MODELREFERENCEADAPTIVETECHNIQUEFORSENSORLESS
SPEED CONTROLOF INDUCTION MOTOR

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Abstract - This paper describes a Model Reference Adaptive System (MRAS) for speed control of the Induction Motor drive (IM) without a speed sensor. In this scheme an Adaptive Pseudoreduced-order Flux Observer (APFO) is used instead of the Adaptive Full-order Flux Observer (AFFO), an APFO is used for estimate the IM rotor speed and stator resistance, and these are used as feedback signals for the Field Oriented Control (FOC), which is a widely used control method for Induction Motor drive (IM). Simulation results show that the proposed scheme can estimate the motor speed under various adaptive PI gains and estimated speed can replace to measured speed in sensorless induction motor drives, this scheme is more efficient at very low speed, and also observed line currents, torque and speed under no-load and load conditions.

Keywords - Adaptive speed estimation, Induction Motor, Model reference adaptive control.

1. Introduction
Indirect field-oriented control (IFOC) method is widely used for IM drives. Within this scheme, a rotational transducer such as a tacho generator, an encoder, was often mounted on the IM shaft. However, a speed sensor cannot be mounted in some cases, such as motor drives in a hostile environment. Also such sensors lower the system reliability and require special attention to noise. Therefore, sensorless induction motor (IM) drives are widely used in industry for their reliability and flexibility, particularly in hostile environment [5]. Various sensorless field-oriented control (FOC) methods for induction motor (IM) drives have been proposed using software instead of hardware speed sensor [1-4, 7]. Adaptive full-order flux observers (AFFO) for estimating the speed of IM were developed using Popov’s and Lyapunov’s stability criteria [1,3,7]. While these schemes are not computationally intensive, an AFFO with a non-zero gain matrix may become unstable. However, large speed errors may occur under heavy loads and steady-state disturbances affecting light loads. An adaptive pseudoreduced-order flux observer (APFO) for sensorless FOC was proposed in using the Lyapunov’s method [2]. The performance of the estimator using APFO was shown to be superior compared to that using AFFO scheme only at medium speed.

In the MRAS-based technique for sensorless induction motor drives the rotor speed is estimated with an APFO and is used as the feedback signal for the FOC. The rotor flux is estimated through a closed-loop observer, thus eliminating the need for auxiliary variables related to the flux and need for the pure integration for flux calculations. As a result, the drive has a wider adjustable speed range and can be operated at zero and very low speeds.

2. Model Reference Adaptive System
The main drawback of this algorithm is its sensitivity to inaccuracies in the reference model, and difficulties of designing the adaptation mechanism block in MRAS. Selection of adaptive mechanism gains is a compromise between achieving a high speed of response and high robustness to noise and disturbances affecting the system. With the large PI gains for rotor speed identification in adaptive mechanism, \( k_p \) and \( k_i \) the convergence speed for speed estimation is fast; however, high order harmonic components and noises are present in the estimated speed.

3. Adaptive Flux Observer

For an induction motor, if the stator current \( i_s \) and rotor flux \( \Phi_r \) are selected as the state variables, the state equations can be expressed as eq.(1) in the stationary reference frame [1].

\[
\frac{d}{dt} [i_s, \Phi_r] = [A_{11} i_s, A_{12} i_s + B_1 v_s] + [B_0 v_s, 0] v_s = A x + B v_s
\]

Where \( i_s = [i_s, i_{qs}]^T \) is stator current
\[
A_{11} = \begin{pmatrix}
R_1 + \frac{1}{\sigma L_1} & \frac{1}{\sigma L_1} \\
\frac{1}{\sigma L_1} & \frac{1}{\sigma L_1}
\end{pmatrix} - \begin{pmatrix}
1 & 0 \\
0 & 1
\end{pmatrix}
\]
\[
A_{12} = \begin{pmatrix}
-\frac{1}{L_2} & \frac{1}{L_2}
\end{pmatrix}
\]
\[
B_1 = \begin{pmatrix}
\frac{1}{\sigma L_1}
\end{pmatrix}
\]
\[
v_s = [v_{ds}, v_{qs}] \text{ is stator voltage}
\]
\[
\sigma = \frac{L_2}{\sigma L_1 + \frac{1}{L_2}} \text{ is the leakage coefficient}
\]
\[
1 = \begin{pmatrix}
1 & 0 \\
0 & 1
\end{pmatrix}
\]
\[
& J = \begin{pmatrix}
0 & 1 \\
1 & 0
\end{pmatrix}
\]

Where \( R_1, R_2, \) and \( L_1, L_2 \) are stator and rotor resistances and self-inductances, respectively, \( L_m \) is mutual inductance, \( T_r \) is the rotor time constant \( L_2/R_2 \) and \( W_r \) is electrical motor angular speed.

The APFO flux observer can be written as follows

\[
\frac{di_s}{dt} = [A_{11} i_s + B_1 v_s] + (A_{12} i_s + B_1 v_s) + G(i_s - i_s)
\]
\[
\frac{d\Phi_r}{dt} = [A_{22} \Phi_r + B_{12} v_s] + G(\Phi_r - \Phi_r)
\]

Where \( i_s \) and \( v_s \) are measured values of stator current vector and stator voltage vector, respectively, \( G \) is the reduced-order observer gain matrix which is also determined to make eq.(3) stable and \( ^\wedge \) denotes the estimated values. The observer is a closed-loop system, which is obtained by driving the estimated model of the induction motor by the residual of the current measurement \( (e_{is}) \).

\[
e_{is} = i_s - \hat{i}_s
\]

The estimation of stator currents is conducted by a closed-loop observer with a \([2x2]\) feedback gain matrix \( G \), as in eq. (3), whereas the estimation of rotor fluxes is carried out by an open-loop observer of eq.(4) without the flux error. Therefore, the real and estimated rotor fluxes are assumed the same.

\[
\Phi_r = \hat{\Phi}_r
\]

Where the observer gain matrix \( G \) is calculated based on the pole placement technique.

\[
G = \begin{pmatrix}
g_1 & g_2 \\
g_2 & g_1
\end{pmatrix}^T
\]

Let us choose,

\[
g_1 = (k - 1) \sigma R_1 \\
g_2 = k_p, \quad k_p \geq 1
\]

Where \( g_1 \) is proportional to the IM parameters, \( g_2 \) is an arbitrary gain, \( k \) is an arbitrary positive constant value, and \( k_p \) is an arbitrary value \( (k_p > 1) \).

4. Adaptive Scheme for Speed Estimation

The error equation of state variables can be driven write from eq (1) and eq (3) as follows.

\[
\frac{de_{is}}{dt} = (A_{11} + G)e_{is} + \Delta A_{11} i_s + \Delta A_{12} \Phi_r
\]

Where \( W \) is the non linear block and is defined as follows:

\[
W = -\Delta A_{11} i_s - \Delta A_{12} \Phi_r
\]

The error matrix are represented by

\[
\Delta A_{1} = A_{11} - \bar{A}_{11} = -\Delta A_{11} \Delta R_1 = -\frac{\Delta R_1}{\sigma L_1} - \frac{\Delta R_1}{\sigma L_1} - \frac{\Delta R_1}{\sigma L_1}
\]

\[
\Delta A_{2} = A_{12} - \bar{A}_{12} = -\Delta A_{12} \Delta R_1 = -\frac{L_m}{\sigma L_1 L_2} \Delta R_1
\]

Where \( \Delta \omega_r \) is estimated speed error, and \( \Delta R_1 \) is the estimated error of stator resistance, using eq.(7), a mras representation of the system is shown in fig. 2, \( e_{is}(e_{is}), e_{is}(e_{is}) \) are the identifying mechanisms for the motor speed and stator resistance estimators[5], respectively, the system is hyper stable, if the forward path transfer matrix is strictly positive real and the input output of the nonlinear feedback block satisfies popov’s integrality of eq. (9) [6].

![Fig 2. MRAS representation for identifying the speed and stator resistance](image-url)

**This section describes the case where the primary resistance and rotor speed are set incorrectly. In such a case, the popov’s integral inequality as follows:**

\[
\int_{0}^{T} e_{is}^T W e_{is} dt \geq -\gamma^2
\]
Where $\rho = \frac{I_m}{\sigma_1 I_2}$ and $\tau^2$ is fine positive constant which is independent of $to$.

It is verified that the popov’s inequality of eq.(9) is satisfied if the estimate of the resistance is chosen to be a linear function of an inner product of the current estimate and the estimation error. The estimate of the rotor speed is chosen to be a linear function of an inner product of the current estimate and the estimation error. Starting in the form of a theorem we get;

Theorem: If the estimate of the stator resistance $R_1$ and the estimate of the speed $\dot{\omega}$ satisfy;

$$\dot{R}_1 = K_p^\prime \dot{I}_{is} + K_{i2} \int_0^t \dot{I}_{is} dt$$

$$\dot{\omega}_r = K_p^\prime \dot{I}_{is} \Phi_r + K_{i3} \int_0^t \dot{I}_{is} \Phi_r dt$$

Then popov’s criterion of eq.(9) will be satisfied $KP_2$, $Ki_2$, $kp_3$ and $Ki_3$ are the stator resistance and motor speed identification gains respectively.

5. Simulation Results

The basic configuration of speed estimation of sensorless induction motor drive is shown in Fig.3. All reference or command preset values are superscripted with a ‘*‘ in the diagram. IM speed will be estimated by eq. (11) and will be compared with the set point in order to create speed error. The error between the estimated and command values of speed drives the speed pi controller which in turn generates the required command value for the torque current component ($I_t$).

The current controlled voltage source inverter with field orientation control provides a fast time response and a smoother inverter current output. Although many current control algorithms have been proposed in recent years. Hysteresis band current control is still a preferred method. This algorithm is especially suitable for implementing the field orientation control. As a result this control algorithm offers a higher quality dynamical torque control.

Estimated rotor speed $\dot{\omega}_r$ and estimated rotor flux angle $\theta_r$ are achieved by the MRAS-based pseudoreduced- order flux observer. $I_m$ And $I_1$ are the magnetizing and torque components of the stator current, respectively. These components are the equivalent dc values in the synchronously rotating reference frame. By the application of inverse Clarke and park transformation in “Vector Rotator” block the command values will be compared with the measured or sensed currents $i_a$, $i_b$ and $i_c$ to generate proper pulsating sequence in order to fire the IGBT switching devices of the inverter.

For investigation of the MRAS’s behaviour underloading condition a load of 0.5 pu is applied to the IM at time 0.5s Fig.8 shows that after a small speed drop both estimated and real speeds converge very well. Meanwhile this simulation indicates that both speeds follow reference one with negligible error. This simulation results given better performance in both the cases i.e. under no-load condition which are shown in below

A. Case-1: No-Load Condition

Reference speed = 100 rad/sec

![Graph showing speed and current for no-load condition](image)

B. Case-2: Step Change in Load

Reference speed = 100 rad/sec; Load torque of 15 N-m is applied at $t = 0.25$ sec

![Graph showing speed, torque, and current for load change](image)
Fig. 4 shows that the actual speed of induction motor and estimated speed using MRAS are same. Fig. 5 shows no-load line currents, speed and torque wave forms under load condition. First the motor is started under no load and at t=0.25sec a load of 15n-M is applied. It can be seen that at t=0.25sec, the values for currents, torque will inc increase to meet the load demand and at the same time speed of the motor is slightly falls. The motor is started under no load condition and speed reversal command is applied at t=0.5sec. at 0.5sec the motor speed decays from 100 rad/sec and within 0.1sec it reaches its final steady state in the opposite direction. At 0.5sec torque will increase negatively and reaches to steady state speed value.

Fig. 8 (a) Show direct and quadrature axes currents (I_{dq}). From the graph it is observed that both currents are displaced by 90°. Hence the coupling effect can be eliminated.

VII. Conclusion

In this paper, Sensorless control of induction motor using Model Reference Adaptive System (MRAS) technique has been proposed. Sensorless control gives the benefits of Vector control without using any shaft encoder. In this thesis the principle of vector control and Sensorless control of induction motor is given elaborately. The mathematical model of the drive system has been developed and results have been simulated. Simulation results of Vector Control and Sensorless Control of induction motor using MRAS technique were carried out by using Matlab/Simulink and from the analysis of the simulation results, the transient and steady state performance of the drive have been presented and analyzed. From the simulation results, it can be observed that, in steady state there are ripples in torque wave and also the starting current is high. We can also increase the ruggedness of the motor as well as fast dynamic response can be achieved.

REFERENCES

