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Vector Control of Faulty Three-Phase Induction Motor with an Adaptive Sliding Mode Control

***Abstract**. This paper presents an adaptive sliding mode control with an adaptive switching gain in order to vector control of Three-Phase Induction Motor (TPIM) based on Rotor Flux Oriented Control (RFOC) method under open-phase fault (faulty TPIM). This method can be utilized to control of IMs in some critical applications which require fault-tolerant scheme. To confirm the good performance of the proposed method, simulation results have been presented. The Simulation results confirm a completely satisfactory effectiveness of the proposed method.*

*Streszczenie. W artykule zaprezentowano adaptacyjne sterowanie ślizgowe trójfazowego silnika indukcyjnego bazujące na metodzie RFOC w przypadku gdy jedna z faz wykazuje błąd. (**Adaptacyjne ślizgowe Sterowanie wektorowe trójfazowego silnika indukcyjnego z błędem fazy**)*

Keywords: adaptive sliding mode control, faulty three-phase induction motor, rotor flux oriented control.
Słowa kluczowe: sterowanie ślizgowe, sterowanie adaptacyjne, silnik indukcyjny trójfazowy.

Introduction

Induction Motor (IM) drives is commonly employed in industrial applications such as Heating, Ventilation and Air Conditioning (HVAC), compressors, fans and etc. In the resent years, Field Oriented Control (FOC) is one of the most common approaches to control of IMs. Open-phase fault is one of usual failures in the IM stator windings [1, 2]. This fault is caused by blown fuse, mechanical shakings of the machine and etc. Several techniques have been proposed for detection of the fault in electrical motors [3, 4]. A fast open-phase fault diagnosis can be achieved by a technique based on high-frequency, low-amplitude signals injection [4]. Therefore for further considerations in this paper it is assumed that the fault detection procedure is immediate. Fault-tolerant operations during open-circuit faults of the IM are significant, in particular in some critical applications, such as railway or space technology. The model structure of the faulty TPIM is similar to balanced TPIM model as reported in [5]. To obtain fault-tolerant control scheme of the IM the classical FOC method should be modified. In this paper a new control technique for a faulty TPIM based on FOC is presented. In the proposed FOC strategy for faulty TPIM, based on equivalent circuit of Single-Phase Induction Motor (SPIM), unbalanced rotational transformations are utilized. It is proved by using of these rotational transformations the equations of the unbalanced TPIM (faulty TPIM) become like the balanced TPIM equations. As a result, by using proposed rotational transformations a new FOC system developed for TPIM under open-phase fault from the conventional vector control. It is shown in the presented control system for faulty TPIM control, the speed PI controller has to be substituted by an adaptive sliding mode controller. The performance of the classical FOC structure (with PI controllers) for both balanced and unbalanced TPIM (such as [1, 5]) strongly depends on uncertainties, which are usually due to unknown parameters, external load disturbances, parameter variations, nonlinear dynamics and etc [6]. Many studies have been presented to protect the performance of the drive system under external load disturbances and parameter variations (such as: genetic algorithm [7], fuzzy control [8], neural control [9] and adaptive control [10]). Because of fast dynamic response, insensitivity to variations of parameter, good performance against un-modelled and nonlinear dynamics, rejection of external load disturbance and etc, the sliding mode control can be used in the speed control of electric drives [11]. In this paper, to compensate the system uncertainties an adaptive sliding mode control with on-line sliding gain estimation is proposed. The adaptive sliding mode control stability has been also proved thought Lyapunov stability theory.

Model of faulty TPIM

The equations of faulty TPIM in the stationary reference frame (superscript "*s*") can be shown as follows [5]:

Stator and rotor voltage equations:

$$(1)\quad \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} r_s + L_{ds}\frac{d}{dt} & 0 & M_d\frac{d}{dt} & 0 \\ 0 & r_s + L_{qs}\frac{d}{dt} & 0 & M_q\frac{d}{dt} \\ M_d\frac{d}{dt} & \omega_r M_q & r_r + L_r\frac{d}{dt} & \omega_r L_r \\ -\omega_r M_d & M_q\frac{d}{dt} & -\omega_r L_r & r_r + L_r\frac{d}{dt} \end{bmatrix} \begin{bmatrix} i_{ds}^s \\ i_{qs}^s \\ i_{dr}^s \\ i_{qr}^s \end{bmatrix}$$

Stator and rotor flux equations:

$$(2)\quad \begin{bmatrix} \lambda_{ds}^s \\ \lambda_{qs}^s \\ \lambda_{dr}^s \\ \lambda_{qr}^s \end{bmatrix} = \begin{bmatrix} L_{ds} & 0 & M_d & 0 \\ 0 & L_{qs} & 0 & M_q \\ M_d & 0 & L_r & 0 \\ 0 & M_q & 0 & L_r \end{bmatrix} \begin{bmatrix} i_{ds}^s \\ i_{qs}^s \\ i_{dr}^s \\ i_{qr}^s \end{bmatrix}$$

Electromagnetic torque equation:

$$(3)\quad \tau_e = \frac{Pole}{2}(M_q i_{qs}^s i_{dr}^s - M_d i_{ds}^s i_{qr}^s)$$

$$\frac{Pole}{2}(\tau_e - \tau_l) = J\frac{d\omega_r}{dt} + F\omega_r$$

where,

$$L_{ds} = L_{ls} + L_{md}, L_{qs} = L_{ls} + L_{mq}, L_{md} = \frac{3}{2}L_{ms}$$

$$L_{mq} = \frac{1}{2}L_{ms}, M_d = \frac{3}{2}L_{ms}, M_q = \frac{\sqrt{3}}{2}L_{ms}$$

Moreover, v^s_{ds}, v^s_{qs}, i^s_{ds}, i^s_{qs}, i^s_{dr}, i^s_{qr}, λ^s_{ds}, λ^s_{qs}, λ^s_{dr} and λ^s_{qr} are the d-q axes voltages, currents, and fluxes of the stator and rotor in the stator reference frame. r_s and r_r denote the stator and rotor resistances. L_{ds}, L_{qs}, L_r, M_d and M_q denote the stator, and the rotor self and mutual inductances. ω_r is the machine speed. τ_e, τ_l, J and F are electromagnetic torque, load torque, inertia and viscous friction coefficient. It is shown in [5], in the faulty mode, the normalized transformation matrix (normalized transformation matrix is the matrix for transformation of variables from a-b-c to d-q frame) for stator variables is as follows:

(4) $$\left[T_s^{fault}\right] = \frac{\sqrt{2}}{2}\begin{bmatrix} 1 & -1 \\ 1 & 1 \end{bmatrix}$$

In the rotor flux oriented control method, the equation of the machine is transferred to the rotor flux oriented reference frame. For this purpose, rotational transformation (conventional rotational transformation) must be applied to the IM equations as follows:

(5) $$\left[T_s^e\right] = \begin{bmatrix} \cos\theta_e & \sin\theta_e \\ -\sin\theta_e & \cos\theta_e \end{bmatrix}$$

In equation (5), θ_e means the angle between the stationary reference frame and the rotating reference frame. By employing of conventional rotational transformation, the faulty TPIM equations are obtained as forward and backward components (The backward components are generated because of unequal inductances; $M_d \neq M_q$ and $L_{ds} \neq L_{qs}$ in the faulty TPIM equations). In this research, it is attempted to introduce novel transformation matrixes to the faulty TPIM equations, which guarantee a balance of the system.

RFOC of faulty TPIM

The main idea of using transformation matrixes is gained from the steady-state equivalent circuit of the SPIM. Fig.1 shows the steady state equivalent circuit of SPIM.

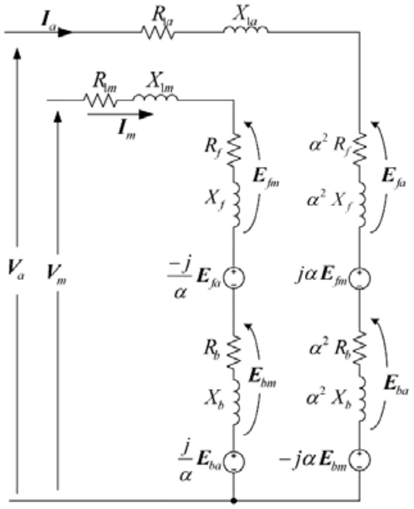

Fig.1. Steady state equivalent circuit of SPIM

In this figure, V_m, V_a, I_m and I_a are the main and auxiliary voltages and currents, "α" is the turn ratio ($\alpha = N_a / N_m$) and "j" is the square root of "-1". E_{fm}, E_{fa}, E_{bm} and E_{ba} are the forward and backward voltage of magnetizing branch of the main and auxiliary windings. R_f, R_b, X_f and X_b are the forward and backward stator resistance and inductance in main winding. R_{lm}, R_{la}, X_{lm} and X_{la} are the leakage resistance and inductance of the main and auxiliary winding. Based on Fig.1, the motor main and auxiliary windings voltage can be written as follows:

(6) $$V_m = Z_{lm} I_m + E_{fm} - \frac{j}{\alpha} E_{fa} + E_{bm} + \frac{j}{\alpha} E_{ba}$$
$$V_a = Z_{la} I_a + E_{fa} + j\alpha E_{fm} + E_{ba} - j\alpha E_{bm}$$

where,

$$E_{fm} = Z_f I_m \quad , \quad E_{bm} = Z_b I_m$$
$$E_{fa} = \alpha^2 Z_f I_a \quad , \quad E_{ba} = \alpha^2 Z_b I_a$$
$$Z_f = R_f + jX_f \quad , \quad Z_b = R_b + jX_b$$
$$Z_{lm} = R_{lm} + jX_{lm} \quad , \quad Z_{la} = R_{la} + jX_{la}$$

By following substitutions:

(7) $$I_m = \frac{1}{2}(I_1 + I_2) \quad , \quad I_a = \frac{j}{2\alpha}(I_1 - I_2)$$

(8) $$V_1 = Z_3 V_m + jZ_4 V_a$$

Equivalent circuit of SPIM (Fig.1) can be simplified as Fig.2. In equations (7) and (8), Z_3 and Z_4 are the functions in terms of inductances. By selecting every value of Z_3 and Z_4, Fig.1, can be simplified as a balanced circuit.

(9) $$Z = (Z_3 + Z_4)\frac{Z_1(Z_{lm} + 2Z_f) + Z_2(Z_{lm} + 2Z_b)}{2Z_1}$$
$$Z_1 = \alpha Z_{lm} + jZ_{la} + 2\alpha Z_b(\alpha j + 1)$$
$$Z_2 = -\alpha Z_{lm} + jZ_{la} + 2\alpha Z_f(\alpha j - 1)$$

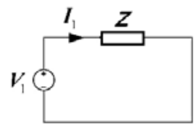

Fig.2. Simplified equivalent circuit of SPIM

As can be seen, by using equations (7) and (8), the equivalent circuit of SPIM changed into a balanced circuit. Equations (7) and (8) can be written as:

(10) $$\begin{bmatrix} jV_1 \\ V_1 \end{bmatrix} = \begin{bmatrix} -Z_4 & jZ_3 \\ -jZ_4 & Z_3 \end{bmatrix}\begin{bmatrix} V_a \\ V_m \end{bmatrix} \, , \, \begin{bmatrix} jI_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} \frac{N_a}{N_m} & j \\ -j\frac{N_a}{N_m} & 1 \end{bmatrix}\begin{bmatrix} I_a \\ I_m \end{bmatrix}$$

Equation (10) is the transformation matrixes for changing variables from unbalanced mode to the balanced mode. With following change of variables:

(11) $$j \to \sin\theta_e \, , 1 \to \cos\theta_e \, , \frac{N_m}{N_a} \to \frac{N_q}{N_d} = \frac{M_q}{M_d}$$
$$jV_1 \to v_{ds}^e \, , V_1 \to v_{qs}^e \, , \; V_a \to v_{ds}^s \; , \; V_m \to v_{qs}^s$$
$$jI_1 \to i_{ds}^e \; , I_1 \to i_{qs}^e \; , I_a \to i_{ds}^s \; , I_m \to i_{qs}^s$$

And with substation of equation (11) in equation (10), the proposed rotational transformations for stator voltage and current variables are obtained as following equations:

Rotational transformation for stator voltage variables:

(12) $$\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix} = \left[T_{vs}^e\right]\begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix} = \begin{bmatrix} -Z_4\cos\theta_e & Z_3\sin\theta_e \\ Z_4\sin\theta_e & Z_3\cos\theta_e \end{bmatrix}\begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix}$$

Rotational transformation for stator current variables:

(13) $$\begin{bmatrix} i_{ds}^e \\ i_{qs}^e \end{bmatrix} = \left[T_{is}^e\right]\begin{bmatrix} i_{ds}^s \\ i_{qs}^s \end{bmatrix} = \begin{bmatrix} \frac{M_d}{M_q}\cos\theta_e & \sin\theta_e \\ -\frac{M_d}{M_q}\sin\theta_e & \cos\theta_e \end{bmatrix}\begin{bmatrix} i_{ds}^s \\ i_{qs}^s \end{bmatrix}$$

It is expected by using presented rotational transformations (equations (12) and (13)), the equations of faulty TPIM become similar to the balanced motor equations. By employing (12) and (13) to the faulty TPIM equations (equations (1)-(3)) and after simplifying, we have:

Rotor voltage equations:

(14) $$\begin{bmatrix} 0 \\ 0 \end{bmatrix} = \begin{bmatrix} M_q\frac{d}{dt} & (\omega_r - \omega_e)M_q \\ -(\omega_r - \omega_e)M_q & M_q\frac{d}{dt} \end{bmatrix}\begin{bmatrix} i_{ds}^e \\ i_{qs}^e \end{bmatrix} + \begin{bmatrix} r_r + L_r\frac{d}{dt} & (\omega_r - \omega_e)L_r \\ -(\omega_r - \omega_e)L_r & r_r + L_r\frac{d}{dt} \end{bmatrix}\begin{bmatrix} i_{dr}^e \\ i_{qr}^e \end{bmatrix}$$

Rotor flux equations:

(15) $$\begin{bmatrix} \lambda_{dr}^e \\ \lambda_{qr}^e \end{bmatrix} = \begin{bmatrix} M_q & 0 \\ 0 & M_q \end{bmatrix}\begin{bmatrix} i_{ds}^e \\ i_{qs}^e \end{bmatrix} + \begin{bmatrix} L_r & 0 \\ 0 & L_r \end{bmatrix}\begin{bmatrix} i_{dr}^e \\ i_{qr}^e \end{bmatrix}$$

Electromagnetic torque equation:

(16) $$\tau_e = \frac{Pole}{2} M_q (i_{qs}^e i_{dr}^e - i_{ds}^e i_{qr}^e)$$

In equation (14), ω_e is the angular velocity of the Rotor Flux Oriented reference frame. As can be seen, by applying proposed rotational transformations, equations of rotor voltages, rotor fluxes and electromagnetic torque are obtained like balanced equations (the only difference between these equations and balanced TPIM equations is that in these equations it is obtained: $M=M_q=\sqrt{3}/2L_{ms}$ but in the balanced TPIM equations, we have: $M=3/2L_{ms}$). In the RFOC technique, the rotor flux vector is aligned with d-axis ($\lambda_{dr}^e = |\lambda_r|$ and $\lambda_{qr}^e = 0$). With this assumption and from equations (14)-(16) equations of RFOC for TPIM under open-phase fault are obtained as follows:

(17) $$|\lambda_r| = \frac{M_q i_{ds}^e}{1+T_r \frac{d}{dt}}, \omega_e = \omega_r + \frac{M_q i_{qs}^e}{T_r|\lambda_r|}, \tau_e = \frac{Pole}{2}\frac{M_q}{L_r}|\lambda_r| i_{qs}^e$$

In equation (17), T_r is rotor time constant. As mentioned before the rotational transformation for stator variables can be considered as following equation:

(18) $$\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix} = \left[T_{vs}^e\right] \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix} = \begin{bmatrix} -Z_4\cos\theta_e & Z_3\sin\theta_e \\ Z_4\sin\theta_e & Z_3\cos\theta_e \end{bmatrix} \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix} = \begin{bmatrix} a_v & b_v \\ c_v & d_v \end{bmatrix} \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix}$$

By employing equations (13) and (18) and after simplifying, di_{qs}^s/dt, di_{qr}^s/dt in the equation of v_{ds}^e and di_{ds}^s/dt, di_{dr}^s/dt in the equation of v_{qs}^e will be appeared. For generating of di_{qs}^s/dt, di_{qr}^s/dt terms in the equation of v_{ds}^e and di_{ds}^s/dt, di_{dr}^s/dt in the equation of v_{qs}^e, PI controllers are needed. Using of these PI controllers cause the controlling system to be complex. For solving of this problem, the coefficients of di_{qs}^s/dt, di_{qr}^s/dt in the equation of v_{ds}^e and the coefficients of di_{ds}^s/dt, di_{dr}^s/dt in the equation of v_{qs}^e are considered equal to zero. Therefore, we have:

(19) $$a_v = -b_v.Z_5.\cot\theta_e \quad , \quad c_v = d_v.Z_5.\tan\theta_e$$

where:

$$Z_5 = \left(L_{qs} - \frac{M_q^2}{L_r}\right) \Big/ \left(-\frac{L_{ds}M_q}{M_d} + \frac{M_d M_q}{L_r}\right)$$

With supposition $a_v = -\cos\theta_e$ and $c_v = \sin\theta_e$, b_v and d_v is obtained as:

(20) $$b_v = \left(\frac{-L_{ds}L_r M_q + M_d^2 M_q}{L_{qs}L_r M_d - M_d M_q^2}\right)\sin\theta_e$$

$$d_v = \left(\frac{-L_{ds}L_r M_q + M_d^2 M_q}{L_{qs}L_r M_d - M_d M_q^2}\right)\cos\theta_e$$

Therefore, equation (18) can be re-written as:

(21) $$\left[T_{vs}^e\right] = \begin{bmatrix} -\cos\theta_e & \left(\frac{-L_{ds}L_r M_q + M_d^2 M_q}{L_{qs}L_r M_d - M_d M_q^2}\right)\sin\theta_e \\ \sin\theta_e & \left(\frac{-L_{ds}L_r M_q + M_d^2 M_q}{L_{qs}L_r M_d - M_d M_q^2}\right)\cos\theta_e \end{bmatrix}$$

By considering of $L_{ds} / L_{qs} = (Md / Mq)^2$, (in the faulty TPIM: $M_q = \sqrt{3} / 2\, L_{ms}$, $M_d = 3 / 2\, L_{ms}$, $L_{qs} = L_{ls} + 1 / 2\, L_{ms}$, $L_{ds} = L_{ls} + 1 / 2\, L_{ms}$, $L_{ms} >> L_{ls}$) the proposed rotational transformations for stator voltage variables are obtained as following equation:

Rotational transformation for stator voltage variables:

(22) $$\begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix} = \left[T_{vs}^e\right] \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix} = \begin{bmatrix} -\cos\theta_e & -\frac{M_d}{M_q}\sin\theta_e \\ \sin\theta_e & -\frac{M_d}{M_q}\cos\theta_e \end{bmatrix} \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix}$$

By applying equations (13) and (22) and after simplifying, RFOC equations for stator voltages are obtained as following equations:

(23) $$v_{ds}^e = \left(\frac{r_s M_q^2 + r_s M_d^2}{2M_d^2}\right) i_{ds}^e + \left(L_{qs} - \frac{M_q^2}{L_r}\right)\frac{di_{ds}^e}{dt} - \omega_e i_{qs}^e \left(L_{qs} - \frac{M_q^2}{L_r}\right) + \left(\frac{M_q}{L_r}\right)\left(\frac{M_q i_{ds}^e - |\lambda_r|}{T_r}\right) + v_{ds}^{-e}$$

(24) $$v_{qs}^e = \left(\frac{r_s M_q^2 + r_s M_d^2}{2M_d^2}\right) i_{qs}^e + \left(L_{qs} - \frac{M_q^2}{L_r}\right)\frac{di_{qs}^e}{dt} + \omega_e i_{ds}^e \left(L_{qs} - \frac{M_q^2}{L_r}\right) + \omega_e M_q \frac{|\lambda_r|}{L_r} + v_{qs}^{-e}$$

where,

$$\begin{bmatrix} v_{ds}^{-e} \\ v_{qs}^{-e} \end{bmatrix} = \left(\frac{r_s M_q^2 - r_s M_d^2}{2M_d^2}\right) \begin{bmatrix} \cos 2\theta_e & -\sin 2\theta_e \\ -\sin 2\theta_e & -\cos 2\theta_e \end{bmatrix} \begin{bmatrix} i_{ds}^e \\ i_{qs}^e \end{bmatrix}$$

As you can see the structure of equations (23) and (24) are like the structure of balanced TPIM equations (based on (23) and (24), the only difference between these equations and balanced TPIM equations is that in these equations it is obtained: $r_s = ((M_q^2 + M_d^2)/2M_d^2) r_s$); $M=M_q=\sqrt{3}/2L_{ms}$; $L_s = L_{qs} = L_{ls} + 1/2L_{ms}$; v_{ds}^{-e} and v_{qs}^{-e} as above but in the balanced TPIM equations, we have: $M=3/2L_{ms}$; $L_s = L_{ls} + 3/2L_{ms}$). In summery, essential modifications on the conventional vector control, to make it suitable to control of faulty TPIM, are summarized in Table 1. For simplifying implementation of proposed method in this work, the current control loop with standard hysteresis controllers is used.

Adaptive sliding mode control

In the controlling faulty TPIM, the coefficients of PI controller need to modify from balanced mode to faulty mode. For this purpose, in this paper, an adaptive sliding mode is replaced instead of speed PI controller. Equation (3) can be shown as:

(25) $$\frac{d\omega_m}{dt} + a\omega_m + c = b i_{qs}^e$$

where:

$$\omega_m = \frac{2}{Pole}\omega_r \quad , \quad a = \frac{F}{J} \quad , \quad b = \frac{Pole}{2}\frac{M_q}{L_r}|\lambda_r| \quad , \quad c = \frac{\tau_l}{J}$$

Equation (25) can be considered with uncertainties as follows:

(26) $$\frac{d\omega_m}{dt} = -(a+\Delta a)\,\omega_m + (b+\Delta b)\, i_{qs}^e - (c+\Delta c)$$

where, the terms Δa, Δb and Δc denote the uncertainties of the terms *a, b* and *c* which these uncertainties are unknown and calculation of an upper bound is rather difficult to obtain.

The speed error can be represented as:

(27) $$e(t) = \omega_m(t) - \omega_m^*(t)$$

where, $\omega_m^*(t)$ is the reference speed. Taking derivative of equation (27) yields:

(28) $$\dot{e}(t) = \frac{e(t)}{dt} = -ae(t) + u(t) + d(t)$$

where,

$$u(t) = b i_{qs}^e(t) - a\omega_m^*(t) - \dot{\omega}_m^*(t) - c(t)$$

$$d(t) = \Delta b i_{qs}^e(t) - \Delta a \omega_m(t) - \Delta c(t)$$

The switching surface with integral component for the sliding mode speed control is considered as:

(29) $$S(t) = e(t) + \int_0^t (a+k)\, e(\tau)\, d\tau = 0$$

where, *k* is constant gain. In this work, the speed controller is considered is as follows:

(30) $$u(t) = -ke(t) - \tilde{\rho}(t)\,\alpha\,\mathrm{sgn}(S(t))$$
where,
$$\dot{\tilde{\rho}}(t) = \alpha \left| S(t) \right| \quad , \quad \mathrm{sgn}(S(t)) = \begin{cases} +1 & S(t) \rangle 0 \\ -1 & S(t) \langle 0 \end{cases}$$

Moreover, $\rho(t)$ is estimated switching gain and α is a positive constant.

Table 1. Comparison between two vector control methods

Conventional vector control for the balanced TPIM	Proposed Modified vector control for the faulty TPIM
3 to 2 transformation for the stator currents: $\begin{bmatrix} i_{ds}^s \\ i_{qs}^s \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} +1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_{as} \\ i_{bs} \\ i_{cs} \end{bmatrix}$	2 to 2 transformation for the stator currents according to (4): $\begin{bmatrix} i_{ds}^s \\ i_{qs}^s \end{bmatrix} = \frac{\sqrt{2}}{2} \begin{bmatrix} 1 & -1 \\ 1 & 1 \end{bmatrix} \begin{bmatrix} i_{as} \\ i_{bs} \end{bmatrix}$
Balanced rotational transformation for the stator currents according to (5): $\begin{bmatrix} i_{ds}^e \\ i_{qs}^e \end{bmatrix} = \begin{bmatrix} \cos\theta_e & \sin\theta_e \\ -\sin\theta_e & \cos\theta_e \end{bmatrix} \begin{bmatrix} i_{ds}^s \\ i_{qs}^s \end{bmatrix}$	Unbalanced rotational transformation for the stator currents according to (13): $\begin{bmatrix} i_{ds}^e \\ i_{qs}^e \end{bmatrix} = \begin{bmatrix} \frac{M_d}{M_q}\cos\theta_e & \sin\theta_e \\ -\frac{M_d}{M_q}\sin\theta_e & \cos\theta_e \end{bmatrix} \begin{bmatrix} i_{ds}^s \\ i_{qs}^s \end{bmatrix}$
Inverse of balanced rotational transformation for the stator voltages according to (5): $\begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix} = \left(\begin{bmatrix} \cos\theta_e & \sin\theta_e \\ -\sin\theta_e & \cos\theta_e \end{bmatrix} \right)^{-1} \begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix}$	Inverse of unbalanced rotational transformation for the stator voltages according to (22): $\begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix} = \left(\begin{bmatrix} -\cos\theta_e & -\frac{M_d}{M_q}\sin\theta_e \\ \sin\theta_e & -\frac{M_d}{M_q}\cos\theta_e \end{bmatrix} \right)^{-1} \begin{bmatrix} v_{ds}^e \\ v_{qs}^e \end{bmatrix}$
2 to 3 transformation for the stator voltages: $\begin{bmatrix} v_{as} \\ v_{bs} \\ v_{cs} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} +1 & 0 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix}$	2 to 2 transformation for the stator voltage according to (4): $\begin{bmatrix} v_{as} \\ v_{bs} \end{bmatrix} = \left(\frac{\sqrt{2}}{2} \begin{bmatrix} 1 & -1 \\ 1 & 1 \end{bmatrix} \right)^{-1} \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix}$
Stator self inductance: $L_s = L_{ls} + \frac{3}{2} L_{ms}$	Stator self inductance according to (23) and (24): $L_s = L_{qs} = L_{ls} + \frac{1}{2} L_{ms}$
Stator and rotor mutual inductance: $M = \frac{3}{2} L_{ms}$	Stator and rotor mutual inductance according to (17), (23) and (24): $M = M_q = \frac{\sqrt{3}}{2} L_{ms}$
Stator resistance: r_s	Stator resistance according to (23) and (24): $\frac{r_s M_q^2 + r_s M_d^2}{2 M_d^2} = \frac{2}{3} r_s$
----------	Stator backward components according to (23) and (24): $\begin{bmatrix} v_{ds}^{-e} \\ v_{qs}^{-e} \end{bmatrix} = \left(\frac{r_s M_q^2 - r_s M_d^2}{2 M_d^2} \right) \times \left(\begin{bmatrix} \cos 2\theta_e & -\sin 2\theta_e \\ -\sin 2\theta_e & -\cos 2\theta_e \end{bmatrix} \begin{bmatrix} i_{ds}^e \\ i_{qs}^e \end{bmatrix} \right)$
---------	Modification of speed PI controller coefficients according to Fig. 3

Theorem 1. The adaptive structure speed controller with the adaptation algorithm (30) makes the controlled system convergent to the switching surface $S(t) = 0$ and the stability for the speed control can be guaranteed.

Proof: Choosing a Lyapunov function candidate:

(31) $$V(t) = \frac{1}{2}\left(S^2(t) + \hat{\rho}^2(t)\right)$$

where,
$$\hat{\rho}(t) = \tilde{\rho}(t) - \rho$$
Taking the derivative of the Lyapunov function:

(32)
$$\begin{aligned}
\dot{V}(t) &= S(t)\dot{S}(t) + \hat{\rho}(t)\dot{\hat{\rho}}(t) \\
&= S(t)\left(\dot{e}(t) + (a+k)e(t)\right) + \hat{\rho}(t)\dot{\tilde{\rho}}(t) \\
&= S(t)\left(-ae(t) + u(t) + d(t) + (a+k)e(t)\right) + \alpha\hat{\rho}(t)\left| S(t) \right| \\
&= S(t)\left(u(t) + d(t) + ke(t)\right) + \alpha\left(\tilde{\rho}(t) - \rho\right)\left| S(t) \right| \\
&= S(t)\left(d(t) - \tilde{\rho}(t)\alpha\,\mathrm{sgn}(S(t))\right) + \alpha\left(\tilde{\rho}(t) - \rho\right)\left| S(t) \right| \\
&= S(t)\left(d(t) - \tilde{\rho}(t)\alpha \left| S(t) \right|\right) + \alpha\left(\tilde{\rho}(t) - \rho\right)\left| S(t) \right| \\
&= S(t)d(t) - \alpha\rho\left| S(t) \right| \\
&\le \left| S(t) \right|\left| d(t) \right| - \alpha\rho\left| S(t) \right|
\end{aligned}$$

Assumption 1: $\rho > d_{max}$

Therefore, from equation (32):

(33)
$$\begin{aligned}
S(t)d(t) - \alpha\rho\left| S(t) \right| &\le \left| S(t) \right|\left| d(t) \right| - d_{\max}\alpha\left| S(t) \right| \\
&= \left| S(t) \right|\left(\left| d(t) \right| - d_{\max}\alpha\right)
\end{aligned}$$

Assumption 2: $\alpha > 1$

Therefore, from equation (33):

(34) $$\left| S(t) \right|\left(\left| d(t) \right| - d_{\max}\alpha\right) \le 0 \qquad \square$$

Using Lyapunov theorem, the controlled system is stable. Since $S(t)$ is bounded, $e(t)$ is also bounded. From equation (29):

(35) $$\dot{S}(t) = S(t) = \dot{e}(t) + (a+k)e(t) = 0 \Rightarrow \dot{e}(t) = -(a+k)e(t)$$

Assumption 3: $a + k > 0$

From equation (35) and **Assumption 3**, It is obvious the tracking error $e(t)$ converges to zero. Therefore, from equation (28) and (30):

(36) $$i_{qs}^e(t) = \frac{1}{b}\left(-ke(t) - \tilde{\rho}(t)\alpha\,\mathrm{sgn}(S(t)) + \alpha\omega^*(t) + \dot{\omega}^*(t) + c\right)$$

Based on equation (36), the block diagram for the speed control with adaptive sliding mode can be presented, like in Fig.3.

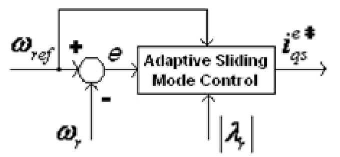

Fig.3. Scheme of the speed controller with adaptive sliding mode

Results and simulations

A motor which is fed from a SPWM (Sine Pulse Width Modulation) Voltage Source Inverter (VSI) was simulated using Matlab. Motor data have been given as follows: *Voltage:*125*V, f* = 50*Hz, no. of poles* = 4*, Power* = 475*W,* r_s = 20.6Ω*,* r_r = 19.15Ω*,* L_{lr} = L_{ls} = 0.0814*H,* L_{ms} = 0.851*H, J* = 0.0038*kg.m2*

The controller, which has been employed to the speed control of IM, is a conventional RFOC. To verify the effectiveness of the proposed control system for faulty TPIM, vector control drive system based on Table 1 and Fig.3 is also simulated. For showing the better performance of the proposed drive system, an uncertainty around 15% after fault occurrence is supposed.

Fig.4 shows the simulation results of the conventional RFO controller. In starting and loading the motor is healthy. At time t=0.4s the load torque steps from 0N.m to 1N.m, and as before, it is supposed that there is an uncertainty around 15% after fault occurrence in the load torque (In TPIM and under open phase fault, the maximum permissible torque is about 30% of the rated motor torque as mentioned in [12]). At time t=1s an open-phase fault occurs and the motor becomes unbalanced. Results show that the conventional controller can not control the faulty motor appropriately. Especially, we can see considerable oscillations in the torque and speed of motor after fault.

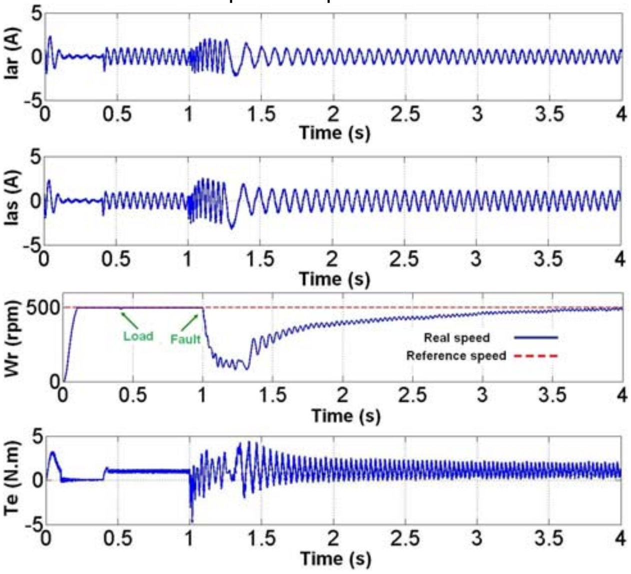

Fig.4. Simulation results of the conventional controller (Rotor a-axis current, Stator a-axis current, Rotor speed, Electromagnetic torque)

In Fig.5, the identical procedure is done again but this time after the open-phase fault, the proposed modified controller is applied. Simulation results show that the proposed system drive reduces the torque and rotor speed oscillation noticeably (e.g., rotor speed oscillation, by using conventional RFO controller, after open-phase fault and at steady state is ~ 8 rpm but by using proposed modified RFO controller the speed oscillation decreased outstandingly ~ 0.5 rpm at rotor speed of 500 rpm).

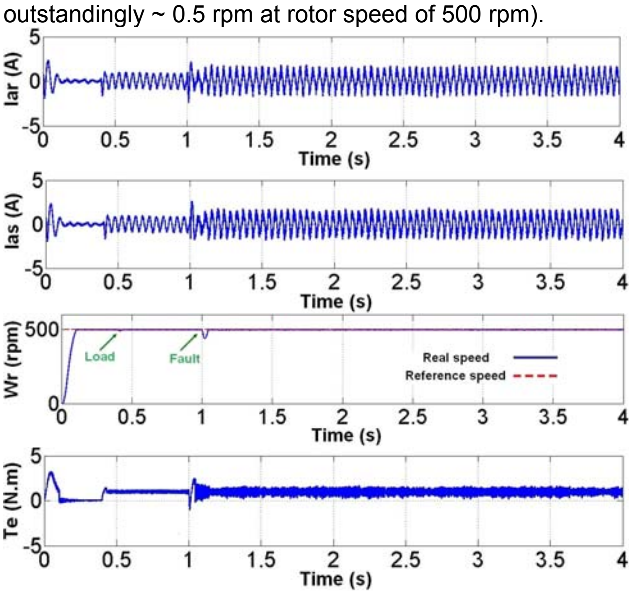

Fig.5. Simulation results of the proposed modified controller (Rotor a-axis current, Stator a-axis current, Rotor speed, Electromagnetic torque)

Using sliding mode control in the proposed faulty TPIM drive, the controlled speed is insensitive to variations in the load torque disturbances and parameters of motor. Due to the variable structure control nature, this proposed control system is robust under uncertainties caused by changes in the load and parameter errors.

Conclusion

This work shows vector control of faulty TPIM with an adaptive sliding mode control Based on RFOC. The theory and analysis for the proposed vector control procedure has been explained as above. It can be seen from presented results, the dynamic performance of the proposed approach for vector control of faulty TPIM is extremely acceptable. This method can be used for high critical industrial applications where we need to have fault-tolerant drive system and also can be utilized for vector control of SPIM, because SPIM with two main and auxiliary windings is like an unbalanced IM.

REFERENCES

[1] M.Jannati, E.Fallah, "Modeling and Vector Control of Unbalanced induction motors (faulty three phase or single phase induction motors)," *1st. Conference on Power Electronic & Drive Systems & Technologies (PEDSTC)*, pp. 208 – 211, May 2010.

[2] M. S. Ballal, D. M. Ballal, H. M. Suryawanshi and M. Kumar Mishra, "Wing Technique: A Novel Approach for the Detection of Stator Winding Inter-Turn Short Circuit and Open Circuit Faults in Three Phase Induction Motors," *Journal of Power Electronics*, Vol. 12, No. 1, pp. 208-214, January 2012

[3] Ghazal, M. and J. Poshtan, "Robust stator winding fault detection in induction motors," In Power Electronics*, Drive Systems and Technologies Conference (PEDSTC)*, 2011 2nd, pp. 163-168. IEEE, 2011

[4] Alberto Gaeta, Giacomo Scelba and Alfio Consoli, "Modelling and Control of Three-Phase PMSMs under Open-Phase Fault," *IEEE Transaction on Ind. Apl.*, Vol. 49, No. 1, pp. 74-83, Jun/Feb 2013

[5] M. Jannati, N. R. N. Idris and Z. Salam, "A New Method for Modelling and Vector Control of Unbalanced Induction Motors", *Energy Conversion Congress and Exposition (ECCE)*, Sep 2012

[6] S. Jena and K. B. Mohanty, ROBUST FIELD ORIENTED INDUCTION MACHINE CONTROL USING SMC, *The 5th PSU-UNS International Conference on Engineering and Technology (ICET-2011)*, Phuket, May 2-3, 2011

[7] M. Montazeri-Gh, A. Poursamad, B. Ghalichi, Application of genetic algorithm for optimization of control strategy in parallel hybrid electric vehicles, *Journal of the Franklin Institute* 343 (2006) 420–435.

[8] M.A. Fnaiech, F. Betin, G.A. Capolino, F. Fnaiech, Fuzzy logic and sliding-mode controls applied to six-phase induction machine with open phases, *IEEE Transactions on Industrial Electronics* 57 (1) (2010) 354–364.

[9] T-J. Ren, T-C. Chen, Robust speed-controlled induction motor drive based on recurrent neural network, *Electric Power System Research* 76 (2006) 1064–1074.

[10] R. Marino, P. Tomei, C.M. Verrelli, An adaptive tracking control from current measurements for induction motors with uncertain load torque and rotor resistance, *Automatica* 44 (2008) 2593–2599.

[11] V.I. Utkin, Sliding mode control design principles and applications to electric drives, *IEEE Transactions on Industrial Electronics* 40 (1993) 26–36.

[12] Saleh, A.; Pacas, M.; Shaltout, A., "Fault Tolerant Field Oriented Control of the Induction Motor for Loss of One Inverter Phase, *IEEE Industrial Electronics, IECON 2006 - 32nd Annual Conference on* , vol., no., pp.817,822, 6-10 Nov. 2006

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