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적분형 가변구조 제어를 이용한 인버터 구동 유도전동기의 전류제어

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Current Control of Inverter-Fed Induction Motor Using Integral Variable Structure Control

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ABSTRACT

A new current control technique for inverter-fed vector controlled induction motor drives is presented. Using a integral variable structure control(IVSC) approach, the current controller is designed in the stationary rotating reference frame. The chattering reduction technique is also considered by the full state flux observer. By employing the proposed control scheme, a good control performance is achieved in the transient and steady state. The effectiveness of the proposed scheme is demonstrated through the comparative simulations.

I. INTRODUCTION

In order to achieve high performance torque control of induction motor drives by the field oriented control technique, it is required that the stator current controller has a good control performance in wide operating conditions. The PI control in the synchronous reference frame, called synchronous PI control, is widely used as one of the effective ways[4]. It is, however, known that the implementation of synchronous frame control is more complex than that of the stationary frame control, and furthermore the PI control does not provides the desirable performance in wide operating condition.

In this paper, a new stationary current control technique is proposed using the integral variable structure control(IVSC). In addition, the reduction of the chattering is considered by the feed forward control using the full state flux observer. Since it is well verified that the VSC has a robustness and fast tracking performance, the proposed control scheme provides a good transient and steady state performance in wide operating conditions. The effectiveness of the proposed control scheme is well demonstrated through the comparative simulations.

II. DESIGN OF IVSC FOR CURRENT CONTROL OF INDUCTION MOTOR

A. Modeling of induction motor

The state space model of the induction motor in the stationary reference frame is expressed as follows[1];

$$\begin{pmatrix} \dot{i}_s \\ \dot{\lambda}_r \end{pmatrix} = \begin{pmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{pmatrix} \begin{pmatrix} i_s \\ \lambda_r \end{pmatrix} + \begin{pmatrix} B_1 \\ 0 \end{pmatrix} v_s \tag{1.a}$$

$$i_s = C \cdot \begin{pmatrix} i_s \\ \lambda_r \end{pmatrix} \tag{1.b}$$

where

$$i_s = \begin{pmatrix} i_{qs} \\ i_{ds} \end{pmatrix},\ \lambda_r = \begin{pmatrix} \lambda_{qr} \\ \lambda_{dr} \end{pmatrix},\ v_s = \begin{pmatrix} v_q \\ v_d \end{pmatrix}$$

$$A_{11} = -\frac{1}{\sigma}\left(\frac{1}{\tau_s} + \frac{1-\sigma}{\tau_r}\right) \cdot I,\ A_{12} = \frac{1-\sigma}{\sigma L_m}\left(\frac{1}{\tau_r} I + \omega_r J\right)$$

$$A_{21} = \frac{L_m}{\tau_r} I,\ A_{22} = -\frac{1}{\tau_r} I + \omega_r J$$

$$B_1 = \frac{1}{\sigma L_s} I,\ \sigma = 1 - \frac{L_m^2}{L_s L_r},\ \tau_r = \frac{L_r}{R_r},\ \tau_r = \frac{L_s}{R_s}$$

$$I = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix},\ J = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix},\ C = [I\ \ 0].$$

The meanings of the parameters are as follows:

R_s, R_r	:	stator and rotor resistance
L_s, L_r	:	stator and rotor inductance
L_m	:	mutual inductance
σ	:	leakage coefficient
ω_r	:	electrical rotor angular velocity
v_{qs}, v_{ds}	:	q-axis and d-axis stator voltages
i_{qs}, i_{ds}	:	q-axis and d-axis stator currents
λ_{qr}, λ_{dr}	:	q-axis and d-axis rotor fluxes.

In (1), it can be regarded that the rotor flux is an unknown disturbance. Therefore, the first row of (1) is rewritten as follows:

$$\dot{i}_{qs} = a_q i_{qs} + b_q v_q + f_q \tag{2.a}$$

$$\dot{i}_{ds} = a_d i_{ds} + b_d v_d + f_d \tag{2.b}$$

where

$$\begin{pmatrix} f_q \\ f_d \end{pmatrix} = \begin{pmatrix} a_r \lambda_{qr} - a_i \lambda_{dr} \\ a_i \lambda_{qr} + a_r \lambda_{dr} \end{pmatrix}$$

$$a_q = a_d = -\frac{1}{s}\left(\frac{1}{t_s} + \frac{1-s}{t_r}\right),\ b_q = b_d = \frac{1}{\sigma L_s}$$

$$a_r = \frac{1-\sigma}{\sigma L_m}\left(\frac{1}{\tau_r}\right),\ a_r = \frac{1-\sigma}{\sigma L_m}\omega_r.$$

B. Design of sliding surface

The sliding surface of the proposed scheme is given as[6][7]

$$s = X + c\int_{-\infty}^{t} X d\tau = 0 \tag{3}$$

where

$$s = \begin{pmatrix} s_q \\ s_d \end{pmatrix},\quad c = \begin{pmatrix} c_q & 0 \\ 0 & c_d \end{pmatrix},\quad X = \begin{pmatrix} X_q \\ X_d \end{pmatrix} = \begin{pmatrix} i_{qs} - i_{qs}^{*} \\ i_{ds} - i_{ds}^{*} \end{pmatrix}$$

and the superscript ' * ' denotes the reference value. In (3), the design parameters of the sliding surface are the coefficient of the sliding surface c and the initial conditions of the integrator which is defined by

$$I_0 = \int_{-\infty}^{0} X d\tau\ . \tag{4}$$

Based on the equivalent condition ($s = 0, \dot{s} = 0$), the dynamic characteristics of the system can be represented as

$$\dot{X} = -cX. \tag{5}$$

In (5), the state X exponentially tends to zero with time constant $1/c$. Therefore, c is chosen to be satisfy the exponential stability by the pole placement method.

With different from the conventional VSC, the determination of I_0 is very important in the IVSC. By properly choosing I_0, the complete robustness is guaranteed without the reaching problem. From (3), the initial condition of the integrator to achieve the sliding mode at $t = 0$ is derived as

$$I_0 = -c^{-1}X(0) \tag{6}$$

where $X(0)$ is the initial condition of the state X.

C. Choice of control input

The input structure of the proposed scheme can be given as follows:

$$u = \Delta u + u_{ff}. \tag{7}$$

where

$$u = \begin{pmatrix} v_q \\ v_d \end{pmatrix},\quad \Delta u = \begin{pmatrix} \Delta v_q \\ \Delta v_d \end{pmatrix},\quad u_{ff} = \begin{pmatrix} v_{qff} \\ v_{dff} \end{pmatrix}.$$

The switching control input Δu is can be designed as the following structure:

$$\Delta v_q = \Psi_{q1} X_q + \Psi_{q2} \tag{8.a}$$

$$\Delta v_d = \Psi_{d1} X_d + \Psi_{d2} \tag{8.b}$$

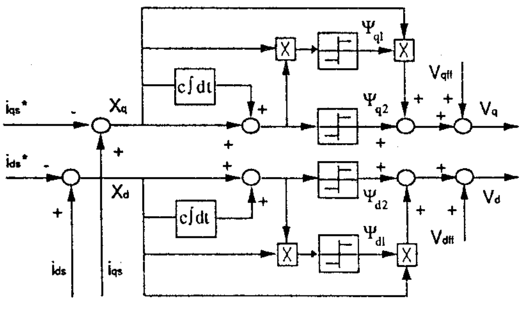

Fig. 1 Control input of the proposed scheme

In (8), the gains of the switching control input are determined by the sliding mode existence condition which is given as follows:

$$s_q \dot{s}_q < 0,\quad s_d \dot{s}_d < 0. \tag{9}$$

From (10), the gains of the switching control input are derived as follows:

$$\Psi_{q1} = \begin{cases} \alpha_{q1} > \max[-(a_q + c_q)/b_q] & : \text{if } s_q X_q < 0 \\ \beta_{q1} < \min[-(a_q + c_q)/b_q] & : \text{if } s_q X_q > 0 \end{cases} \tag{10.a}$$

$$\Psi_{q2} = \begin{cases} \alpha_{q2} > \max[-(a_r\lambda_{qr} - a_i\lambda_{dr})/b_q] & : \text{if } s_q < 0 \\ \beta_{q2} < \min[-(a_r\lambda_{qr} - a_i\lambda_{dr})/b_q] & : \text{if } s_q > 0 \end{cases} \tag{10.b}$$

$$\Psi_{d1} = \begin{cases} \alpha_{d1} > \max[-(a_d + c_d)/b_d] & : \text{if } s_d X_d < 0 \\ \beta_{d1} < \min[-(a_d + c_d)/b_d] & : \text{if } s_d X_d > 0 \end{cases} \tag{10.c}$$

$$\Psi_{d2} = \begin{cases} \alpha_{d2} > \max[-(a_i\lambda_{qr} + a_r\lambda_{dr})/b_d] & : \text{if } s_d < 0 \\ \beta_{d2} < \min[-(a_i\lambda_{qr} + a_r\lambda_{dr})/b_d] & : \text{if } s_d > 0 \end{cases} \tag{10.d}$$

The feed forward control input u_{ff} is obtained from the full state flux observer, which is discussed in later section. If the estimated flux is assumed as $\hat{\lambda}_{qr}$ and $\hat{\lambda}_{dr}$, the u_{ff} is given as

$$v_{qff} = -(a_r\hat{\lambda}_{qr} - a_i\hat{\lambda}_{dr})/b_q \tag{11.a}$$

$$v_{dff} = -(a_i\hat{\lambda}_{qr} + a_r\hat{\lambda}_{dr})/b_d \tag{11.b}$$

Hereafter, the symbol '^' denotes the estimated value. Fig.1 shows the block diagram of the proposed control scheme.

D. Chattering reduction by the flux observer

By the feed-forward control of the estimated rotor flux, the input chattering can be reduced. The full state observer to estimate the rotor flux is given as follows[3]:

$$\begin{pmatrix} \dot{\hat{i}}_s \\ \dot{\hat{i}}_r \end{pmatrix} = \begin{pmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{pmatrix}\begin{pmatrix} \hat{i}_s \\ \hat{i}_r \end{pmatrix} + \begin{pmatrix} B_1 \\ 0 \end{pmatrix} v_s + G(i_s - \hat{i}_s) \tag{12}$$

where the G is the gain matrix of the flux observer. In (12), the error dynamic equation of the flux observer is derived as

$$\dot{e} = (A - GC)\cdot e \tag{13}$$

where the A is the system matrix in (1) and the estimation error e is defined by

$$e = \begin{pmatrix} i_s & - & \hat{i}_s \\ \lambda_r & - & \hat{\lambda}_r \end{pmatrix}.$$

In (13), the closed loop poles of the flux observer is chosen by the pole placement technique. With choice of the gain matrix G as given in (14), it is possible to locate the poles at any specified complex conjugate pairs in the s domain.

$$G = \begin{pmatrix} g_1 I + g_2 J \\ g_3 + g_4 J \end{pmatrix} = \begin{pmatrix} g_1 & -g_2 \\ g_2 & g_1 \\ g_3 & -g_4 \\ g_4 & g_3 \end{pmatrix} \tag{14}$$

III. SIMULATIONS

The simulations are carried out to the induction motor with the following parameters:

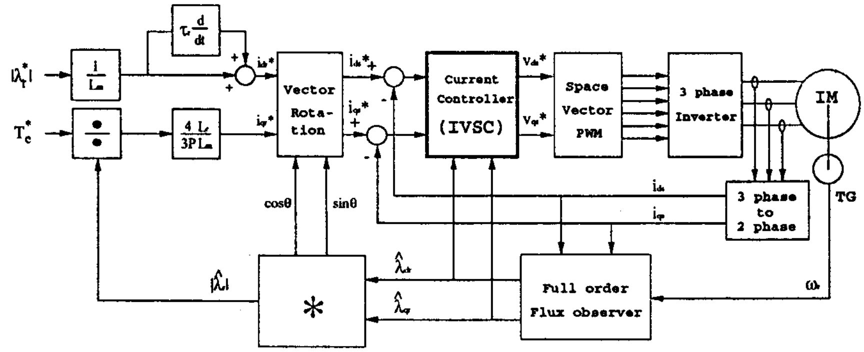

Fig. 2 Overall block diagram of the direct vector control system with the proposed current control scheme

$r_s = 1.0\ [\Omega]$, $r_r = 0.7[\Omega]$

$L_s = 116[\text{mH}]$, $L_r = 116[\text{mH}]$, $L_m = 100[\text{mH}]$.

The space vector PWM is used as a PWM technique in the proposed control scheme. The overall configuration of the direct field oriented control system with the proposed current control is shown in Fig. 2. Fig. 3 shows the comparative simulation results. In the steady state, there exist the undesirable phase delay and steady state error in the stationary PI control. Whereas, the synchronous PI control and proposed control provide a good steady state performance. In transient state, the proposed controller has the best tracking performance. Moreover, since the proposed controller is a stationary frame controller, its implementation is more simple than that of the synchronous PI controller.

IV. CONCLUSIONS

Using the IVSC, a simple and high performance current controller is designed in the stationary reference frame. The reduction of the chattering is also achieved using the full state flux observer. Through the comparative simulation studies, it is shown that the proposed current control has a better transient and steady state performance than the PI control. Furthermore, in view of implementation, the proposed controller is more simple than that of the synchronous PI control.

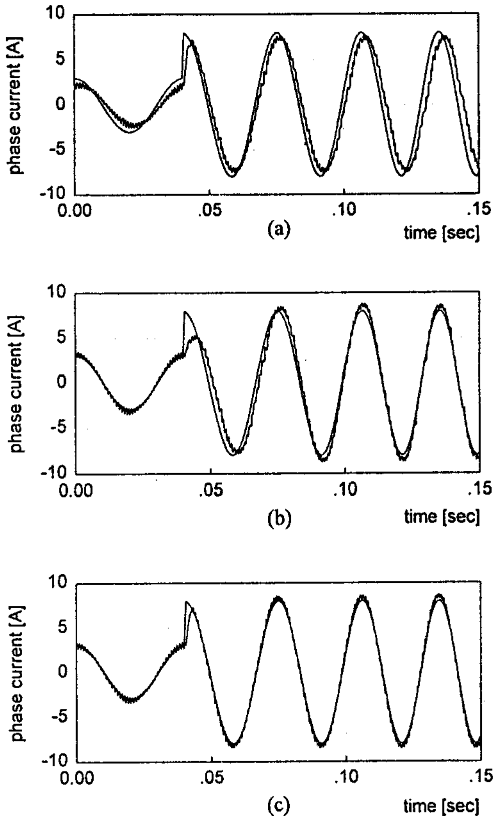

Fig. 3 Simulation results (a) Staionary PI control (b) Synchronous PI control (c) Proposed control

REFERENCES

[1] P.C. Krause, *Analysis of Electric Machinary*, McGraw-Hill, 1986.

[2] V.I. Utkin, *Sliding Modes and Their Applications in Variable Structure Approach*, MIR, 1978.

[3] G.F. Franklin, J.D. Powell, and A. Emami-Naeini, *Feedback Control of Dynamic Systems*, Addison Wesley, 1986.

[4] T.M. Rowan and R.J. Kerkman, "A new synchronous current regulator and an analysis of current regulated PWM inverters," IEEE Trans. Ind. Appl., Vol. IA-22, No. 4, pp. 678-690, 1986.

[5] D.C. Lee, S.K. Sul, and M.H. Park, "High performance current regulator for a field oriented controlled induction motor drives," IEEE IAS Conf. Rec., pp. 538-544. 1992.

[6] T.L. Chern and Y.C. Wu, 'Design of brushless DC position servo systems using integral variable structure approach," IEE proc.-B, Vol. 140, No. 1, pp. 27-34, 1993.

[7] G.T. Park, J.K. Choi, and D.S. Kim, "Design of variable structure control system having the first order dynamics," Trans. of KIEE, Vol. 41, No. 4, pp. 392-399, 1992.