kanev, 2006; Zhang and Jiang, 2008; Campos-Delgado et al., 2008): active fault tolerant control (AFTC) and passive fault tolerant control (PFTC). The AFTC technique usually needs a fault detection and isolation (FDI) scheme that has the task to detect and localize faults that eventually occur in the system. The PFTC is based on robust control techniques which makes system insensitive to certain faults that taken into account in the design step. This technique requires no redesign and online detection of faults, and therefore is computationally more interesting.

Three phase induction motors (IMs) are very important in industry because of their reliability, high performance and low costs (Gaied, 2014b). However, due to mechanical, environmental, thermal and electrical stresses, IMs are subjected to multiple faults. Stator and rotor failures including eccentricity and broken bars are the most famous asymmetric faults. The eccentricity faults can be categorized into two types: static eccentricity and dynamic eccentricity. The static and dynamic eccentricities can cause stator and rotor asymmetries, respectively (Vas, 1994; Bonivento et al., 2004).

From AFTC point of view, different FDI techniques have been proposed for dynamical systems such as IMs for designing reliable controllers. In (Cho et al., 1992; Bachir et al., 2006; Karami et al., 2010), the authors have detected the broken bar fault by model-based parameter estimation techniques. Multiple signal processing techniques like fast Fourier transform (FFT) are studied in (Hedayati Kia et al., 2004; Vaimann and Kallaste, 2011). In (Filippetti et al., 2000; Benbouzid et al., 2007; Zidani et al., 2008), artificial intelligence techniques have discussed for fault detection. Also, some other sliding mode (Yan and Edwards, 2007), adaptive (Najafabadi et al., 2011), and sensorless methods (Holtz, 2006; Aurora and A. Ferrara, 2007; Ghanes and Zheng, 2009a; Ghanes et al., 2010b; Gaeid et al., 2012a) are proposed in the recent years.

From PFTC point of view, in recent years, many methods have been employed and combined with each other's for providing reliable controllers for linear and nonlinear systems. A reliable H_∞ controller has been proposed in (Shi et al., 2014). In (Jin et al., 2013), a non-fragile H_∞ compensation filter has been designed in the framework of linear matrix inequalities (LMIs) technique. An output feedback fuzzy controller has been developed in (Wang et al, 2013) for T–S fuzzy systems with sensor faults based on LMI. Passivity technique is discussed in (Benosman and Lum, 2010a, 2010b) with respect to loss of actuator effectiveness. Moreover, SMC technique have received great attentions for spacecraft and quadrotor vehicles control due to its robust behaviour against model uncertainties (Li et al., 2013; Mirshams et al., 2014; Merheb et al., 2015).

In recent two decades, some studies have been developed for IMs control with PFTC (Bonivento *et al.,* 2004, Fekih, 2007, Djeghali *et al.,* 2011, 2013), and the most interesting approach is the SMC. In (Fekih, 2007), the author have proposed an approach which detects the occurrence of fault and switch itself between a nominal operation controller and sliding mode faulty controller. In (Djeghali *et al.,* 2011, 2013), a second-order sliding mode controller (SOSMC) is designed based of backstepping technique. The SOSMC is utilized to avoid the chattering phenomenon, also the backstepping technique is applied to bring the IM model nonlinearities into the control inputs direction. But both of the mentioned literatures only have studied the rotor broken bar fault and there is no attention on other faults like the stator eccentricity.

In this paper, two passive fault tolerant sliding mode controllers are developed for IM in order to attenuate the effects of the stator and rotor faults (broken bar plus eccentricity) and the load torque variations. The first controller is developed based on NBC technique which is reported in (Loukianov, 2002a; Loukianov *et al.,* 2002b). This NBC technique is a powerful tool to handle nonlinearities of IM model. But it is computationally burden and it is not able to remove the complete effects of non-vanishing unmatched faults/uncertainties. To solve the mentioned problem an improved NBC technique is proposed, which is the first novelty of this study. On the other hand, since the SMC with linear sliding surface presents slow transient responses and considerable tracking errors (Shoja Majidabad and Toosian Shandiz, 2012), then a novel nonlinear sliding surface is suggested that shows fast transient response with small tracking errors. This technique is known as nonlinear sliding mode control (NSMC).

This paper is organized as follows: In Section 2, the IM healthy and faulty dynamical models are presented. Two NBC transformation techniques are implemented for IM faulty model in Section 3. In Section 4, two robust fault tolerant sliding mode controllers are developed based on the NBC transformed models. In Section 5, the merits of the designed controllers are verified by the simulations on IM subjected to the matched/unmatched faults and load torque disturbances. Finally, conclusions are given in Section 6.

2. DYNAMIC MODEL OF IM AND VECTROR CONTROL

In this section, healthy and faulty models of IM are expressed. In addition, the model simplification based on field-oriented control (FOC) technique is explained.

2.1. Healthy Model

Under balance operation, the state space model of IM is presented in the synchronously rotating reference frame (d-q) by the following equations (Fekih, 2007; Djeghali et al., 2013; Krause et al., 2013):

$$
\begin{aligned}
v_{ds} &= R_s i_{ds} + \frac{d\psi_{ds}}{dt} - \omega_s \psi_{qs} \\
v_{qs} &= R_s i_{qs} + \frac{d\psi_{qs}}{dt} + \omega_s \psi_{ds} \\
v_{dr} &= 0 = R_r i_{dr} + \frac{d\psi_{dr}}{dt} - (\omega_s - \omega_r)\psi_{qr} \\
v_{qr} &= 0 = R_r i_{qr} + \frac{d\psi_{qr}}{dt} + (\omega_s - \omega_r)\psi_{dr} \\
\\
\psi_{ds} &= L_s i_{ds} + L_m i_{dr} \\
\psi_{qs} &= L_s i_{qs} + L_m i_{qr} \\
\psi_{dr} &= L_r i_{dr} + L_m i_{ds} \\
\psi_{qr} &= L_r i_{qr} + L_m i_{qs} \\
\\
\frac{d\omega_m}{dt} &= \frac{1}{J}\left(T_e - F\omega_m - T_L\right), \qquad \omega_r = p\omega_m \\
T_e &= p\frac{L_m}{L_r}\left(\psi_{dr} i_{qs} - \psi_{qr} i_{ds}\right)
\end{aligned}
\tag{1}
$$

where i_{ds}, i_{qs} are the components of the stator current, ψ_{dr}, ψ_{qr} are the components of the rotor flux, v_{ds}, v_{qs} are the stator voltage components, v_{dr}, v_{qr} are the rotor voltage components, ω_s is the stator pulsation, ω_r is the electrical angular speed, ω_m is the mechanical speed, R_s and R_r are stator and rotor resistances, L_s and L_r are stator and rotor inductances and L_m is the mutual inductance, p is the number of pole pairs, T_e is the electromagnetic torque, T_L is an unknown load torque, J is the moment of inertia coefficient, and F is the friction coefficient.

Based on the FOC strategy, the coupling between the rotor flux vector $[\psi_{dr}, \psi_{qr}]^T$ and the electromagnetic torque T_e should be reduced. This coupling is reducible by selecting the rotor flux orientation as follows (Fekih, 2007; Djeghali et al., 2013):

$$
\psi_{dr} = \psi_r \quad , \psi_{qr} = 0 \tag{2}
$$

From the above assumption, the IM model will be simplified as:

$$
\begin{aligned}
\frac{d\omega_m}{dt} &= \frac{pL_m}{L_r J} i_{qs}\psi_{dr} - \frac{F}{J} - \frac{T_L}{J} \\
\frac{d\psi_{dr}}{dt} &= \frac{L_m}{\tau_r} i_{ds} - \frac{1}{\tau_r}\psi_{dr} \\
\frac{di_{ds}}{dt} &= -a i_{ds} + \omega_s i_{qs} + \frac{L_m}{\sigma L_s L_r \tau_r}\psi_{dr} + \frac{v_{ds}}{\sigma L_s} \\
\frac{di_{qs}}{dt} &= -a i_{qs} - \omega_s i_{ds} - \frac{L_m Pp}{\sigma L_s L_r}\omega_m \psi_{dr} + \frac{v_{qs}}{\sigma L_s}
\end{aligned}
\tag{3}
$$

With

$$\omega_s = p\omega_m + \frac{L_m}{\tau_r \psi_{dr}} i_{qs} \quad (4)$$

where $\tau_r = \frac{L_r}{R_r}$ is the rotor time constant, $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ is the coefficient of dispersion, and $a = \left(\frac{R_s}{\sigma L_s} + \frac{1-\sigma}{\sigma \tau_r} \right)$ is a constant.

2.2. Faulty Model

In this study, two classes of faults in the IM model are considered (Vas, 1994; Bonivento et al., 2004):

1- Rotor asymmetries, mainly due to broken rotor bars or dynamic eccentricity. Broken rotor bars lead to the rotor resistance variation. This variation can be modeled by substituting $R_r + \Delta R_r$ instead of R_r in the IM dynamic model (3).

2- Stator asymmetries, mainly due to static eccentricity.

From [6], it turns out that in the presence of static and dynamic eccentricities, faults generates asymmetries in the induction motor, yielding some slot harmonics in the stator winding. In the (d− q) reference frame, it is possible to model this effect by adding a sinusoidal corruption term to the stator currents values. Specifically, letting $i_d^{uf}(t)$ and $i_q^{uf}(t)$ denote the values of the stator current in the absence of faults and $i_d^f(t)$ and $i_q^f(t)$ are corresponding values in the presence of faults, the current can be expressed in the form

$$\begin{aligned} i_{sd}^f(t) &= i_{sd}^{uf} + A\sin(\omega_c(t) + \theta_s(t) + \phi), \\ i_{sq}^f(t) &= i_{sq}^{uf} + A\cos(\omega_c(t) + \theta_s(t) + \phi) \end{aligned} \quad (5)$$

Where the amplitude A and the phases ϕ are unknown, they depend on the stator or rotor faults entity.

Supposing that faults only occur when induction motor is working in a steady operation, then $\omega_c(t) + \theta_s(t)$ the faults concerning static eccentricity are concerned,

$$\omega_c(t) + \theta_s(t) = 2\pi f t + (p\omega_{m,ref} + \omega_{sl})t + \theta_{s0} \quad (6)$$

The faults concerning dynamic eccentricity are concerned,

$$\omega_c(t) + \theta_s(t) = 2(\pi f \pm 2k\omega_{sl})t + (p\omega_{m,ref} + \omega_{sl})t + \theta_{s0} \quad (7)$$

where $\omega_{sl} = \omega_s - p\omega_{m,ref}$ is the slip angular frequency, $\omega_{m,ref}$ is the rotor speed reference, f is the supply frequency, θ_{s0} is the unknown position of the reference frame once the fault occurs, $k = 1,....N$ is the finite integer.

From the above mentioned faults, the IM faulty model can be described as

$$\begin{aligned} \frac{d\omega_m}{dt} &= \frac{pL_m}{L_r J} i_{qs}\psi_{dr} - \frac{F}{J} + h_1(t) \\ \frac{d\psi_{dr}}{dt} &= \frac{L_m}{\tau_r} i_{ds} - \frac{1}{\tau_r}\psi_{dr} + h_2(t) \\ \frac{di_{ds}}{dt} &= -ai_{ds} + \omega_s i_{qs} + \frac{L_m}{\sigma L_s L_r \tau_r}\psi_{dr} + \frac{v_{ds}}{\sigma L_s} + h_3(t) \\ \frac{di_{qs}}{dt} &= -ai_{qs} - \omega_s i_{ds} - \frac{L_m Pp}{\sigma L_s L_r}\omega_m \psi_{dr} + \frac{v_{qs}}{\sigma L_s} + h_4(t) \end{aligned} \quad (8)$$

Where $h_1(t) = -\frac{T_L}{J} + h_1(\Delta i_{sdq})$, $h_2(t) = h_2(\Delta i_{sdq}, \Delta R_r)$, $h_3(t) = h_3(\Delta i_{sdq}, \Delta R_r)$ and $h_4(t) = h_4(\Delta i_{sdq}, \Delta R_r)$ represent the fault and uncertainty terms due to rotor resistance variation, harmonics in the stator winding ($\Delta i_{sdq} = i_{sdq}^f - i_{sdq}^{uf}$) and load torque variation. It is worthy to notify that the load torque T_L is added with the fault terms due to its typically uncertain nature.

The control objective is to design sliding mode controllers for governing the rotor mechanical speed ω_m and the rotor flux $\psi_r = [\psi_{dr}, \psi_{qr}]^T$ in the presence of the faults and uncertainties. By taking a glance in the IM faulty model (8), it is evident that the model is not in the companion form. Therefore, SMC technique is not applicable directly. To solve this problem, the NBC transformation technique will be developed in the next section.

3. NBC TRANSFORMATION

In this section, two types of NBC transformation techniques are expressed. The first one is the conventional NBC technique which is presented in (Loukianov *et al.,* 2002; Loukianov, 2002). But because of non-vanishing uncertainty problem and high computational characteristic, we proposed the second and new method. More details are presented in below:

To make the IM faulty notations (8) less complex, consider the following new variables:

$$x_{11} = \omega_m, \quad x_{12} = \psi_{dr}, \quad x_{21} = i_{ds}, \quad x_{22} = i_{qs} \quad (9)$$

Hence, the motor dynamics can be written as

$$\begin{aligned} \frac{d}{dt}x_{11} &= \frac{pL_m}{L_r J} x_{12}x_{22} - \frac{F}{J}x_{11} + h_1(t) \\ \frac{d}{dt}x_{12} &= \frac{L_m}{\tau_r}x_{21} - \frac{1}{\tau_r}x_{12} + h_2(t) \\ \frac{d}{dt}x_{21} &= -ax_{21} + \omega_s x_{22} + \frac{L_m}{\sigma L_s L_r \tau_r}x_{12} + \frac{1}{\sigma L_s}v_{ds} + h_3(t) \\ \frac{d}{dt}x_{22} &= -ax_{22} - \omega_s x_{21} - \frac{L_m Pp}{\sigma L_s L_r}x_{11}x_{12} + \frac{1}{\sigma L_s}v_{qs} + h_4(t) \end{aligned} \quad (10)$$

The system model (10), can be represented in the following NBC-form as

$$\dot{X}_1 = f_1(X_1) + B_1(X_1)X_2 + H_1(t)$$
$$\dot{X}_2 = f_2(X_1, X_2) + B_2 u(t) + H_2(t) \tag{11}$$

where

$$X_1 = \begin{bmatrix} x_{11} \\ x_{12} \end{bmatrix}, \quad f_1(X_1) = \begin{bmatrix} -\frac{F}{J}x_{11} \\ -\frac{1}{\tau_r}x_{12} \end{bmatrix}$$

$$X_2 = \begin{bmatrix} x_{21} \\ x_{22} \end{bmatrix}, \quad f_2(X_1, X_2) = \begin{bmatrix} -ax_{21} + \omega_s x_{22} + \frac{L_m}{\sigma L_s L_r \tau_r}x_{12} \\ -ax_{22} - \omega_s x_{21} - \frac{L_m p}{\sigma L_s L_r}x_{11}x_{12} \end{bmatrix}$$

$$B_1(X_1) = \begin{bmatrix} 0 & \frac{pL_m}{L_r}x_{12} \\ \frac{L_m}{\tau_r} & 0 \end{bmatrix}, \quad B_2 = \begin{bmatrix} \frac{1}{\sigma L_s} & 0 \\ 0 & \frac{1}{\sigma L_s} \end{bmatrix}$$

$$u(t) = \begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix}, \quad H_1(t) = \begin{bmatrix} h_1(t) \\ h_2(t) \end{bmatrix}, \quad H_2 = \begin{bmatrix} h_3(t) \\ h_4(t) \end{bmatrix}$$

The NBC-form (11) is composed of two blocks, where X_2 is the fictitious control for the first block, and $H_1(t)$ and $H_2(t)$ are the unmatched and matched fault vectors, respectively.

Assumption 1: Assume that the fault vectors be bounded as

$$\|H_1(t)\| < H_{1\max} \tag{12}$$
$$\|H_2(t)\| < H_{2\max} \tag{13}$$

where $H_{1\max}$ and $H_{2\max}$ are known positive constants.

3.1. Conventional NBC

Now, let us to define the following tracking errors

$$Z_1 = \begin{bmatrix} z_{11} \\ z_{12} \end{bmatrix} = X_{1,ref} - X_1 = \begin{bmatrix} x_{11,ref} - x_{11} \\ x_{12,ref} - x_{12} \end{bmatrix} = \begin{bmatrix} \omega_{m,ref} - \omega_m \\ \psi_{dr,ref} - \psi_{dr} \end{bmatrix} \tag{14}$$

where $X_{1,ref}$ is the desired value of X_1.

The time derivative of (14) is

$$\dot{Z}_1 = \dot{X}_{1,ref} - \dot{X}_1 \tag{15}$$

Substituting the first block of (11) in (15), yields

$$\dot{Z}_1 = -f_1(X_1) + \dot{X}_{1,ref} - B_1(X_1)X_2 - H_1(t) \tag{16}$$

Let the fictitious control input X_2 for (16) be chosen as

$$X_2 = B_1^{-1}(X_1)[-f_1(X_1) + \dot{X}_{1,ref} + K_1 Z_1 - Z_2] \tag{17}$$

where Z_2 is a new variable, and $K_1 = \begin{bmatrix} k_{11} & 0 \\ 0 & k_{12} \end{bmatrix}$, $k_{11} > 0$, $k_{12} > 0$ is the control gain matrix.

Inserting the virtual control law (17) in (16), results in the first transformation block in the new coordinates

$$\dot{Z}_1 = -K_1 Z_1 + Z_2 - H_1(t) \tag{18}$$

From (17), Z_2 can be derived as follows

$$Z_2 = f_1(X_1) - \dot{X}_{1,ref} - K_1 Z_1 + B_1(X_1)X_2 \tag{19}$$

The time derivative of (19) is

$$\dot{Z}_2 = \frac{d}{dt}(f_1(X_1)) - \ddot{X}_{1,ref} - K_1\dot{Z}_1 + \frac{d}{dt}(B_1(X_1))X_2 + B_1(X_1)\dot{X}_2 \tag{20}$$

The above equation can be represent as

$$\dot{Z}_2 = F(Z_1, Z_2, X_1, X_{1,ref}, X_2) + B(X_1)u(t) + H(t) \tag{21}$$

with the following details

$$F(Z_1, Z_2, X_1, X_{1,ref}, X_2) = \begin{bmatrix} -\frac{F}{J} & 0 \\ 0 & -\frac{1}{\tau_r} \end{bmatrix}(K_1 Z_1 - Z_2 + \dot{X}_{1,ref}) - \ddot{X}_{1,ref} - K_1(-K_1 Z_1 + Z_2) + \begin{bmatrix} 0 & \frac{pL_m}{L_r J}(k_{12}z_{12} - z_{22} + \dot{x}_{12,ref}) \\ 0 & 0 \end{bmatrix}X_2 + B_1(X_1)f_2(X_1, X_2)$$

$$B(X_1) = B_1(X_1)B_2$$

$$H(t) = \left(K_1 + \begin{bmatrix} -\frac{F}{J} & 0 \\ 0 & -\frac{1}{\tau_r} \end{bmatrix}\right)H_1(t) + B_1(X_1)H_2(t) + \begin{bmatrix} \frac{pL_m}{L_r J} \\ 0 \end{bmatrix}h_{12}$$

From (18) and (21), the IM first transformed model can be expressed in the new coordinates (Z_1, Z_2) as follows

$$\dot{Z}_1 = -K_1 Z_1 + Z_2 - H_1(t)$$
$$\dot{Z}_2 = F(Z_1, Z_2, X_1, X_{1,ref}, X_2) + B(Z_1, X_{1,ref})u(t) + H(t) \tag{22}$$

Remark 1: The above transformed model contains three main drawbacks:

1- Deriving the second block dynamic is troublesome due to the high derivational computation (20). Also, for high dimensional systems this problem will be a real challenge, and needs to be solved.

2- In designing controllers, low tracking error ($Z_1 \to 0$) is

hard to achieve due to the non-vanishing uncertainties like T_L in the unmatched vector $H_1(t)$. To avoid this problem, the load torque T_L should be known by measuring or approximating it.
3- There is no reference value for X_2. In the other word, governing the virtual state X_2 is not considered so important, while this inconsideration cases some problems in practical applications.
The next transformation technique is able to improve the mentioned problems.

3.2. Improved NBC

Consider the following tracking errors

$$
\begin{aligned}
Z_1 &= X_{1,ref} - X_1 = \begin{bmatrix} z_{11} \\ z_{12} \end{bmatrix} = \begin{bmatrix} x_{11,ref} - x_{11} \\ x_{12,ref} - x_{12} \end{bmatrix} = \begin{bmatrix} \omega_{m,ref} - \omega_m \\ \psi_{dr,ref} - \psi_{dr} \end{bmatrix} \\
Z_2 &= X_{2,ref} - X_2 = \begin{bmatrix} z_{21} \\ z_{22} \end{bmatrix} = \begin{bmatrix} x_{21,ref} - x_{21} \\ x_{22,ref} - x_{22} \end{bmatrix} = \begin{bmatrix} i_{ds,ref} - i_{ds} \\ i_{qs,ref} - i_{qs} \end{bmatrix}
\end{aligned} \tag{23}
$$

Taking time derivative of Z_1 and substituting the first block of (11) in it, yields

$$
\begin{aligned}
\dot{Z}_1 &= \dot{X}_{1,ref} - \dot{X}_1 \\
&= \dot{X}_{1,ref} - f_1(X_1) - B_1(X_1)(X_{2,ref} - Z_2) - H_1(t) \\
&= -f_1(X_1) + \dot{X}_{1,ref} - B_1(X_1)X_{2,ref} + B_1(X_1)Z_2 - H_1(t)
\end{aligned} \tag{24}
$$

In (24), we replaced X_2 by $X_{2,ref} - Z_2$. Hence, $X_{2,ref}$ will be the fictitious control input instead of X_2. This replacement is the main reason for the derivational calculation reduction.

Now, let propose the fictitious control input $X_{2,ref}$ for (24) as

$$
X_{2,ref} = B_1^{-1}(X_1)\left(-f_1(X_1) + \dot{X}_{1,ref} + K_1 Z_1 + K_{sw1}\tanh(Z_1/\rho_1)\right) \tag{25}
$$

where $\tanh(\cdot)$ is the hyperbolic tangent function and is the approximation of $\mathrm{sgn}(\cdot)$, ρ_1 is an enough small positive constant, and $K_{sw1} = \begin{bmatrix} k_{sw11} & 0 \\ 0 & k_{sw12} \end{bmatrix}$ is the sliding gain and should be selected in such a way that $\|K_{sw1}\| > H_{1\max}$.
By substituting the virtual control law (25) in (24), we have

$$
\dot{Z}_1 = -K_1 Z_1 + B_1(X_1)Z_2 - K_{sw1}\tanh(Z_1/\rho_1) - H_1(t) \tag{26}
$$

Moreover, calculating time derivative of Z_2 and inserting the second block of (11) into it, yields

$$
\dot{Z}_2 = \dot{X}_{2,ref} - \dot{X}_2 = -f_2(X_1, X_2) + \dot{X}_{2,ref} - B_2 u(t) - H_2(t) \tag{27}
$$

From (26) and (27), the IM second transformation model can be expressed in the new coordinates (Z_1, Z_2) as follows

$$
\begin{aligned}
\dot{Z}_1 &= -K_1 Z_1 + B_1(X_1)Z_2 - K_{sw1}\tanh(Z_1/\rho_1) - H_1(t) \\
\dot{Z}_2 &= -f_2(X_1, X_2) + \dot{X}_{2,ref} - B_2 u(t) - H_2(t)
\end{aligned} \tag{28}
$$

It is worthy to notify that compared to the conventional NBC:
1- The second block dynamic (27) is determined easily with no more derivational calculations.
2- The non-vanishing unmatched fault and uncertainty effects can be attenuated by the robust term $K_{sw1}\tanh(Z_1/\rho_1)$.
3- Converging $Z_2 \to 0$, will result $X_2 \to X_{2,ref}$.

4. PASSIVE FAULT TOLERANT SLIDING MODE CONTROL

In this section, two passive fault tolerant sliding mode controllers are designed for the conventional and the improved NBC transformed models.

4.1. SMC with the conventional NBC

Let us define the following sliding manifold

$$
s(t) = Z_2(t) = 0 \tag{29}
$$

By taking time derivative from (29) and using (22), we can get

$$
\dot{s}(t) = F(Z_1, Z_2, X_1, X_{1,ref}, X_2) + B(X_1)u(t) + H(t) \tag{30}
$$

Theorem 1: Consider the IM first transformed NBC model (22) with the linear sliding surface (29) and Assumption 1, then the following control law

$$
\begin{aligned}
u(t) = B^{-1}(X_1)\big[&-F(Z_1, Z_2, X_1, X_{1,ref}, X_2) - K_2 Z_2 \\
&- K_{sw2}\,\mathrm{sgn}(s(t))\big]
\end{aligned} \tag{31}
$$

guarantees zero convergence of Z_2, and boundedness of Z_1 around zero. Hence, the IM speed and flux tracking errors convergence around the origin will be assured. Where $K_2 = \begin{bmatrix} k_{21} & 0 \\ 0 & k_{22} \end{bmatrix}$, $k_{21} > 0$, $k_{22} > 0$, $K_{sw2} = \begin{bmatrix} k_{sw21} & 0 \\ 0 & k_{sw22} \end{bmatrix}$, and $K_{sw} > H_{\max} > \|H(t)\|$.

Stability proof: In order to assure the closed-loop system stability, consider the following Lyapunov candidate

$$
V = \frac{1}{2} s_2^T s_2 \tag{32}
$$

Taking time derivative from (32) and using (30) and (31), results in

$$
\begin{aligned}
\dot{V} &= s_2^T \dot{s}_2 = Z_2^T \dot{Z}_2 \\
&= Z_2^T (F(Z_1, Z_2, X_1, X_{1,ref}, X_2) + B(X_1)u(t) + H(t)) \\
&= Z_2^T (-K_2 Z_2 - K_{sw2}\,\mathrm{sgn}(Z_2) + H(t)) \\
&= -K_2 Z_2^T Z_2 + Z_2^T\left(-K_{sw2}\,\mathrm{sgn}(Z_2) + H(t)\right)
\end{aligned} \tag{33}
$$

Selecting $K_{sw} > H_{\max}$, yields

$$\dot{V} = -K_2 Z_2^T Z_2 < 0 \tag{34}$$

which fulfills $Z_2 \to 0$. Then, the first transformation model (22) can be reduced as

$$\dot{Z}_1 = -K_1 Z_1 + 0 - H_1(t) \tag{35}$$

From (35), it is obvious that Z_1 will converge to zero for vanishing uncertainties, while $H_1(t)$ have the non-vanishing term T_L in $h_1(t)$. Therefore, the flux convergence is possible ($\psi_{dr} \to \psi_{dr,ref}$), but the mechanical speed will have deviations from the reference speed ($\omega_{m,ref} - \omega_m \to \varepsilon \neq 0$).

It is worthy to note that choosing large values for K_1, can reduce the speed tracking error. But it is not able to remove the error permanently.

4.2. NSMC with the improved NBC

In this part, the NSMC is applied for the second transformed NBC model (28). In the other word, a nonlinear sliding surface is proposed to improve the conventional SMC slow responses.

Consider the following nonlinear sliding manifold

$$s(t) = Z_2(t) + c\int_0^t Z_2^{\mu}(\tau)d\tau \tag{36}$$

where $s(t) = [s_1(t), \ \ s_2(t)]^T$, $0 < \mu = \dfrac{p}{q} < 1$, and p,q are rational odd numbers, also c is a positive constant.

Taking time derivative from (36), yields

$$\dot{s}(t) = \dot{Z}_2 + cZ_2^{\mu} \tag{37}$$

Inserting (28) in (37), results in

$$\dot{s}(t) = -f_2(X_1, X_2) + \dot{X}_{2,ref} + cZ_2^{\mu} - B_2 u(t) - H_2(t) \tag{38}$$

Theorem 2: Consider the IM second transformed NBC model (28) with the nonlinear sliding surface (36) and Assumption 1, then the following control law

$$\begin{aligned} u(t) = B_2^{-1}\big(&- f_2(X_1, X_2) + \dot{X}_{2,ref} + cZ_2^{\mu} + K_2 s(t) \\ &+ K_{sw2}\,\mathrm{sgn}(s(t))\big) \end{aligned} \tag{39}$$

guarantees Z_1 and Z_2 zero convergence. Hence, the IM speed and flux tracking errors convergence is assured. Where $K_2 = \begin{bmatrix} k_{21} & 0 \\ 0 & k_{22} \end{bmatrix}$, and $\|K_{sw2}\| > H_{2\max}$.

Stability analysis: Let us define the following Lyapunov function

$$V(t) = 0.5 s^T(t) s(t) \tag{40}$$

Taking time derivative from (40) and using (38) and (39), we can get

$$\begin{aligned} \dot{V}(t) &= s^T(t)\dot{s}(t) \\ \dot{V}(t) &= s^T(t)\big(- f_2(X_1, X_2) + \dot{X}_{2,ref} + cZ_2^{\mu} - B_2 u(t) - H_2(t)\big) \\ \dot{V}(t) &= s^T(t)\big(- K_2 s(t) - K_{sw2}\,\mathrm{sgn}(s(t)) - H_2(t)\big) \\ \dot{V}(t) &= -s^T(t) K_2 s(t) + s^T(t)\big(- K_{sw2}\,\mathrm{sgn}(s(t)) - H_2(t)\big) \end{aligned} \tag{41}$$

Selecting $\|K_{sw2}\| > H_{2\max}$, yields

$$\dot{V}(t) = -s^T(t) K_2 s(t) < 0 \tag{42}$$

which guarantees the second block state vector zero convergence ($Z_2 \to 0$). Then, the equation (28) can be rewritten as

$$\dot{Z}_1 = -K_1 Z_1 + 0 - K_{sw1}\tanh(Z_1/\rho_1) - H_1(t) \tag{43}$$

To analyze (43) stability, consider the following Lyapunov function

$$V_1(t) = 0.5 Z_1^T Z_1 \tag{44}$$

Calculating the time derivative of (44) and substituting (43) in it, results in

$$\begin{aligned} \dot{V}_1(t) = Z_1^T \dot{Z}_1 &= Z_1^T\big(-K_1 Z_1 - K_{sw1}\tanh(Z_1/\rho_1) - H_1(t)\big) \\ &= -Z_1^T K_1 Z_1 + Z_1^T\big(-K_{sw1}\tanh(Z_1/\rho_1) - H_1(t)\big) \end{aligned} \tag{45}$$

By selecting the sliding gain $\|K_{sw1}\| > H_{1\max}$, we have

$$\dot{V}_1(t) = -Z_1^T K_1 Z_1 < 0 \tag{46}$$

which assures $Z_1 \to 0$. The inequality (46), is the main advantage of the second transformation form (28) and the control law (39) against the first transformation form (22) with the control law (31).

5. SIMULATION RESULTS

To testify the performance of the sliding mode block controllers, series of MATLAB/SIMULINK simulation tests for a 4 kW cage rotor induction machine with 50 Hz and 220 V power supply is carried out. The nominal electrical and mechanical parameters of the IM are shown in Table 1. Also, the induction motor control block diagram based on FOC technique is illustrated in Fig. 1. This figure contains rotor flux estimation block which is applied to avoid the use of flux sensor. The flux estimation block dynamic is selected as:

$$\hat{\psi}_{dr}(s) = \frac{L_m}{1 + \tau_r s} i_{ds}(s)$$

Comparative simulations of the discussed PFTC controllers (31) and (39) (SMC with the conventional NBC vs. NSMC with the improved NBC) are presented in Figs. 2-7. The IM speed response, electromagnetic torque, rotor flux, stator currents, speed and flux tracking errors and new transformation variables are showed in these figures.

Table. 1 Nominal parameters of the IM adopted for the simulation (Benbouzid et al., 2008).

Description	Parameter	Value	Units
stator resistance	R_s	1.2	Ω
rotor resistance	R_r	1.8	Ω
stator inductance	L_s	0.1554	H
rotor resistance	L_r	0.1566	H
mutual inductance	L_m	0.15	H
rotor inertia	J	0.024	$Kg.m^2$
friction coefficient	F	0.011	$N.m.s/rad$
number of pole pairs	p	2	-
rated speed	ω_{mN}	1440	rpm

The simulation test involves the following operating sequences:

1- To test the robustness of the proposed controllers with respect to rotor faults, we have considered the responses of the IM in the presence of rotor resistance variation $+100\%R_r$, this resistance variation is taken into account at $t=2s$.

2- The motor shaft is subject to a step load torque variation of $4Nm$ at $t=3s$ which is considered unknown.

3- To investigate the proposed controller's performance under stator faults, simulations are performed in the presence of static eccentricity. To simulate stator fault, the stator currents have been destroyed by assuming that $A=0.5$, $\phi=0$ and $\omega_c(t)+\theta_s(t)$ is computed using (6). The time of stator fault occupancy is considered at $t=9s$.

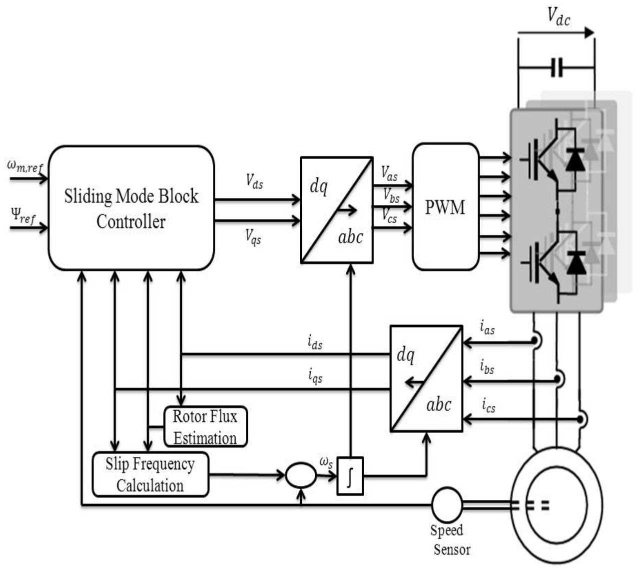
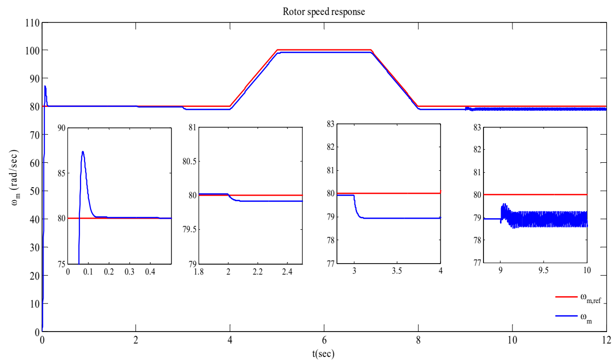
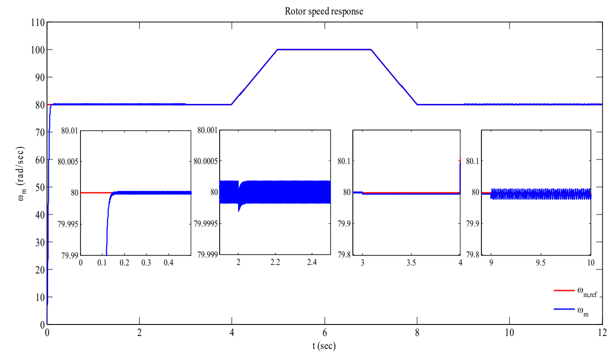


Fig. 1. Block diagram of the designed fault-tolerant controllers for IM.

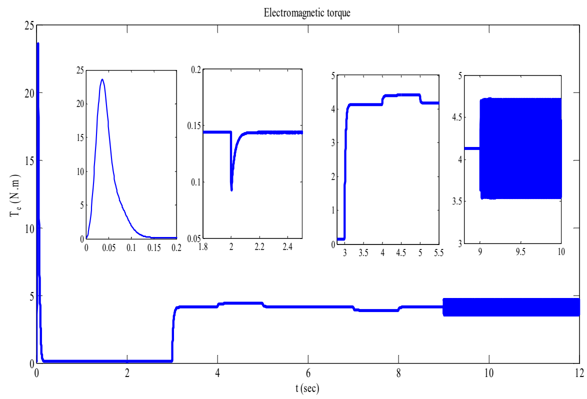


a) SMC with the conventional NBC.

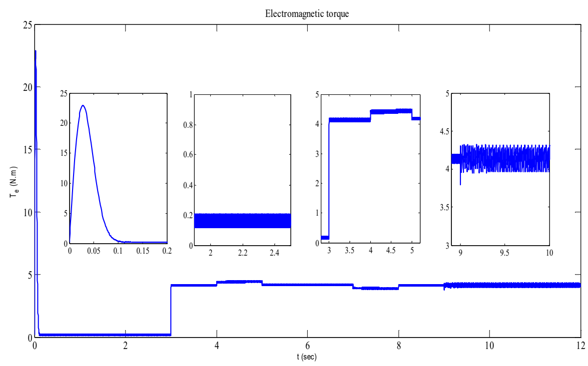


b) NSMC with the improved NBC.

Fig. 2. IM speed response.

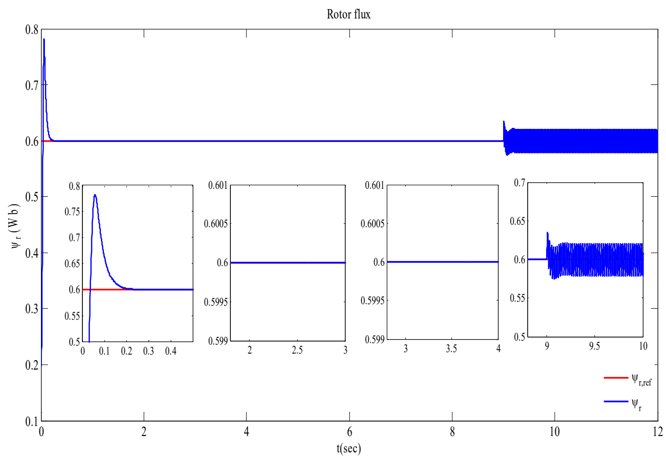


a) SMC with the conventional NBC.

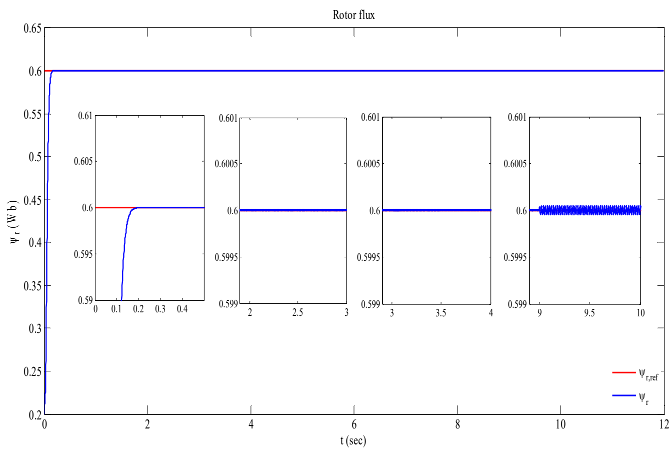


b) NSMC with the improved NBC.

Fig. 3. Electromagnetic torque.

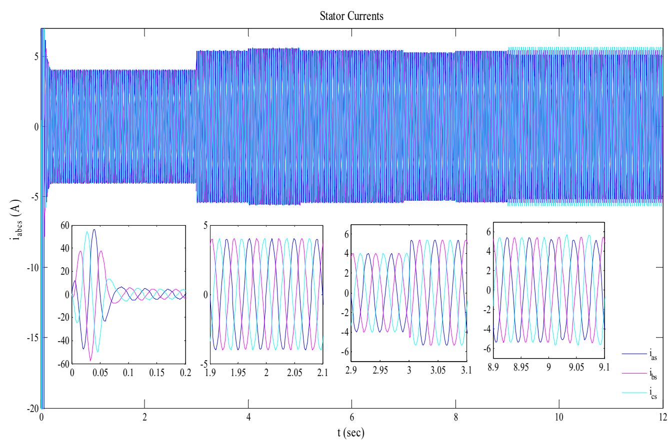


a) SMC with the conventional NBC.

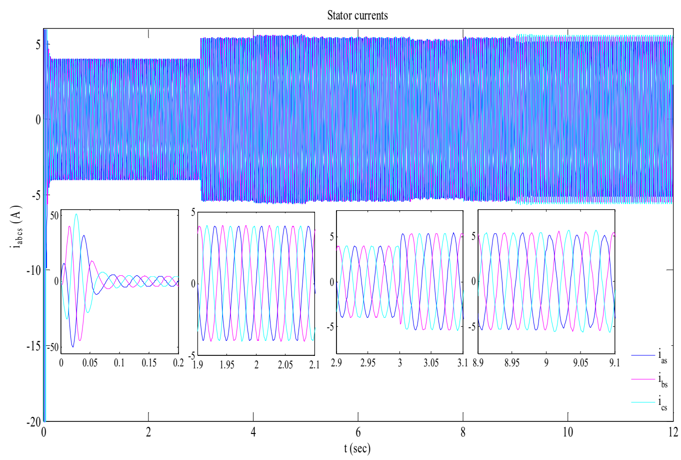


b) NSMC with the improved NBC.

Fig. 4. Rotor flux.

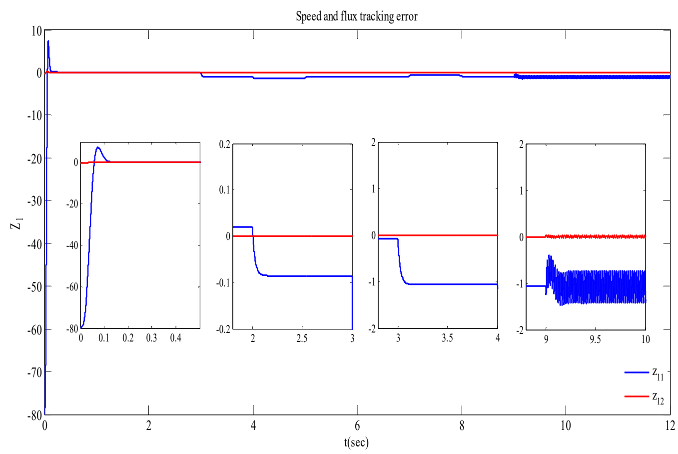


a) SMC with the conventional NBC.

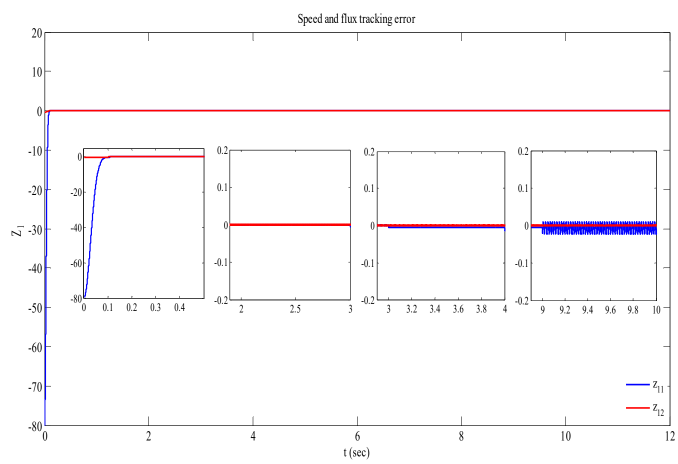


b) NSMC with the improved NBC.

Fig. 5. Stator currents.

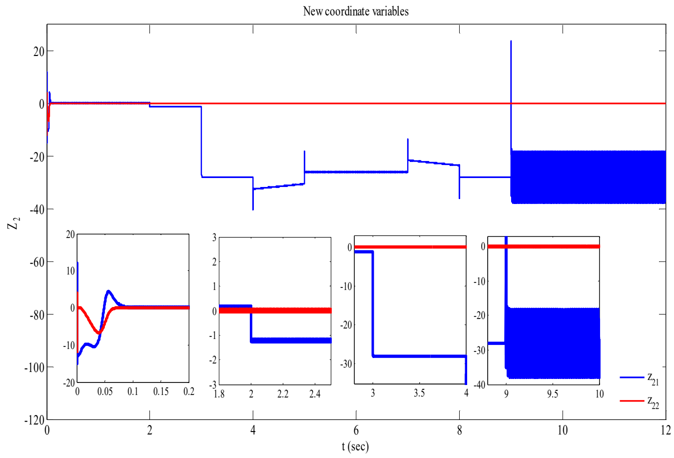


a) SMC with the conventional NBC.

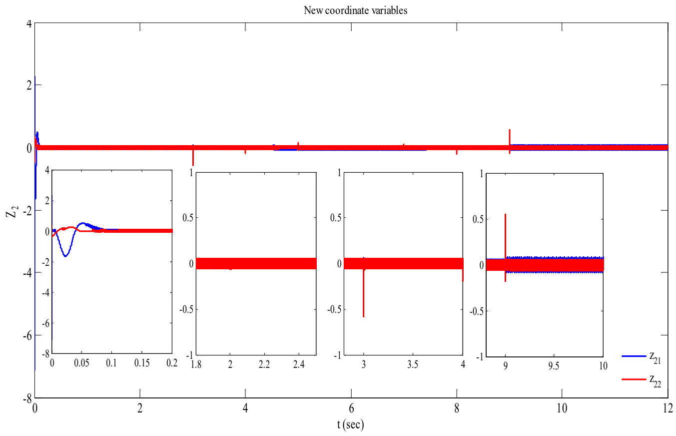


b) NSMC with the improved NBC.

Fig. 6. Speed and flux tracking errors.



a) SMC with the conventional NBC.



b) NSMC with the improved NBC.

Fig 7. New coordinate variables.

From Figs. 2-7, in is evident that the NSMC with the improved NBC technique has low tracking error and low variation against different faults in comparison with the SMC with the conventional NBC approach. İn addition, the unmatched fault effects are removed effectively. Also, NSMC with the improved NBC method chattering is less than the SMC with the conventional NBC method.

6. CONCLUSIONS

In this study, passive fault tolerant sliding mode control of three phase induction motors is investigated. First, the SMC with the conventional NBC technique is designed. But the conventional NBC technique has some flaws like unmatched fault/uncertainty rejection problem. İn addition, the SMC controller shows a slow response with considerable tracking error. Then in the second step, to solve the mentioned problems, the NSMC with the improved NBC technique has been proposed. Comparative simulation results revealed the effectiveness of the proposed method for multiple rotor and stator faults.

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