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Two-Mode-Dependent Controller Design for Networked Markov System With Time-Delay in Both S/C Link and C/A Link

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ABSTRACT The two-mode-dependent controller design problem for networked Markov system with time-delay in both S/C link and C/A link is investigated in this paper. Two independent Markov chains are used to describe the time-delay in S/C link and C/A link. A two-mode-dependent state feedback controller is proposed that depends on both the S/C time-delay and the mode of the Markov controlled plant. The sufficient conditions on the stochastic stability of the closed-loop system are established. The design method of the controller is also proposed on condition that the transition probability matrices of S/C time-delay and mode of the controlled plant are completely known and partly unknown respectively. A numerical example is exploited to illustrate the effectiveness and superiority of the proposed method.

INDEX TERMS Markov jump system, stochastic stability, two-mode-dependent, closed-loop system, stabilization.

I. INTRODUCTION

Networked control system (NCS) has gained great attentions during the past decades and it is applied widely in real-time industrial control, environmental monitoring, military, telemedicine and other fields [1]–[4]. The stability analysis and controller design for NCS with time-delay and data packet dropout has become a hot research filed on account of its essential impact on modern control theory, and a great many literatures have been reported [5]–[7].

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One of the focuses of research for NCS is the time-delay, which may degrade the performance of the system or even cause instability [8], [9]. How to explicitly incorporate the time-delay into the controller design is the object of the related research. Modeling the time-delay as a random sequence of Bernoulli distribution is one of the common methods in research of NCS. However this method can only deal with one-step time-delay [10], [11]. The time-delay was modeled as a random sequence of Bernoulli distribution, and the robust $\mathcal{H}_\infty$ filtering problem for NCS with both random time-delay and packet dropout was researched. The sufficient condition on making the filtering error system exponentially stable was given [10]. The optimal linear estimation problem of NCS with random time-delay and packet dropout was researched. Two random variables that satisfied Bernoulli distribution were used to describe the one-step random time-delay and multiple packet loss that may exist in network data transmission. The optimal linear state filter, predictor and smoother under linear minimum variance were proposed [11].

Another method to deal with the time-delay is to model the time-delay as a Markov chain. The Markov chain can not only describe the dependency between the current time-delay and the previous time-delay, but also include packet dropout, and it is an effective method to describe the time-delay in NCS [12]–[16]. A time-delay compensation control scheme was proposed, in which the random time-delay was modeled as a Markov chain, thus the closed-loop system was modeled as a Markovian jump linear system (MJLS). To perform the stability analysis, a new necessary and sufficient condition was established [12]. The time-delay τ_k from sensor to controller (S/C) and the time-delay γ_k from controller to actuator (C/A) were modeled as two Markov chains. An asymptotic mean-square stability criterion was established to compensate for the random time-delay and packet losses in both S/C link and C/A link [13]. The state feedback control problem for a class of nonlinear NCS with S/C time-delay τ_k was researched. The state augmentation method was used to obtain the model of the closed-loop system based on that τ_k was modeled as a Markov chain. The sufficient conditions on stochastic stability of the closed-loop system were given [14]. Considering the S/C time-delay τ_k and C/A time-delay γ_k, based on the free weight matrix method, the sufficient conditions on the stochastic stability of closed-loop systems under $\mathcal{H}_\infty$ performance constraints were obtained. The design method of mode-dependent $\mathcal{H}_\infty$ state feedback controller was given [15]. Both the S/C time-delay and C/A time-delay were modeled as Markov processes, and the resulting closed-loop system was modeled as a MJLS. A state feedback controller that made the closed-loop system stochastically stable was designed, which could be solved by the proposed algorithm [16].

In all the aforementioned references, the controlled plant was described by the deterministic model which was linear or certainty kind of nonlinear system. However, the deterministic model cannot be used to represent the behavior of many actual systems, because the phenomena embodied by these systems are not specific, and their structures and parameters have characteristics of random changes. These random changed are often caused by system jumps, such as random failures and repairs of system components, changes in internal interconnected systems, sudden changes in the environment. In this regard, Markov controlled plant exists widely in communication, power systems, and aircraft control systems [17]–[19]. Therefore, the design and research of NCS controller based on Markov controlled plant has important theoretical and practical significance.

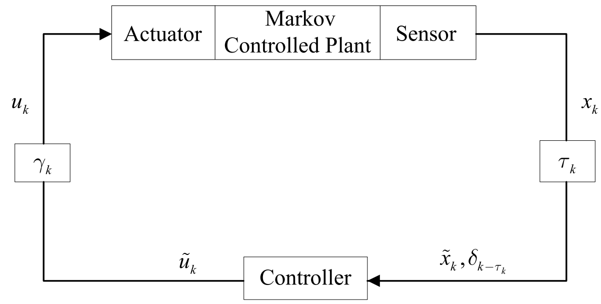

FIGURE 1. Structure of networked Markov system with time-delay.

Some literatures for Markov controlled plant based NCS have been reported [20], [21]. The event-triggered $\mathcal{H}_\infty$ control problem for networked Markov jump system subject to repeated scalar nonlinearities was researched. An event-triggered transmission scheme was adopted and an event generator was presented between the controller and the sensor [20]. The stabilization problem for a kind of networked Markovian jump systems with random time-delay was researched. The closed-loop system model was established through the state augmentation technique and the necessary and sufficient conditions on the stochastic stability were derived [21]. However, the controller in [20] and [21] was designed without the consideration of the system mode or even independent of the time-delay. To the best of the authors' knowledge, involving time-delay and system mode to design the state feedback controller has not been researched, which motivate this investigation. The main contributions of this paper can be exhibited in the following two aspects:

1) Through the analysis of time-delay and system mode information, a two-mode-dependent state feedback controller that simultaneously depends on both τ_k and $\delta_{k-\tau_k}$ is proposed for the networked Markov system.
2) By constructing proper Lyapunov-Krasovskii functional, the design method of state feedback controller gain matrix is derived on condition that the transition probabilities are completely available and partly unavailable, respectively.

The rest content of the paper is organized as follows. In Section II, the available time-delay and system mode information is analyzed and a two-mode-dependent controller is proposed. The sufficient conditions on the stochastic stability of the closed-loop system are presented first and the equivalent conditions with constrains are derived in section III. In Section IV, a simulation example is given to illustrate the effectiveness of the proposed controller. The conclusions are addressed in Section V.

II. PROBLEM FORMULATION AND PRELIMINARIES

The structure of the NCS with random time-delay considered in this paper is shown in Figure 1, where the controlled plant is a MJLS and the state equation of which is as follows:

$$x_{k+1} = A_{\delta_k} x_k + B_{\delta_k} u_k \tag{1}$$

where x_k is the system state vector, u_k is the control input vector, A_{δ_k} and B_{δ_k} are known real constant matrices with appropriate dimensions. δ_k takes value from the set $\mathcal{W} = \{1, \cdots, D\}$, and the transition probability matrix of δ_k is $\Theta = [\rho_{pq}]$, where ρ_{pq} is defined as $\rho_{pq} = \Pr\{\delta_{k+1} = q|\delta_k = p\}$, $\sum_{q=1}^{D} \rho_{pq} = 1, \rho_{pq} \geq 0, p, q \in \mathcal{W}$.

τ_k and γ_k stands for the time-delay in S/C link and the time-delay in C/A link and takes value from the finite set $\mathcal{M} = \{0, \cdots, \tau\}$, $\mathcal{N} = \{0, \cdots, \gamma\}$, respectively. The transition probability matrix of τ_k and γ_k is $\Xi = [\omega_{ij}]$ and $\Pi = [\pi_{rs}]$, respectively, where ω_{ij} and π_{rs} is defined as $\omega_{ij} = \Pr\{\tau_{k+1} = j|\tau_k = i\}$, $\pi_{rs} = \Pr\{\gamma_{k+1} = s|\gamma_k = r\}$, respectively, where $\sum_{j=0}^{\tau} \omega_{ij} = 1, \sum_{s=0}^{\gamma} \pi_{rs} = 1, \omega_{ij} \geq 0, \pi_{rs} \geq 0, i, j \in \mathcal{M}, r, s \in \mathcal{N}$.

On one hand, due to the existence of S/C time-delay τ_k, the system state obtained at the controller node at time instant k is as follows:

$$\tilde{x}_k = x_{k-\tau_k} \tag{2}$$

On the other hand, at time instant k, the information of the S/C time-delay τ_k and the system mode $\delta_{k-\tau_k}$ is available to the controller node. Hence, the state feedback controller that depends both τ_k and $\delta_{k-\tau_k}$ can be designed as follows:

$$\tilde{u}_k = K_{\tau_k,\delta_{k-\tau_k}} x_{k-\tau_k} \tag{3}$$

Apparently, the controller in (3) is two-mode-dependent.

Due to the existence of C/A time-delay γ_k, the control input acting on the controlled plant at time instant k is as follows:

$$u_k = \tilde{u}_{k-\gamma_k} \tag{4}$$

The state equation of the closed-loop system can be obtained from (1)-(4):

$$x_{k+1} = A_{\delta_k} x_k + B_{\delta_k} K_{\tau_k,\delta_{k-\tau_k}} x_{k-\tau_k-\gamma_k} \tag{5}$$

Remark 1: By applying the two-mode-dependent controller in (3), the resulting closed-loop system (5) is not a standard MJLS, due to the fact that the closed-loop system depends on δ_k, τ_k, $\delta_{k-\tau_k}$ and $\tau_k + \gamma_k$. Furthermore, $\delta_{k-\tau_k}$ is related with both δ_k and τ_k, which makes the system stability analysis and controller design more complex.

The objective of this paper is to design the two-mode-dependent controller in (3) to guarantee the stochastic stability of the closed-loop system (5).

The notion of stochastic stability is introduced as follows:

Definition 1 [22]: The closed-loop system (5) is stochastically stable if for every initial state x_0 and initial mode $\tau_0 \in \mathcal{M}$, $\delta_{-\tau_0} \in \mathcal{W}$, there exists a positive-definite matrix $R > 0$ such that $E\left\{\sum_{k=0}^{\infty} \| x_k\|^2 | x_0, \tau_0, \delta_{-\tau_0}\right\} < x_0^T R x_0$ holds.

Before the main results are given, two related lemmas are introduced as follows:

Lemma 1 [23]: If the transition probability matrix from δ_k to δ_{k+1} is Θ, then the transition probability matrix from $\delta_{k-\tau_k}$ to $\delta_{k+1-\tau_{k+1}}$ is $\Theta^{1+\tau_k-\tau_{k+1}}$, which is still a transition probability matrix.

Lemma 2 [24]: $(\theta - \theta_0 + 1) \sum_{\rho=\theta_0}^{\theta} \upsilon_\rho^T X \upsilon_\rho \geq \sum_{\rho=\theta_0}^{\theta} \upsilon_\rho^T X \sum_{\rho=\theta_0}^{\theta} \upsilon_\rho$ holds for any positive-definite matrix $X > 0$ and arbitrary vector υ, where θ and θ_0 two scalars which satisfies $\theta \geq \theta_0 \geq 1$.

Remark 2: If an embedded processor is placed at the actuator node, by comparing the current time with the time-stamp of the control input received by the embedded processor, the C/A time-delay γ_k can be calculated, and the information of $\gamma_{k-\tau_k}$ at time instant k would be available to the controller node. Further, the controller $\tilde{u}_k = K_{\tau_k,\gamma_{k-\tau_k},\delta_{k-\tau_k}} x_{k-\tau_k}$ which simultaneously depends on τ_k, $\delta_{k-\tau_k}$ and $\gamma_{k-\tau_k}$ can be designed. However, this will increase the system cost and make the system structure more complicated.

III. MAIN RESULTS

In this section, the sufficient conditions on the stochastic stability for the closed-loop systems (5) will be presented and the equivalent conditions of linear matrix inequalities (LMIs) with nonconvex constraints will be derived.

Theorem 1: Under the state feedback control law (3), the resulting closed-loop system (5) is stochastically stable if there exist positive-definite matrices $P_{i,r} > 0$, $P_{j,q} > 0$, $S_1 > 0, S_2 > 0, Z > 0$ and matrix $K_{i,r}$ such that the following matrix inequality

$$\Lambda = \begin{bmatrix} \Lambda_{11} & * & * \\ \Lambda_{21} & \Lambda_{22} & * \\ 0 & Z & -S_2 - Z \end{bmatrix} < 0, \tag{6}$$

where

$$\begin{aligned}
\Lambda_{11} &= A_{\delta_k}^T \tilde{P}_{j,q} A_{\delta_k} + (\tau+\gamma)^2 \left(A_{\delta_k} - I\right)^T Z \left(A_{\delta_k} - I\right) \\
&\quad + (\tau + \gamma + 1) S_1 + S_2 - Z - P_{i,p}, \\
\Lambda_{21} &= A_{\delta_k}^T \tilde{P}_{j,q} B_{\delta_k} K_{i,p} + (\tau+\gamma)^2 \left(A_{\delta_k} - I\right)^T Z B_{\delta_k} K_{i,p} \\
&\quad + Z, \\
\Lambda_{22} &= \left(B_{\delta_k} K_{i,p}\right)^T \tilde{P}_{j,q} B_{\delta_k} K_{i,p} - 2Z - S_1 \\
&\quad + (\tau+\gamma)^2 \left(B_{\delta_k} K_{i,p}\right)^T Z B_{\delta_k} K_{i,p}, \\
\tilde{P}_{j,q} &= \sum_{j=0}^{\tau} \sum_{q=1}^{D} \omega_{ij} \Theta_{pq}^{i-j+1} P_{j,q},
\end{aligned}$$

holds for all $i, j \in \mathcal{M}, p, q \in \mathcal{W}$. ■

Proof: For the closed-loop system (5), construct the following Lyapunov-Krasovskii functional:

$$V(x_k, \tau_k, \delta_{k-\tau_k}) = \sum_{l=1}^{4} V_l(x_k, \tau_k, \delta_{k-\tau_k}) \triangleq x_k^T \Omega_{\tau_k,\delta_{k-\tau_k}} x_k,$$

where

$$V_1(x_k, \tau_k, \delta_{k-\tau_k}) = x_k^T P_{\tau_k,\delta_{k-\tau_k}} x_k,$$

$$V_2(x_k,\tau_k,\delta_{k-\tau_k}) = \sum_{n=-\tau-\gamma+1}^{0}\sum_{m=k+n}^{k-1} x_m^T S_1 x_m + \sum_{l=k-\tau_k-\gamma_k}^{k-1} x_l^T S_1 x_l,$$

$$V_3(x_k,\tau_k,\delta_{k-\tau_k}) = \sum_{l=k-\tau-\gamma}^{k-1} x_l^T S_2 x_l,$$

$$V_4(\chi_k,\tau_k,\delta_{k-\tau_k}) = \sum_{n=-\tau-\gamma+1}^{0}\sum_{m=k+n}^{k-1}(\tau+\gamma)\chi_m^T Z\chi_m,$$

$$\chi_m = x_{m+1}-x_m.$$

It is noted that $\Omega_{\tau_k,\delta_{k-\tau_k}} > 0$.

$$E\{\Delta V_1\} = E\left\{x_{k+1}^T P_{\tau_{k+1},\delta_{k+1-\tau_{k+1}}} x_{k+1} \,|\tau_k=i,\,\delta_{k-\tau_k}=p\right\} - x_k^T P_{\tau_k,\delta_{k-\tau_k}} x_k.$$

From Lemma 1, one can obtain that the transition probability matrix from $\delta_{k-\tau_k}$ to $\delta_{k+1-\tau_{k+1}}$ is Θ^{i-j+1}, and the transition probability from $\delta_{k-\tau_k}$ to $\delta_{k+1-\tau_{k+1}}$ under the transition probability matrix Θ^{i-j+1} is denoted as Θ_{pq}^{i-j+1}. Hence, one has:

$$\begin{aligned}
&E\left\{x_{k+1}^T P_{\tau_{k+1},\delta_{k+1-\tau_{k+1}}} x_{k+1}\,|\tau_k=i,\,\delta_{k-\tau_k}=p\right\} - x_k^T P_{\tau_k,\delta_{k-\tau_k}} x_k\\
&= E\left\{\left(A_{\delta_k}x_k + B_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\right)^T\right\}\sum_{j=0}^{\tau}\sum_{q=1}^{D}\omega_{ij}\Theta_{pq}^{i-j+1}\\
&\quad P_{j,q}\left(A_{\delta_k}x_k + B_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\right) - x_k^T P_{i,p}x_k\\
&= x_k^T A_{\delta_k}^T \tilde{P}_{j,q}A_{\delta_k}x_k + x_k^T A_{\delta_k}^T\tilde{P}_{j,q}B_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad + x_{k-\tau_k-\gamma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T\tilde{P}_{j,q}B_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad + x_{k-\tau_k-\sigma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T\tilde{P}_{j,q}A_{\delta_k}x_k - x_k^T P_{i,p}x_k.
\end{aligned} \tag{7}$$

$$\begin{aligned}
&E\{\Delta V_2\}\\
&= (\tau+\gamma)x_k^T S_1 x_k - \sum_{l=k+1-\tau-\gamma}^{k} x_l^T S_1 x_l + x_k^T S_1 x_k\\
&\quad - x_{k-\tau_k-\gamma_k}^T S_1 x_{k-\tau_k-\gamma_k} + \sum_{l=k+1-\tau_{k+1}-\gamma_{k+1}}^{k-1} x_l^T S_1 x_l\\
&\quad - \sum_{l=k+1-\tau_k-\gamma_k}^{k-1} x_l^T S_1 x_l\\
&= (\tau+\gamma)x_k^T S_1 x_k - \sum_{l=k+1-\tau-\gamma}^{k} x_l^T S_1 x_l + x_k^T S_1 x_k\\
&\quad - x_{k-\tau_k-\gamma_k}^T S_1 x_{k-\tau_k-\gamma_k} + \sum_{l=k+1-\tau_k-\gamma_k}^{k-1} x_l^T S_1 x_l\\
&\quad + \sum_{l=k+1-\tau_{k+1}-\gamma_{k+1}}^{k-\tau_k-\gamma_k} x_l^T S_1 x_l - \sum_{l=k+1-\tau_k-\gamma_k}^{k-1} x_l^T S_1 x_l\\
&\le (\tau+\gamma)x_k^T S_1 x_k - \sum_{l=k+1-\tau-\gamma}^{k} x_l^T S_1 x_l + x_k^T S_1 x_k\\
&\quad - x_{k-\tau_k-\gamma_k}^T S_1 x_{k-\tau_k-\gamma_k} + \sum_{l=k+1-\tau_k-\gamma_k}^{k-1} x_l^T S_1 x_l\\
&\quad + \sum_{l=k+1-\tau-\gamma}^{k} x_l^T S_1 x_l - \sum_{l=k+1-\tau_k-\gamma_k}^{k-1} x_l^T S_1 x_l\\
&= (\tau+\gamma)x_k^T S_1 x_k + x_k^T S_1 x_k - x_{k-\tau_k-\gamma_k}^T S_1 x_{k-\tau_k-\gamma_k}.
\end{aligned} \tag{8}$$

$$E\{\Delta V_3\} = x_k^T S_2 x_k - x_{k-\tau-\gamma}^T S_2 x_{k-\tau-\gamma}. \tag{9}$$

$$\begin{aligned}
&E\{\Delta V_4\}\\
&= E\left\{(\tau+\gamma)^2\chi_k^T Z\chi_k\right\} - \sum_{l=k-\tau-\gamma}^{k-1}(\tau+\gamma)\chi_l^T Z\chi_l\\
&= E\left\{(\tau+\gamma)^2\left(\left(A_{\delta_k}-I\right)x_k + B_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\right)^T\right.\\
&\quad \left. Z\left(\left(A_{\delta_k}-I\right)x_k + B_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\right)\right\}\\
&\quad - \sum_{l=k-\tau-\gamma}^{k-1}(\tau+\gamma)\chi_l^T Z\chi_l\\
&= (\tau+\gamma)^2 x_k^T\left(A_{\delta_k}-I\right)^T Z\left(A_{\delta_k}-I\right)x_k\\
&\quad + (\tau+\gamma)^2 x_k^T\left(A_{\delta_k}-I\right)^T ZB_{u_{\delta_k}}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad + (\tau+\gamma)^2 x_{k-\tau_k-\gamma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T Z\left(A_{\delta_k}-I\right)x_k\\
&\quad + (\tau+\gamma)^2 x_{k-\tau_k-\gamma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T ZB_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad - \sum_{l=k-\tau-\gamma}^{k-1}(\tau+\gamma)\chi_l^T Z\chi_l\\
&= (\tau+\gamma)^2 x_k^T\left(A_{\delta_k}-I\right)^T Z\left(A_{\delta_k}-I\right)x_k\\
&\quad + (\tau+\gamma)^2 x_k^T\left(A_{\delta_k}-I\right)^T ZB_{u_{\delta_k}}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad + (\tau+\gamma)^2 x_{k-\tau_k-\gamma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T Z\left(A_{\delta_k}-I\right)x_k\\
&\quad + (\tau+\gamma)^2 x_{k-\tau_k-\gamma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T ZB_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad - \sum_{l=k-\tau_k-\gamma_k}^{k-1}(\tau_k+\gamma_k)\chi_l^T Z\chi_l\\
&\quad - \sum_{l=k-\tau-\gamma}^{k-\tau_k-\gamma_k-1}(\tau+\gamma-\tau_k-\gamma_k)\chi_l^T Z\chi_l.
\end{aligned}$$

By Lemma 2, one has:

$$\begin{aligned}
&E\{\Delta V_4\}\\
&\le (\tau+\gamma)^2 x_k^T\left(A_{\delta_k}-I\right)^T Z\left(A_{\delta_k}-I\right)x_k\\
&\quad + (\tau+\gamma)^2 x_k^T\left(A_{\delta_k}-I\right)^T ZB_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad + (\tau+\gamma)^2 x_{k-\tau_k-\gamma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T Z\left(A_{\delta_k}-I\right)x_k\\
&\quad + (\tau+\gamma)^2 x_{k-\tau_k-\gamma_k}^T\left(B_{\delta_k}K_{i,p}\right)^T ZB_{\delta_k}K_{i,p}x_{k-\tau_k-\gamma_k}\\
&\quad - \left[x_k - x_{k-\tau_k-\gamma_k}\right]^T Z\left[x_k - x_{k-\tau_k-\gamma_k}\right]\\
&\quad - \left[x_{k-\tau_k-\gamma_k} - x_{k-\tau-\gamma}\right]^T Z\left[x_{k-\tau_k-\gamma_k} - x_{k-\tau-\gamma}\right].
\end{aligned} \tag{10}$$

From (7)-(10), one can obtain:

$$E\left\{\Delta V\left(x_k,\tau_k,\delta_{k-\tau_k}\right)\right\}\leq \xi_k^T\Lambda\xi_k, \tag{11}$$

where $\xi_k^T=\begin{bmatrix}x_k^T & x_{k-\tau_k-\gamma_k}^T & x_{x_{k-\tau-\gamma}}^T\end{bmatrix}$.
Hence, if $\Lambda<0$, one has

$$\begin{aligned}E\left\{\sum_{k=0}^{T}\|x_k\|^2\right\}&\leq-\lambda_{\min}(-\Lambda)\xi_k^T\xi_k\\&\leq-\lambda_{\min}(-\Lambda)x_k^Tx_k\\&=-\lambda_{\min}(-\Lambda)\|x_k\|^2.\end{aligned}\tag{12}$$

For any positive integer $T\geq 1$, the follows holds:

$$\begin{aligned}&E\left\{\sum_{k=0}^{T}\|x_k\|^2\right\}\\&\leq\frac{1}{\lambda_{\min}(-\Lambda)}\left(E\left\{V\left(x_0,\tau_0,\delta_{-\tau_0}\right)\right\}\right.\\&\quad\left.-E\left\{V\left(x_{T+1},\tau_{T+1},\delta_{T+1-\tau_{T+1}}\right)\right\}\right)\\&\leq\frac{1}{\lambda_{\min}(-\Lambda)}E\left\{V\left(x_0,\tau_0,\delta_{-\tau_0}\right)\right\}\\&=\frac{1}{\lambda_{\min}(-\Lambda)}x_k^T\Omega_{\tau_0,\delta_{-\tau_0}}x_k.\end{aligned}\tag{13}$$

From the Definition 1, the closed-loop system (5) is stochastically stable, which completes the proof. □

Theorem 1 presents the sufficient conditions on the existence of the state feedback controller. To get a feasible solution for controller gain matrix $K_{\tau_k,\delta_{k-\tau_k}}$, the equivalent LMIs conditions with nonconvex constraints will be given in Theorem 2.

Theorem 2: There exists a controller (3) such that the closed-loop system (5) is stochastically stable if there exist positive-definite matrices $P_{i,p}>0, L_{j,q}>0, S_1>0, S_2>0$, $Z>0$, $Y>0$ and matrix $K_{i,r}$ such that

$$\begin{bmatrix}\Sigma_{11} & * & *\\ \Sigma_{21} & -Y & *\\ \Sigma_{31} & 0 & \Sigma_{33}\end{bmatrix}<0, \tag{14}$$

$$P_{j,q}L_{j,q}=I,\quad ZY=I, \tag{15}$$

where

$$\begin{aligned}\Sigma_{11}&=\begin{bmatrix}\tilde\Sigma_{11} & * & *\\ Z & S_1-2Z & *\\ 0 & Z & S_2-Z\end{bmatrix},\\ \tilde\Sigma_{11}&=(\tau+\gamma+1)S_1+S_2-Z-P_{i,p},\\ \Sigma_{21}&=(\tau+\gamma)\begin{bmatrix}A_{\delta_k}-I & B_{\delta_k}K_{i,p} & 0\end{bmatrix},\\ \Sigma_{31}^T&=\left[\sqrt{\lambda_{i1}\Theta_{p1}^{i-j+1}}\vartheta^T\cdots\sqrt{\lambda_{i\tau}\Theta_{pD}^{i-j+1}}\vartheta^T\right],\\ \vartheta&=\begin{bmatrix}A_{\delta_k} & B_{\delta_k}K_{i,p} & 0\end{bmatrix},\\ \Sigma_{33}&=\text{Diag}\left\{L_{01},\cdots,L_{\tau D}\right\},\end{aligned}$$

hold for all $i,j\in\mathcal{M}, p,q\in\mathcal{W}$. ■

Proof: Letting $Y=Z^{-1}, L_{j,q}=P_{j,q}^{-1}, j\in\mathcal{M}, q\in\mathcal{W}$ and by applying the Schur complement, the proof can be readily completed. □

The conditions stated in Theorem 2 are in fact a set of LMIs with some matrix inverse constraints. Although they are nonconvex, which bring difficulties in using the existing convex optimization tool to solve them, one can use the cone complementary linearization (CCL) algorithm to transform this problem into the nonlinear minimization problem with LMI constraints as follows:

$$\text{Min tr}\Big(\sum_{j=0}^{\tau}\sum_{q=1}^{D}P_{j,q}L_{j,q}+ZY\Big)\ \text{s.t. (14), (16) and (17).}$$

$$\begin{bmatrix}P_{j,q} & I\\ I & L_{j,q}\end{bmatrix}>0,\quad j\in\mathcal{M},\ q\in\mathcal{W}, \tag{16}$$

$$\begin{bmatrix}Z & I\\ I & Y\end{bmatrix}>0. \tag{17}$$

Further, the procedure for solving the state feedback controller gain matrix $K_{i,p}$ is exhibited in Algorithm 1.

The state feedback controller gain matrix $K_{i,p}$ is derived in Theorem 2 on condition that all elements in Ξ and Θ^{i-j+1} are completely known. However, it is usually difficult to obtain the full transition probabilities, and the controller gain matrix $K_{i,p}$ will be derived on Theorem 3 on condition that there are some unknown elements in Ξ and Θ^{i-j+1}.

For notational clarity, $\forall j\in\mathcal{M}$, let $\mathcal{M}=\mathcal{M}_k^i+\mathcal{M}_{uk}^i$ with $\mathcal{M}_k^i=\{j:\omega_{ij}\text{ is known}\}$, $\mathcal{M}_{uk}^i=\{j:\omega_{ij}\text{ is unknown}\}$. If $\mathcal{M}_k^i$ is not empty, it can be further described as $\mathcal{M}_k^i=\{\mathcal{M}_{k_1^i},\mathcal{M}_{k_2^i},\cdots,\mathcal{M}_{k_a^i}\}$, where $\mathcal{M}_{k_a^i}$ represents the column index of the a th known transition probability in the i th row of Ξ. $\mathcal{M}_{uk}^i$ can be described as $\mathcal{M}_{uk}^i=\{\mathcal{M}_{\bar k_1^i},\mathcal{M}_{\bar k_2^i},\cdots,\mathcal{M}_{\bar k_{\tau-a}^i}\}$, where $\mathcal{M}_{\bar k_{\tau-a}^i}$ represents the column index of the $(\tau-a)$ th unknown transition probability in the i th row of Ξ.

Similarly, $\forall q\in\mathcal{W}$, let $\mathcal{W}=\mathcal{W}_k^p+\mathcal{W}_{uk}^p$ with $\mathcal{W}_k^p=\{q:\omega_{pq}\text{ is known}\}$, $\mathcal{W}_{uk}^p=\{q:\omega_{pq}\text{ is unknown}\}$. If $\mathcal{W}_k^p$ is not empty, it can be further described as $\mathcal{W}_k^p=\{\mathcal{W}_{k_1^p},\mathcal{W}_{k_2^p},\cdots,\mathcal{W}_{k_b^p}\}$, where $\mathcal{W}_{k_b^p}$ represents the column index of the b th known transition probability in the p th row of matrix Θ^{i-j+1}. $\mathcal{W}_{uk}^p$ can be described as $\mathcal{W}_{uk}^p=\{\mathcal{W}_{\bar k_1^p},\mathcal{W}_{\bar k_2^p},\cdots,\mathcal{W}_{\bar k_{D-b}^p}\}$, where $\mathcal{W}_{\bar k_{D-b}^p}$ represents the column index of the $(D-b)$ th unknown transition probability in the p th row of matrix Θ^{i-j+1}.

Theorem 3: If there exist positive-definite matrix $P_{i,p}>0$, $L_{j,q}>0$, $S_1>0$, $S_2>0$, $Z>0$, $Y>0$ and matrix $K_{i,r}$ satisfying

$$\begin{bmatrix}\tilde\omega\sum_{q\in\mathcal{W}_k^p}\Theta_{pq}^{i-j+1}\Sigma_{11} & * & *\\ \tilde\omega\sum_{q\in\mathcal{W}_k^p}\Theta_{pq}^{i-j+1}\Sigma_{21} & -\tilde\omega\sum_{q\in\mathcal{W}_k^p}\Theta_{pq}^{i-j+1}Y & *\\ \Upsilon_{\mathcal{M}_k^i,\mathcal{W}_k^p} & 0 & \Sigma_{\mathcal{M}_k^i,\mathcal{W}_k^p}\end{bmatrix}<0, \tag{18}$$

$$\begin{bmatrix}\sum_{q\in\mathcal{W}_k^p}\Theta_{pq}^{i-j+1}\Sigma_{11} & * & *\\ \sum_{q\in\mathcal{W}_k^p}\Theta_{pq}^{i-j+1}\Sigma_{21} & -\sum_{q\in\mathcal{W}_k^p}\Theta_{pq}^{i-j+1}Y & *\\ \Upsilon_{\mathcal{M}_{uk}^i,\mathcal{W}_k^p} & 0 & \Sigma_{\mathcal{M}_{uk}^i,\mathcal{W}_k^p}\end{bmatrix}<0,$$

Algorithm 1 Procedure for Solving the Controller Gain Matrix $K_{i,p}$

1: Set the maximum number of iterations $R_{\max}$
2: Find a set of feasible solution $(P^0_{j,q}, L^0_{j,q}, S^0_1, S^0_2, Z^0, Y^0, K^0_{i,p})$ satisfying (14), (16) and (17), and let $k = 0$
3: Solve the following optimization problem for variables:
$$\text{Min tr}\Big(\sum_{j=0}^{\tau}\sum_{q=1}^{D} P_{j,q}L_{j,q} + ZY\Big),\ \text{s.t. (14), (16) and (17)}$$
4: Set $(P^k_{j,q} = P_{j,q}, L^k_{j,q} = L_{j,q}, S^k_1 = S_1, S^k_2 = S_2, Z^k = Z, Y^k = Y, K^k_{i,p} = K_{i,p})$
5: **while** number of iterations $< R_{\max}$ **do**
6:   **if** (14), (15) is satisfied **then**
7:     break
8:   **else**
9:     $k = k + 1$, go to step 3.
10:   **end if**
11: **end while**

$$j \in \mathcal{M}^i_{uk}, \tag{19}$$

$$\begin{bmatrix} \tilde{\omega}\Sigma_{11} & * & * \\ \tilde{\omega}\Sigma_{21} & -\tilde{\omega}Y & * \\ \Upsilon_{\mathcal{M}^i_k,\mathcal{W}^p_{uk}} & 0 & \Sigma_{\mathcal{M}^i_k,\mathcal{W}^p_{uk}} \end{bmatrix} < 0, \quad q \in \mathcal{W}^p_{uk}, \tag{20}$$

$$\begin{bmatrix} \Sigma_{11} & * & * \\ \Sigma_{21} & -Y & * \\ \Upsilon_{\mathcal{M}^i_{uk},\mathcal{W}^p_{uk}} & 0 & \Sigma_{\mathcal{M}^i_{uk},\mathcal{W}^p_{uk}} \end{bmatrix} < 0,\ j \in \mathcal{M}^i_{uk},\ q \in \mathcal{W}^p_{uk}, \tag{21}$$

$$P_{j,q}L_{j,q} = I,\ ZY = I, \tag{22}$$

where

$$\Upsilon^T_{\mathcal{M}^i_k,\mathcal{W}^p_k} = \Big[\sqrt{\lambda_{i1}\Theta^{i-j+1}_{p1}}\vartheta^T \cdots \sqrt{\lambda_{ia}\Theta^{i-j+1}_{pb}}\vartheta^T\Big],$$
$$\Sigma_{\mathcal{M}^i_k,\mathcal{W}^p_k} = \text{Diag}\{-L_{\mathcal{M}_{k^i_1},\mathcal{W}_{k^p_1}}, \ldots, -L_{\mathcal{M}_{k^i_a},\mathcal{W}_{k^p_b}}\},$$
$$\Upsilon^T_{\mathcal{M}^i_{uk},\mathcal{W}^p_k} = \Big[\sqrt{\Theta^{i-j+1}_{p1}}\vartheta^T \cdots \sqrt{\Theta^{i-j+1}_{pb}}\vartheta^T\Big],$$
$$\Sigma_{\mathcal{M}^i_{uk},\mathcal{W}^p_k} = \text{Diag}\{-L_{j,\mathcal{W}_{k^p_1}}, \ldots, -L_{j,\mathcal{W}_{k^p_b}}\},$$
$$\Upsilon^T_{\mathcal{M}^i_k,\mathcal{W}^p_{uk}} = [\sqrt{\lambda_{i1}}\vartheta^T \cdots \sqrt{\lambda_{ia}}\vartheta^T],$$
$$\Sigma_{\mathcal{M}^i_k,\mathcal{W}^p_{uk}} = \text{Diag}\{-L_{\mathcal{M}_{k^i_1},q}, \ldots, -L_{\mathcal{M}_{k^i_a},q}\},$$
$$\Upsilon^T_{\mathcal{M}^i_{uk},\mathcal{W}^p_{uk}} = [\vartheta^T \cdots \vartheta^T],$$
$$\tilde{\omega} = \sum_{j\in\mathcal{M}^i_k} \omega_{ij},$$

for all $i, j \in \mathcal{M}, p, q \in \mathcal{W}$, the closed-loop system (5) under the controller (3) is stochastically stable. ■

Proof: By the Schur complement, $\Lambda < 0$ is equivalent to

$$\begin{bmatrix} \Sigma_{11} & * \\ \Sigma_{21} & -Y \end{bmatrix} + \begin{bmatrix} \vartheta \\ 0 \end{bmatrix}^T \tilde{P}_{j,q}\,[\vartheta\, 0] < 0, \tag{23}$$

which can be written as

$$\Big(\sum_{j\in\mathcal{M}^i_k}\sum_{q\in\mathcal{W}^p_k} \omega_{ij}\Theta^{i-j+1}_{pq} + \sum_{j\in\mathcal{M}^i_{uk}}\sum_{q\in\mathcal{W}^p_k} \omega_{ij}\Theta^{i-j+1}_{pq} + \sum_{j\in\mathcal{M}^i_k}\sum_{q\in\mathcal{W}^p_{uk}} \omega_{ij}\Theta^{i-j+1}_{pq} + \sum_{j\in\mathcal{M}^i_{uk}}\sum_{q\in\mathcal{W}^p_{uk}} \omega_{ij}\Theta^{i-j+1}_{pq}\Big)$$
$$\left(\begin{bmatrix} \Sigma_{11} & * \\ \Sigma_{21} & -Y \end{bmatrix} + \begin{bmatrix} \vartheta \\ 0 \end{bmatrix}^T \tilde{P}_{j,q}\begin{bmatrix} \vartheta & 0 \end{bmatrix}\right)$$
$$= \tilde{\omega}\sum_{q\in\mathcal{W}^p_k}\Theta^{i-j+1}_{pq}\left(\begin{bmatrix} \Sigma_{11} & * \\ \Sigma_{21} & -Y \end{bmatrix} + \begin{bmatrix} \vartheta \\ 0 \end{bmatrix}^T \tilde{P}_{j,q}\begin{bmatrix} \vartheta & 0 \end{bmatrix}\right)$$
$$+\bar{\omega}\left(\sum_{q\in\mathcal{W}^p_k}\Theta^{i-j+1}_{pq}\left(\begin{bmatrix} \Sigma_{11} & * \\ \Sigma_{21} & -Y \end{bmatrix} + \begin{bmatrix} \vartheta \\ 0 \end{bmatrix}^T \tilde{P}_{j,q}\begin{bmatrix} \vartheta & 0 \end{bmatrix}\right)\right)$$
$$+\sum_{q\in\mathcal{W}^p_{uk}}\Theta^{i-j+1}_{pq}\left(\tilde{\omega}\left(\begin{bmatrix} \Sigma_{11} & * \\ \Sigma_{21} & -Y \end{bmatrix} + \begin{bmatrix} \vartheta \\ 0 \end{bmatrix}^T \tilde{P}_{j,q}\begin{bmatrix} \vartheta & 0 \end{bmatrix}\right)\right)$$
$$+\bar{\omega}\sum_{q\in\mathcal{W}^p_{uk}}\Theta^{i-j+1}_{pq}\left(\begin{bmatrix} \Sigma_{11} & * \\ \Sigma_{21} & -Y \end{bmatrix} + \begin{bmatrix} \vartheta \\ 0 \end{bmatrix}^T \tilde{P}_{j,q}\begin{bmatrix} \vartheta & 0 \end{bmatrix}\right),$$

where $\bar{\omega} = \sum_{j\in\mathcal{M}^i_{uk}} \omega_{ij}$.

Appling Schur complement again, one can get:

$$\sum_{j\in\mathcal{M}^i_k}\sum_{q\in\mathcal{W}^p_k} \omega_{ij}\Theta^{i-j+1}_{pq}\left(\begin{bmatrix} \Sigma_{11} & * \\ \Sigma_{21} & -Y \end{bmatrix} + \begin{bmatrix} \vartheta \\ 0 \end{bmatrix}^T \tilde{P}_{j,q}\,[\vartheta 0]\right) < 0 \tag{24}$$

is equivalent to (18).

Therefore, if (18) holds, then (24) holds. Because $\omega_{ij} \geq 0$, $\pi_{rs} \geq 0$, if (18)-(22) hold, then $\Lambda < 0$ holds, that is, the closed-loop system (5) is stochastically stable. This completes the proof. □

Remark 3: If there are some unknown elements in matrix Θ, then matrix Θ^{i-j+1} has more unknown elements than Θ, for example, if $\Theta = \begin{bmatrix} ? & ? \\ 0.5 & 0.5 \end{bmatrix}$, then all the elements are unknown in Θ^2. In this case, only (20)-(22) should be satisfied to guarantee the stochastic stability of the closed-loop system (5).

Remark 4: Similar to Theorem 2, (18)-(22) in Theorem 3 can also be solved by the CCL algorithm, the detail procedure is omitted here.

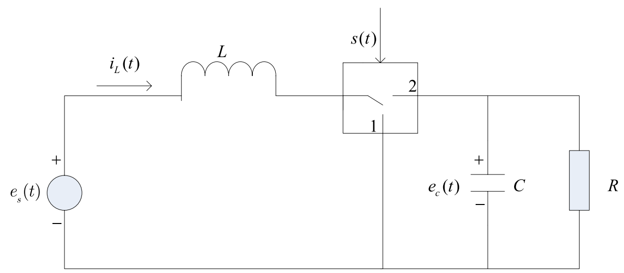

FIGURE 2. The boost converter.

IV. NUMERICAL EXAMPLE

To illustrate the effectiveness of the proposed method, the results in this paper are applied to the pulse-width-modulation (PWM)-driven boost converter as shown in Figure 2, where the switch $s(t)$ is controlled by a PWM device, R is the resistance, L is the inductance, C is the capacitance, and $e_s(t)$ is the power source [25]. The converter is used to transform the source voltage into a higher voltage. With different closed positions of the $s(t)$, the state space equation of the converter is also different, which can be modeled as a typical Markov jump system. The state space model of the converter is as follows:

$$x_{k+1} = A_{\delta_k} x_k + B_{\delta_k} u_k, \delta_k \in \{1, 2\},$$

where

$$A_1 = \begin{bmatrix} 0.94 & 0.1 & 0.06 \\ -0.3 & 0.95 & -0.3 \\ -0.25 & -0.06 & 0.63 \end{bmatrix}, B_1 = \begin{bmatrix} -0.3 \\ 0.2 \\ 0.1 \end{bmatrix},$$
$$A_2 = \begin{bmatrix} 0.93 & 0.08 & 0.07 \\ -0.14 & 0.66 & -0.2 \\ -0.16 & -0.4 & 0.66 \end{bmatrix}, B_2 = \begin{bmatrix} -1.4 \\ 0.3 \\ 0.2 \end{bmatrix}.$$

The transition probability matrix between the two subsystems is $\Theta = \begin{bmatrix} 0.8 & 0.2 \\ 0.3 & 0.7 \end{bmatrix}$. Assume S/C time-delay $\tau_k \in \mathcal{M} = \{0, 1\}$ and C/A time-delay $\gamma_k \in \mathcal{N} = \{0, 1\}$. The transition probability matrix of τ_k and γ_k is as follows, respectively:

$$\Xi = \begin{bmatrix} 0.7 & 0.3 \\ 0.3 & 0.7 \end{bmatrix}, \Pi = \begin{bmatrix} 0.6 & 0.4 \\ 0.2 & 0.8 \end{bmatrix}.$$

According to Theorem 2, the mode-dependent controller gain matrix is obtained as follows:

$$\begin{aligned} K_{01} &= \begin{bmatrix} 0.2422 & -0.1052 & -0.0149 \end{bmatrix}, \\ K_{11} &= \begin{bmatrix} 0.2586 & -0.1055 & -0.0253 \end{bmatrix}, \\ K_{02} &= \begin{bmatrix} 0.0958 & 0.0166 & -0.0157 \end{bmatrix}, \\ K_{12} &= \begin{bmatrix} 0.0959 & 0.0155 & -0.0307 \end{bmatrix}. \end{aligned}$$

By the traditional Lyapunov-Krasovskii method, the mode-independent controller gain matrix can be obtained: $K = \begin{bmatrix} 0.1191 & -0.0025 & -0.0265 \end{bmatrix}$. The system mode δ_k, the time-delay τ_k and the time-delay γ_k is shown in Figure 3, Figure 4 and Figure 5, respectively. Assuming the initial state of the system $x_0^T = \begin{bmatrix} 1 & -0.5 & 0.5 \end{bmatrix}$, Figure 6, Figure 7 and Figure 8 illustrate the state response of the closed-loop system (5) using the mode-dependent controller proposed in this paper and the mode-independent controller.

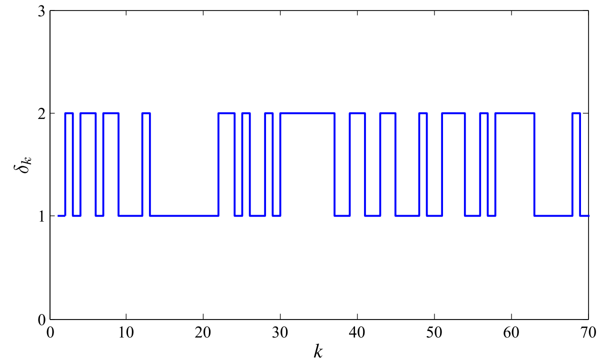

FIGURE 3. The system mode δ_k.

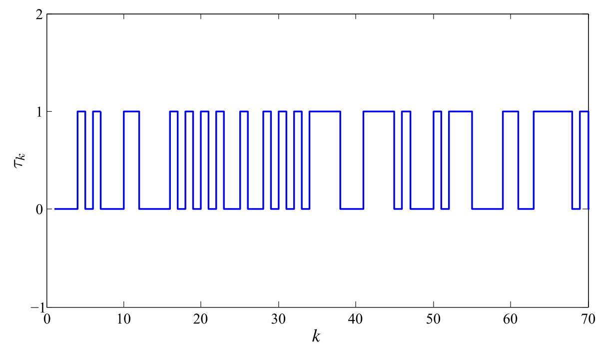

FIGURE 4. The S/C time-delay τ_k.

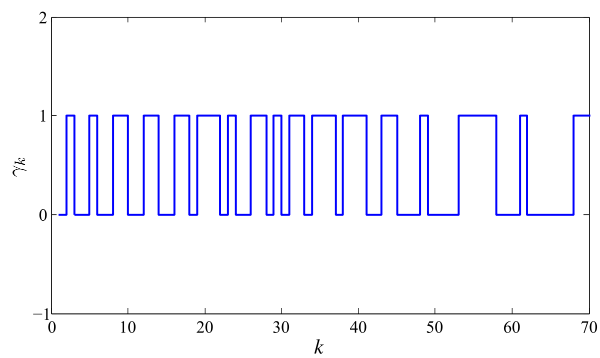

FIGURE 5. The C/A time-delay γ_k.

From Figure 6, Figure 7 and Figure 8, it can be seen that the proposed two-mode-dependent controller outperforms the mode-independent one.

Assuming that all the elements in Θ and Ξ are completely unknown, the mode-dependent controller gain matrix can also be obtained by Theorem 3 as follows:

$$\begin{aligned} K_{01} &= \begin{bmatrix} 0.2962 & -0.1245 & -0.0147 \end{bmatrix}, \\ K_{11} &= \begin{bmatrix} 0.2962 & -0.1245 & -0.0147 \end{bmatrix}, \\ K_{02} &= \begin{bmatrix} 0.1106 & -0.0215 & -0.0176 \end{bmatrix}, \\ K_{12} &= \begin{bmatrix} 0.1106 & -0.0215 & -0.0176 \end{bmatrix}. \end{aligned}$$

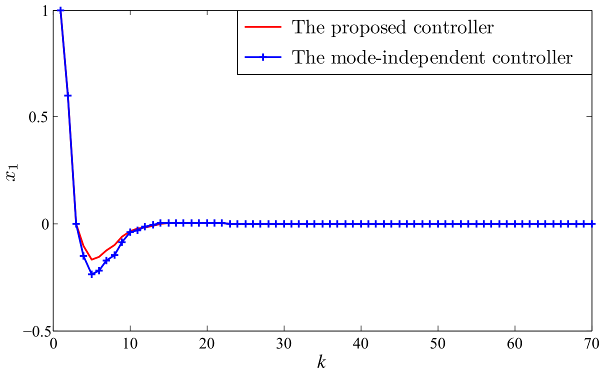

FIGURE 6. The closed-Loop system status x_1.

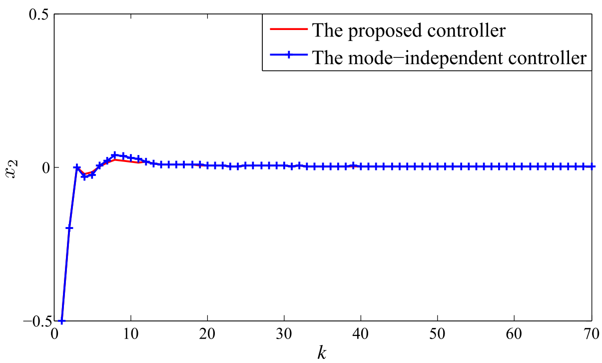

FIGURE 7. The closed-Loop system status x_2.

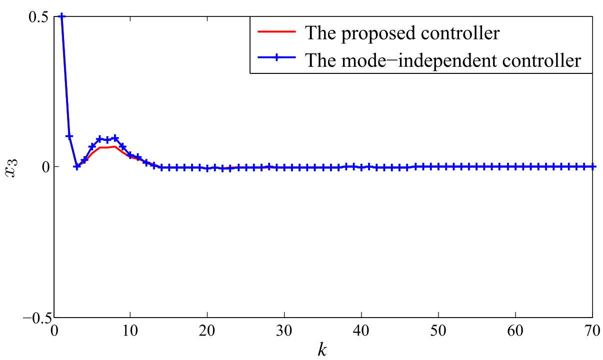

FIGURE 8. The closed-Loop system status x_3.

V. CONCLUSION

The two-mode-dependent state feedback controller design method for a kind of networked Markov system with random time-delay is researched in this paper. The S/C time-delay and the C/A time-delay are both considered, the sufficient conditions on the stochastic stability of the closed-loop system and the solution of the controller gain matrix are given. The numerical simulation shows that the controller designed in this paper is superior to the mode-independent controller.

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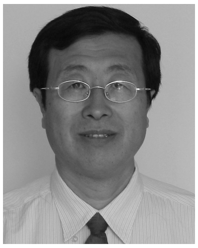

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