Multi-Objective Controller Design for Stochastic Model Reference Systems via Sliding Mode Control Concept and LMI Approach

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**Abstract**

In this paper, we design a multi-objective controller \(u(t)\) to achieve the following three objectives simultaneously. They are pole placement, \(H_\infty\) norm constraint and individual error state variance constraint for stochastic model reference systems. By using the invariance property of sliding mode control, the matched model reference input and the plant error term will disappear on the sliding mode of the error system. Combining the upper bound covariance control theory, pole placement skill, \(H_\infty\) norm control theory and linear matrix inequality approach, the controller \(u(t)\) is derived in which the control feedback gain matrix \(G\) is synthesized for achieving the above multiple objectives.

**Keywords:** stochastic model reference systems, LMI, \(H_\infty\) norm constraint, individual error state variance constraint and pole placement constraint.

**1. Introduction**

In practice, we always require to develop some ways for designing controllers to achieve multi-objective. \([1]\) and \([14]\) have discussed the \(H_\infty\) norm and variance constrained problem simultaneously. However, the Riccati equation approach applied by them, which minimizes a scalar cost index, does not ensure satisfying the individual variance constraints. A more straightforward method for designing controllers to achieve variance constraints of individual states is developed in \([10]\), \([12]\) and \([13]\). However, the approach described in \([12]\) does not consider the presence of system perturbations; the system may be unstable when it suffers from perturbations. An improved control method, called Upper Bound Covariance Control (UBCC), which satisfies variance constraints with perturbations is proposed in \([10]\) and \([13]\). Nevertheless, the drawback of the direct UBCC approach is that the state feedback gain designed in \([10]\) and \([13]\) will become very large when the systems suffer from large perturbations.

Pole location is directly associated with performance specifications, such as the setting time a -nd rise time of a control system. For the regional pole constraint, a typical rule for evaluating the relative stability of closed-loop systems is to judge whether all of the poles are located within a prescribed circular region. This specified circular region with the center at \(-q + j0\) \((q > 0)\) and the radius \(\rho\), \((\rho < q)\) is denoted by \(D(-q, \rho)\). This constraint is one of the most frequently employed performance requirements in system control design problems. In this paper, we will consider three system performance requirements: pole placement constraints, \(H_\infty\) norm constraints, and individual variance constraints. To satisfy these three performance constraints, the linear matrix inequality (LMI) approach is a good way to be considered. Because LMI’s intrinsically reflect constraints rather than optimality, many papers tend to offer more flexibility for combining several constraints on the system \([3]\), \([8]\) and \([9]\). Moreover, software like MatLab LMI Control Toolbox is now available to solve such LMI’s in a fast and user-friendly manner.

Owing to the following advantages: simple design, easy implementability and insensitivity to system perturbations, sliding mode control (SMC) has become a successful synthesis method for a system control design and has been applied to many complex systems. The main advantage of SMC system is that the system dynamics in the sliding mode are invariant if parameter uncertainties and/or perturbations satisfy a certain matching condition. Using the concept of SMC, \([4]\) and \([5]\) discuss the covariance control problems in large-scale stochastic systems. In \([7]\), the authors have successfully extended the above approach to linear perturbed systems. Moreover, the authors also have applied this combined technique to deal with the covariance controller design problems for stochastic model reference systems \([6]\). Extending the results of \([6]\), this paper will design a useful controller that can force the systems into simultaneously achieving specified pole location constraints, \(H_\infty\) norm constraints, and individual variance constraints. Therefore, the main contribution of our proposition enables a quick and accurate state tracking response, perturbation rejection, noise attenuation, and robust stability. Here, we pre-define some notations which will be used in the consequent sections, \(\|z(t)\|\) and

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\[ M \] are the 2-norm and induced 2-norm of the vector \( z(t) \) and the matrix \( M \) respectively. \( (\cdot)^T \) denotes the transposition of \( (\cdot) \); \((\cdot)^*\) denotes the conjugate transpose of \( (\cdot) \) and \( \lambda(M) \) denotes the eigenvalue of the matrix \( M \).

2. System Description and Problem Formulation

2.1 System Description

A linear time invariant stochastic plant system (1a), (1b) and the corresponding desired reference model (1c), (1d) are established on a filtered probability space \( (\Omega, F, (F)_t \in \mathbb{R}^r, P) \) and they are described respectively as follows:
The plant:
\[
\begin{align*}
\dot{x}(t) &= A_p x(t) + B_p u(t) + D_p w(t), \\
y(t) &= Fx(t).
\end{align*}
\]

The reference model:
\[
\begin{align*}
\dot{\hat{x}}(t) &= A_m \hat{x}(t) + B_m r(t), \\
\dot{\hat{y}}(t) &= \hat{F} \hat{x}(t),
\end{align*}
\]
where \( x(t) \) and \( \hat{x}(t) \in \mathbb{R}^{nx1} \) are the state vectors of the plant and the reference model, respectively. \( u(t) \in \mathbb{R}^{mx1}, r(t) \in \mathbb{R}^{mx1} \) and \( w(t) \in \mathbb{R}^{nx1} \) are the control input, reference input and the white noise signal input, respectively; \( y(t) \) and \( \hat{y}(t) \in \mathbb{R}^{mx1} \) are the output of the plant and reference model respectively; \( A_m \in \mathbb{R}^{mxm}, F \in \mathbb{R}^{mxm}, B_p \in \mathbb{R}^{nxm}, D_p \in \mathbb{R}^{nxm} \) and \( B_m \in \mathbb{R}^{mxm} \) are the coefficient matrices but \( A_p \in \mathbb{R}^{nxn} \) is a bounded uncertainty. It is assumed that the pair \((A_p, B_p)\) is stabilizable and that the desired reference model is stable.

Define a tracking error state \( e(t) \) and error output \( y_e(t) \) as follows:
\[
\begin{align*}
e(t) &= x(t) - \hat{x}(t), \\
y_e(t) &= y(t) - \hat{y}(t).
\end{align*}
\]
The dynamic of the error system is now directly from (1) and (2) in the following:
\[
\begin{align*}
\dot{e}(t) &= A_p e(t) + (A_p - A_m)x(t) + B_p d(t) + D_p w(t) - B_p r(t), \\
y_e(t) &= F e(t).
\end{align*}
\]

Here, the white noise \( w(t) \) satisfies (4),
\[
E(w(t)) = 0, E(x(0)w(t)) = 0, E(e(0)w(t)) = 0, E(w(t)w^T(t)) = I, \tag{4}
\]
where \( x(0) \) and \( e(0) \) denote the known initial condition of \( x(t) \) and \( e(t) \) respectively; \( I \) denotes the identity matrix. And we also assume that \( A_p, B_p, A_m \) and \( B_m \) belong to the class of matrices that satisfy perfect model matching conditions as follows (see [11])
\[
\begin{align*}
\text{rank} \{ B_p : A_p - A_m \} &= \text{rank} \{ B_p \}, \\
\|A_p - A_m\| &\leq \eta, \text{ where } \eta \text{ is a positive constant.}
\end{align*}
\]

Moreover, we also assume that the range space of matrix \( B_p \) intercepts the range space of the matrix \( D_p \) only at the origin.

2.2 Sliding Phase of the System

First, we define the switching function \( S(t) \) corresponding to \( e(t) \) in the error space as follows:
\[
S(t) = C e(t) - \int_0^t (CA_m + CB_p G)e(t) d\tau, \tag{6}
\]
where \( S(t) = [S_1(t) \cdots S_i(t) \cdots S_m(t)]^T \in \mathbb{R}^{mx1} \); \( C \) and \( G \in \mathbb{R}^{mxn} \) are constant matrices to be designed. \( C \) is chosen such that \( CB_p \neq 0 \) and \( CD_p = 0 \), and \( G \) is the control feedback gain matrix to be determined to deal with the above three problems.

Differentiating (6) with respect to time and using (3) and choosing \( CD_p = 0 \), we obtain
\[
\dot{S}(t) = C(A_p - A_m)x(t) - CB_m r(t) - CB_p G e(t) + CB_p d(t) - (7)
\]
In the sliding mode, \( S(t) = 0 \) holds, we can get the equivalent control as follows:
\[
u_e(t) = G e(t) - (CB_p)^{-1}C (A_p - A_m)x(t) - B_m r(t). \tag{8}
\]
By substituting (8) into (3), we have
\[
\dot{e}(t) = (A_p + B_p G)e(t) + D_p w(t) - (8a)
\]
\[
y_e(t) = F e(t). \tag{9b}
\]

Let the plant error term and the reference model input be regarded as the perturbation to the error system. With the aid of the matching condition (5) and the invariance property of SMC, the dynamics (9) is insensitive to these perturbations. Thus, (9) is reduced to
\[
\dot{e}(t) = (A_p + B_p G)e(t) + D_p w(t), \tag{10a}
\]
\[
y_e(t) = F e(t). \tag{10b}
\]

2.3 Hitting Phase of the System

This subsection tries to find a controller \( u(t) \) such that the error states of the system (3) can be forced to the sliding surface. Let us define a Lyapunov function
\[
V(S(t)) = S^T(t) S(t) = S_1^2(t) + \cdots + S_m^2(t) + \cdots + S_m^2(t). \tag{11}
\]
We have the following lemma and theorem.

Lemma 2.1 [6]

Consider the system (3) and let the Lyapunov function \( V(S(t)) \) be assigned as (11). If \( w(t) \) is satisfied with (4) and \( CD_p = 0 \) holds, then we can obtain
\[
\frac{d}{dt} V(S(t)) = 2 S^T(t) S(t). \tag{12}
\]

Theorem 2.1 [6]
For the system (3), let \( CD_p = 0 \) and the controller \( u(t) \) be
\[
u(t) = G(t) - (CR_p)^{-1}\left[k_1 E(t) + \frac{k_2}{\eta} + \frac{k_3}{\eta} + \frac{k_4}{\eta} + \alpha \right] \text{sgn}(S(t)), \tag{13}\]
where \( k_1 > \eta \), \( \eta \) is bounded for uncertain \( A_p - A_m \); \( \alpha \) is an arbitrary positive number, and
\[
\text{sgn}(S(t)) = \left[ \text{sgn}(S_1(t)) \cdots \text{sgn}(S_\ell(t)) \cdots \text{sgn}(S_m(t)) \right]^T
\]
in which \( \text{sgn}(S_i(t)) = \begin{cases} 1 & S_i(t) > 0 \\ 0 & S_i(t) = 0 \\ -1 & S_i(t) < 0 \end{cases} \). Then the error state of system (3) will converge to the sliding surface.

2.4 Problem Formulation of the System

The goal of this paper is to design the control \( u(t) \) for the error system (3) to satisfy the following objectives

Objective (i): Performance Level of Noise Attenuation

In the error system (3), the effect of the noise input \( w(t) \) on the output \( y_e(t) \) should be kept small for the system. Under the assumption, the error system (3) is controlled to be stable, let \( H(s) \) denote the closed-loop transfer function from \( w(t) \) to \( y_e(t) \). The desired \( H_{\infty} \) performance level is described as
\[
\|H(s)\|_\infty = \sup_{\|w(t)\| = 1} \left\| \frac{y_e(t)}{w(t)} \right\| < \gamma \tag{14}\]
where the performance level upper bound \( \gamma \) can be implemented as a constraint to be met or a parameter to be minimized during the controller construction.

Objective (ii): Constraints on Pole Placement Region

In this paper, the region of the complex \( z \)-plane is described by the LMI condition \[ D = \{z \in C : f_D(z) = U + zN + \Sigma N^T < 0 \} \tag{15}, \]
where \( C \) denotes the set of complex numbers, \( U = U^T \) and \( N \) are real matrix parameters for choosing a suitable convex region by defining the characteristic function \( f_D(z) \). Specifically, we consider the region of the disk \( D(-q, \rho) \) with center at \( (-q, 0) \) and radius \( 0 < \rho < q \) for the closed-loop pole of the error system. The disk region \( D(-q, \rho) \) in the complex plane \( z = x + jy \) can be described as
\[
\left( q + e_x \right)^2 + e_y^2 = (q + z)(q + \bar{z}) < \rho^2. \tag{16}\]
By the property of Schur’s complement \[3\], we have the characteristic function \( f_D(z) \) of (16) within the disk region \( D(-q, \rho) \) as follows:
\[
f_D(z) = \begin{bmatrix} -\rho & q + z \\ q + \bar{z} & -\rho \end{bmatrix} < 0. \tag{17}\]
In comparison with the defined LMI condition (15), the matrix parameters for the disk region \( D(-q, \rho) \) are
\[
U = \begin{bmatrix} -\rho & q \\ q & -\rho \end{bmatrix}, \quad N = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix}. \tag{18}\]
The considered error system is called \( D \)-stable if the eigenvalues of the error system are located in the disk region, i.e.,
\[
\lambda(A_m + B_p G) \in D(-q, \rho). \tag{19}\]
The closed-loop poles of the error system are specified in terms of the system matrix \( A_m + B_p G \) and required to lie within the disk region \( D(-q, \rho) \) with suitable chosen parameters \( q > \rho > 0 \).

Objective (iii): Constraints on Upper Bound Error State Covariance

The individual error state variance constraints can be required to satisfy the following upper bound constraints:
\[
\left[ \frac{x_i}{\sigma_i} \right]_t = \text{Var}(e_i(t)) \leq \left[ \frac{x_i}{\sigma_i} \right]_0 \leq \sigma_i^2, \quad i = 1, 2, \ldots, n, \tag{20}\]
where \( \text{Var}(e_i(t)) \) and \( \sigma_i \) denote the variance value and root mean square (RMS) constraints for the \( \ell \)-th state of the error system, \( \left[ \frac{x_i}{\sigma_i} \right]_0 \) denotes the \( \ell \)-th diagonal element of upper bound error state covariance matrix \( \tilde{\Sigma} \) and \( \left[ \frac{x_i}{\sigma_i} \right]_t \) denotes the \( \ell \)-th diagonal element of the matrix \( \xi \) which is defined as
\[
\xi = \lim_{t \to \infty} E(e(t)e^T(t)). \tag{21}\]

3. Design of the Gain Matrix \( G \) for Satisfying Multi-objective Constraints

In this section, the design of control feedback gain matrix \( G \) is conducted for the error system (10) to achieve the multi-objective performance constraints in terms of the LMI conditions. It is known that state covariance \( \tilde{\Sigma} \) defined in Section 2 satisfies the following Lyapunov equation
\[
(A_m + B_p G)\tilde{\Sigma} + \tilde{\Sigma}(A_m + B_p G)^T + D_p D_p^T = 0. \tag{22}\]
For the moment, let us consider some useful lemmas as follows.

Lemma 3.1

In the error system (10), let \( \gamma > 0 \) be a fixed scalar. If there exist positive definite matrix \( \tilde{\Sigma} \) and matrix \( L \) such that the following LMI condition holds
\[
\begin{bmatrix}
    A_m \tilde{\Sigma} + B_p L + \tilde{\Sigma} A_m^T + L^T B_p^T + D_p D_p^T & \tilde{\Sigma} \\
    L \tilde{\Sigma} & -\gamma I
\end{bmatrix} < 0 \tag{23}
\]
where \( L = G\tilde{\Sigma} \). Then, the \( H_{\infty} \) norm constraint (14) is satisfied. Furthermore, in this case, we have
\(\xi \leq \tilde{\xi}. \quad (24)\)

**Proof:**

First, we define the Lyapunov function for the system dynamics (10) as following:
\[
V(e(t)) = e^T(t) \tilde{\xi}^{-1} e(t),
\]
where \(\tilde{\xi} = \xi^T > 0\) is the upper bound of \(\xi\). According to Itô’s differential rule, the time derivative of the quadratic Lyapunov function is
\[
d\frac{d}{dt} V(e(t)) = e^T(t) \tilde{\xi}^{-1} \frac{d}{dt} (A_e + B_p G) e(t) + e^T(t) (A_e + B_p G)^T \tilde{\xi}^{-1} e(t) + \text{trace} \left(D_p \tilde{z}^{-1} D_p (w(t)) w(t) \right).
\]
(26)

One knows that the error system (10) is asymptotically stable due to \(d\frac{d}{dt} V(e(t)) < 0\). From equation (29), letting \(t \to \infty\) and combining expression (30) is equivalent to (17). The feasibility of pole region (17) is equal to the matrix inequality \(M_p(\hat{A}, \hat{\xi}) = U \otimes \hat{\xi} + N \hat{\theta}(\hat{A} \hat{\xi}^T + \hat{\xi} \hat{A}^T)\), which was proven in [8] and as a counterpart of Gutman’s theorem for LMI regions. Since the expressions of \(M_p(\hat{A}, \hat{\xi})\) in (37) and \(f_D(z)\) in (17) are related by the substitution \((\hat{\xi}, \hat{A} \hat{\xi}^T, \hat{\xi} \hat{A}^T) \mapsto (1, z, \bar{z})\), the matrix inequality condition for the disk region \(D(-q, \rho)\) as shown in (17) can be written as
\[
\begin{bmatrix}
-\rho \hat{\xi}^2 & q \hat{\xi} \hat{A}^T + A \hat{\xi} \\
q \hat{\xi}^T A^T & -\rho \hat{\xi}^2
\end{bmatrix} < 0.
\]
(38)

By the substitution of \(\hat{A} = A_e + B_p G\) and \(L = G \hat{\xi}\), we obtain the LMI condition as (33). The proof is completed.

**Lemma 3.3**

Consider the desired upper bound error state covariance constraint on the error system as described in (20). Let \(\sigma_\ell > 0\) is given. If there exist positive definite matrix \(\hat{\xi}\) such that the following LMI condition holds
\[
\begin{bmatrix}
\sigma^2 & I_{\ell} \xi \\
\xi I_{\ell}^T & \xi 
\end{bmatrix} \geq 0, \quad \ell = 1, 2, \ldots, n, 
\]
(39)

where \( I_{\ell} = [0 \cdots 1 \cdots 0] \in R^{1 \times n} \) denotes a row vector with the \( \ell \)-th element is 1 and others are 0. Then, the upper bound error state covariance constraint can be achieved.

**Proof:**

Rewriting (20), one has
\[
\sigma^2 - I_{\ell} \xi \xi^T - \frac{1}{\xi^T I_{\ell}^T} \xi I_{\ell}^T \geq 0, \quad \ell = 1, 2, \ldots, n, 
\]
(40)

Using the property of Schur’s complement [3], (40) can be reformulated as (39). The proof is completed.

Now, we synthesize those above performance constraints that have composed the related linear matrix inequalities to the following theorem 3.1.

**Theorem 3.1**

In the error system (10), given \( \gamma > 0, \sigma > 0, \rho > 0 \). If there exist positive definite matrix \( \xi \) and matrix \( L \) such that the following LMIs (23), (33) and (39) hold. Then, the feasible solution of the control feedback gain \( G \) can achieve multi-objective (i) ~ (iii).

**Proof**

From the proofs of Lemma 3.1, Lemma 3.2 and Lemma 3.3, one knows that multi-objective (i) ~ (iii) can be achieved by a convex optimization problem with constraints (23), (33) and (39). In other word, if matrices \( \xi \) and \( L \) exist and satisfy LMIs, then the control feedback gain \( G \) can achieve multi-objective (i) ~ (iii) and will be obtained by
\[
G = L \xi^{-1}. 
\]
(41)

The proof is completed.

Here, we should check whether \( \|H(s)\|_{\infty} \leq \gamma \) holds or not, the following lemma will be helpful.

**Lemma 3.4 [2]**

Consider the system (10). There exists a positive scalar \( \gamma \) to satisfy \( \|H(s)\|_{\infty} \leq \gamma \) if and only if \( M_{\gamma} \) has no eigenvalues on the imaginary axis, where
\[
M_{\gamma} = \begin{bmatrix}
A_p + B_p G & \gamma^{-1} D_p D_p^T \\
-A_p D_p^{-1} F^T F^{-1} & -(A_p + B_p G)^T
\end{bmatrix}. 
\]

4. A numerical example

Suppose the plant and the reference model are given as (1) with \( A_p = \begin{bmatrix} 0 & 1 \\ 3\delta & \delta \end{bmatrix}, \quad B_p = \begin{bmatrix} 0 \\ 1 \end{bmatrix} \),
\[
D_p = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, \quad A_m = \begin{bmatrix} 0 & 1 \\ -1 & -3 \end{bmatrix}, \quad B_m = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad \text{and} 
F = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, \quad \text{in which} \quad -1 \leq \delta \leq 1. \] The design goals are to find the controller \( u(t) \) such that the steady error state of the error system satisfies the following requirements.

\[
q = 36, \quad \rho = 35, 
\]
(42)
\[
\|H(s)\|_{\infty} \leq 1, 
\]
(43)
\[
\text{Var}(e_1(t)) < 1.2, \quad \text{Var}(e_2(t)) < 2.5. 
\]
(44)

Suppose that the \( r(t) = 1 \) and the white noise \( w(t) \) satisfies (4) with identity covariance. Moreover, we also choose the initial values of \( x(t) \) and \( \dot{x}(t) \) as follows: \( x(0) = [x_0(0), x_{\hat{0}}(0)]^T = [5, 4]^T, \quad \dot{x}(0) = [\dot{x}_0(0), \dot{x}_{\hat{0}}(0)]^T = [-4, 3]^T \). Then, the proposed design procedure may be obtained as follows.

Step 1: Choosing \( C = [0 \ 1] \) such that \( CD_p = 0 \) and \( CB_p \neq 0 \). Then the sliding mode is expressed by equation (10).

Step 2: From the main theorem, we get the feasible solutions of upper bound error state covariance matrix \( \xi = \begin{bmatrix} 0.6405 & -0.8780 \\ -0.8780 & 2.2238 \end{bmatrix} \) which has diagonal elements satisfying the performance constraints (44), and matrix \( L = [-1.3861, -43.2667] \). Therefore, the control feedback gain \( G \) can be obtained from (41) as follows.
\[
G = [-62.851, -44.271]. 
\]
(45)

Step 3: From (6), the switching function has the following form.
\[
S(t) = [0 \ 1] e(t) - \int_0^t [-63.851, -47.271] e(\tau) d\tau. 
\]
(46)

Step 4: From (13), the controller \( u(t) \) becomes
\[
u(t) = [-62.851, -44.271] e(t) - \frac{5.8}{1 + \|t\| + 1} \text{sgn}(S(t)) \text{sgn}(S(t)) 
\]
(47)

where \( k_3 = 5.8 \) and \( \alpha = 1 \) are chosen.

From the above design procedure, we can conclude that the upper bound error state covariance \( \xi \) will be achieved if the system is driven by the controller (47). In the simulation, we use random number, which value varies between -1 and 1, to replace \( \sigma \). The simulation results for the time responses of \( x_1(t) \), \( \dot{x}_1(t) \), \( x_2(t) \) and \( \dot{x}_2(t) \) are shown in Fig. 1~Fig. 2, respectively. It is easy to check that the matrix \( M_{\gamma} \), which is defined in Lemma 3.4, have no eigenvalues on the imaginary axis; hence the \( H_{\infty} \) norm constraint (43) is satisfied. Moreover, the variances of \( e_1(t) \), and \( e_2(t) \) are 0.2964 and 0.5696 , respectively. Therefore, the individual variance constraints (44) are also achieved. We also check the poles of error system (3), which locate at \(-1.3917 \) and \(-45.8793 \) that satisfy the poles location constraint in (42).

5. Conclusion

This paper has applied the invariance property of SMC to the stochastic model reference systems such that the plant error term and reference model
input can be ignored for the systems. Since the utilization of SMC and UBCC, the designed control feedback gain matrix $G$ not only achieves the multi-objective performance constraints for the system but also determines the sliding surface of the system. Furthermore, in order to obtain the feasible solution of $G$, the LMI approach is applied. Finally, a numerical example is used to illustrate our proposed method. In this paper, it has been shown that the present approach is a new trial to combine SMC and LMI methods for achieving multi-objective of a system. Moreover, the new scheme to high performance complex system will be developed in the future.

6. References


Acknowledgement

The authors are grateful to the National Science Council of Taiwan, ROC, for partial financial support under the Grant NSC 93-2218-E270-003.