Sensorless Vector Controller for Induction Motor Drives with Parameter Identification

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Abstract - A sensorless vector controller for induction motor drives [1] based on the theory of Model Reference Adaptive System (MRAS) [2] has been presented, which identifies synchronous speed of motor drives using the error of reactive power as the reference input of adaptation mechanism. Moreover, as shown in [1] the MRAS-based synchronous speed identification scheme requires *neither* rotor time constant *nor* stator resistance in both adjustable and reference models. However, rotor resistance is still used for the estimation of shaft speed. In this paper, a stator resistance identification scheme based upon MRAS using the error of active power as the reference input of adaptation mechanism and details about the stability of the identification scheme will be presented. Moreover, the rotor resistance is adjusted using the variation rate of stator resistance [3]. Simulation and experimental results will be presented to confirm the theoretical analysis.

Nomenclature

X_a^u

U:

- α referring to a stationary frame linked to phase A of stator windings
- r referring to the rotor flux frame

D:

- d d-axis component
- q q-axis component
- s stator component
- r rotor component

x^*

reference or command value of x

$\hat{x}$

estimated (calculated) or feedback value of x

$\bar{x}$

conjugate of x

x_0

equilibrium point of x

Δx

deviation of x from the equilibrium point x_0

$\text{Re}(x)$

real part of x

$\text{Im}(x)$

imaginary part of x

Symbols

ω_e

synchronous speed

ω_r

rotor speed

ω_{sl}

slip frequency

p

differential operator

R_r

rotor resistance

R_s

stator resistance

L_r

rotor self-inductance

L_s

stator self-inductance

L_m

mutual inductance

σ

total leakage factor; $\sigma = 1 - (L_m^2 / L_s L_r)$

τ_r

rotor time constant; $\tau_r = L_r / R_r$

T_s

sampling period of digital implementation

J_m

inertial of motors

s

variable of the Laplace Transform

e_m'

back electromagnetic force (EMF)

Q

reactive power of machine

P

actual power of machine

$L_m^a = L_m^2 / (L_r \tau_r \sigma L_s)$

value of parameter

$e_R = e_m^a - \hat{e}_m^a$

state error of MRAS

$e_R' = p \hat{R}_s = -\text{Re}(\hat{e}_s^a)$

output error of MRAS

$e_R^* = P - P'$

output error of MRAS

$e_{\omega}^* = Q - Q'$

output error of MRAS

$v_s^r = v_{ds}^r + jv_{qs}^r$

stator voltage

$i_r^r = i_{dr}^r + ji_{qr}^r$

rotor current

$i_s^r = i_{ds}^r + ji_{qs}^r$

stator current

$i_m^r = \frac{L_r}{L_m} i_r^r + i_s^r$

magnetizing current

$\lambda_r^r = L_m i_m^r + L_r i_r^r$

rotor flux

$\lambda_s^r = L_s i_s^r + L_m i_m^r$

stator flux

I. INTRODUCTION

It is well known that the advantages of speed sensorless motor drives over conventional closed-loop system with shaft sensors include the reduction of cost and size of drives. A variety of methods have been proposed which heavily rely upon plant parameters. A comprehensive review of sensorless control of induction motor drives can be found in reference [4].

Sensorless induction motor control based upon the theory of Model Reference Adaptive System (MRAS) [2] provides an alternative way for the development. However, the method shown in [5] requires both rotor and stator resistors, and integral operation which involves initial value problem. Although, integration operation is not required in [6], rotor time constant which is sensitive to temperature variation, is still necessary for the adjustable model.

To deal with this problem, a sensorless controller [1] based on MRAS has been presented by the authors. The proposed MRAS scheme requires *neither* rotor time constant *nor* stator resistance in both adjustable and reference models. Moreover, *no* integral operation is required and therefore provides wider bandwidth for speed

control. However, as shown in [1] rotor resistance is still required for the estimation of *mechanical speed*. In this paper, a stator resistance identification scheme based upon the theory of MRAS using the error of *active power* as the reference input of adaptation mechanism is presented. Therefore, the rotor resistance is adjusted using the variation rate of stator resistance [3].

The paper is organized as follows. The theory for the development and stability analysis of the proposed MRAS-based identification scheme are presented, and followed by the simulation and experimental results confirming the theoretical analysis.

II. THEORY

A. MRAS-based Synchronous Speed Identification [1]

The motor voltage referring to rotor flux frame can be derived as follows.

$$v_s^r = R_s i_s^r + p \lambda_s^r + j \omega_e \lambda_s^r \quad (1)$$

$$v_{ds}^r = R_s i_{ds}^r + p \lambda_{ds}^r - \omega_e \lambda_{qs}^r \quad (2)$$

$$v_{qs}^r = R_s i_{qs}^r + p \lambda_{qs}^r + \omega_e \lambda_{ds}^r \quad (3)$$

Equations (2) and (3) are substituted into (4), which is the reactive power of machine, thereby resulting in (5).

$$Q = v_{qs}^r i_{ds}^r - v_{ds}^r i_{qs}^r \quad (4)$$

$$Q = \omega_e (\lambda_{qs}^r i_{qs}^r + \lambda_{ds}^r i_{ds}^r) + i_{ds}^r p \lambda_{qs}^r - i_{qs}^r p \lambda_{ds}^r \quad (5)$$

Substituting the stator flux equation shown in (6) into (5), the equation of reactive power shown in (5) can be rewritten as shown in (7).

$$\lambda_s^r = \lambda_{ds}^r + j \lambda_{qs}^r = L_s i_{ds}^r + j \sigma L_s i_{qs}^r \quad (6)$$

$$Q = \omega_e \left[\sigma L_s (i_{qs}^r)^2 + L_s (i_{ds}^r)^2 \right] + i_{ds}^r \sigma L_s p i_{qs}^r - i_{qs}^r L_s p i_{ds}^r \quad (7)$$

According to the theory of MRAS, (4) and (7) can be used as the reference and adjustable models, respectively, to identify the synchronous speed. The block diagram for the new MRAS synchronous speed identification scheme is shown in Fig. 1. Noting that the reactive power shown in (4) and (7) does *not* contain *either* rotor resistance *nor* stator resistance. Moreover, the calculations of reactive power for both reference and adjust models shown in (4) and (7), respectively do *not* require any integration operation.

It is important to note that the error signal, ε_ω'' , defined in Fig. 1, of the new MRAS synchronous speed identification can be derived from Fig. 3 and is shown as follows.

$$\varepsilon_\omega'' = i_{ds}^r v_{qs}^r - i_{qs}^r v_{ds}^r - \omega_e \left[\sigma L_s (i_{qs}^r)^2 + L_s (i_{ds}^r)^2 \right] - i_{ds}^r \sigma L_s p i_{qs}^r + i_{qs}^r L_s p i_{ds}^r \quad (8)$$

More details about the proof of stability of the MRAS-based synchronous speed identification scheme can be found in [1].

B. New MRAS-based Stator Resistance Identification

The active power of machine can be shown by (9) and rewritten as shown in (10) by substituting (2) and (3) into (9).

$$P = v_{ds}^r i_{ds}^r + v_{qs}^r i_{qs}^r \quad (9)$$

$$P = R_s [(i_{ds}^r)^2 + (i_{qs}^r)^2] + \omega_e (\lambda_{ds}^r i_{qs}^r - \lambda_{qs}^r i_{ds}^r) + i_{ds}^r p \lambda_{ds}^r + i_{qs}^r p \lambda_{qs}^r \quad (10)$$

Substituting (6) into (10), the active power can be derived as shown in (11).

$$P = R_s [(i_{ds}^r)^2 + (i_{qs}^r)^2] + \omega_e (1 - \sigma) L_s i_{ds}^r i_{qs}^r + \sigma L_s i_{qs}^r p i_{ds}^r \quad (11)$$

By the theory of MRAS [2], the stator resistance can be identified using (9) and (11) as reference and adjustable models respectively, as shown in Fig. 2.

C. Rotor Resistance Adaptation

It has been shown [7] that rotor resistance and shaft speed can not be estimated simultaneously without injecting additional exciting signal for inverter control. One of the simplest approaches to estimate both rotor resistance and shaft speed is to adjust rotor resistance using the variation rate of stator resistance [3] as shown in (12).

Fig. 1 MRAS-based synchronous speed identification scheme [1]

Fig. 2 New MRAS-based stator resistance identification scheme

$$\hat{R}_r = \frac{R_{ro}}{R_{so}} \hat{R}_s \quad (12)$$

where R_{r0} and R_{s0} are initial values of rotor and stator resistors respectively.

D. Stability Analysis

The voltage and flux equations referring to the stationary frame are as follows.

$$v_s^\alpha = R_r i_s^\alpha + p \lambda_s^\alpha \quad (13)$$

$$0 = R_r i_r^\alpha + p \lambda_r^\alpha + j \omega_r \lambda_r^\alpha \quad (14)$$

$$\lambda_s^\alpha = L_s i_s^\alpha + L_m i_r^\alpha \quad (15)$$

$$\lambda_r^\alpha = L_m i_s^\alpha + L_r i_r^\alpha = L_m i_m^\alpha \quad (16)$$

Substituting the flux equations shown in (15) and (16), into voltage equations, (13) and (14), the machine equations become

$$v_s^\alpha = R_s i_s^\alpha + \sigma L_s p i_s^\alpha + \frac{L_m^2}{L_r} p i_m^\alpha \quad (17)$$

$$\frac{L_m^2}{L_r} p i_m^\alpha = \frac{L_m^2}{L_r} \left[\frac{1}{\tau_r} (i_s^\alpha - i_m^\alpha) + j \omega_r i_m^\alpha \right] \quad (18)$$

Moreover, the associated equivalent circuit can be derived from (17) and (18), as shown in Fig. 3 and the induced electromagnetic force (EMF), $e_m^\alpha \equiv \frac{L_m^2}{L_r} p i_m^\alpha$, across the magnetizing inductance shown in Fig. 3 is defined as follows.

$$e_m^\alpha = v_s^\alpha - [R_s i_s^\alpha + \sigma L_s p i_s^\alpha] \quad (19)$$

$$e_m^\alpha = \frac{L_m^2}{L_r} \left[\frac{1}{\tau_r} (i_s^\alpha - i_m^\alpha) + j \omega_r i_m^\alpha \right] \quad (20)$$

The following equations can be derived from (19) and (20) respectively.

$$p i_s^\alpha = \frac{1}{\sigma L_s} (v_s^\alpha - R_s i_s^\alpha - e_m^\alpha) \quad (21)$$

$$\begin{aligned} p e_m^\alpha &= \frac{L_m^2}{L_r} \left[\frac{1}{\tau_r} (p i_s^\alpha - p i_m^\alpha) + j \omega_r p i_m^\alpha \right] \\ &= \frac{L_m^2}{L_r} \frac{1}{\tau_r} p i_s^\alpha - \frac{1}{\tau_r} e_m^\alpha + j \omega_r e_m^\alpha \end{aligned} \quad (22)$$

Moreover the derivative of the induced EMF can be derived by substituting (21) into (22) and is shown as follows.

$$p e_m^\alpha = -\left(L_m'' + \frac{1}{\tau_r}\right) e_m^\alpha + j \omega_r e_m^\alpha + L_m'' v_s^\alpha - L_m'' i_s^\alpha R_s \quad (23)$$

where $L_m'' = \frac{L_m^2}{L_r} \frac{1}{\tau_r} \frac{1}{\sigma L_s}$

Similar results for the estimated EMF, $\hat{e}_m^\alpha$, can also be derived by the same process and is shown as follows.

$$p \hat{e}_m^\alpha = -\left(L_m'' + \frac{1}{\tau_r}\right) \hat{e}_m^\alpha + j \omega_r \hat{e}_m^\alpha + L_m'' v_s^\alpha - L_m'' i_s^\alpha \hat{R}_s \quad (24)$$

Let $\varepsilon_R \equiv e_m - \hat{e}_m$, the state error equation, derived from (23) and (24), is given by

$$\begin{aligned} p \varepsilon_R &= -\left(L_m'' + \frac{1}{\tau_r}\right) \varepsilon_R + j \omega_r \varepsilon_R - L_m'' i_s^\alpha (R_s - \hat{R}_s) \\ &= A \varepsilon_R - W \end{aligned} \quad (25)$$

where

$$A = \left[-\left(L_m'' + \frac{1}{\tau_r}\right) + j \omega_r\right], \quad W = L_m'' i_s^\alpha (R_s - \hat{R}_s)$$

The equivalent nonlinear feedback system is shown in Fig. 4. As shown in Fig. 4, the stator resistance is identified using the adaptation mechanism and (25). Moreover, the system is hyperstable [2] if the linear time-invariant forward-path transfer matrix is strictly positive real and the nonlinear feedback block, including the adaptation mechanism, satisfies the Popov's criterion [8] as shown in (26).

$$\int_0^{t_1} \text{Re}(\varepsilon_R^T W) dt = \int_0^{t_1} (\varepsilon_{dr} i_{ds}^\alpha + \varepsilon_{qr} i_{qs}^\alpha) L_m'' (R_s - \hat{R}_s) dt \geq -r_o^2 \quad \forall t_1 \geq 0 \quad (26)$$

where r_o^2 is a positive constant.

By letting

$$\begin{aligned} f(t) &\equiv (R_s - \hat{R}_s) \\ \frac{df(t)}{dt} &\equiv (\varepsilon_{dr} i_{ds}^\alpha + \varepsilon_{qr} i_{qs}^\alpha) L_m'' \end{aligned} \quad (27)$$

and using the well-known inequality,

$$\int_0^{t_1} f(t) \frac{df(t)}{dt} dt = \frac{f^2(t_1) - f^2(0)}{2} \geq \frac{-f^2(0)}{2}$$

The Popov's inequality shown in (26) can be satisfied. Therefore the solution to the Popov's inequality can be derived from (27) and the stator resistance can be identified using adaptation mechanism referring to the synchronous frame fixed on the axis of rotor-flux frame, as shown in (28).

Fig. 3 Machine model referring to the stationary frame

$$\begin{aligned}
\hat{R}_s(t) &= \int_0^t p \hat{R}_s d\tau + \hat{R}_s(0) \\
&= (k_p + \frac{k_i}{p}) \left[-\text{Re}(L_m^* i_s^\alpha \bar{\epsilon}_R^\alpha) \right] \\
&= (k_p + \frac{k_i}{p}) \left\{ -\text{Re} \left[(L_m^* (i_s^r e^{j\delta}) (\bar{\epsilon}_R^r e^{-j\delta})) \right] \right\} \\
&= (k_p + \frac{k_i}{p}) \left[-\text{Re}(L_m^* i_s^r \bar{\epsilon}_R^r) \right]
\end{aligned} \quad (28)$$

Furthermore, it is possible to demonstrate that $[F(j\omega) + F^T(-j\omega)]$ is a strictly positive real Hermitian matrix, where $F(s) = [sI - A]^{-1}$.

It has been shown that [8] if all poles of $F(s)$ are on the left half plane of s-plane and $[F(j\omega) + \bar{F}^T(-j\omega)]$ is strictly positive definite Hermitian matrix, then the linear time-invariant forward-path transfer matrix shown in Fig. 5 is strictly positive real. As shown in Fig. 4 that these two conditions are satisfied since the linear time-invariant forward-path transfer matrix $F(s)$ is

$$F(s) = [sI - A]^{-1} = \frac{1}{s + \bar{L}_m + \frac{1}{\tau_r} - j\omega_r} \quad (29)$$

; $F(s)$ does not have any pole on the right half plane of s-plane and the imaginary axis. Moreover,

$$\begin{aligned}
&[F(j\omega) + \bar{F}^T(-j\omega)] \\
&= \frac{1}{j\omega + \bar{L}_m + \frac{1}{\tau_r} - j\omega_r} + \frac{1}{-j\omega + \bar{L}_m + \frac{1}{\tau_r} + j\omega_r} \\
&= \frac{2(L_m^* + \frac{1}{\tau_r})}{(L_m^* + \frac{1}{\tau_r})^2 + (\omega - \omega_r)^2}
\end{aligned} \quad (30)$$

and $[F(j\omega) + \bar{F}^T(-j\omega)]$ is an one by one square matrix with positive real entry which demonstrates that $[F(j\omega) + \bar{F}^T(-j\omega)]$ is strictly positive definite Hermitian matrix.

E. Effect of Parameter Variation

In this subsection, the effects of the variations parameters other than stator and rotor resistors, on the identification results of stator resistance and rotor resistance will be investigated.

Under steady state, i_{qs}^r and i_{ds}^r are regarded as constants, and the output quantity of the reference model shown in Fig. 2 is

$$P_0 = R_s [(i_{ds}^r)^2 + (i_{qs}^r)^2] + \omega_{e0} (1 - \sigma_0) L_{s0} i_{qs}^r i_{ds}^r \quad (31)$$

where σ_0 and L_{s0} nominated values of total leakage factor and stator self-inductance, respectively.

When the output of the adjustable model converges to P_0 , the identified stator resistance can be rewritten as

$$\hat{R}_s = \frac{P_0 - \omega_{e0} (1 - \sigma_0) L_{s0} i_{qs}^r i_{ds}^r}{(i_{ds}^r)^2 + (i_{qs}^r)^2} \quad (32)$$

The variations of stator resistance caused by parameter variations of stator self-inductance and total leakage factor can thus be derived and shown as (33) and (34), respectively.

$$\Delta \hat{R}_s = \frac{\partial \hat{R}_s}{\partial L_s} \Delta L_s = \frac{-\omega_{e0} (1 - \sigma_0) i_{qs}^r i_{ds}^r}{(i_{ds}^r)^2 + (i_{qs}^r)^2} \Delta L_s \quad (33)$$

$$\Delta \hat{R}_s = \frac{\partial \hat{R}_s}{\partial \sigma} \Delta \sigma = \frac{\omega_{e0} L_{s0} i_{qs}^r i_{ds}^r}{(i_{ds}^r)^2 + (i_{qs}^r)^2} \Delta \sigma \quad (34)$$

Fig. 4 MRAS representation of a non-linear feedback system

Fig.5 Influence of parameter variation on $\hat{R}_s$ estimation illustration of simulation results

(a) $L_s = 1.1 L_{s0}, T_e = 6 Nt - m$;

(b). $\sigma = 1.5\sigma_0$, $\omega_{e0} = 50(rad/sec)$

As shown in (33), the error caused by the variation of stator self-inductance is independent on the torque current as illustrated in Fig. 5(a) using $L_s = 1.1 L_{s0}$ as an example. As shown in Fig. 5(a) the parameter sensitivity to stator self inductance is proportional to the synchronous speed. In contrast, the stator resistance error caused by the variation of total leakage factor is dependent on the load condition. As the load increases, the stator resistance becomes less sensitive to total leakage factor as illustrated in Fig. 5(B) using $\sigma = 1.5 \sigma_0$ as an example.

F. Sensorless Vector Controller with Parameter Identification

The following equations for a vector-controlled induction motor drives can be derived under well-orientation condition.

$$\tau_r p |i_m^r| + |i_m^r| = i_{ds}^r \quad (35)$$

$$\omega_e = \omega_r + \frac{i_{qs}^r}{\tau_r |i_m^r|} \quad (36)$$

$$\omega_{sl} = \frac{i_{qs}^r}{\tau_r |i_m^r|} \quad (37)$$

$$|\lambda_r^r| = L_m |i_m^r| = \frac{L_m i_{ds}^r}{1 + \tau_{rs}} \quad (38)$$

Furthermore the associated block diagram of this so-called rotor flux model can be derived from (35)-(38) and is shown in Fig. 6.

Fig. 7 shows the system configuration of the sensorless vector controller with parameter identification. As shown in Fig. 7, only current sensors are required for the implementation.

Moreover, two MRAS-based identification schemes are used to identify the synchronous speed and stator resistance respectively, thereby calculating the shaft speed and adjusting the rotor resistance. Noting of course, to reduce the effect of harmonics caused by inverter PWM control, low-pass filters (LPF) are used for current measurement as shown in Fig. 7. The specifications for the

sensorless induction motor drive system are shown in Table 1.

III. SIMULATION AND EXPERIMENTAL RESULTS

A. Hardware configuration

The MRAS-based sensorless induction motor drive shown in Fig. 7 has been realized using a single-chip-based control board which consists of an INTEL 80196 MC single-chip micro-processor, protection circuits, current detection circuits, and decoding circuit. Fig. 8 shows the block diagram of hardware configuration. As shown in Fig. 8, the object code generated by the cross-assembler using personal computer is downloaded into the in-circuit emulator (ICE) which emulates the target system in real time for on-line debug.

One of the features of the 80196 MC single-chip micro-processor is its embedded PWM generator which generates the three-phase PWM gating signals according to the command voltage using traditional natural-sampled sinusoidal pulse-width modulation (SPWM) technique.

Fig. 6 Block diagram of rotor flux model under well orientation

Table 1 Specifications

cut-off frequency of LPF	15 Hz
Bandwidth of current loop	300 rad/sec
Bandwidth of flux controller	30 rad/sec
Bandwidth of speed loop	30 rad/sec

Fig. 7. System configuration of the vector controller based on the proposed MRAS schemes

Fig. 8. Hardware configuration of the implementation of the MRAS-based controller

Several protection circuits are included in the single-chip microprocessor board and will cause trip whenever one of the events, including over voltage, under voltage, regeneration voltage, over current, and over load, occurs.

The current detection circuits transform the load current signals into the acceptable voltage level, -0.5 $\sim$ $+5.5V$, of the embedded A/D converter of 80196 MC single-chip micro-processor. It is worthy of note that the sampling time of A/D is around 15μ sec/channel. Although the output pulses of incremental encoder of induction motor is not fed into the controller, a decoding circuit is built in the single-chip microprocessor board for measurement purpose and further applications. More details about the hardware can be found in [9].

Fig. 9. Illustration of experimental and simulation results, ± 200 rpm, 0.8 Nt-m; (a). speed response, simulation; (b). speed response, experiment; (c). reference and identified stator resistance, simulation; (d). reference and identified stator resistance, experiment; (e). current of phase "A", simulation; (e). current of phase "A", experiment.

B. Experimental and Simulation Results

Fig. 9 illustrates the simulation and experimental results using the case with a step change of stator resistance at $t = 1.5$ sec. and 6.5 sec. as an example. As shown in Fig. 9, the estimated values of rotor speed and stator resistance converge to their reference values asymptotically. Moreover, the simulation results agree with experimental results very well confirming the theoretical analysis.

IV. CONCLUSION

A sensorless vector controller for induction motor drives with *stator* and *rotor* resistance identification is presented, which identifies stator resistance based upon MRAS scheme using the error of *active* power and adjusts the rotor resistance using the variation rate of stator resistance. Simulation and experimental results are presented confirming the theoretical analysis.

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