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Robust Adaptive Precision Position Control of PMSM

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ABSTRACT

A new control method for precision robust position control of a permanent magnet synchronous motor (PMSM) is presented. In direct drive motor systems, a load torque disturbance directly affects the motor shaft. The application of the load torque observer is published in [1] using a fixed gain to solve this problem. However, the motor flux linkage cannot be determined precisely for a load torque observer. Therefore, an asymptotically stable adaptive observer base on a deadbeat observer is considered to overcome the problems of unknown parameters, torque disturbance and a small chattering effect. To find the critical parameters the system stability analysis is carried out using the Liapunov stability theorem.

Keywords: Adaptive control, Precision position control, PMSM control

1. Introduction

Recently, precision position control has become more important in chip mount machines, semiconductor production machines, precision milling machines, high resolution CNC machines, precision assembly robots, high speed hard disk drivers and other areas. Additionally, it is also important for direct drive systems. BLDC motors have replaced many DC motors since the industry applications require more powerful actuators in smaller sizes. The advantage of using a BLDC motor is that it can be manipulated to have similar speed-torque characteristics of a permanent magnet DC motor. Furthermore, the BLDC motor has low inertia, large power-to-volume ratio, and low noise compared to a permanent magnet DC servomotor having the same output rating [2]. However, the disadvantages of the BLDC motor are high cost and the need for a more complex controller because of the nonlinear characteristic [3]. The P-I (proportional-integral) controller usually used in BLDC motor control is simple to realize but difficult in obtaining sufficiently high performance in the tracking application.

It is, however, well known that the tracking controller problem can be simply solved by using state variable feedback with the augmentation of the state variables using the output error [4,5]. For the unknown and inaccessible inputs, the observer technique was studied by [6]. Also, the DC motor application of the load torque observer with fixed gain was published [7]. In this case, noisy speed information was used and the analysis was done by the in-out model. A new systematic approach was done in state space using digital position information in the BLDC motor system [1]. However, the machine flux linkage was not exactly known for a load torque observer creates problems of uncertainty. The cogging effect, damage on a permanent magnet over current, can affect the value of k_t. This caused a small position or speed errors and increased the chattering effect, which should be

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reduced as much as possible. Therefore, a model reference adaptive observer based on dead-beat control was considered to overcome the problem of the unknown parameters [8]. However, that case was limited to bounded input and bounded output [BIBO] stability. To reduce the position error, a new asymptotically stable load torque is needed. This paper proposes a deadbeat torque to solve this problem with an adaptive (DTOWA) algorithm. The comparison between the responses of two systems, which are fixed gain and DTOWA gain, has been done in detail. A directly coupled inertial load made of copper is used as load.

2. MODELING OF BLDC MOTOR

Generally, the behavior of a small horsepower BLDC motor used for position control is essentially the same as the permanent magnet synchronous machine. By means of field-oriented control, it is possible to make i_{ds} become zero [3]. The vector control method is used to align the rotor frame to the stator frame. In this case absolute position information is needed. Therefore, the system equations of a BLDC motor model can be described as:

$$\dot{\omega}_r = \frac{3}{2}\frac{1}{J}\left(\frac{p}{2}\right)^2 \lambda_m i_{qs} - \frac{B}{J}\omega_r - \frac{p}{2J}T_L \quad (1)$$

$$T_e = \frac{3}{2}\frac{p}{2}\lambda_m i_{qs} \quad (2)$$

$$= k_t i_{qs} \quad (3)$$

$$\dot{y} = \omega_r \quad (4)$$

where, $k_t = \frac{3}{2}\frac{p}{2}\lambda_m$. For the implementation of the field-oriented control, each of the three phase current control commands must be generated separately.

3. CONTROL ALGORITHM

The control reference in the drive is a step value as in a tracking servo problem. The dynamic equation of a given system can be expressed as follows:

$$\dot{x} = Ax(t) + bu(t) \quad (5)$$

$$y = cx(t) \quad (6)$$

where,

the dimensions of the matrices A, b and **c** are $n \times n$, $n \times 1$ and $1 \times n$, respectively. Usually, a linear quadratic controller is used to solve the regulator problem resulting in a state variable feedback. A new state is defined for the tracking controller as $\dot{z} = y - y_r$ where y_r is the rotor position reference [5].

The control input is calculated as: $u = -kx - k_1 z$. From this equation, the optimal control law minimizing the performance index with the weighting matrices Q and R of nominal values can obtain the state feedback controller gain. However, a large feedback gain, which results in a very large current command, is needed for a fast reduction of the error caused by the disturbance. Generally, the measured current is much too noisy to be used in the digital controller or the observer. If the load torque T_L is known, an equivalent current command i_{qc2} can be expressed as $T_L = k_t i_{qc2}$.

The feed forward equivalent q-axis current command to the output controller compensates the load torque effect. However, in the real system, there are many cases where some of the input is unknown or inaccessible. For simplicity, a 0-observer is selected. Thus, T_L can be considered as an unknown and assumed to be a constant. The system equation can be expressed as:

$$\begin{pmatrix} \dot{\hat{\omega}} \\ \dot{\hat{y}} \\ \dot{\hat{T}}_L \end{pmatrix} = \begin{pmatrix} -\frac{B}{J} & 0 & -\frac{p}{2}\frac{1}{J} \\ 1 & 0 & 0 \\ 0 & 0 & 0 \end{pmatrix} \begin{pmatrix} \hat{\omega} \\ \hat{y} \\ \hat{T}_L \end{pmatrix} + \begin{pmatrix} k_t \frac{p}{2}\frac{1}{J} \\ 0 \\ 0 \end{pmatrix} i_{qs} + L\left(y - \begin{pmatrix} 0 & 1 & 0 \end{pmatrix} \begin{pmatrix} \hat{\omega} \\ \hat{y} \\ \hat{T}_L \end{pmatrix} \right) \quad (7)$$

To guarantee the time required for calculating the load torque to be less than the overall system response time and to compensate the load torque at a transient state, a deadbeat observer is desirable. It then follows from the Cayley-Hamilton theorem that $\Phi_c^n = 0$ where, $\Phi_c = \hat{\Phi} - \hat{L}c$. The pole placement using Ackermann's formula can obtain the feedback gain L as follows: $L = P(\Phi)W_o^{-1}[0 \;\; 0 \;\; .... \;\; 1]$ where c, $P(\Phi) = \Phi_c^n$, and W_o are output, polynomial, and observable matrix, respectively.

Even though the observer feedback gain is obtained by using the nominal parameter value, there is a certain variation or uncertainty of the parameter, such as the machine flux linkage. To overcome this problem, an adaptive observer is considered [10].

In this system, the reference model is the real plant with the augmented state variable feedback controller. In a similar way, the adaptive load torque observer is considered as an adjustable system. The augmented system can be partitioned as follows:

$$\dot{x}_1 = Ax_1 + B_1 u - B_2 T_L \tag{8}$$

$$\dot{z} = c_1 x_1 - y_r \tag{9}$$

where,

$$u = -K_1 x_1 - k_2 z + k_3 \hat{T}_L$$

$$x_1 = \begin{pmatrix} \omega_r \\ y \end{pmatrix} \quad A = \begin{pmatrix} -\frac{B}{J} & 0 \\ 1 & 0 \end{pmatrix} \quad B_1 = \begin{pmatrix} k_t \frac{p}{2J} \\ 0 \end{pmatrix} \quad B_2 = \begin{pmatrix} \frac{p}{2J} \\ 0 \end{pmatrix} \quad c_1 = \begin{pmatrix} 0 & 1 \end{pmatrix}$$

$K_1, k_2, \text{and } k_3$ are $1 \times n$ vector and two scalars, respectively. Similarly, the adaptive torque observer of (7) is described by

$$\dot{\hat{x}}_1 = A\hat{x}_1 + \hat{B}_1 u - B_2 \hat{T}_L + L_1(c_1 x_1 - c_1 \hat{x}_1) \tag{10}$$

$$\dot{\hat{T}}_L = l_3(c_1 x_1 - c_1 \hat{x}_1) \tag{11}$$

where, the ^ means estimated values and $L_1 = (l_1 \;\; l_2)^T$ In this case, l_1, l_2, and l_3 are elements of L. Since k_t is the dominant value for the parameter variation, this variable is selected for the adaptation value. In order to derive the adaptive scheme, the Lyapunov theorem is utilized. From (8) - (10), the estimated errors of the rotor speed and rotor positions are described by the following equation:

$$\begin{aligned} \dot{e}_1 &= (A - L_1 c_1)e_1 + (B_1 - \hat{B}_1)u - B_2(T_L - \hat{T}_L) \\ &= Ge_1 + (B_1 - \hat{B}_1)u - B_2(T_L - \hat{T}_L) \end{aligned} \tag{12}$$

where, $e_1 = x_1 - \hat{x}_1$ and $G = A - L_1 c_1$. Now, a new Lyapunov function candidate V is defined as follows:

$$V = e_1^T P e_1 + \frac{1}{a}(B_1 - \hat{B}_1)^T (B_1 - \hat{B}_1) + \frac{1}{\beta}(T_L - \hat{T}_L)^2 \tag{13}$$

where, P and α are a positive definite matrix and a positive constant, respectively. The time derivative of V becomes

$$\begin{aligned} \dot{V} &= e_1^T\left((A - L_1 c_1)^T P + P(A - L_1 c_1)\right)e_1 \\ &+ 2\left(e_1^T u - \frac{1}{a}\dot{\hat{B}}_1^T\right)\Delta B_1 - 2\left(e_1^T B_2 + \frac{1}{\beta}\dot{\hat{T}}_L\right)\Delta T_L \end{aligned} \tag{14}$$

where, $\Delta B_1 = B_1 - \hat{B}_1$ and $\Delta T_L = T_L - \hat{T}_L$. From (14), the adaptive mechanism can be obtained by equalizing the second term to zero as follows:

$$\dot{\hat{B}}_1^T = \alpha e_1^T u \tag{15}$$

The third term can be decreased to zero using the following equation:

$$\dot{\hat{T}}_L = -\beta e_1^T B_2 = -\beta'(\omega_r - \hat{\omega}_r) \tag{16}$$

where, $\beta' = \frac{p}{2J}\beta > 0$.

Therefore, the new adaptive torque observer can be obtained as

$$\dot{\hat{T}}_L = l_3(c_1 x_1 - c_1 \hat{x}_1) - \beta'(\omega_r - \hat{\omega}_r) \tag{17}$$

If we determine the observer gain matrix L_1 using the optimal theory, the first term of (14) can be seen as negative semi-definite. Assuming that there exists a positive definite matrix R such that

$$G^T P + PG = -R \tag{18}$$

The derivative of the Lyapunov function candidate can be written as

$$\dot{V} = -e_1^T (R) e_1 \le 0 . \tag{19}$$

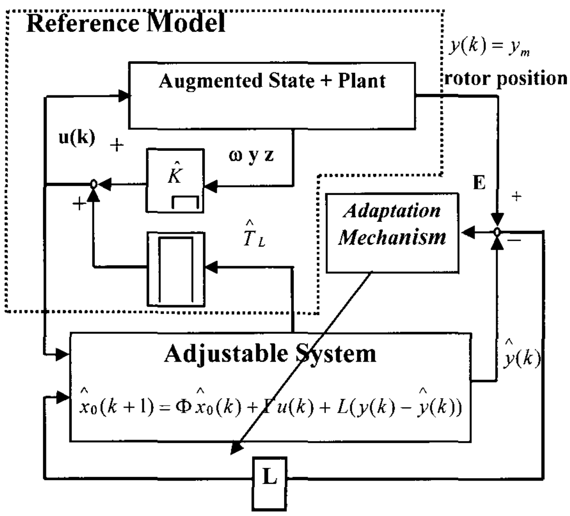


Fig. 1 Configuration of the model reference adaptive torque observer

Hence, e_1 is uniformly asymptotically stable. Reducing the estimated load torque error and selecting the gain L_1 properly to make e_1 equal to zero can decrease the maximum error. The discrete motor equation can be written as follows:

$$\begin{aligned} y(k+1) &= a_3 \omega_r(k) + y(k) + b_2 i_{qc}, \\ \omega_r &= a_1 \omega(k-1) + b_1 i_{qc}(k-1) \end{aligned} \tag{20}$$

where, a_1, a_3, b_1, and b_2 are parameters, which are obtained by converting the system to digital domain [11].Using approximation of $\omega_r(k) = (y(k) - y(k-1))/h$, the ARMA model can be obtained as follows:

$$\begin{aligned} y(k+1) = &[y(k) \quad y(k-1) \quad y(k-2)][A_1 \quad A_2 \quad A_3]^T \\ &+ [u(k) \quad u(k-1)][B_1 \quad B_2]^T \end{aligned} \tag{21}$$

where, A_1, A_2, A_3, B_1 and B_2 are 1, $a_1 a_3 / h$, $-a_1 a_3 / h$, b_2 and $a_3 b_1$, respectively. In this case, B_1 and B_2 are not exact. Therefore, only these terms are expressed as the quasi term. Employing the definition of the new quasi -output can reduce the order of the estimated matrix. The quasi-output Y and the model output Y_m can be obtained as the difference of the real output and the known values as follows:

$$Y(k+1) = (y(k+1) - \Phi_1 \Theta_1) = \Phi_2 \hat{\Theta}_1 \tag{22}$$

$$Y_m(k+1) = b_2 i_{qc}(k) + a_3 b_1 i_{qc}(k-1) \tag{23}$$

where,
$\Phi_1 = [y(k) \quad y(k-1) \quad y(k-2)]$,
$\Phi_2 = [u(k) \quad u(k-1)]$, $\Theta_1 = [A_1 \quad A_2 \quad A_3]$, $\hat{\Theta}_2 = [\hat{B}_1 \quad \hat{B}_2]$.
Using, $E = Y(k+1) - Y_m(k+1)$ the gradient can be obtained as

$$\hat{\Theta}_2(k+1) = \hat{\Theta}_2(k) - h \begin{pmatrix} \alpha_1 i_{qc}(k) \\ \alpha_2 a_3 i_{qc}(k-1) \end{pmatrix} E \tag{24}$$

where, α_1 and α_2 are the elements of the vector α . The resultant block diagram of the model reference adaptive observer is shown in Fig. 1. It places Current Reference- Pulse Width Modulation (CRPWM) on the track of the stator three-phase current obtained by this result. The switching frequency has been adapted as 20 kHz.

4. CONFIGURATION OF OVERALL SYSTEM

The block diagram of the proposed controller is shown in Fig. 2 where the controller is composed of two parts. The position controller is composed of the augmented state feedback. For the realization of the augmented state z(k+1), the discrete form of this state is approximately obtained by using a trapezoidal rule. DSP TMS320C31 is used as a digital controller with C-Language.

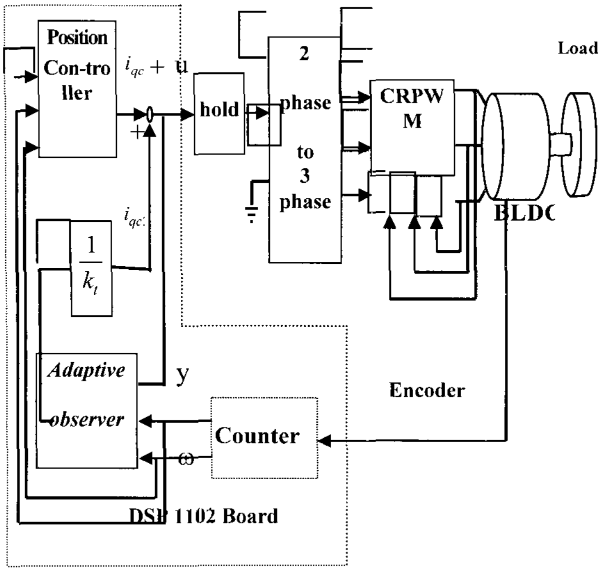


Fig. 2 Block diagram of the proposed precision position control system

An asymptotically stable load torque observer is used in the second part. The estimated torque is used for compensation of the position error. The vector controller that makes the system a linear system drives the motor system. In this experiment, inertia load is considered.
This load, shaped as an annular cylinder, is made of brass for time varying torque. Inertia J_L is $0.372\times10^{-3} kgm^2$ which corresponds to approximately 10 times the motor inertia.

This inertia load is directly connected to the motor shaft as shown in Fig. 3. The load equation is as follows:

$$(J + J_L)\dot{\omega}_r + B\omega + T_L = k_t i_{qc}. \tag{25}$$

We write (25) as $J\dot{\omega}_r + B\omega + (J_L\dot{\omega}_r + T_L) = k_t i_{qc}$. Since speed is time varying in position control, inertia variation makes the time varying load torque as follows:

$$T_{LNEW}(t) = J_L\dot{\omega}_r(t)\ (= J_L \frac{d}{dt}\omega_r(t)) \tag{26}$$

This direct coupled inertia load creates a time varying load torque in a position control system. Fig.4 shows this real system which has DS1102 and Servo driver.

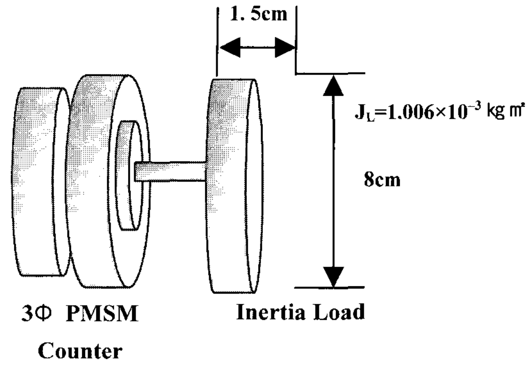


Fig. 3 Configuration of inertia load

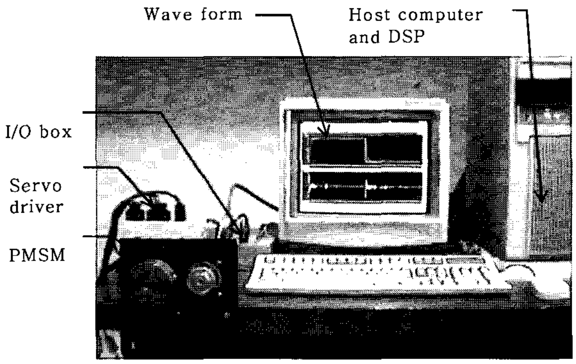


Fig. 4 The configuration of the real system

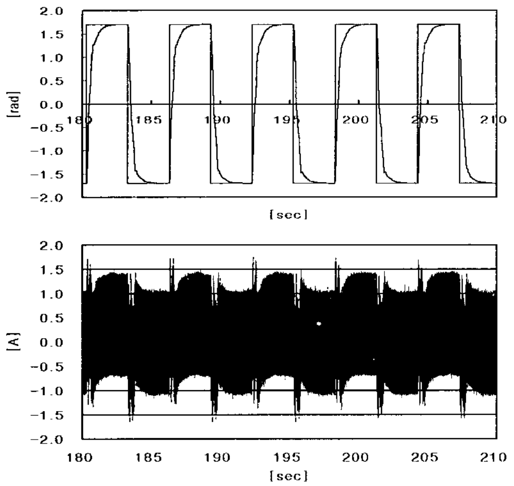
5. SIMULATION AND EXPERIMENTAL RESULTS

The parameters of a PMSM used in this experiment are given as follows:

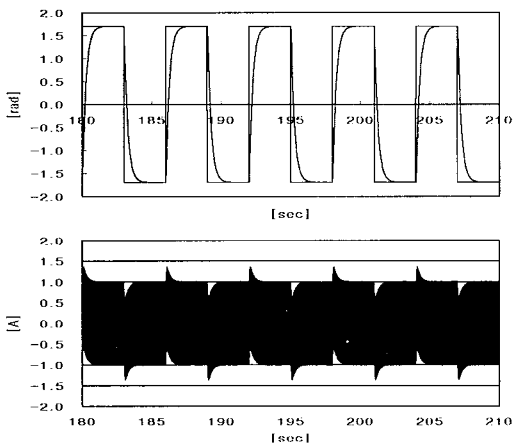
Table 1 Parameters of the BLDC motor

Power	400 *watt*
Inertia3	0.363×10^{-4} kgm^2
Rated torque	1.3 *Nm*
Rated current	2.7 *A*
Stator resistance	1.07 Ω / *phase*
Time constant	0.5 *ms*

The hysteresis-band gap is chosen as $0.05\,A$, and the sampling time Ts is determined as $0.1\,ms$. Since nameplate values are not exact in real value, we assume that each parameter value of J, B, λ_m, R, and L change about 20% in the real system. It is commonly accepted that even if we receive the nameplate values of the motor, there is observable error up to 20%. Furthermore, if the temperature increases, resistance R may increase. K_t also will vary as the operating point of the B-H curve. Equivalent L will vary following the operating frequency. We cannot know the real viscous friction of B.

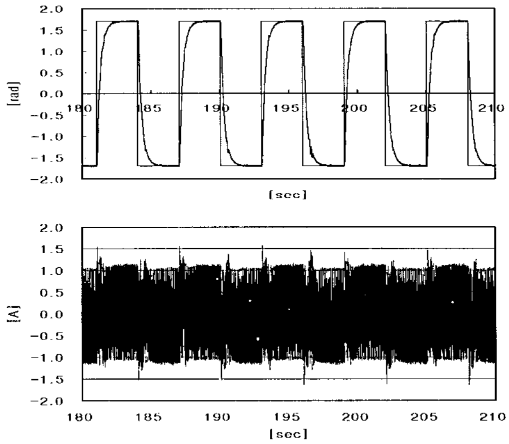


(a) Dead beat control with inertial load

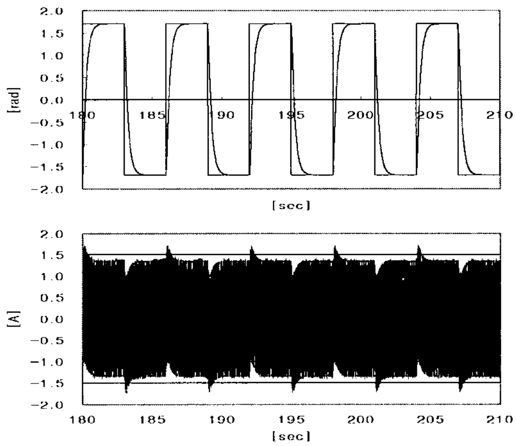


(b) Adaptive observer with inertial load

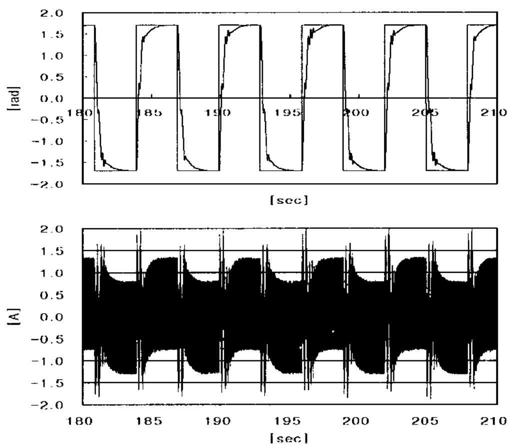
Fig. 5 Simulation results of the rotor position and q phase current command



(a) Dead beat control with inertial load



(b) Adaptive observer without inertial load



(c) Adaptive observer with inertial load

Fig. 6 Experimental results of the rotor position and q phase current command

Nominal value must be determined. We change 20% of these values only in simulation.

After some trial and error, the weighting matrix is selected as R=1, $Q = diag[0.1 \quad 60 \quad 1000]$ and the optimal gain matrix becomes $k = [0.0598 \quad 2.0810 \quad 7.3540]$.

The nominal deadbeat observer gain becomes $L = [21762 \quad 2.8187 \quad -1000.2]$. In the adaptation mechanism, the adaptation rates α_1, α_2 and β' are obtained as 0.004, 0.002 and 0.025 respectively by trial and error. The simulation results are in Fig. 5.

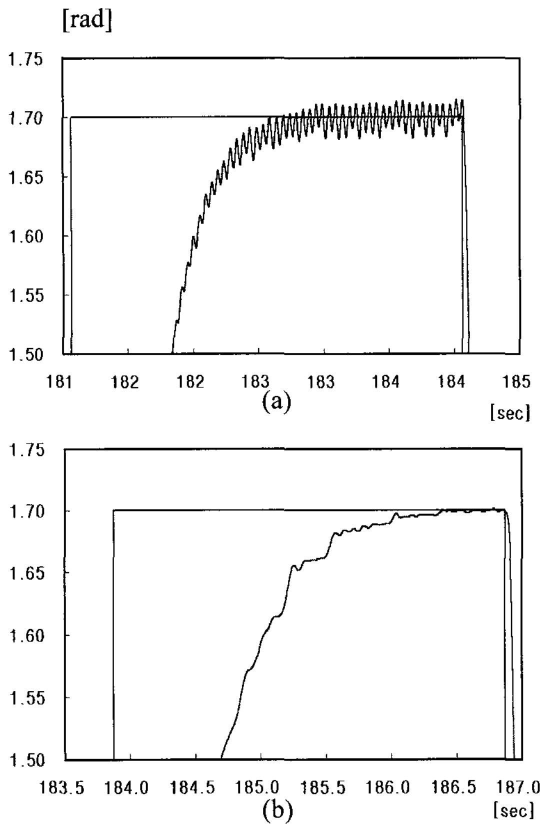


Fig. 7 Performance comparison with deadbeat and adaptive observers
(a) Results of conventional deadbeat observer
(b) Results of proposed adaptive observer

The deadbeat observer seems to provide positive results in large scales as shown in Fig. 5(a). However, in precision scales some position errors are large as shown as Fig. 7, which is the experimental result. In the experimental system, the nominal control parameter is

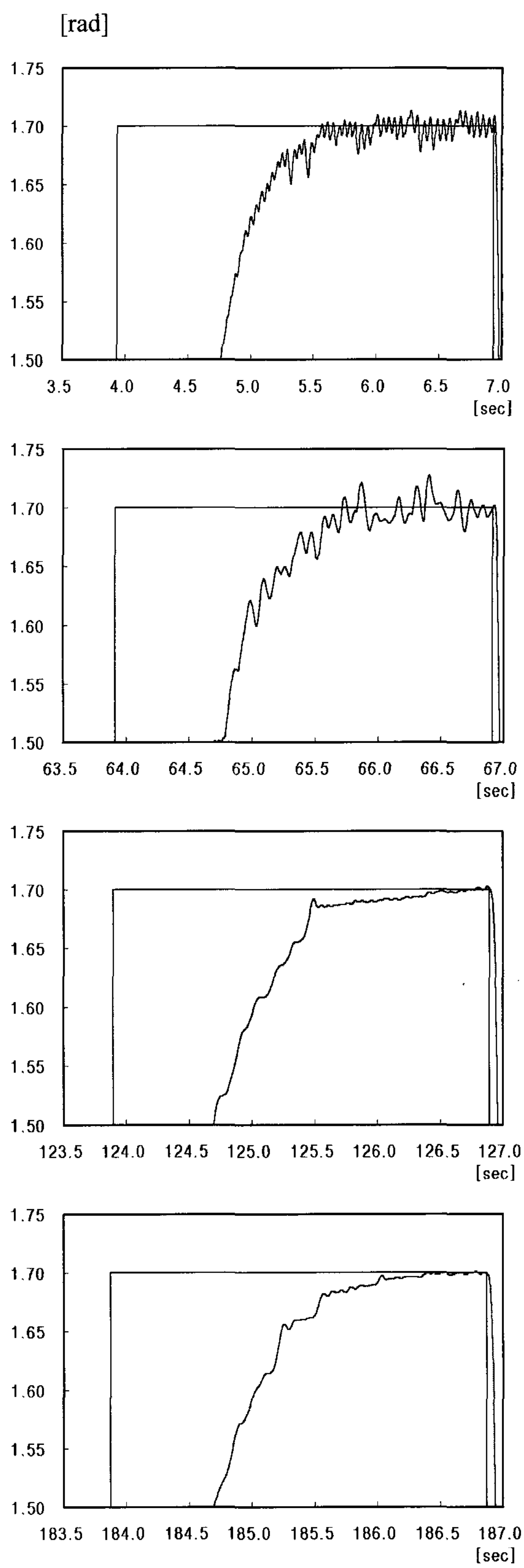


Fig. 8 Process of successive adaptation with

used for what is obtained in nameplate value. The chattering increases because of the variation of parameters compared with the result in [1]. That current shows adaptive control action is greater than the deadbeat system. The experimental results are in Fig. 5, Fig. 6 and Fig. 7 with position and q-phase current command. This position and current command can be obtained in the memory of the DSP system. The deadbeat observer seems to show a

positive result in a large scale view inertial load as shown in Fig. 6(a) as observed in simulation. We can see many current ripples at steady state caused by the dead beat controller.

However on a precision scale we found precision position errors as large as an estimated 0.025 peak-to-peak radian. This is quite small but will create problems in chip mount machines, semiconductor production machines, precision milling machines, high resolution CNC machines, precision assembly robots and other areas. If some parameter variations increase or decrease, the chattering phenomena increase when compared with the result in [1]. From that current we can see a control action for the reduction of error.

However, there is little chattering in the position at the bottom side. In Fig. 6(b), a different type control action is depicted. There is very small position error in not only the upper side but also the lower side. This indicates good adaptation to different conditions. The current pattern changes as a position response. The difference cannot be detected in the position result between the unloaded case of Fig. 6(b) and the loaded case of Fig. 6(c). As shown in Fig. 6(c), the magnitude of the current in transient increases up to 2[A] which has a more controlled effect to reduce the inertia load.

Zoomed results are depicted in Fig. 7. In the conventional case, error magnitude is not increased as the load is increased. However, there is a little position chattering. The effects of the parameter uncertainty and increased J or load are compared with those results of [1]. This small chattering decreases as shown in Fig 7(b) by the proposed algorithm. This new algorithm has the deadbeat gain and the new adaptive algorithm. Processing of this algorithm is shown in Fig 7. It is started from Fig. 6(a) as a nominal gain. The adaptive algorithm changes Γ to reduce error. During this time, the deadbeat gain is not changed. When we initially use this system, we can use this controller after about 3 minutes, which will allow time for adapting. The next system start is from saved data Γ, which will reduce time lost.

6. CONCLUSION

A new deadbeat load torque observer with an adaptive algorithm is proposed to obtain better performance from the BLDC motor in a precision position control system. Also, the augmented state variable feedback is used in the digital experimental control system with optimal gain. The system response comparison between the fixed deadbeat gain observer and the adaptive observer has been performed.

The position error caused by the problem of unknown parameter, torque disturbances and a small chattering effect decreased to zero. The load torque compensator based on the adaptive observer and the feedforward injection was used to cancel out the steady state and the transient position error due to the external disturbances, such as various friction, load torque and small chattering effect of deadbeat control. The stability analysis is carried out using the Liapunov stability theorem. Under this analysis, the new adaptive torque observer is proposed. The total adaptive control system is realized by a digital controller DS1102 (TMS320c31) with 0.1ms sampling time and the gain is obtained in z-domain using the optimal theory. The chattering effect of the deadbeat controller is also reduced although its advantage is held. Therefore advantages such as robustness of deadbeat control and asymptotically stable adaptation are obtained.

List of symbols

r_s : Stator resistance

L_q, L_d : q, d axis inductance

J : Inertia moment of rotor and load

J_L : Load inertia

B : Viscous friction coefficient

k_t : Motor torque constant

λ_m : Flux linkage of permanent magnet

p : Number of poles

i_{qc}, i_{dc} : q, d axis stator current commands
i_{qs}, i_{ds} : q, d axis stator current
V_q, V_d : q, d axis voltage
ω_r : Electrical angular speed of rotor
T_L : Load torque
$y, \hat{y}, y_r$: Rotor position, estimated rotor position and reference rotor position
$\Theta_1, \hat{\Theta}$: System parameter vectors of known and unknown coefficients
E : Estimated output error
e : Rotor position error
h : Sampling time
T_{LNEW} : Time-varying load torque by inertia variation
α, β : Adaptation rate

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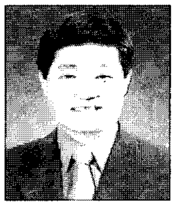

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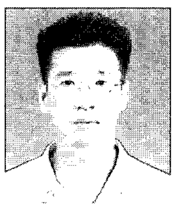

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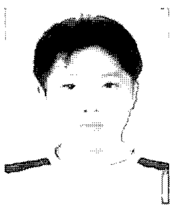

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