Proceedings of the 17th World Congress
The International Federation of Automatic Control
Seoul, Korea, July 6-11, 2008

Multirate Digital Servo Drive Based on Acceleration Observer and Disturbance Compensator

Chin-Sheng Chen*, Ying-Tsung Teng**

**Graduate Institute of Automation Technology, National Taipei University of Technology, Taipei, Taiwan (e-mail: saint@ntut.edu.tw)*
***Graduate Institute of Automation Technology, National Taipei University of Technology, Taipei, Taiwan (e-mail: t5618027@ntut.edu.tw)*

Abstract: This paper presents a digital servo driver that realizes a novel multirate feedback controller based on position, velocity and acceleration feedback. The velocity and acceleration signals are firstly estimated by state observer using optical encoder information, and the estimated velocity is fed into discrete disturbance observer (DDOB) to estimate and compensate the external disturbance. In this scheme, controller with acceleration feedback can be realized by replacing the current loop with acceleration loop. When the DC servo motor is controlled by the proposed acceleration feedback control with disturbance compensation, the total servo system from acceleration to position becomes the acceleration controlled system which is fixed to a nominal double integral dynamics in the presence of parameter variation and torque disturbance. Hence, the fast and precise position control can be carried out easily. The proposed acceleration feedback controller and PD position feedback controller are evaluated experimentally on a DSP, which is implemented multirate control scheme, controlled DC servo motor positioning system. The experimental results show that this digital servo system is robust and remarkably sustains the same performance compared to the single fast rate control scheme.

1. INTRODUCTION

The mechanical systems must be supported by motion controllers, which ensure robust, high speed and high accuracy tracking performance. To control the system adequately, the controller algorithm will be implemented on a microprocessor using digital control theory.

The conventional cascade control of servo drives, shown in Fig. 1(a), inner current control loop has a high-gain to minimize the current over the operating range of the system. In this scheme, I_{CMD} and I_f are current command and current response respectively; T_d is disturbance; $\omega(s)$ is velocity response; E is the stator input voltage; J is rotor inertia constant; B is viscous damping constant; L is stator inductance; R is stator resistance; K_e is electromotive force constant and K_t is torque constant of motor; K_A is gain of pulse width modulation(PWM) amplifier; K_p and K_i are proportional gain and integral gain of current loop controller respectively. According to high-gain current control, the velocity dynamics still depend on inertia, viscous damping and disturbance, shown in Fig. 1(b). Some researchers employed acceleration feedback to solve this problem. Acceleration control is, however, seldom implemented in practical drive systems due to unsatisfactory results of most acceleration measurement methods. Kalman filter is a general estimator of velocity and acceleration in servo systems. Polynomial predictive filtering methods have been successfully applied in motion control. S. H. Lee proposed a low-acceleration estimator to accurately estimate the acceleration in very low velocity and/or very low acceleration.

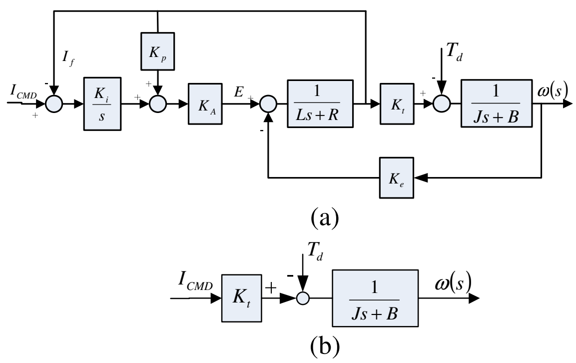


Fig. 1. Scheme of servo drive: (a) Current control loop (b) Simplified structure.

In motion control, the major sources of uncertainties are inertia, damping friction and torque disturbance. Kempf *et al.* proposed the disturbance observer and feed-forward design to improve the tracking performance for a high-speed direct-drive positioning table with torque disturbance. Hsu *et al.* use an ARX mathematical model to identify the velocity loop of the motion system and propose a disturbance observer to decrease the perturbations caused by external disturbances.

Recently, a computationally efficient multirate control scheme is successful used in servo control. H. Fujimoto *et al.* realized perfect tracking control based on multirate feedforward control with generalized sampling periods.

Furthermore, they successfully applied multirate adaptive robust control in linear motor. S. H. Lee proposed multrate digital control scheme to improve the performance of disk drive actuator. Mizuochi *et al.* realized a high performance motion controller based on acceleration control.

In this paper, a digital state observer is firstly proposed to estimate the velocity and acceleration of DC motor, and the estimated velocity is fed into the DDOB to estimate and compensate the external disturbance. Then, an inner acceleration control loop and a PD position loop feedback controller are proposed. When the DC servo motor is controlled by the proposed acceleration feedback control system, the total servo system from acceleration to position becomes the acceleration controlled system which is fixed to a nominal double integral dynamics in the presence of parameter variation and torque disturbance. Hence, the fast and precise position control can be carried out easily. The proposed acceleration feedback controller and PD position feedback controller are evaluated experimentally on a DSP, which is implemented multirate control scheme, controlled DC servo motor positioning system. The experimental results show that this digital servos ystem is robust and remarkably sustains the same performance compared to the single fast rate control scheme.

2. Multirate Digital Control Design

The reasons of multirate control scheme applied in our proposed digital servo driver are that: 1) the acceleration feedback loop takes a higher bandwidth than position feedback loop; 2) the position controller does not require to output control effort every control period of acceleration loop; 3) it is computationally efficient.

The inner integral acceleration control loop and outer PD position control loop are proposed in this paper, shown in Fig. 2, in which the discrete-time equivalent acceleration feedback controller and position feedback controller are transformed by backward different method. In this scheme, θ_{CMD} and $\theta(z)$ are position command and position response respectively; K_{ai} is integral gain of acceleration loop controller; K_{pp} and K_{pd} are proportional gain and derivative gain of position loop controller. S_{T_s} and H_{T_c} represent the encoder sampling operator and control effort holding operator, with time period T_s and T_c, respectively. Here we represent the sampling update period, $T_s = T_c R_c$, where R_c is the multiplicity of control update rate.

2.1 Acceleration Feedback Control Design

The electrical time constant of the brushed DC motor is often ignored due to its relatively low inductance. Hence, the dynamic equation of DC motor should be as follows

$$R \cdot J \cdot \alpha + (R \cdot B + K_t \cdot K_e)\omega = K_t \cdot E \tag{1}$$

The transfer function from input voltage to angular velocity is

$$G(s) = \omega(s)/E(s) = K_t/(J_{eq} \cdot s + B_{eq}) \tag{2}$$

where

$B_{eq} = (R \cdot B + K_t \cdot K_e)$,

$J_{eq} = J \cdot R$.

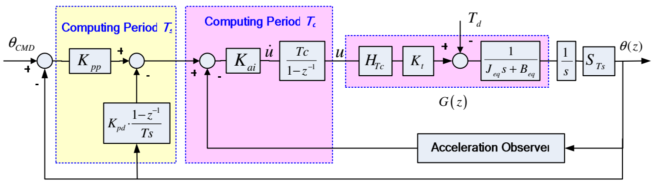


Fig. 2. Multirate control scheme.

According to Fig. 2, the z transform of the plant model with zero order hold (ZOH) is as follows:

$$G(z) = \omega(z)/u(z) = b/(z+a) \tag{3}$$

where

$a = -\exp[-(B_{eq} \cdot T_s)/J_{eq}]$,

$b = (K_t/B_{eq}) \cdot \left\{1 - \exp\left[-(B_{eq} \cdot T_s)/J_{eq}\right]\right\}$,

and the discrete control effort $u(z)$ passing the ZOH corresponds to the input voltage $E(s)$.

Fig. 3 shows the acceleration control loop with a simple integral control gain K_{ai}, where α_{CMD} and $\alpha(z)$ are acceleration command and acceleration response respectively. Assuming $T_d = 0$, the transfer function from acceleration command to acceleration response is described as

$$TF_{acc}(z) = b \cdot K_t \cdot K_{ai}/(z + a + b \cdot K_t \cdot K_{ai}) \tag{4}$$

For this first order system, the bandwidth is design as ω_{BW}, then the pole is located as

$$z = \exp(s \cdot T_c) = \exp(-\omega_{BW} \cdot T_c) \tag{5}$$

From (4), the integral gain K_{ai} can designed as

$$K_{ai} = -(z+a)/(b \cdot K_t) \tag{6}$$

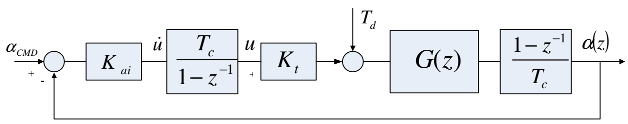


Fig. 3. Acceleration feedback control.

Acceleration feedback can be realized by replacing the current loop with acceleration loop. Then the cut-off frequency of $TF_{acc}(z)$ must be chosen to match to the bandwidth of conventional inner current loop and the bandwidth must be at 5~10 times of outer loop. Thus the dynamics is so fast that $TF_{acc}(z)$ can be treated as a constant acceleration gain with respect to outer position loop.

2.2 PD Position Feedback Control Design

This paper proposes a PD position loop feedback controller, shown in Fig. 4. In the PD controller, the proportional gain K_{pp} provides the stiffness of the system; the differential gain K_{pd} provides the damping for stability; K_{acc} is a acceleration loop transfer function. According to the cascade control philosophy, we design the bandwidth of acceleration loop 5~10 times of position loop. Then the $TF_{acc}(z)$ could be depicted as a DC gain K_{acc}.

$$TF_{acc}(z) \approx K_{acc} = b \cdot K_t \cdot K_{ai} / (1 + a + b \cdot K_t \cdot K_{ai}) \tag{7}$$

From Fig. 4, the transfer function can be easily obtained as the following equation

$$TF_p(z) = \frac{k_{pp} K_{acc} T_s^2}{1 - \frac{(2 + k_{pd} K_{acc} T_s)}{1 + K_{acc}(k_{pp} T_s^2 + k_{pd} T_s)} z^{-1} + \frac{1}{1 + K_{acc}(k_{pp} T_s^2 + k_{pd} T_s)} z^{-2}} \tag{8}$$

For this second order system, we can design a compensator by the pole-placement method. Let ξ, ω_n be the two parameters similar to damping ratio and natural frequency of a standard second order system, the parameters of this controller are

$$K_{pd} = \frac{\left(\frac{2 \cdot \exp(-\xi \cdot \omega_n \cdot T_s) \cdot \cos(\omega_d \cdot T_s)}{\exp(-2\xi \cdot \omega_n \cdot T_s)}\right) - 2}{K_{acc} T_s} \tag{9a}$$

$$K_{pp} = \frac{1/(\exp(-2\xi \cdot \omega_n \cdot T_s)) - 1 - K_{pd} \cdot K_{acc} \cdot T_s}{T_s^2} \tag{9b}$$

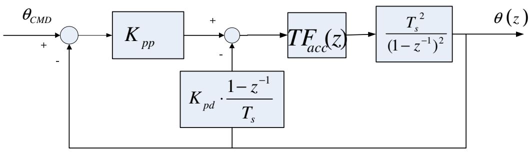


Fig. 4. PD position feedback control.

3. Multirate Digital Observer Design

In this paper, a multirate two stages observer design is proposed. The velocity and acceleration are estimated in the first stage. In the second stage, the estimated velocity and control effort are fed into the DDOB to estimate the external disturbance. Fig. 5 shows the multirate digital state observer scheme. In this scheme, the computing period is T_c in acceleration observer and T_s in DDOB.

3.1 Acceleration Observer Design

For implementation of acceleration and velocity feedback control using optical encoder information, a reliable acceleration and velocity signal must be obtained. The direct differentiation method is based on the assumption of a low noise velocity signal obtained from a high-performance pulse encoder. Unfortunately, accurate pulse encoders with high pulse rate are expensive and low noise velocity signals are not often encountered in industrial environments. A novel state observer is proposed to estimate the velocity and acceleration of DC motor. According to (1), the differential of acceleration is obtained as

$$\dot{\alpha}(t) = -(B_{eq} / J_{eq}) \cdot \alpha(t) + (K_t / J_{eq}) \cdot \dot{u}(t) \tag{10}$$

where $\dot{u}(t)$ is the control force of acceleration feedback loop.

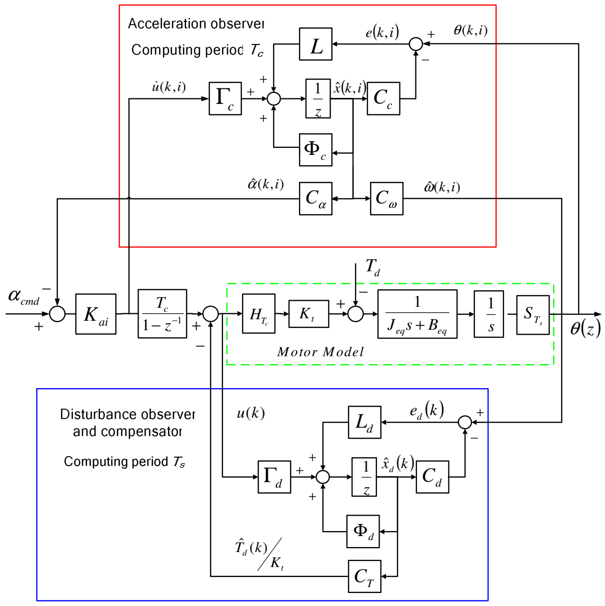


Fig. 5. Multirate digital state observer scheme.

Then the state-space dynamic equation are described as

$$\begin{aligned} \dot{X}(t) &= A \cdot X(t) + B \cdot \dot{u}(t) \\ Y(t) &= C \cdot X(t) \end{aligned} \tag{11}$$

where vectors *X(t)* and matrix *A*, *B* and *C* are given by

$$X(t) \triangleq [\theta(t) \quad \omega(t) \quad \alpha(t)]^T,$$

$$A \triangleq \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & -J_{eq}/B_{eq} \end{bmatrix},$$

$$B \triangleq [0 \quad 0 \quad K_t / J_{eq}]^T,$$

$$C \triangleq = [1 \quad 0 \quad 0]$$

From (11), a multirate discrete-time system can be expressed as

$$\begin{aligned} x(k, i+1) &= \Phi_c \cdot x(k,i) + \Gamma_c \cdot \dot{u}(k,i) \\ y(k,i) &= C_c \cdot x(k,i) \end{aligned} \tag{12}$$

for $i = 0, 1, \ldots, R_c - 1$, where

$$\Phi_c = e^{AT_c}\ ,\ \Gamma_c = \int_0^{T_c} e^{As} ds \cdot B\ ,\ C_c = C$$

The indexes k and i indicate measurement update and control update instants, respectively, such that $t = (k + i / R_c)T_s$ for $i = 0,1,..., R_c - 1$.

From (12), the triplet (Φ_c, Γ_c, C_c) is controllable and observable. Hence a digital observer is proposed to estimate the acceleration and velocity signals, as shown in Fig. 5. It is described by

$$\begin{aligned} \hat{x}(k,i+1) &= \Phi_c \cdot \hat{x}(k,i) + \Gamma_c \cdot \dot{u}(k,i) + L(y(k,0) - C_c \hat{x}(k,0)) \\ \hat{y}(k,i) &= C_c \cdot \hat{x}(k,i) \end{aligned} \tag{13}$$

If the difference between the status $x(k,i)$ of the real motor and the status $\hat{x}(k,i)$ of the observer is assumed to be $e(k,i)$. Subtracting (13) from (12), we obtain following equation

$$e(k,i+1) = \Phi_c \cdot e(k,i) - LC_c e(k,0) \tag{14}$$

For Lemma 1, (14) can be represented as

$$e(k+1,0) = \left[\Phi_c^{\ R_c} - L_c C_c\right] \cdot e(k,0) \tag{15}$$

where

$$L_c = \sum_{i=0}^{R_c - 1} \Phi_c^{\ i} \cdot L$$

If an L_c exists which makes the eigenvalues of $(\Phi_c^{\ Rc} - L_c C_c)$ are all inside unit circle, then $\hat{x}(k,0)$ will approach $x(k,0)$ asymptotically. Then the observer gain can be obtained by

$$L = \left[\sum_{i=0}^{R_c - 1} \Phi_c^{\ i}\right]^{-1} \cdot L_c \tag{16}$$

Thus the estimated acceleration $\hat{\alpha}$ and velocity $\hat{\omega}$ are obtained from $\hat{x}(k,i)$ multiplied by vectors $C_\alpha = [0 \ \ 0 \ \ 1]$ and $C_\omega = [0 \ \ 1 \ \ 0]$ respectively.

Lemma 1: Consider a multirate discrete-time system (12), the dynamics of reconstruction error $e(k,0)$ can be represented as (16).

Proof: Proof by induction begins by showing that equation hold for $R_c = 2$:

$$\begin{aligned} e(k+1,0)\big|_{R_c=2} &= \Phi_c\left[\Phi_c \cdot e(k,0) - LC_c \cdot e(k,0)\right] - LC_c \cdot e(k,0) \\ &= \left[\Phi_c^{\ 2} - \Phi_c LC_c - LC_c\right] \cdot e(k,0) \\ &= \left[\Phi_c^{\ 2} - \sum_{i=0}^{1} \Phi_c^{\ i} LC_c\right] \cdot e(k,0)\big|_{R_c=2} \end{aligned} \tag{17}$$

Next, we assume that the equation hold for n.

$$e(k+1,0)\big|_{R_c=n} = \left[\Phi_c^{\ n} - \sum_{i=0}^{n-1} \Phi_c^{\ i} LC_c\right] \cdot e(k,0)\big|_{R_c=n} \tag{18}$$

Then, we are required to prove that they also are true for $n+1$.

$$\begin{aligned} e(k+1,0)\big|_{R_c=n+1} &= e(k,n+1)\big|_{R_c=n+1} \\ &= \Phi_c \cdot e(k,n)\big|_{R_c=n+1} - LC_c \cdot e(k,0)\big|_{R_c=n+1} \\ &= \Phi_c \cdot e(k+1,0)\big|_{R_c=n} - LC_c \cdot e(k,0)\big|_{R_c=n+1} \\ &= \Phi_c\left[\left(\Phi_c^{\ n} - \sum_{i=0}^{n-1} \Phi_c^{\ i} LC_c\right) \cdot e(k,0)\big|_{R_c=n}\right] \\ &\quad - LC_c \cdot e(k,0)\big|_{R_c=n+1} \\ &= \left[\Phi_c^{\ n+1} - \sum_{i=0}^{n} \Phi_c^{\ i} LC_c\right] \cdot e(k,0)\big|_{R_c=n+1} \end{aligned} \tag{19}$$

which completes the proof.

3.2 Disturbance Observer and Compensator Design

The external disturbance is estimated and compensated by the DDOB. From $G(s)$ in the Fig. 2, the dynamic equation includes external disturbance is represented as

$$J_{eq} \cdot \dot{\omega}(t) + B_{eq} \cdot \omega(t) = K_t \cdot u(t) + T_d(t) \tag{20}$$

Then the state-space dynamic equation are described as

$$\begin{aligned} \dot{X}_d(t) &= A_d \cdot X_d(t) + B_d \cdot u(t) \\ Y_d(t) &= C_d \cdot X_d(t) \end{aligned} \tag{21}$$

where vectors $X_d(t)$ and matrix A_d, B_d and C_d are given by

$$X_d(t) \triangleq \left[\omega(t) \quad T_d(t)\right]^T,$$

$$A_d \triangleq \begin{bmatrix} -B_{eq}/J_{eq} & 1/J_{eq} \\ 0 & 0 \end{bmatrix},$$

$$B_d \triangleq \begin{bmatrix} K_t/J_{eq} & 0 \end{bmatrix}^T,$$

$$C_d = [1 \quad 0].$$

A multirate discrete-time system can be expressed as

$$\begin{aligned} x_d(k+1) &= \Phi_d \cdot x_d(k) + \Gamma_d \cdot u(k) \\ y_d(k+1) &= C_{dd} \cdot x_d(k) \end{aligned} \tag{22}$$

where

$$\Phi_d = e^{A_d T_s}\ ,\ \Gamma_d = \int_0^{T_s} e^{A_d s} ds \cdot B_d\ ,\ C_{dd} = C_d\ .$$

From (22), the triplet (Φ_d, Γ_d, C_d) is uncontrollable, but observable. Hence a digital observer is proposed to estimate and compensate the disturbance, as shown in Fig. 5. It is described by

$$\begin{aligned}\hat{x}_d(k+1) &= \Phi_d \cdot \hat{x}_d(k) + \Gamma_d \cdot u(k) + L_d(y_d(k) - C_d\hat{x}_d(k)) \\ \hat{y}_d(k+1) &= C_d \cdot \hat{x}_d(k)\end{aligned} \quad (23)$$

If the difference between the status $x_d(k)$ of the real motor and the status $\hat{x}_d(k)$ of the observer is assumed to be $e_d(k)$. Subtracting (29) from (28), we obtain following equation

$$e_d(k+1) = (\Phi_d - L_d C_d) \cdot e_d(k) \quad (24)$$

If an L_d exists which makes the eigenvalues of $(\Phi_d - L_d C_d)$ are all inside unit circle, then $\hat{x}_d(k,0)$ will approach $x_d(k,0)$ asymptotically. Thus the estimated disturbance $\hat{T}_d$ are obtained from $\hat{x}_d(k,i)$ multiplied by vectors $C_T = [0 \quad 1]$. Therefore, the compensated force $\hat{T}_d / K_t$ is fed into the acceleration control loop.

4. Multirate Implementation and Experimental Results

The control scheme mentioned in the pervious section will be implemented in a multirate environment.

4.1 Implemented Scheme

The schematic diagram of the multirate experimental system is depicted in Fig. 6, in which the DSP is TI TMS320C6711 and FPGA is XILINX Spartan-II XC2S50. The DSP takes the main task of control firmware and FPGA takes the tasks of communication handshake to print port, PWM interface to send the control input to motor driver and encoder interface to receive the encoder feedback signals from DC motor encoder. The proposed acceleration feedback controller, PD position feedback controller and observer are implemented on this DSP, which is implemented multirate control scheme, controlled DC servo motor positioning system. The motor driver includes a PWM amplifier to drive DC motor. The PC utilizes the interface of print port to send high level language to this embedded DSP motion controller.

4.2 Experimental Results

A mechanism of LED wire bonding head is used to test the tracking performance of our proposed controllers. The motion system shown in Fig. 7 comprises ball-screws and guide-ways, motor drivers and a DSP based motion controller. This experimental servo system is built up in a so-called semi-closed loop, i.e., feedback signal is from the encoder of DC motor. The system identification followed the previous work by the authors, in which the parameters K_t, J_{eq} and B_{eq} are obtained as 0.0159795 $N \cdot m / volt$, 0.00006473 $N \cdot m \cdot \sec^2$ and 0.0003494 $N \cdot m \cdot \sec$ respectively. For multirate experiments, let us choose a sampling time of $T_s = 1$ msec and try two multiplicities $R_c = 1$ and 10 such that $T_c = 1$ msec (slow single rate) and 100 $\mu\sec$, respectively. The experimental results are compared with fast single rate, 100 $\mu\sec$, to validate the proposed design.

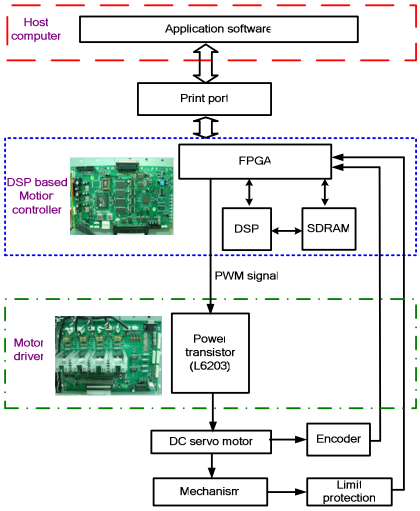


Fig. 6. Implemented scheme.

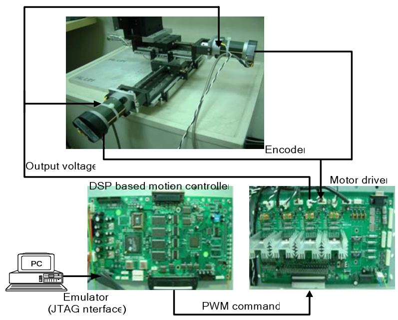
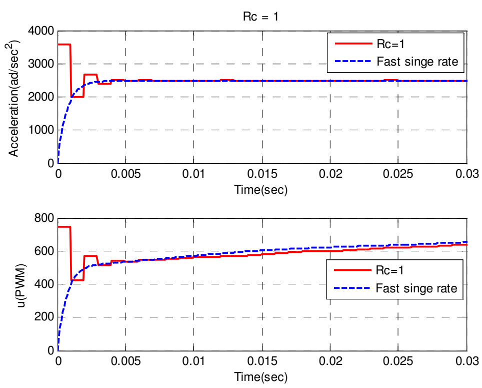
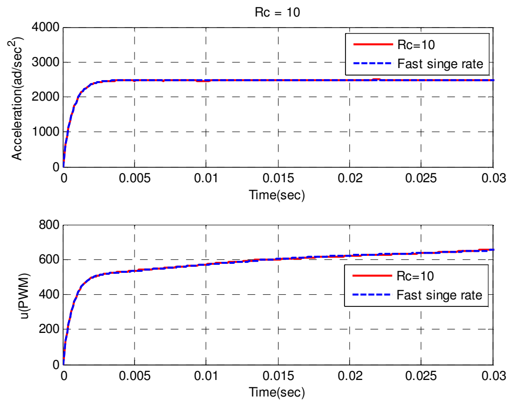


Fig. 7. The schematic diagram of the experimental setup.

In acceleration feedback control, we design the cut-off frequency to be 220 Hz. Experimental step responses are compared in Fig. 8. The slow single rate digital controller (with $R_c = 1$) results in unacceptable resonance excitation because the stepwise discrete-time control effort stimulates the nonmodeled resonance, shown in Fig. 8(a). Obviously, Fig. 8(b) shows that the multirate control with $R_c = 10$ produces a smoother response than the slow single rate control scheme and thereby reduce the resonance excitation. The multirate control produces very similar step response with the fast single rate control, $T_s^* = 100\mu\sec$.



(a)



(b)

Fig. 8. Experimental step response of acceleration feedback control: (a) $R_c = 1$. (b) $R_c = 10$.

In position feedback control, we design the cut-off frequency to be 20 Hz, and a constant torque disturbance 0.1 $N \cdot m$ is added at time instant 0.2 sec. Fig. 9 shows that the disturbance is suppress after 0.08 sec and the position response converges to 50 pulses for multirate digital control with disturbance compensator.

6. CONCLUSIONS

This study has presented a multirate feedback controller based on position, velocity and acceleration feedback using optical encoder information. Firstly, the acceleration and velocity state are estimated using a novel digital state observer, and the estimated velocity is fed into DDOB to estimate and compensate the external disturbance. Then, the proposed cascade feedback control, acceleration-position feedback loop, is applied to improve servo performance without using current feedback sensor. Compared with the conventional cascade control system, current-velocity-position feedback loop, this novel control scheme is computationally efficient, robust and has high bandwidth. Finally, the experimental results validate the feasibility of the proposed multirate digital servo system.

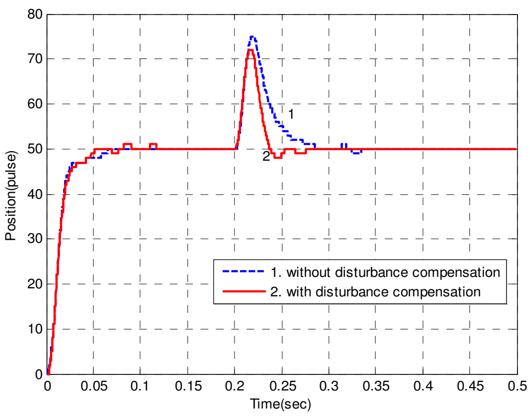


Fig. 9. Results of position feedback control with disturbance.

REFERENCES

Belanger P.R., P. Dobrovolny, A. Helmy, and X. Zhang (1998). Estimation of angular velocity and acceleration from shaft-encoder measurements. *Int. J. Robot. Res.*, **Vol. 17**, **No. 11**, pp. 1225-1233.

Chen C. S. and Y. R. Lee (2006). Auto-tuning digital control based on parameters identification and disturbance observer. *Material Science Forum*, **Vol. 505-507**, pp. 529-534.

Fujimoto H., and Y. Bin (2005). Multirate adaptive robust control for discrete-time non-minimum phase systems and application to linear motors. *IEEE Trans. on Industrial Electronics*, **Vol. 48**, **No. 3**, pp. 636-644.

Fujimoto H., Y. Hori, and A. Kawamura (2001). Perfect tracking control based on multirate feedforward control with generalized sampling periods. *IEEE Trans. on Mechatronics*, **Vol. 10**, **No. 4**.

Gao X.Z. and S. J. Ovaska (2001). Acceleration signal estimation using neural networks. *Measurement Science and Technology*, **Vol. 12**, pp. 1611-1619.

Hsu P. L., Y. C. Houng and S. S. Yeh (2001). Design of an optimal Unknown input observer for load compensation in motion systems. *Asian Journal of Control* , **Vol. 3**, **No. 3**, pp. 204-215.

Kollmorgen (2002). *PLATINUM DDL and SERVOSTAR setup Guide*, Document number: M-LN-016-0702. a Danaher Motion Company.

Kempf C. J. and S. Kobayashi (1999). Disturbance Observer and Feed-forward Design for a High-Speed Direct-Drive Positioning Table. *IEEE Transactions on Control Systems Technology*, **vol. 7**, **no. 5**,, pp. 513-526.

Lee S. H. (2006). Multirate digital control system design and its application to computer disk drives. *IEEE Trans. on Control System Technology*, **Vol. 14**, **No. 1**, pp. 124-133.

Lee S. H. and J. B. Song (2001). Acceleration estimator for low-velocity and low-acceleration regions based on encoder position data. *IEEE/ASME Trans. on Mechatronics*, **Vol. 6**, **No. 1**, pp. 58-64.

M. Mariko, T. Toshiaki, N. Hiroaki, and O. Kouhei (2008). Realization of acceleration control using multirate sampling method. Electrical engineering in Japan. **Vol62**, **No. 4**.

Payette K. (1999). Synchronized motion control with the virtual shaft control algorithm and acceleration feedback. *Proc. of the American Control Conference (ACC 1999)*, pp.2102-2106.

Schmidt P. B. and R. D. Lorenz (1992). Design principles and implementation of acceleration feedback to improve performance of DC drives. *IEEE Trans. On Industry Application*, **Vol. 28**, pp. 594-599.

Väliviita S., S. J. Ovaska and O. Vainio (1999). Polynomial predictive filtering in control instrumentation: a review. *IEEE Trans. On Industry Electronics*, **Vol. 46**, **No. 5**, pp. 876-888.