Decoupled current control and sensor fault detection with second-order sliding mode for induction motor

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Abstract: This study proposes a new decoupled control method for induction motor (IM) based on higher order sliding mode (HOSM) controller. The proposed controller serves dual purpose by offering decoupled control and sensor fault detection. In this scheme, the decoupled control of \( d \) – \( q \) currents does not require the knowledge of the speed. The HOSM controllers play the same role as the compensation voltages produced by a decoupling compensator. From the compensation voltages, the speed can be estimated accurately through algebraic calculations. The estimated and actual speeds are then employed for fault detection algorithm to detect the fault. Simulations on a 1/4-hp three-phase IM in the presence of random measurement noise highlight the performance of the proposed approach for decoupling current control and robust sensor fault detection.

1 Introduction

The significance of field-oriented control (FOC) in induction motor (IM) drives for obtaining torque dynamics equivalent to separately excited dc motor has gained prominence over time [1, 2]. The torque generated in IM under field oriented control is a product of the rotor flux and the \( q \)-axis current in the synchronous frame of reference. The only hindrance in efficient operation of FOC arises from the coupling between the currents of the \( d \)-axis and \( q \)-axis, respectively. The coupling effect results in distortion of torque, as a change in one current produces a change in the other current. Decoupled current control in [3] is widely regarded as one of the optimal solutions to this problem. In the decoupled FOC of an IM, the \( q \)-axis current reference is generated from the speed dynamics, while the \( d \)-axis current is kept constant to obtain a controlled torque response. On the other hand, the occurrence of faults in speed, current and voltage sensors will degrade the drive performance. Therefore it is important for the drive system to be sensor fault resilient.

Recently, fault detection and identification (FDI) has emerged as a necessary practice in industrial applications on improving the factors of safety and reliability [4–6]. In [4], an easy and fast FDI algorithm is proposed, keeping the system performances unchanged under certain faulty sensor conditions when reconfigurations are available. This algorithm is derived from a parity space approach based on temporal redundancies. Further, various fault tolerant control (FTC) methods are also available in the literature for IMs in [7–9]. In [9], a multi-sensor switching for sensor FTC of IMs is proposed. In this approach, faulty sensor is avoided by the switching mechanism through pre-checkable conditions. Recently, in [10], an observer-based sensor FTC using maximum likelihood voting approach is proposed for IM drives. This approach is composed of two observers (an extended Kalman filter and a Luenberger observer) and compares the estimated speeds with the sensor measurements to achieve the sensor FTC.

Sliding mode theory has been used in the design of nonlinear controllers for IMs in [11, 12]. Similarly, sliding mode observers which are robust to parametric uncertainties and disturbances have been implemented for various applications [13, 14]. First-order sliding mode (FOSM) observer-based approaches have been developed to tackle actuator and sensor faults in [15–18]. In [15], the sliding mode ideas for the purpose of fault detection and isolation have been explored. This paper considers the practical situation when the system states are not available. The concept of equivalent output injection is applied to explicitly reconstruct the fault signals. FOSM observer for FTC in IMs is proposed in [19]. Similar approaches involving the use of FOSM-based observers for fault detection and control are analysed in [20, 21]. In [22], an integral sliding mode (ISM) current controller for the IM is implemented to decouple the \( d \) – \( q \) currents and compensate the parameter variations in the current loops of the machine. Moreover, in [23], an estimation of the rotor speed based on ISM current control scheme is proposed. However, the ISM current control involves the use of
low-pass filters to remove the chattering effect to extract the unknown components from ISM. This introduces an unwarranted delay into the estimation procedure. The introduction of higher order sliding mode (HOSM) theory effectively countered this problem without requiring the low-pass filters. Also, it can reduce the well-known chattering phenomenon.

Various approaches on HOSM-based observers/controllers have been investigated [17, 24–27]. Although there are several approaches available in the literature, the problem of decoupled control design based on HOSM is yet to be addressed.

In this paper, a HOSM current-control scheme is proposed to meet the following two objectives: (i) decoupled control of $d-q$ currents for robust tracking and (ii) speed sensor fault detection. The second objective is a by-product of the decoupled control scheme that enables the fault detection. The HOSM current-control scheme comprises of nominal control obtained using proportional–integral (PI) controller and the HOSM controller output. The nominal PI controller acts on the ideal plant dynamics, devoid of coupling terms. And, the HOSM controller generates an equivalent continuous approximation of the compensation voltages that cancels out the coupling terms in the current dynamics without the use of low-pass filtering. Further, the unknown compensation voltages estimated using the HOSM controller are used to calculate an accurate estimate of the rotor speed. In a typical speed sensor fault scenario, the fault is detected on the basis of a threshold. Simulation results in the presence of sensor noise show the effectiveness of the proposed control approach.

The rest of the paper is organised as follows: preliminaries and classical decoupled control are discussed in Section 2. The proposed HOSM controller design is discussed in Section 3. Section 4 presents the speed estimation and fault detection approach. Simulations results are presented in Sections 5 and 6 concludes the paper.

2 Classical decoupled current-control for IM

The model of the IM currents, in the rotational reference frame in combination with rotor flux, can be described as the following [2, 23]

$$\frac{di_d}{dt} = -\gamma i_d + \eta \beta \lambda_i + n_p \omega_i i_q + \eta L_m \frac{i_q^2}{\lambda_i} + \frac{1}{\sigma L_d} V_d$$

$$\frac{di_q}{dt} = -\gamma i_q - \beta n_p \omega_i \lambda_i + n_p \omega_i i_d - \eta L_m \frac{i_d i_q}{\lambda_i} + \frac{1}{\sigma L_q} V_q$$

with the motor parameters $\eta$, $\sigma$, $\beta$ and $\gamma$ given by

$$\eta = \frac{1}{T_i}; \quad \sigma = 1 - \frac{L_m^2}{L_s L_d}; \quad \beta = \frac{L_m}{\sigma L_d L_q}; \quad \gamma = \frac{1}{\sigma L_s} \left( \frac{L_m^2}{L_s L_d R_i} + R_s \right)$$

where $\omega_i$ is the rotor electrical speed, $(i_d, i_q)$ are the currents and $(V_d, V_q)$ are the voltages in rotational reference frame. And, $\lambda_i$ is a magnitude of the rotor flux vector, $(\omega_{i_d}, \omega_{i_q})$ are the projections of the flux vector on the stationary reference frame, $(R_s, R_d)$ are the stator and rotor resistances, $(L_m, L_s, L_d)$ are the magnetising, stator and rotor inductances, $T_i$ is the rotor time constant and $n_p$ is the number of pole pairs.

The current dynamics $(i_d, i_q)$ in (1) and (2) are dependent on each other; and also they have additional terms which are functions of motor parameters, rotor flux magnitude, and speed. Fig. 1 shows the classical decoupled current-control scheme using PI controllers. The PI current controllers are designed for a certain type of reference inputs to regulate the currents with zero steady-state error.

In classical decoupled control, output of the current controllers provides a command voltage which comes from the sum of two terms: the one is a PI controller output $(V_d')$ and $V_q'$ and the other one is a compensation voltage $(V_d^\text{comp}$ and $V_q^\text{comp})$ [28]. The compensation voltages that cancel the additional terms coupling and unknown terms shown in Fig. 1 are dependent on the motor speed and are given by

$$V_d^\text{comp} = -\sigma L_d \left( n_q \omega_i i_q + \eta L_m \frac{i_q^2}{\lambda_i} \right)$$

$$V_q^\text{comp} = \sigma L_q \left( \beta n_p \omega_i \lambda_i + n_p \omega_i i_d + \eta L_m \frac{i_d i_q}{\lambda_i} \right)$$

For instance, decoupled control of current $i_d$ is achieved by inserting (3) in (1), this control cancels the unknown disturbances that depends on $i_q$. Once the cancellation takes place, the overall dynamics of $i_d$ is simplified, and it is easy to design the corresponding PI controller gains.

Further, the classical decoupled-current control requires the speed signal in order to compute the compensation voltages $(V_d^\text{comp}$ and $V_q^\text{comp})$. For example, in some sensorless schemes [23, 29], depending on the control objective, speed may not be available; consequently, the control algorithm requires an additional structure to obtain the speed estimate. Therefore it is difficult to compute the compensation voltages (3) and (4) in sensorless control schemes. Because of this disadvantage, some implementations [30, 31] use only PI controllers (the compensation voltages are omitted). These implementations cannot achieve the current control when the reference currents are step functions [30]. However, the steady-state current errors can be reduced by increasing the proportional gains of the PI controllers [32]. The objective is to design a robust current controller, which can decouple the dependent currents $(i_d, i_q)$ without requiring the knowledge of the speed. An accurate decoupled control of the currents will result in accurate estimation of speed. In addition, in a typical speed sensor fault scenario, the objective is also to detect the fault by comparing the measured and estimated speeds.
3 HOSM decoupled controller design

Let us note that the unknown disturbance terms of the plant dynamics in (1) and (2) act like extra dynamics. The idea is to use sliding mode controller on both the $d$-axis and $q$-axis to cancel the extra dynamics of the plant. In the following, a HOSM controller will be designed to cancel the disturbance dynamics.

The current dynamics in (1) and (2) can be expressed as

$$\begin{align*}
\frac{di_d}{dt} &= -\gamma i_d + \eta \beta \lambda_i + \Delta_d(\omega_i, i_q, \lambda_i) + \frac{1}{\sigma_i} V_d, \\
\frac{di_q}{dt} &= -\gamma i_q + \Delta_q(\omega_i, i_q, i_d, \lambda_i) + \frac{1}{\sigma_i} V_q,
\end{align*}$$

(5)

where $\Delta_d(\omega_i, i_q, \lambda_i)$ and $\Delta_q(\omega_i, i_q, i_d, \lambda_i)$ are the cross-coupling and unknown terms for the $d$-axis and $q$-axis, respectively.

$$\begin{align*}
\Delta_d(\omega_i, i_q, \lambda_i) = n_\delta \omega_i l_q + \eta L_m \frac{d^2}{\lambda_r} \\
\Delta_q(\omega_i, i_q, i_d, \lambda_i) = -\beta a_\delta \omega_i l_q - n_\eta \omega_i l_d - \eta L_m \frac{d^2}{\lambda_r}
\end{align*}$$

(6)

The disturbance terms consist of some known terms and some unknown terms which result in coupling of the current dynamics. The objective is to design a controller that performs decoupling action by cancelling the coupling and unknown terms. The control input provided to the plant comprises of nominal control and disturbance rejection control. The nominal plant dynamics which is devoid of the disturbances can be expressed as

$$\begin{align*}
\frac{di_d}{dt} &= -\gamma i_d + \frac{1}{\sigma_i} u_{0,d} \\
\frac{di_q}{dt} &= -\gamma i_q + \frac{1}{\sigma_i} u_{0,q}
\end{align*}$$

(7)

The nominal control inputs ($u_{0,d}, u_{0,q}$) are generated by the use of PI controllers for the ideal current dynamics shown in Fig. 2. The output of the ideal plant dynamics are the fictitious currents ($i^{\text{non}}_d, i^{\text{non}}_q$), called the model currents. Note that the measured currents ($i_d, i_q$) are not used in the generation of nominal control inputs ($u_{0,d}, u_{0,q}$), they are only used by the proposed HOSM controllers to provide for the tracking of the reference currents. However, it is atypical control scheme because the error term ($i^{\text{non}}_d - i^{\text{non}}_q$) does not appear directly in Fig. 2 like in a classical decoupled-control scheme shown in Fig. 1. To achieve the rejection of the unknown terms $\Delta_d(\omega_i, i_q, \lambda_i)$ and $\Delta_q(\omega_i, i_q, i_d, \lambda_i)$, and to achieve decoupled control, HOSM controllers are employed to control inputs ($\Psi_d, \Psi_q$).

For the design of HOSM controllers, the following sliding surfaces are chosen

$$\begin{align*}
\dot{s}_d(t) &= i_d - i^{\text{ref}}_d \\
\dot{s}_q(t) &= i_q - i^{\text{ref}}_q
\end{align*}$$

(8)

where $i_d, i_q$ are the actual currents and $i^{\text{ref}}_d, i^{\text{ref}}_q$ are the reference currents. Applying the same design principles as for variable structure control, the controller trajectories are constrained to evolve after a finite time on the sliding manifold by the use of a discontinuous output injection signal. Hence, the sliding motion provides a cancellation of extra terms (in finite time) of the system states. The dynamics of the sliding surfaces, employing the modified second-order sliding mode algorithm [33] can be obtained for the $d$-axis and $q$-axis, respectively, as

$$\begin{align*}
\dot{s}_d &= -\gamma i_d + \eta \beta \lambda_i + \Delta_d(\omega_i, i_q, \lambda_i) + \frac{1}{\sigma_i} v_d(t) - \frac{d}{dt} i^{\text{ref}}_d \\
\dot{s}_q &= -\gamma i_q + \Delta_q(\omega_i, i_q, i_d, \lambda_i) + \frac{1}{\sigma_i} v_q(t)
\end{align*}$$

(9)

and define the terms of $\dot{s}_d$ and $\dot{s}_q$ as

$$\begin{align*}
\Omega_d &= -\gamma i_d + \Delta_d(\omega_i, i_q, \lambda_i) \\
\Omega_q &= -\gamma i_q + \Delta_q(\omega_i, i_q, i_d, \lambda_i)
\end{align*}$$

(10)

and define the terms of $\dot{v}_d(t)$ and $\dot{v}_q(t)$ are the sliding mode terms, $K_1, K_2$ and $K_3$ are properly designed positive constants.

3.1 Convergence in sliding mode

The HOSM controller employed in this paper provides finite-time convergence and chattering free estimation of compensation voltages. For illustration, one considers the $d$-axis controller and prove the finite time convergence to
the selected sliding surface, \( s_d \). The sliding surface dynamics in (9) can be written as

\[
\dot{s}_d = -\gamma s_d + \eta \beta \lambda_i + \Omega_d + \frac{1}{\sigma L_s} v_d(t)
\]

Here, \( \eta \beta \lambda_i \) is a bounded quantity which comprises of known system parameters and the flux. The boundedness of this term can be established as \( \eta \beta \lambda_i \leq \rho_d \). The terms of \( \Omega_d \) in (11) are functions of the physical quantities, \( \omega_r \), \( i_q \) and \( \lambda_i \) and other known system parameters. Hence, the boundedness of this function can also be established at least locally as

\[
\dot{\Omega}_d + \eta \beta \frac{d\lambda_i}{dt} \leq \rho_d
\]

for some positive constant \( \rho_d \). The above condition (12) is not restrictive since \( \omega_r \), \( i_q \) and \( \lambda_i \) are continuous on a compact set.

**Theorem 1:** With the condition (12) satisfied, the controller (9) and (10) will ensure that the sliding surfaces \((s_d, s_q)\) converges to zero in finite time.

**Proof:** For simplicity of exposition, we only prove the convergence of sliding surface \((s_d)\). With the robust terms of the controller defined as in (9) and the perturbations bounded as in (12), the convergence of the sliding surface dynamics can be proved similar to [33] by consideration of the following Lyapunov function [34]

\[
V(s_d) = \xi^T Q \xi
\]

where \( \xi = [\phi_1(s_d) \ s_d \ s_d \int_0^t \phi_2(s_d) \ dt]^T \) and \( Q = Q^T > 0 \) is a positive definite matrix defined as

\[
Q = \frac{1}{2} \begin{bmatrix}
(4K_2 + K_1^2) & K_1 \gamma & -K_1 \\
K_1 \gamma & (1 + K_1^2) & -K_3 \\
-K_1 & -K_3 & 2
\end{bmatrix}
\]

The considered Lyapunov function satisfies

\[
\lambda_{\min} \|\xi\|^2 \leq V(s_d) \leq \lambda_{\max} \|\xi\|^2
\]

where \( \lambda_{\max} \) represents the maximum eigenvalue and \( \lambda_{\min} \) represents the minimum singular value. The time derivative of the Lyapunov function along the trajectories of the system can thus be obtained as

\[
\dot{V}(s_d) = 2\xi^T Q \xi
\]

Simplifying further \( \dot{V}(s_d) \) can now be written as

\[
\dot{V}(s_d) = -\frac{1}{\|s_d\|^2} \xi^T Q \xi - \xi^T Q_2 \xi
\]

where

\[
Q_1 = \begin{bmatrix}
q_1 & 0 & -\frac{K_1^2}{2} \\
0 & q_{12} & -\frac{3K_2K_1}{2} \\
-\frac{K_1^2}{2} & -\frac{3K_2K_1}{2} & K_3
\end{bmatrix}
\]

\[
Q_2 = \begin{bmatrix}
q_1 & 0 & -\frac{K_1 \gamma}{4} \\
0 & q_2 & q_3 \\
-\frac{K_1 \gamma}{4} & q_3 & K_3
\end{bmatrix}
\]

![Flowchart of the proposed fault detection algorithm](image1)

**Table 1** Motor specifications and parameters

| Rating | 1/4 hp |
| Speed | 1732 r/min |
| Stator resistance | \( R_s \) | 10.9 \( \Omega \) |
| Stator, rotor leakage inductance | \( L_{ls}, L_{lr} \) | 0.015 H |
| Magnetising inductance | \( L_{mi} \) | 0.30 H |
| Rotor resistance | \( R_r \) | 5.57 \( \Omega \) |
| Number of pole pairs | \( n_p \) | 2 |

![Combined speed estimation and fault detection approach](image2)

![Proposed HOSM-based decoupled control scheme](image3)
with
\[
q_{11} = (2K_2 + K_2^2)\frac{K_1}{2} - K_1\rho_d,
\]
\[
q_{12} = K_1\left(2 + 5K_2^2\right) - \frac{3K_1K_2\gamma}{2},
\]
\[
q_{21} = (K_3 + 2K_1^2)K_2 - K_3\rho_d - \gamma\left(\frac{K_1^2}{2} + 2K_2\right),
\]
\[
q_{22} = K_3(K_3 + 1) - \gamma(K_3 + 2) \quad \text{and} \quad q_{23} = -K_3^2 - \frac{\gamma K_3}{2}.
\]

Similar to the arguments in [33], it can be shown that if the gains $K_1$, $K_2$ and $K_3$ satisfy the following inequalities
\[
\left\{
\begin{array}{l}
K_1 > \left(\frac{2K_2\gamma + K_2\rho_d - K_1K_3}{2K_3 - \frac{\gamma}{2}}\right)^{1/2} \\
K_2 > \frac{2\rho_d - K_3^2}{2} \\
K_3 > \gamma\left(I\left(K_1^2/2\right) + 2K_2\right) \\
\end{array}
\right.
\]

then, it can be shown from (13) that
\[
\dot{V}(s_d) = -\frac{1}{\|s_d\|^2}\lambda_{min}(Q_1)\|\xi\|^2 - \lambda_{max}(Q_2)\|\xi\|^2
\]
\[
\|s_d\|^2 \leq \|\xi\|^2 \leq \frac{V^2(s_d) + s_1V(s_d)}{\lambda_{min}(Q)}
\]

It can be thus shown that
\[
\dot{V}(s_d) \leq s_1V(s_d) - s_2V(s_d)
\]

where
\[
s_1 = \frac{\lambda_{min}(Q)\lambda_{min}(Q_1)}{\lambda_{max}(Q)}, \quad s_2 = \frac{\lambda_{min}(Q_2)}{\lambda_{max}(Q)}
\]

With the proper selection of the gains, $K_i > 0$, $i = 1, 2, 3$ such that (14) is satisfied ensures that $\dot{V}(s_d)$ is negative definite. Thus the sliding surface can be reached in finite time and maintained thereafter. Similarly, the convergence analysis for the $q$-axis current can be established. □

4 Fault detection algorithm

In the sliding mode, we have $s_d = \dot{s}_d = 0$ and $s_q = \dot{s}_q = 0$. The reduced order dynamics of the system (5) can be
obtained as

\[
\begin{align*}
0 &= \eta \beta \lambda_i + \Omega_d - \frac{1}{\omega_r} (K_2 \int_0^\tau \phi_2(s_d) \, dt) \\
0 &= \Omega_q - \frac{1}{\omega_s} (K_2 \int_0^\tau \phi_2(s_q) \, dt)
\end{align*}
\]

The estimated robust terms correspond to the continuous compensation voltages ($V_{d}^{\text{comp}}$ and $V_{q}^{\text{comp}}$) and are given by

\[
\begin{align*}
\Psi_d &= \sigma L_s (\beta n_p \omega_r \lambda_i - \gamma i_{\text{ref}} - d i_{\text{ref}} + \eta \beta \lambda_i) - K_2 \int_0^\tau \phi_2(s_d) \, dt \\
\Psi_q &= \sigma L_s (\gamma i_{\text{ref}} + d i_{\text{ref}} - \eta \beta \lambda_i + n_p \omega_r \lambda_i) - K_2 \int_0^\tau \phi_2(s_q) \, dt
\end{align*}
\]

(15)

One can re-write the above (15) into the following form

\[
\begin{bmatrix}
-n_p i_q - L_m \frac{\omega_r}{\omega_s} \\
\beta n_p \lambda_i + n_p i_d + L_m \frac{\omega_r}{\omega_s}
\end{bmatrix}
\begin{bmatrix}
\omega_r \\
\omega_s
\end{bmatrix}
= -\frac{1}{\sigma L_s}
\begin{bmatrix}
\Psi_d \\
\Psi_q
\end{bmatrix}
\]

(16)

The determinant of the left-hand side matrix is $\Pi = \beta n_p L_m i_q^2$ and is different from zero when $i_q \neq 0$. Then, the solution exists and the speed can be algebraically computed as

\[
\hat{\omega}_r = \frac{\Psi_d i_d + \Psi_q i_q}{\beta n_p i_q + i_d \sigma L_s}
\]

(17)

The speed estimate in (17) requires the real $d - q$ currents, motor parameters and also the estimated robust terms ($\Psi_d, \Psi_q$). The proposed HOSM controller provides smooth estimation of the rotor speed compared to the ISM control obtained using a low-pass filter [23].
The control voltages \((V_d, V_q)\) in (5) are generated by summing two components as

\[
\begin{align*}
V_d &= u_{0,d} + \Psi_d \\
V_q &= u_{0,q} + \Psi_q
\end{align*}
\tag{18}
\]

The terms \((u_{0,d}, u_{0,q})\) are the nominal inputs while \((\Psi_d, \Psi_q)\) are the continuous estimations of the unknown compensation voltages.

Moreover, the fault detection method that relies on the speed estimate obtained from the proposed de-coupled controller is discussed. The measured and estimated speeds are fed into a decision making unit to detect the faulty speed sensor. The combination of speed estimation and fault detection scheme is depicted in Fig. 3a.

Speed estimation only relies on the current dynamics \(i_d, i_q\) and system parameters \(\beta, n_p, \sigma, L_s\) and \(\lambda_r\) that are known. As long as the current dynamics are free from faults in their respective channels, the estimation of speed \(\hat{\omega}_r\) is not influenced by fault in the speed sensor. One should also note that, because of the sensor noise and system parameters, there will be a bounded error between the measured and the estimated speeds. The threshold selection should also depend on the noise and the estimation error to avoid false alarms. For fault identification, we choose the speed index as

\[
\omega_{\text{ind}} = \omega_r - \hat{\omega}_r \tag{19}
\]

The index value in the absence of fault will be a bounded error that will be less than the threshold value \(\omega_{\text{th}}\). In the case of fault, the index value will be greater than the threshold and fault can be identified. A proper threshold selection is crucial to avoid false alarms. Therefore it is easy to identify the speed sensor fault by comparing the speed index \(\omega_{\text{ind}}\) with the threshold value \(\omega_{\text{th}}\). The complete flowchart for the speed sensor fault detection is depicted in Fig. 3b. In practice, the threshold value is chosen as the minimum operating condition of the drive [35]. The speed threshold \(\omega_{\text{th}}\) has to be set slightly smaller than the minimum operating speed.

5 Simulation results

Simulations are performed with the parameters of the three-phase 1/4-hp squirrel cage IM [22]. The specifications and parameters are given in Table 1. The IM drive is operated using direct field orientation (DFO). For the simulation, the drive is operated in speed control mode. The reference current in the \(d\)-axis, \(i_d^*\) is fixed to 0.9 A, and the reference current in the \(q\)-axis is generated by a speed PI controller. The block diagram for the field oriented control-based decoupled current control scheme is shown in Fig. 4. In which, the \(d\)-axis and \(q\)-axis current controllers based on proposed HOSM design are shown in Fig. 2.
To show the motor performance effectively, the motor is started towards a speed of 180 rpm, and after it has stabilised, at $t = 0.4 \text{s}$, the speed reference is changed to 540 rpm. The load torque is chosen as constant, $T_L = 0.1 \text{N.m}$. For the implementation of the HOSM controller, the sliding mode gains are selected as: $K_1 = 0.7, K_2 = 550$ and $K_3 = 70$. The threshold value of the speed for the given speed reference is selected to be $\omega_{\text{r,th}} = 50 \text{rpm}$.

### 5.1 Decoupled control with HOSM controller

Fig. 5a shows the waveforms of $i_d$ and $i_q$ against the simulated model currents $i_d^{\text{sim}}$ and $i_q^{\text{sim}}$. It is clear that the real currents accurately match with the model currents, which conforms the theoretical claim of the HOSM method that the addition of the sliding controllers (robust terms) will result in the disturbed system behave like the ideal system.

The bottom of Fig. 5b shows the HOSM outputs that are equal to the compensation voltages ($V_d^{\text{comp}}, V_q^{\text{comp}}$) produced with the classic decoupled-current control method in the top (Fig. 5b). The HOSM voltages obtained are similar in magnitude, have identical shapes and confirm the validity of the theoretical approach. The compensation voltages are only shown for comparison with the proposed approach.

Fig. 5c shows the IM fluxes and rotor position ($\theta_r$) in the stationary reference frame. The obtained fluxes are smooth and the rotor position is of good quality. Fig. 5d shows the real speed and the HOSM estimated speed. The estimated speed (17) from the HOSM matches well with the real-speed.

### 5.2 In the absence of speed sensor fault

Fig. 6a shows the performance of the fault detection algorithm under normal operation in the absence of speed sensor fault. The speed index is well below its threshold value, which indicates that there is no fault in the speed sensor. In the speed flag shown, the value of 1 signifies a fault and vice versa.

To test the robustness of the proposed scheme in real-applications, 10% random noise is added to the current sensor measurements. The results obtained are shown in Figs. 6b and c. The effect of noise can be observed in the speed estimate shown in Fig. 6c. The speed index and its flag in the presence of noise are shown in Fig. 6b. The speed index is well below the threshold value and no fault is detected. Fig. 6d shows the speed estimate of the IM in the presence of +10% variation of rotor resistance $R_r$. It can be seen that the rotor resistance variation has less affect on the performance of the proposed approach. In other words, we can say that the proposed controller is robust to bounded parametric uncertainties.
5.3 In the presence of speed sensor fault

To simulate for the sudden occurrence of speed sensor fault, the speed of the sensor is changed abruptly at \( t = 1.2 \) s during the transient period between 180 and 540 rpm as shown in Fig. 7d. Fig. 7a shows the real currents against the model currents in the presence of sudden speed sensor fault. It can be clearly observed that the real and model currents are affected when the speed measurement is faulty. This is because of the erroneous \( q \)-axis reference current used for the tracking. Fig. 7b shows the compensation voltages calculated using faulty speed measurement together with the estimated HOSM outputs shown in bottom. The HOSM outputs are affected by the faulty speed sensor because of the erroneous reference values of \((i_d, i_q)\).

Fig. 7c depicts the performance of the fault detection algorithm because of a sudden speed sensor fault. The speed index exceeds its threshold value after fault occurs, as shown in Fig. 7c. Therefore, a speed sensor fault is detected, as shown by the speed sensor fault flag in the bottom of Fig. 7c. The measured and the corresponding estimated speeds are shown in Fig. 7d. It should be noted that the measured rotor speed follows the reference speed because of the PI controller. The corresponding estimated speed is affected by the speed sensor fault, as it requires the \( q \)-axis reference in the calculation of speed estimation.

Moreover, an incipient fault is applied at \( t = 1.2 \) s to validate the robustness of the proposed approach as shown in Figs. 8a and 8b. The speed index exceeds its threshold value and the corresponding flag is shown in Fig. 8a. When the speed index exceeds its threshold value, sensor fault is detected. The measured and the estimated speeds are shown in Fig. 8b.

Further, to test the robustness of the proposed fault detection method, 10% random noise is added in the sensor measurements. The presence of measurement noise affects the estimation of unknown compensation voltages \((\Psi_q, \Psi_d)\). Moreover, the estimated speed is also affected by the random noise, as the speed estimation is a function of the unknown compensation voltages shown in Figs. 8c and 8d. The speed sensor fault detection in the presence of random noise is shown in Figs. 8c and 8d. In the same way, the speed index exceeds its threshold and fault flag is 1, when fault occurs. The measured and estimation of rotor speeds in the presence of fault are shown in Fig. 8d. However, the proposed approach remains robust even in the presence of 10% random noise and gives the similar performance.

6 Conclusion

This paper has presented a new decoupled current control method based on HOSM for IM. The proposed HOSM controller is designed to ensure the robust tracking performance, to reduce the chattering phenomenon, and to avoid the use of low-pass filtering. Based on the decoupled current control, the proposed HOSM controller also provides an accurate estimation of the rotor speed. Further, the measured and estimated speeds are employed for fault detection when the speed sensor is faulty. Simulation results are presented to highlight the performance of the proposed approach. The robustness of the proposed method in the presence of sensor noise is also shown.

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8 References


