Feedforward control for SPMSM with Final-State Control
Based on Voltage Limit Circle with Transient Term

I. INTRODUCTION

Permanent magnet synchronous motors (PMSMs) are widely employed in many industrial applications. SPMSM drive systems should achieve quick torque response and wide operating range. For quick current response, authors proposed final-state control considering voltage limit. Final-state control drives an initial state to a final state with a finite-time. The proposed method improves current response and voltage limit. Therefore, the manipulated variable is voltage phase directly by feedback (FB) controller [5][6].

In this paper, a new feedforward control method which improves current response is proposed. Simulations and experiments are performed to compare two types of feedforward control method. In addition, current response under voltage limit is analyzed by the voltage limit circle of SPMSM with transient term.

II. MODEL AND DISCRETIZATION

A. dq Model of SPMSM

The voltage equation of SPMSM in dq-axis is represented in the form of state equation as

\[ \dot{x}(t) = A_e(\omega_e)x(t) + B_e \{ u(t) - [0 \omega_e K_e]^T \}, \]

\[ y(t) = C_e x(t), \]

where \( v_d, q \) are the d-axis and q-axis voltage, \( R \) is the stator winding resistance, \( L \) is the inductance, \( \omega_e \) is the electric angular velocity, \( i_d, q \) are the d-axis and q-axis current, and \( K_e \) is the back EMF constant.

The motor torque \( T \) is given by

\[ T = K_{mt}i_q, \]

where \( K_{mt} := PK_e \) and \( P \) is the number of pole pairs. In this paper, 2-phase/3-phase transform is absolute transformation.

B. Discretization Based on PWM Hold

In order to discretize a plant, a zero-order hold is generally applied. However, in the case of single-phase inverter, it cannot output arbitrary voltage but only 0 or \( \pm E \) (\( E \) is the dc-bus voltage of single-phase inverter). Therefore, in order to control instantaneous values precisely, the zero-order hold is not suitable. When the pulse is allocated in the center of control period \( T_u \), the plant in inverter drive system can be discretized as follows [11].
A continuous-time state equation of a plant is given by
\[ \dot{x}(t) = A_s x(t) + B_s u(t), \quad y(t) = C_s x(t). \] (5)

The precise discrete model in which the input is the ON time \( \Delta T[k] \) is obtained as
\[ x[k+1] = A_s x[k] + B_s \Delta T[k], \quad y[k] = C_s x[k], \] (6)
\[ A_s := e^{A_s T_s}, \quad B_s := e^{A_s T_s/2} B_e, \quad C_s := C_e. \] (7)

In derivation of (6) and (7), the voltage amplitude is considered +E only. Therefore, if \( \Delta T[k] \) is negative, the voltage amplitude is -E.

\section*{C. PWM Hold Model of SPMSM}

(1) is discretized based on PWM hold. Here, it is assumed that the speed variation during one control period can be neglected. Hence, the back EMF \( \omega_s K_e \) can be presumed as the zero-order hold. Therefore, the PWM hold model of SPMSM is designed as \( E = V_{dc} \) in (5) as follows:
\[ x[k+1] = A_s(\omega_e) x[k] + B_s(\omega_e) \Delta T[k], \]
\[ y[k] = C_s x[k], \]
\[ A_s(\omega_e) := e^{A_s(\omega_e) T_s}, \quad B_s(\omega_e) := e^{A_s(\omega_e) T_s/2} B_e, \quad C_s := C_e. \] (8)
\[ B_{s2}(\omega_e) := A_s^{-1}(\omega_e) \left( e^{A_s(\omega_e) T_s} - I \right) B_e, \] (10)
where \( V_{dc} \) is the dc-bus voltage of three-phase inverter, \( \Delta T = [\Delta T_d \; \Delta T_q]^T \), and \( \Delta T_d, q \) are the \( d \)-axis and \( q \)-axis ON time. \( A_s(\omega_e), \quad B_s(\omega_e), \) and \( B_{s2}(\omega_e) \) are functions of \( \omega_e \). Thus, these functions should be calculated again when \( \omega_e \) is changed.

Here, input of three-phase can be generated based on space vector modulation (SVM) [12].

\section*{III. CONTROL SYSTEM DESIGN}

\subsection*{A. Conventional method}

Fig. 1 shows the block diagram of the conventional method. The decoupling control which is given by (11), (12) is applied.
\[ v_{d}^{ref}[k] = v_{d}^{ref}[k] - \omega_e[k] L i_q[k] \] (11)
\[ v_{q}^{ref}[k] = v_{q}^{ref}[k] + \omega_e[k](L i_d[k] + K_e) \] (12)

The current PI controller is designed with pole-zero cancellation as follows:
\[ C(s) = \frac{L s + R}{\tau s}, \quad \tau = 10T_u \] (13)

By discretizing (13), \( C[z] \) is obtained by Tustin transformation with control period \( T_u \). Here, if \( V_a := \sqrt{v_d^2 + v_q^2} > V_{max} \) (\( V_a \) is the voltage amplitude and \( V_{max} := \sqrt{\frac{2}{3} V_{dc}} \) is the maximum value of voltage amplitude.), the integrator of \( C[z] \) is stopped.

The current reference generator makes \( d \)-axis and \( q \)-axis current references \( i_d^* \) and \( i_q^* \) from torque reference \( T^* \). It selects the intersection point of the voltage limit circle with the constant torque line as Fig. 2. The generation of current references is represented by
\[ i_d^*[k] = \frac{-\omega_e L_i_q}{R^2 + \omega_e^2 L^2} + \sqrt{(i_{q\text{max}} - i_q^*)(i_{q\text{min}} - i_q^*)}, \] (14)
\[ i_q^*[k] = \begin{cases} i_{q\text{max}} & \text{if } T^*[k] > K_{mt} i_{q\text{max}} \\ i_{q\text{min}} & \text{if } T^*[k] < K_{mt} i_{q\text{min}} \\ \frac{T^*[k]}{K_{mt}} & \text{otherwise} \end{cases} \] (15)
\[ i_{q\text{max}} := -\omega_e L_i_q R - V_a \sqrt{R^2 + \omega_e^2 L^2}, \] (16)
\[ i_{q\text{min}} := -\omega_e L_i_q R + V_a \sqrt{R^2 + \omega_e^2 L^2}. \] (17)

If \( d \)-axis current reference is selected to strengthen field flux, \( i_d^*[k] = 0 \).

\[ \Delta \theta_e(= 0.5 \omega_e T_u) \] is added to \( \theta_e \) in the \( dq \)-2-phase transform for sampling error compensation [13]. The ON time limit in SVM is described by
\[ \Delta T[k] = \begin{cases} \frac{\Delta T[k]}{\Delta T_{\text{max}}} \Delta T_{\text{max}} & \text{if } |\Delta T[k]| > \Delta T_{\text{max}} \\ \frac{\Delta T[k]}{\Delta T_{\text{max}}} \Delta T_{\text{max}} & \text{otherwise} \end{cases} \] (18)
\[ \Delta T_{\text{max}} := \sqrt{\frac{3}{2} \frac{T_u}{\sqrt{3}}} \] is the maximum value of \( dq \)-axis ON time vector amplitude and \( \Delta T[k] \) is the limiter output. In this paper, all limiters are calculated in the same way as (18).

\subsection*{B. Proposed method}

Fig. 3 and 4 show the block diagrams of the proposed method and FF controller \( C_1[z] \), respectively. The proposed method consists of the 2-DOF control system and has two types of FF controllers.
The inverse system of (8) is derived as follows:

\[
\Delta T_{ff}[k] = -B^{-1}_s(\omega_e)A_s(\omega_e)\hat{x}[k] + B^{-1}_s(\omega_e)x_\delta[k] + 1 \]

+ \[B^{-1}_s(\omega_e)B_{s2}x_\delta(\omega_e)[0 \omega e K_e]^T \] \hspace{1cm} (19)

The inverse system used as a feedforward controller assures perfect tracking if the plant is nominal and the feedforward input is not limited. \(\hat{x}[k] := \begin{bmatrix} \hat{i}_d[k] \hat{i}_q[k] \end{bmatrix}^T \) is the nominal output which takes the input limit into consideration.

If the feedforward input from the inverse system is not limited, the inverse system is used as a feedforward controller. On the other hand, if the feedforward input from the inverse system is limited, the feedforward input is recalculated with the FSC input generator to achieve quick current response. Here, the proposed method switches the type of feedforward controller but the inverse system is equal to final-state control whose prescribe time interval is one. Therefore, the switch does not make unstable.

The FB controller \(C_2[z] \) suppresses the current error \(e[k] \) between \(\hat{x}[k] \) and \(\hat{x}[k] \). Here, \(C_2[z] \) is the same as the FB controller of conventional method. However, in the proposed method, anti-windup control is not necessary because \(\hat{x}[k] \) considers the input limit.

1) Proposed method [8]: Final-state control drives an initial state to a final state by prescribed time interval. Here, final-state control considering voltage limit is expressed as a programming whose evaluation function and constraint functions are quadratic function and quadratic inequality in the form of LMI, respectively. The FF input called FSC input is generated by solving these LMIs.

From PWM hold model of SPMSM, a final state \(x[N] \) is represented with an initial state \(x[0] \) as follows:

\[
Y = \Sigma U 
\]

\[
Y := x[N] - A_N^T x[0] - \Sigma_2 U_{EMF} 
\]

\[
\Sigma := \begin{bmatrix} A_{N-1} B_s & A_{N-2} B_s & \cdots & B_s \end{bmatrix} 
\]

\[
\Sigma_2 := \begin{bmatrix} A_{N-1} B_{s2} & A_{N-2} B_{s2} & \cdots & B_{s2} \end{bmatrix} 
\]

\[
U := \begin{bmatrix} \Delta T^T[0] \Delta T^T[1] \ldots \Delta T^T[N-1] \end{bmatrix}^T 
\]

\[
U_{EMF} := -\omega e K_e \begin{bmatrix} 0 & 1 & \cdots & 0 & 1 \end{bmatrix}^T 
\]

Here, it is assumed that \(\omega e \) is constant in this derivation.

The FF input series \(U \) which satisfies (20) is not determined uniquely. Thereupon, the square sum of the current error is minimized. The evaluation function \(J \) is represented by

\[
J = E^T Q E, \quad Q > 0, 
\]

\[
E := \begin{bmatrix} e[1]^T & e[2]^T & \cdots & e[N]^T \end{bmatrix}^T, 
\]

\[
e[k] := i^* - x[k] = \begin{bmatrix} \hat{i}_d[k] \hat{i}_q[k] \end{bmatrix}^T - x[k], 
\]

\[
E := i^* - Ax[0] - BU - B_2 U_{EMF}, 
\]

\[
A := \begin{bmatrix} A_s & A_s^2 & \cdots & A_s^N \end{bmatrix}^T, 
\]

\[
B := \begin{bmatrix} B_s & 0 & \cdots & 0 \\ A_s B_s & B_s & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ A_s^{N-1} B_s & A_s^{N-2} B_s & \cdots & B_s \end{bmatrix}, 
\]

\[
B_2 := \begin{bmatrix} B_{s2} & 0 & \cdots & 0 \\ A_s B_{s2} & B_{s2} & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ A_s^{N-1} B_{s2} & A_s^{N-2} B_{s2} & \cdots & B_{s2} \end{bmatrix}, 
\]

\[
I^* := (i^*)^T (i^*)^T (i^*)^T 
\]

Furthermore, (8) is controllable. Thus, \(\Sigma \) is full row rank. There, \(\Sigma^\perp \in \mathbb{R}^{2N \times (2N-2)} \) and \(\Sigma^\perp \in \mathbb{R}^{2N \times 2} \) which fulfill \(\Sigma^\perp \Sigma^\perp = 0 \) and \(\Sigma^\perp \Sigma^\perp = I \) respectively are defined. In addition, \(U \) is obtained by

\[
U := [\Sigma^\perp \Sigma^\perp] \hat{U}. 
\]

By assigning (34) to (20), \(Y = \begin{bmatrix} I & 0 \end{bmatrix} \hat{U} \) is given. Therefore, \(\hat{U} \) is represented by

\[
\hat{U} = \begin{bmatrix} Y & q \end{bmatrix}^T, 
\]

where \(q \in \mathbb{R}^{(2N-2) \times 1} \) is a free parameter.

By substituting (29), (34), and (35) to (26), the evaluation function \(J \) is transformed as follows:

\[
J = R(q) + S^T(q)QS(q), 
\]

\[
Z := I^* - Ax[0] - B_2 \Sigma^\perp Y - B_2 U_{EMF}, 
\]

\[
R(q) := Z^T QZ - 2Z^T QS(q), 
\]

\[
S(q) := B \Sigma^\perp q. 
\]
With LMI, the condition which satisfies $J < \gamma$ for provided $\gamma$ is given by (40) [9].

$$
\begin{bmatrix}
\gamma - R(q) & S(q)^T \\
S(q) & Q^{-1}
\end{bmatrix} > 0.
$$  \tag{40}

Next, the voltage limit is described with LMI. The voltage limits at each sampling point are described as

$$
\Delta T^T[k] \Delta T[k] = \Delta T_d^2[k] + \Delta T_q^2[k] \\
\leq \Delta T_{max}^2, \quad (k = 0, 1, \cdots, N - 1).
$$  \tag{41}

Here, $g(i) \in \mathbb{R}^{2 \times 2N}$ ($i := 2k + 1$) of which (1, $i$) th and ($2, i + 1$) th entries are 1 and other entries are 0 is defined. $g(i)$ separates input vector $\Delta T[k]$ as follows:

$$
\Delta T[k] = g(i)U(q).
$$  \tag{42}

With $g(i)$, (41) is transformed as

$$
g(i)U(q))^T g(i)U(q) \leq \Delta T_{max}^2.
$$  \tag{43}

(43) is expressed with $N$ LMIs by

$$
\begin{bmatrix}
\Delta T_{max}^2 & U(g(i))^T \\
U(g(i)) & I
\end{bmatrix} \geq 0,
$$  \tag{44}

where $i = 2k + 1$, $k = 0, 1, \cdots, N - 1$.

By minimizing $\gamma$ under (40) and (44), the FF input, which drives an initial state $x[0]$ to a final state $x[N]$ by the prescribed time interval $N$ for the PWM hold model of SPMSM and minimizes the square sum of current error to fulfill the voltage limit, is obtained.

1) Proposed method 1: Proposed method 1 can be caused inverse response of $q$-axis current because $q$-axis current response is untrammled. Generally, inverse response of $q$-axis current is not desirable, so constraint functions of $q$-axis current response are added in proposed method 1.

$$
f(i) \in \mathbb{R}^{1 \times 2N} \ (i := 2k + 1) \ of \ which \ (1, i + 1) \ th \ entry \ is \ 1 \ and \ other \ entries \ are \ 0 \ is \ defined. \ i_q[k + 1] \ is \ described \ with \ f(i) \ as \ follows:
$$

$$
i_q[k + 1] = f(i) \left( A x[0] + B U + B_2 U_{EMF} \right).
$$  \tag{45}

Therefore, the constraint function which wraps inverse response of $q$-axis current is represented by

$$
f(i) \left( A x[0] + B U + B_2 U_{EMF} \right) - i_q[0] \geq 0.
$$  \tag{46}

By minimizing $\gamma$ under (40), (44), and (46), the FSC input which drives an initial state $x[0]$ to a final state $x[N]$ by the prescribed time interval $N$ for the PWM hold model of SPMSM, minimizes the square sum of current error to fulfill the voltage limit, and prevent inverse response of $q$-axis current is obtained.

Here, the FSC input exists in proposed method 2 because it is assumed that $\omega_e$ is constant in this paper. However, if speed increases during prescribed time interval, FSC input of proposed method 2 could not exist.

2) Proposed method 2: Proposed method 2: Proposed method 1 can be caused inverse response of $q$-axis current because $q$-axis current response is untrammled. Generally, inverse response of $q$-axis current is not desirable, so constraint functions of $q$-axis current response are added in proposed method 2.

$$
\delta = \tan^{-1} \left( -\frac{\Delta T_d}{\Delta T_q} \right).
$$  \tag{48}

The dotted lines in Fig. 5(c), 5(g), and 5(k) are described with LMI. The voltage phase variation was slow and quick response was not achieved. In addition, when the voltage amplitude was not limited, current response still is slow because anti-windup control made closed-loop discontinuous and conventional method is only composed with FB controllers.

In the proposed method 1 and 2, the FSC inputs were generated as $Q = I$. The prescribed time interval $N$ was chosen to be the quickest response (proposed method 1: $N = 40$, proposed method 2: $N = 49$). During the beginning of transient response, inverse response of $q$-axis current happened but $d$-axis current variation is large. This made large $d$-axis current quickly. Finally, proposed method 1 achieved the quickest response. On the other hand, proposed method 2 suppressed inverse response of $q$-axis current and maximum overshoot. However, the settling time is larger than one of proposed method 2.

IV. SIMULATION

The parameters of SPMSM for the simulation are shown in Table I. The dc-bus voltage of the three-phase inverter $V_{dc}$ is 36 [V]. The control period $T_u$ is 0.1 [ms].

Fig. 5 shows simulation results when step torque reference was given and the rotor velocity was 1000 [rpm]. Here, the voltage amplitude of current reference generator was set to be $0.98V_{max}$. The dq-axis ON time vector amplitude $\Delta T_a$ and the dq-axis ON time vector phase (voltage phase) $\delta$ are represented by

$$
\Delta T_a = \sqrt{\Delta T_d^2 + \Delta T_q^2},
$$  \tag{47}

$$
\delta = \tan^{-1} \left( -\frac{\Delta T_d}{\Delta T_q} \right).
$$  \tag{48}

The dotted lines in Fig. 5(c), 5(g), and 5(k) are described with $\Delta T_{max}$. Conventional method cannot operate the voltage phase because it consists of current vector control. Therefore, the voltage phase variation was slow and quick response was not achieved. In addition, when the voltage amplitude was not limited, current response still is slow because anti-windup control made closed-loop discontinuous and conventional method is only composed with FB controllers.

In the proposed method 1 and 2, the FSC inputs were generated as $Q = I$. The prescribed time interval $N$ was chosen to be the quickest response (proposed method 1: $N = 40$, proposed method 2: $N = 49$). During the beginning of transient response, inverse response of $q$-axis current happened but $d$-axis current variation is large. This made large $d$-axis current quickly. Finally, proposed method 1 achieved the quickest response. On the other hand, proposed method 2 suppressed inverse response of $q$-axis current and maximum overshoot. However, the settling time is larger than one of proposed method 2.

V. EXPERIMENT

Experimental results are shown in Fig. 6. Experiments were performed in the same condition as simulations. The parameters of SPMSM and the control period were the same too. The load motor which controlled velocity was the same with the controlled motor. Therefore, for quick current control of load motor, $V_{dc}$ was set to be 80 [V]. In the controller calculation, the three-phase inverter dc-bus voltage was $V_{dcn} = 36$ [V] and the input was multiplied by $V_{dcn}/V_{dc}$ to compensate the difference between $V_{dc}$ and $V_{dcn}$ in SVM. $\Delta \theta$ in experiments is $0.9\omega_e T_u$ to consider the computational time $0.4T_u$.

The variations of $d$-axis reference in all experimental results are due to speed variation. In conventional method,
the voltage phase variation was slow and anti-windup control deteriorated current response.

In the proposed methods, the FSC inputs which are the same as simulations were calculated by off-line. When torque reference was obtained, the FF inputs were output from time series table of FSC input. If the response is not settled during prescribed time interval, the FF input was generated from (19). Proposed method 1 achieved the quickest response. Current is not settled during prescribed time interval because of speed variation. The $q$-axis current response of proposed method 2 is slower than one of proposed method 1 but proposed method 2 improved undershoot of $q$-axis current and maximum overshoot of $d$-axis current.

VI. ANALYSIS CURRENT RESPONSE USING VOLTAGE LIMIT CIRCLE WITH TRANSIENT TERM

This section expresses the voltage limit circle with transient term for SPMSM. In addition, the FSC inputs of both proposed methods are analyzed by using it.

A. Voltage limit circle with transient term

The $dq$-axis voltage equation which is represented by (1) is described under voltage limit ($v_d^2 + v_q^2 \leq V_{\text{max}}^2$) as follows:

$$\left(\frac{v_{\text{max}}}{L}\right)^2 \geq (i_d - C_d)^2 + (i_q - C_q)^2,$$

(49)

$$C_d = -\frac{R}{L}i_d + \omega_e i_q,$$

(51)

$$C_q = -\omega_e i_d - \frac{R}{L} i_q - \frac{\omega_e K_e}{L}.$$

(52)

This is the voltage limit circle with transient term. Here, $C_d$ and $C_q$ are the $d$-axis and $q$-axis center coordinates of the voltage limit circle with transient term, respectively. In general, the voltage limit circle of SPMSM represents $d$-axis and $q$-axis current under voltage limit. The voltage limit circle with transient term expresses $d$-axis and $q$-axis current variation $i_d$ and $i_q$ under voltage limit.

B. Current response of SPMSM under voltage limit

$C_d \geq 0$ and $C_q \leq 0$ when $\omega_e \geq 0$ and the SPMSM is powering. Therefore, the center of voltage limit circle with transient term exists on fourth quadrant as shown in Fig. 7.

![Fig. 7. Voltage limit circle with transient term ($\omega_e \geq 0$, powering)](image-url)
Here, the area on fourth quadrant ($i_d \geq 0$, $i_q \leq 0$) is large, so quick response as $i_q \leq 0$ can be achieved easily. On the other hand, quick response as $i_q \geq 0$ is not expected because the area on second quadrant ($i_d \leq 0$, $i_q \geq 0$) is small. That is to say the operation which extends the area on second quadrant is necessary.

Here, center coordinates $C_d$, $C_q$ are partially differentiated by $-i_d$, $i_q$ as follows:

$$\frac{\partial C_d}{\partial (-i_q)} = \frac{R}{L}, \quad \frac{\partial C_q}{\partial (-i_d)} = \omega_c, \quad (53)$$

$$\frac{\partial C_d}{\partial i_q} = \omega_c, \quad \frac{\partial C_q}{\partial i_d} = -\frac{R}{L}. \quad (54)$$

This means that the center of voltage limit circle with transient term is moved to first coordinate and fourth coordinate by $i_d < 0$ and $i_q > 0$, respectively. Especially, in high-speed operation, the variations of $C_d$ and $C_q$ are subject to $i_q$ and $i_d$, respectively, because of $\omega_c \gg \frac{R}{L}$ as shown in Fig. 8.

Namely, large $i_q > 0$ is undesirable at beginning of transient response because it decreases the area on second coordinate. On the other hand, $i_d < 0$ increases the area on second coordinate. If $d$-axis current is preferentially flowed at beginning of transient response, $q$-axis current response can be improved. Moreover, if inverse response of $q$-axis current is acceptable, a point on third coordinate ($i_d \leq 0$, $i_q \leq 0$) can be selected. The point of maximum $d$-axis current variation exists on third coordinate. In addition, if $d$-axis current is negatively larger than $d$-axis current reference, a point on first coordinate ($i_d \geq 0$, $i_q \geq 0$) can be taken. Thus, quick $q$-axis response is prospected because there is a point of maximum $i_q$. By the aforementioned process, quickest current response is realized.

Fig. 9 shows the transit of voltage limit circle with transient term in simulation. Conventional method selected a point on second coordinate only. Therefore, the area on second coordinate decreased and $q$-axis current response is deteriorated. In order to improve $q$-axis current variation, proposed method 1 moves the center of voltage limit circle to the second coordinate by $q$-axis current undershoot and $d$-axis current overshoot and achieve the quickest response. Proposed method 2 does not change $q$-axis current but $d$-axis
current at beginning of transient response. Thus, the center of voltage limit circle is moved to the first coordinate and it increases the area on second coordinate. However, \(d\)-axis current variation is negatively smaller than one of proposed method 2 because \(q\)-axis current undershoot does not happen. This mentions that changing \(q\)-axis current decreases the settling time better than increasing \(d\)-axis current overshoot. Therefore, maximum \(d\)-axis current overshoot decreases.

The reason why proposed methods can act these operations is because they use PWM hold model and consider voltage limit.

VII. CONCLUSION

For quick current response under voltage limit, our previous paper proposed final-state control considering voltage limit but inverse response of \(q\)-axis current happened. This paper proposes a new FSC input which suppresses inverse response of \(q\)-axis current. Simulations and Experiments are performed to show that proposed method 2 can suppress maximum \(d\)-axis current overshoot and \(q\)-axis current undershoot. However, the settling time of proposed method 2 is longer than one of proposed method 1. This reason is explained by using voltage limit circle with transient term. In short, our proposed methods consider voltage limit circle with transient term exactly.

In our future work, online calculation of proposed methods will be realized and robust feedback controller for modeling error is worked out.

REFERENCES