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DSP-based Sequential Parameter Estimation of PWM Inverter-fed IPM Synchronous Machine for Auto-tuning Applications

Kyeong-Hwa Kim

Department of Electrical and Information Engineering
Seoul National University of Science and Technology, 232 Gongneung-ro,
Nowon-gu, Seoul, 139-743, Korea
k2h1@seoultech.ac.kr

Abstract

A DSP-based sequential estimation of electrical parameters of a PWM inverter-fed interior permanent magnet synchronous motor (IPMSM) drive for auto-tuning applications is presented. The flux linkage and stator resistance are considered as main parameters and sequentially estimated using the state observer and model reference adaptive control (MRAC) scheme. As compared with other adaptation schemes such as a recursive least square method (RLSM), the proposed MRAC-based sequential parameter estimation is not sensitive to noise. Also, it does not require a strict condition like the persistent excitation. Considering that the most servo system has a current reference in the step waveform, it can be a very effective way of estimating electrical parameters. The whole control system is implemented using the digital signal processor (DSP) TMS320F28335 based controller and the effectiveness is verified through the comparative simulations and experiments.

Keywords: *Sequential parameter estimation, MRAC, PWM Inverter, IPMSM Drive, Improved stability*

1. Introduction

In general, an interior permanent magnet synchronous motor (IPMSM) drive systems are faced with unavoidable and immeasurable disturbances or some parameter variations [1-2]. Coupling the load to the motor shaft may cause the variations of the inertia and friction coefficient besides the load variation. The flux linkage varies nonlinearly with the temperature rise, and also, with the external field produced by the stator current due to the nonlinear demagnetization characteristics of the magnet. The stator resistance varies according to operating temperature. When these controller parameters are mismatched with the real parameters, a satisfactory performance cannot be obtained [3-5].

To guarantee the robust response against the parameter variation, the controller parameters should be adaptively changed according to the variation of plant parameters. Many research works have been presented for a robust controller design of electric drive systems using adaptive control [4], fuzzy logic control [6], fuzzy adaptive PID speed control [7], and neural network observer [8]. However, these schemes require a quite complex controller design. Using an effective auto-tuning algorithm will be sufficient to achieve a robust control performance with a relatively simple algorithm. Furthermore, the exact information on the parameters is very important to implement a sensorless drive scheme [9] or to diagnose a fault in a drive system [10-12].

In an initial set-up of most servo controllers, a proper tuning process for the parameters and gains is needed since the motor parameters of drive systems may vary according to the load and operating conditions. For this tuning process, a time-consuming trial and error method has been often employed. Thus, executing an auto-tuning process in a short time at initial set-up of servo systems is highly desirable to estimate main motor parameters at one time.

This paper presents a digital signal processor (DSP) based sequential estimation of electrical parameters for a PWM Inverter-fed IPMSM drive where the flux linkage and stator resistance are sequentially estimated using the state observer and model reference adaptive control (MRAC) scheme. The simultaneous estimation of several motor parameters is generally not simple. To avoid the limitation of estimating parameters simultaneously in the conventional method, a sequential parameter estimation algorithm is proposed where dominant motor parameters are estimated one by one, based on the fixed scheduled automatic routine that only one parameter is estimated during each interval. Since various works that effectively estimate only the specific interested parameter of motors are well known in many literatures [2, 5], this can be an effective way of estimating main motor parameters effectively at one time.

As compared with other adaptation schemes such a recursive least square method (RLSM), the proposed MRAC-based parameter estimation is not sensitive to noise. Also, it does not require a strict condition like the persistent excitation [2]. Considering that the most servo control system has a current reference in the step waveform, it can be a very effective way of estimating electrical parameters. The whole control system is implemented by DSP TMS320F28335 for an IPMSM drive and the effectiveness is verified through the comparative simulations and experiments [13].

2. Modeling of an IPMSM and state observer

An IPMSM consists of permanent magnets buried in the rotor core and three phase stator windings. The voltage equations of an IPMSM in the synchronous reference frame are described as follows:

$$v_{qs} = R_s i_{qs} + L_{qs}\dot{i}_{qs} + L_{ds}\omega_r i_{ds} + \lambda_m \omega_r \, . \tag{1}$$

$$v_{ds} = R_s i_{ds} + L_{ds}\dot{i}_{ds} - L_{qs}\omega_r i_{qs} \, . \tag{2}$$

where R_s is the stator resistance, L_{qs} is the q-axis stator inductance, L_{ds} is the d-axis stator inductance, ω_r is the electrical rotor angular velocity, and λ_m is the flux linkage. Using i_{qs} and i_{ds} as state variables, the state equation of the IPMSM can be expressed as follows:

$$\dot{i}_s = Ai_s + Bv_s + d \, . \tag{3}$$

where $i_s = [i_{qs} \quad i_{ds}]^T$, $v_s = [v_{qs} \quad v_{ds}]^T$

$$A = \begin{pmatrix} -\dfrac{R_s}{L_{qs}} & -\dfrac{L_{ds}}{L_{qs}}\omega_r \\ \dfrac{L_{qs}}{L_{ds}}\omega_r & -\dfrac{R_s}{L_{ds}} \end{pmatrix}, \; B = \begin{pmatrix} \dfrac{1}{L_{qs}} & 0 \\ 0 & \dfrac{1}{L_{ds}} \end{pmatrix}, \; d = \begin{pmatrix} -\dfrac{\lambda_m}{L_{qs}}\omega_r \\ 0 \end{pmatrix}.$$

The full state current observer which estimates stator currents can be expressed as

$$\dot{\hat{i}}_s = \hat{A}\hat{i}_s + Bv_s + \hat{d} + G(\hat{i}_s - i_s) . \tag{4}$$

where " $\wedge$ " denotes the estimated quantities and G is an observer gain matrix. $\hat{A}$ and $\hat{d}$ are the matrices in which R_s and λ_m are replaced by the estimated values $\hat{R}_s$ and $\hat{\lambda}_m$, respectively.

3. Sequential parameter estimation of flux linkage and stator resistance

To estimate the stator resistance and flux linkage, an MRAC technique is used. Figure 1 shows the basic structure of an MRAC scheme. The state equation of the IPMSM is used for a reference model and the full state current observer is used for an adjustable model. The errors between the reference model and the adjustable model can be used to drive the adaptation mechanism and to update the parameters in the adjustable model in order to reduce the errors between model outputs.

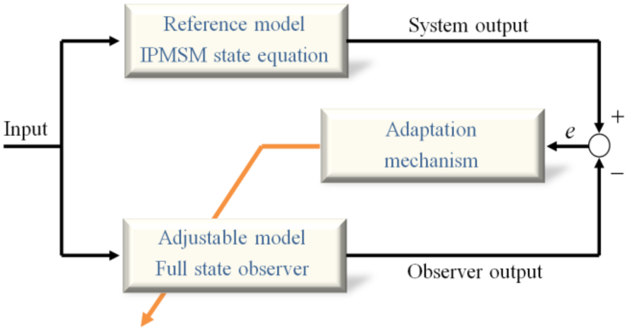


Figure 1. Basic structure of MRAC scheme

When $\hat{R}_s$ and $\hat{\lambda}_m$ are used in the current observer, the error dynamic equation can be obtained by subtracting (4) from (3) as follows:

$$\dot{e} = (A+G)e - W . \tag{5}$$

where $e = i_s - \hat{i}_s$ and W is a nonlinear time-varying block defined as

$$W = -\Delta A \cdot \hat{i}_s - \Delta d . \tag{6}$$

where ΔA and Δd are the error matrices caused by the parameter variations and can be expressed as

$$\Delta A = A - \hat{A} = \begin{pmatrix} -\frac{\Delta R_s}{L_{qs}} & 0 \\ 0 & -\frac{\Delta R_s}{L_{ds}} \end{pmatrix} = -B \cdot \Delta R_s . \tag{7}$$

$$\Delta d = d - \hat{d} = -\begin{pmatrix} \omega_r / L_{qs} \\ 0 \end{pmatrix} \Delta\lambda_m = -d_1 \Delta\lambda_m . \tag{8}$$

$$\Delta R_s = R_s - \hat{R}_s\,,\ \Delta\lambda_m = \lambda_m - \hat{\lambda}_m\,.$$

From (6), the adaptive law is defined as follows [14]:

$$\hat{R}_s(e,t) = \int_0^t \Phi_{R1}(e,t)d\tau + \Phi_{R2}(e,t) + \hat{R}_s(0)\,. \tag{9}$$

$$\hat{\lambda}_m(e,t) = \int_0^t \Phi_{\lambda 1}(e,t)d\tau + \Phi_{\lambda 2}(e,t) + \hat{\lambda}_m(0)\,. \tag{10}$$

where Φ_{R1}, Φ_{R2}, $\Phi_{\lambda 1}$, and $\Phi_{\lambda 2}$ are the adaptive mechanism, respectively, $\hat{R}_s(0)$ and $\hat{\lambda}_m(0)$ are the initial estimated value for the stator resistance and flux linkage. The design objective of the adaptive control for an asymptotic stability becomes as

1. $\lim\limits_{t\to\infty} e(t) = 0$ for any initial conditions $e(0)$, $\Delta R_s(0)$, and $\Delta\lambda_m(0)$

2. Find the adaptation rules which lead to $\lim\limits_{t\to\infty} \hat{R}_s = R_s$ and $\lim\limits_{t\to\infty} \hat{\lambda}_m = \lambda_m$.

Based on (6), the MRAC system is constructed as shown in Figure 2, which consists of a linear time invariant forward block and a nonlinear feedback block.

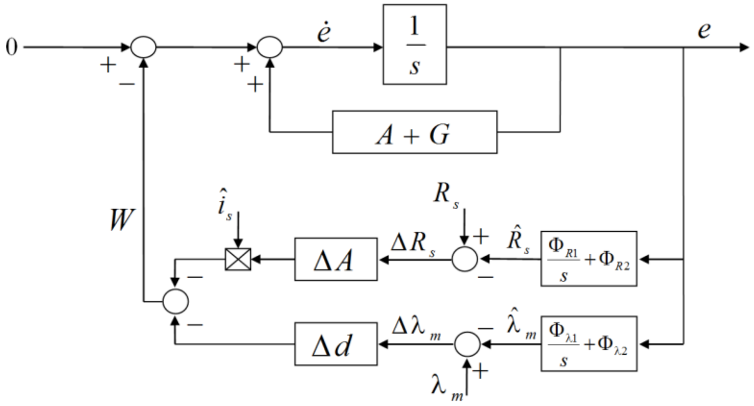


Figure 2. MRAC Structure for estimation of flux linkage and stator resistance

This system is hyperstable if the forward transfer function matrix is strictly positive real and the input-output inner product of nonlinear feedback block satisfies the Popov's integral inequality [14]. With the stable error dynamics, the forward transfer matrix $[sI-(A+G)]^{-1}$ is always strictly positive real. From the Popov's integral inequality, the stator resistance can be estimated as follows:

$$\hat{R}_s = -(k_{PR} + k_{IR}/s)\cdot(e^T B\cdot \hat{i}_s)\,. \tag{11}$$

where k_{PR} and k_{IR} are the PI gains for the stator resistance estimation. Similarly, the flux linkage can be estimated as follows:

$$\hat{\lambda}_m = -\left(k_{P\lambda} + k_{I\lambda}/s\right)\cdot(e_{qs}\omega_r)\,. \qquad \textbf{(12)}$$

where $k_{P\lambda}$ and $k_{I\lambda}$ are the PI gains for the flux linkage estimation.

As compared with other adaptation schemes such as the RLSM, this MRAC-based parameter estimation is not sensitive to noise. Also, it does not require a strict condition like the persistent excitation. Considering that the most servo control system has a current reference of step waveform, it can be an effective way of estimating parameters. When these estimated parameters converge to the real values, the closed loop observer error dynamics becomes

$$\dot{e} = (A+G)e \cong \begin{pmatrix} -k\dfrac{R_{so}}{L_{qs}} & 0 \\ 0 & -k\dfrac{R_{so}}{L_{ds}} \end{pmatrix}\cdot e\,. \qquad \textbf{(13)}$$

which is k-times faster than that of the IPMSM.

4. Simulations and experiments

The overall control system consists of the current controller, pulse width modulator, full state current observer, and parameter estimator. The current controller is achieved by using the synchronous PI control technique with the estimated parameters. Figure 3 shows the configuration of the experimental system.

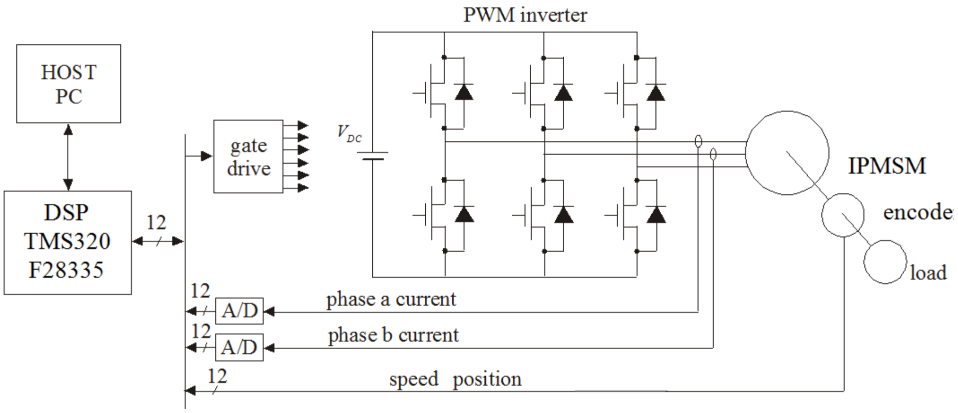
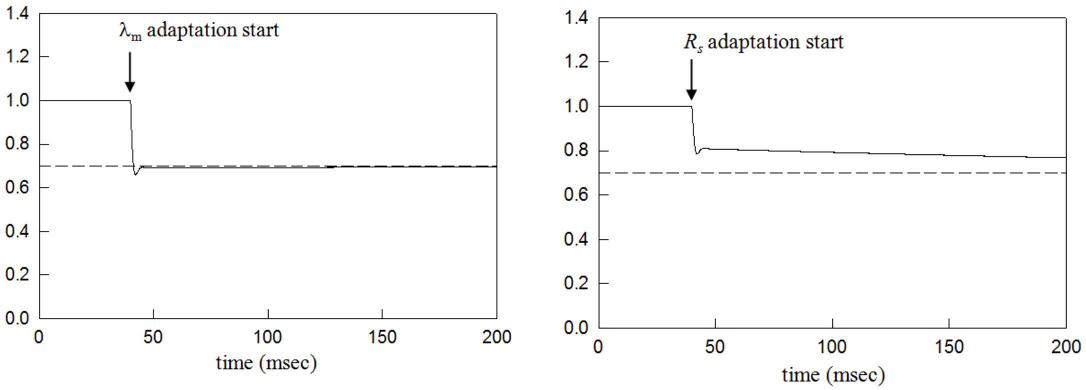


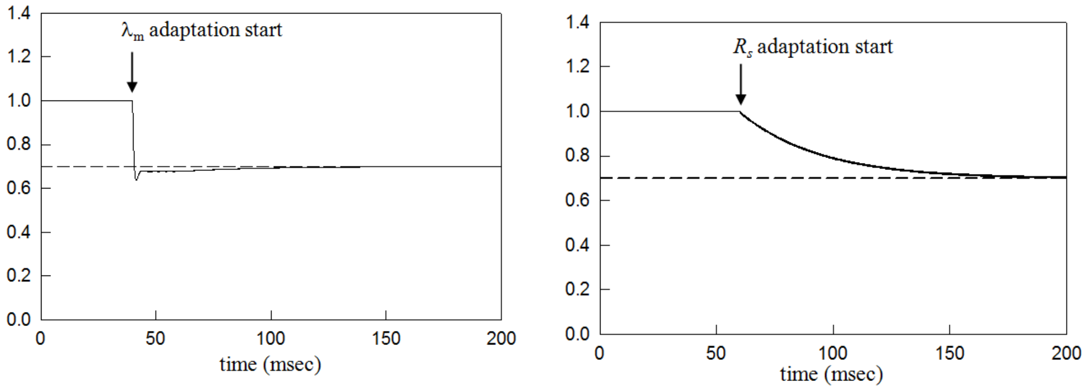
Figure 3. Configuration of the experimental system



(a) estimation of flux linkage (dashed: λ_m/λ_{mo}, solid: $\hat{\lambda}_m/\lambda_{mo}$)

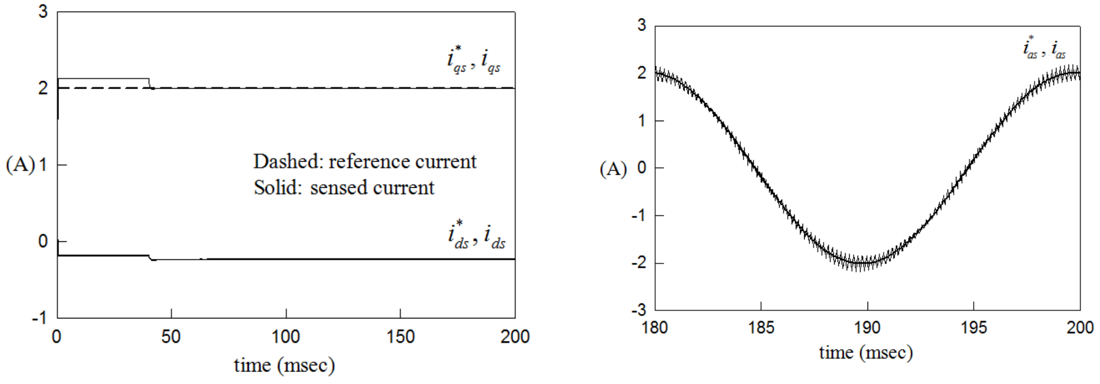
(b) estimation of stator resistance (dashed: R_s/R_{so}, solid: $\hat{R}_s/R_{so}$)

Figure 4. Performance of simultaneous parameter estimation with the initial parameter error of 30%



(a) estimation of flux linkage (dashed: λ_m/λ_{mo}, solid: $\hat{\lambda}_m/\lambda_{mo}$)

(b) estimation of stator resistance (dashed: R_s/R_{so}, solid: $\hat{R}_s/R_{so}$)



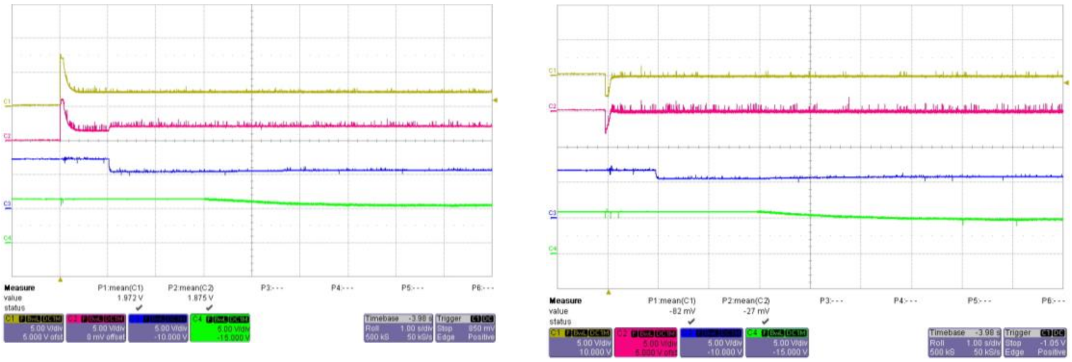
(c) response of the *q*-axis and *d*-axis currents

(d) *a*-phase current response at steady-state

Figure 5. Performance of the proposed sequential parameter estimation with the initial parameter error of 30%

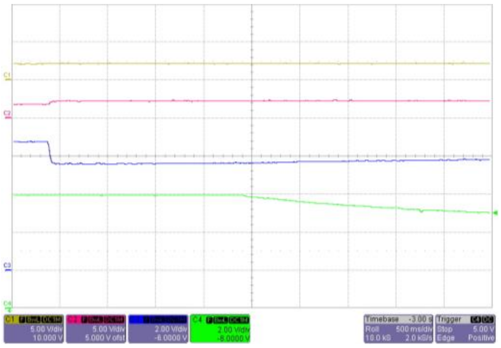
The whole control algorithms are implemented using 32-bit floating-point DSP TMS320F28335 with 150MHz clock [13]. The sampling period is set to 100 [μsec] both in the simulations and experiments, which yields a switching frequency of 10 kHz. The reference voltages are applied through the space vector PWM technique [15]. The rotor speed and absolute rotor position are detected through a rotary encoder and fed to the internal eQEP module within DSP, where the resolution of position is 12 bit/rev. The phase currents are detected by the Hall-effect devices and fed to DSP through the internal 12-bit A/D converters, where the resolution of current is $40/2^{11}$ [A].

Figure 4 shows the simultaneous estimation of the flux linkage and stator resistance with the initial parameter error of 30%. The estimation starts at 40 [msec] at the same time. While the estimation of the flux linkage shows a good convergence within a short period of time, the estimation of the stator resistance is not so good, taking much time for convergence.



(a) current and estimated parameters (time: 1 sec/div, C1: i_{qs}, 2 A/V, C2: $\hat{i}_{qs}$, 2A/V, C3: $\hat{\lambda}_m$, 0.003 Wb/V, C4: $\hat{R}_s$, 0.05 Ω/V)

(b) current and estimated parameters (time: 1 sec/div, C1: i_{ds}, 2 A/V, C2: $\hat{i}_{ds}$, 2 A/V, C3: $\hat{\lambda}_m$, 0.003 Wb/V, C4: $\hat{R}_s$, 0.05 Ω/V)



(c) enlarged waveform of (a) (time: 0.5 sec/div)

Figure 6. Experimental results for the proposed sequential parameter estimation scheme with the initial parameter error of 30%

Figure 5 shows the performance using the proposed sequential parameter estimation for the flux linkage and stator resistance with the initial parameter error of 30%. In this figure, the flux linkage is first estimated at 40 [msec] because it has a dominant influence on current control and observer. Once the flux linkage estimation is accomplished, the estimation of the stator resistance starts at 50 [msec]. Using the proposed scheme, a stable parameter estimating performance can be obtained without an estimation failure or any difficulty in choosing the estimation gains as in the simultaneous estimation technique.

Figure 6 shows the experimental results for the proposed sequential parameter estimation scheme with the initial parameter error of 30%. All results are well coincident with the simulation results in Figure 5.

5. Conclusions

This paper presents a sequential estimation of electrical parameters for a PWM inverter-fed IPMSM drive where the flux linkage and stator resistance are sequentially estimated using the state observer and MRAC scheme. Using this scheme, a stable parameter estimating performance can be obtained without an estimation failure or any difficulty in choosing the estimation gains as in the simultaneous estimation technique. The whole control system is implemented by TMS320F28335 for an IPMSM and the effectiveness is verified through the comparative simulations and experiments.

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Author

Kyeong-Hwa Kim was born in Seoul, Korea, in 1969. He received the B.S. degree from Hanyang University, Seoul, Korea, and the M.S. and Ph.D. degrees from KAIST, Taejon, Korea, in 1991, 1993, and 1998, respectively, all in electrical engineering. From 1998 to 2000, he was a Research Engineer with Samsung Electronics Company, where he was engaged in research and development of AC machine drive systems. From 2000 to 2002, he was a Research Professor with KAIST. Since August 2002, he has been with Seoul National University of Science and Technology, where he is currently an Associate Professor. His current research interests are in the areas of AC machine drive, control, diagnosis, power electronics, renewable energy, and DSP-based control applications. Prof. Kim is a member of the Korean Institute of Power Electronics (KIPE).