Sensorless Approach for Speed Control of Induction Motor using MRAS

Shrinivas P. Ganjewar and Chandulal guguloth

ABSTRACT: This paper presents a model reference adaptive system-based sensorless induction motor drive. In this scheme, an adaptive pseudoreduced-order flux observer is used instead of the adaptive full-order flux observer. Simulation results show that 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.

Index Terms—adaptive speed estimation, Induction motor, Model reference adaptive control.

I. INTRODUCTION

Indirect field-oriented control (IFOC) method is widely used for IM drives. By providing decoupling of torque and flux control demands, the vector control can navigate an AC motor drive similar to a separately excited DC motor drive without sacrificing the quality of the dynamic performance. Within this scheme, a rotational transducer such as a tachogenerator, an encoder or a resolver, 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 or high-speed motor drives. 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.

Various sensorless field-oriented control (FOC) methods for induction motor (IM) drives have been proposed using software instead of hardware speed sensor. Adaptive full-order flux observers (AFFO) for estimating the speed of IM were developed using Popov's and Lyapunov's stability criteria. While these schemes are not computationally intensive, an AFFO with a non-zero gain matrix may become unstable. The proportionality constant in the adaptive algorithm has to be adapted for different speeds. If the gain matrix of the AFFO is set to Zero, no adaptation is required. 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. The performance of the estimator using APFO was shown to be superior compared to that using AFFO scheme only at medium speed. Further, accuracy of speed estimation over the entire operating speed range strongly depends on this mechanical model and the IM torque estimation.

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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.

II. MODEL REFERENCE ADAPTIVE SYSTEM (MRAS):

The model reference adaptive system (MRAS) is one of the major approaches for adaptive control and is one of many promising techniques employed in adaptive control. Among various types of adaptive system configuration, MRAS is important since it leads to relatively easy to- implement systems with high speed of adaptation for a wide range of applications. The basic scheme of the MRAS given in Fig. 1 is called a parallel configuration (output error method) MRAS in order to differentiate it from other MRAS configurations where the relative placement of the reference model and of the adjustable system is not the same. The MRAS scheme presented above are characterized by the fact that the reference model was disposed in parallel with the adjustable system. The series–parallel configuration (equation error method) is used in general for parameter identification. In this configuration, the reference model is partitioned into two parts: one in series with the adjustable system and one in parallel with the adjustable model. The series configuration is also used for identification and is often called the input error method. In this scheme the reference and adjustable models are located in series.

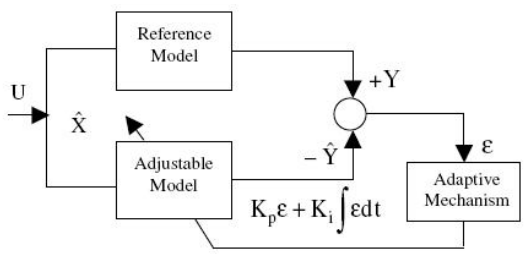


Fig. 1. Basic configuration of a parallel model reference adaptive system.

The use of parallel MRAS is determined by its excellent noise-rejection properties that allow obtaining unbiased parameter estimates, and in this scheme an error vector is derived using the difference between the outputs of two dynamic models, i.e. the reference and adjustable models, where only one of the models includes the estimated parameter as a system parameter, i.e. speed/resistance, and the inputs of two models are the same. The error vector, e, is driven to zero through an adaptive law. As a result, the estimated parameter, X1 , will converge to its true value X. One of the most noted advantages of this type of adaptive system is its high speed of adaptation. This is due to the fact that a measurement of the difference between the outputs of the reference model and adjustable model is obtained directly by the comparison of the states (or outputs) of the reference model with those of the adjustable system. The block ''reference model'' represents demanded dynamics of actual control loop. The block ''adjustable model'' has the same structure as the reference one, but with adjustable parameters instead of the unknown ones.

The main drawbacks of this algorithm are 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.

III. ADAPTIVE FLUX OBSERVER:

For an induction motor, if the stator current i_s and rotor flux Φ_r are selected as the state variables, the state equations can be expressed as (1) in the stationary reference Frame.

$$\frac{d}{dt}\begin{bmatrix} i_s \\ \phi_r \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix}\begin{bmatrix} i_s \\ \phi_r \end{bmatrix} + \begin{bmatrix} B_1 \\ 0 \end{bmatrix} v_s = Ax + Bv_s \tag{1}$$

$$i_s = Cx \tag{2}$$

where

$\boldsymbol{i}_s = [i_{ds}\ i_{qs}]^T$ is stator current

$\phi_r = [\phi_{dr}\ \phi_{qr}]^T$ is rotor flux

$\boldsymbol{v}_s = [v_{ds}\ v_{qs}]^T$ is stator voltage

$\boldsymbol{x} = [\boldsymbol{i}_s\ \phi_r]^T$

$A_{11} = -\{R_1/(\sigma L_1) + (1-\sigma)/(\sigma\tau_r)\}\boldsymbol{I} = a_{r11}\boldsymbol{I}$

$A_{12} = L_m/(\sigma L_1 L_2)\{(1/\tau_r)\boldsymbol{I} - \omega_r \boldsymbol{J}\} = a_{r12}\boldsymbol{I} + a_{i12}\boldsymbol{J}$

$A_{21} = (L_m/\tau_r)\boldsymbol{I} = a_{r21}\boldsymbol{I}$

$A_{22} = -(1/\tau_r)\boldsymbol{I} + \omega_r \boldsymbol{J}$

$B_1 = 1/(\sigma L_1)\boldsymbol{I}, \boldsymbol{C} = [\boldsymbol{I}\ 0]$

$\sigma = 1 - L_m^2/L_1 L_2$ is the inductance leakage coefficient,

$$\boldsymbol{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \text{ and } \boldsymbol{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$$

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

The APFO flux observer can be written as follows

$$\frac{d\hat{i}_s}{dt} = \hat{A}_{11}\hat{i}_s + \hat{A}_{12}\hat{\phi}_r + B_1 v_s + G(\hat{i}_s - i_s) \tag{3}$$

$$\frac{d\hat{\phi}_r}{dt} = \hat{A}_{22}\hat{\phi}_r + A_{21}\hat{i}_s \tag{4}$$

here 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 (3) stable and ^ 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[2 × 2] feedback gain matrix G, as in (3), whereas the estimation of rotor fluxes is carried out by an open-loop observer of (4) without the flux error. Therefore, the real and estimated rotor fluxes are assumed the same.

$$\phi_r = \hat{\phi}_r$$

The observer gain matrix is chosen as:

$$G = \begin{bmatrix} g_1 & g_2 \\ -g_2 & g_1 \end{bmatrix}^T \tag{5}$$

Where the observer gain matrix G is calculated based on the pole placement technique. The selection of the observer poles is a compromise between the rapidity of error responses and the sensitivity to disturbances and measurement noises. In practice, the eigenvalues of the observer are selected to be negative, so that the state of The observer will converge to the state of the observed system, and they are chosen to be somewhat more negative than the eigenvalues of the observed system so that convergence is faster than other system effects. Based on the above mentioned criteria

Let we chose,

$$g_1 = (k-1)a_{r11}$$
$$g_2 = k_p,\ k_p \geqslant 1 \tag{6}$$

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 ($k_p \geq -1$).

IV. Adaptive scheme for speed estimation:

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

$$\frac{de_{is}}{dt} = (A_{11}+G)e_{is} + \Delta A_1\hat{i}_s + \Delta A_2\hat{\phi}_r = (A_{11}+G)e_{is} - W \tag{7}$$

$$W = -\Delta A_1\hat{i}_s - \Delta A_2\hat{\phi}_r \tag{8}$$

Where W is the nonlinear block and is defined as:

W

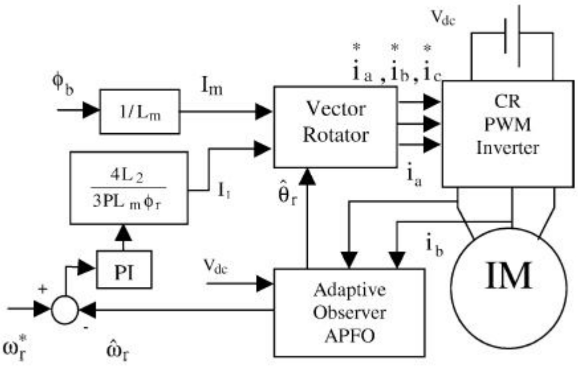
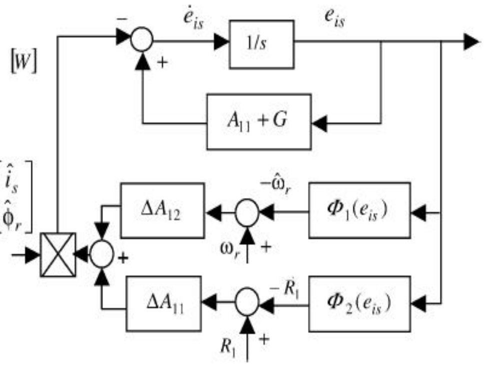


Fig. 3. Block diagram of sensorless IM drive.

The error matrices ΔA_1, and ΔA_2 are represented by:

Where $\Delta\omega_r$ is the estimated speed error, and ΔR_1 is the estimated error of stator resistance. Using (7), a MRAS representation of the system is shown in Fig. 2, which is constructed from a linear time-variant forward path transfer



(9)

Fig. 2. MRAS representation for identifying the speed and stator resistance.

matrix and a nonlinear feedback block. $\emptyset_1(e_{is})$ And $\emptyset_2(e_{is})$ are the identifying mechanisms for the motor speed and stator resistance estimators, respectively. The system is hyper stable, if the forward path transfer matrix is strictly positive real and the input and output of the nonlinear feedback block satisfies Popov's integral inequality of (9).

In this section we discuss the case where the primary resistance and rotor speed are set incorrectly. In such a case, the Popov's integral inequality as follows:

Where $\rho = L_m/\sigma L_1 L_2$ and γ^2 is a finite positive constant, which is independent of t_0.

It is verified that the Popov's inequality of (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 and the estimate of the rotor speed is chosen to be a linear function of an inner product of the flux estimate and the stator current estimation error. Stating in the form of a theorem we get:

V. SIMULATION RESULTS:

The basic configuration of speed estimation of sensorless induction motor drive is shown in Fig. 3. This configuration will be used for both simulation and experimentation. All reference or command preset values are superscripted with a ''*'' in the diagram. IM speed will be estimated by (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

$$\Delta A_1 = A_{11} - \widehat{A}_{11} = -\Delta A_{11}\Delta R_1 = \frac{\Delta R_1}{\sigma L_1} I$$

$$\Delta A_2 = A_{12} - \widehat{A}_{12} = \Delta A_{12}\Delta\omega_r = -\frac{L_m}{\sigma L_1 L_2} J\Delta\omega_r$$

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. It presents a simple and fast control as well as flexibility in maintaining the stator current in a predetermined specified. This algorithm is especially suitable for implementing the field orientation control. Hysteresis band current control can rapidly follow any dynamical current demand change in a shorter time span than any other control method. As a result, this control algorithm offers a higher quality dynamical torque control. Estimated rotor speed $\hat{\omega}_r$ and estimated rotor flux angle $\hat{\theta}_r$ are achieved by the MRAS-based pseudoreduced- order flux observer. I_m And I_t 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 transformations in ''Vector Rotator'' block, the command values i^*_a, i^*_b, and i^*_c can be obtained. These real time values will be compared with the measured or sensed currents i_a, i_b, and i_c to generate proper pulsing sequence in order to fire the IGBT switching devices of the inverter.

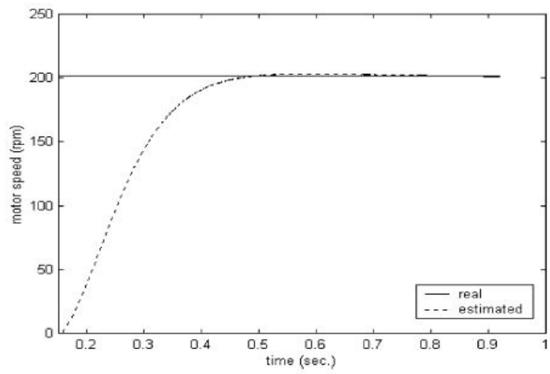


Fig. 4. Behaviour of speed estimation at $K_{p3} = 4$, $K_{i3} = 150$.

Figs. 4and 5 show the behavior of IM speed estimation under various values of adaptive scheme PI gains. These figures show that with the large PI gains for the adaptive scheme, Kp3 and Ki3, the convergence for the speed estimation is fast, however, a lot of high order harmonics are present in the estimated speed. In Fig. 4 where the IM rotates at a constant speed (200 rpm) under no load condition and initial value of the estimated speed is zero, the estimated speed reaches the real one in less than 0.35 s. With larger PI gains the convergence time reduces to 0.25 s as in Fig. 5.

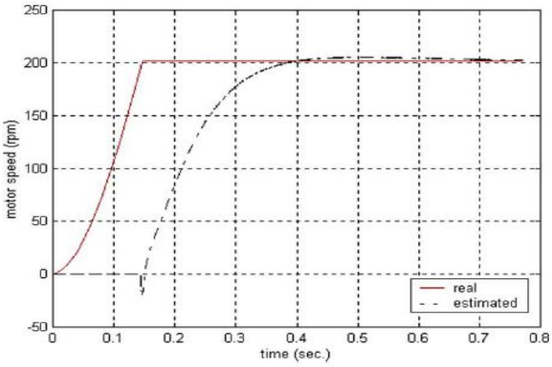


Fig. 5. Behaviour of speed estimation at $K_{p3} = 5$, $K_{i3} = 250$.

VI. CONCLUSION:

This paper presents a MRAS-based APFO sensorless induction motor drive. This method has been applied to a direct field-oriented induction motor control with and without speed sensors. The simulation results demonstrated that with larger PI gains for the adaptive PI regulators, the convergence for the speed estimation is fast, however, higher order harmonics and noises are included in the estimated speed. The validity of the MRAS-based pseudoreduced-order flux observer has been verified by simulation.

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