Analysis and Design of Wireless Networked Control System Utilizing Adaptive Coded Modulation

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Abstract In this paper, we explore the analysis and design of wireless networked control system (WNCS) utilizing adaptive coded modulation (ACM) schemes, which can improve the energy efficiency and increase data rate over a fading channel. To capture the characteristics of varying rate, interference, and routing in wireless transmission channel, we introduce the concept of equivalent delay (ED). Based on the time-varying network condition, the analytic lower and upper bounds of EDs are given. Whereafter, WNCS is modelled as a discrete-time system with time-varying input delay. Sufficient stabilization condition of the closed-loop WNCS is derived by making use of novel techniques of time-delay system. Finally, the numerical result shows the validity of our proposed control strategies.

Key words Wireless networked control system (WNCS), adaptive coded modulation (ACM), equivalent delay (ED), discrete-time system, time-varying input delay, stabilization

In recent years, the field of networked control system (NCS) has received considerable attention[1−14], because of its advantages, such as reduced installation, maintenance costs and so on. However, there are some faults because of the presence of networks in the closed loop. For example, the transmission of a packet may be put off, or this packet is likely to be dropped. Thus, there must be novel techniques to deal with those problems. In [1], the authors provided a new model for NCSs under consideration of both network-induced delay and data packet dropout, where a controller design method was proposed by delay-dependent approach. Then, a series of papers by the same authors and their colleagues studied the H_∞ controller design for linear systems based on this model and methods of delay systems[2−6]. The method for robust state feedback controller design of networked control system with nonlinearity was proposed in [7]. In [8], an observer-based feedback controller was designed for networked system with random communication delay, and the robust one with random packet loss was designed in [9]. In [10], a state-feedback guaranteed cost controller was developed for the networked control systems with state-delay, and a robust H_∞ networked control method for Takagi-Sugeno (T-S) fuzzy systems with uncertainty and time delay was concerned in [11]. The predictive control method for NCS with random network delay and packet loss was discussed in [12]. In [13], the NCS with random packet loss and delay was modeled as a switched system, and an interative method was used to design the controller. The problems of passivity analysis and passification were investigated for NCS in [14]. However, all of them neglected the important role that the characteristic of wireless communication channel (such as signal-to-noise ratio (SNR) and routing) plays in the modeling procedure of the controlled system and the selection of an appropriate controller.

Because of fully mobile operation, flexible installation, and rapid deployment, while reducing maintenance costs, the wireless networked control system (WNCS) is becoming dominant[15−19]. However, in practise, due to the new problems in WNCS, such as mobility of the nodes (the varying routing), signal-attenuation, and interference channels, the delay and loss are much more pronounced than in wired networks. In [15], detailed analysis on how different factors affected the nature of time delays and frame loss (and ultimately, control performance) was provided, but they did not consider the adaptive coded modulation scheme and controller design. In [16], the authors presented an extended version of separation principle of linear quadratic Gaussian control (QGC) for WNCS with time-delay. In [18], a cross-layer framework was presented for the joint design of wireless networks and distributed controllers, and they used the controller of [20] to get good performance. The gain scheduling scheme of discrete plant was employed for tuning the controller′s parameters based on the recorded quality of service (QoS) in [21].

In this paper, to overcome the faults introduced by node mobility (the varying routing), signal-attenuation, and interference channels, we apply a new coded modulation scheme into WNCS. In the traditional communication technology, constant coded modulation was adopted, but this scheme faced serious problems: when the variation of the SNR is great, we will neither guarantee the given bit error rate (BER) nor utilize the full channel capacity sufficiently. At the same time, there will be more packet loss and longer network delay under deep fading condition. Future systems must show high degree adaptability on many levels to support traffic flows with varying information rates. When the condition of the channel is varying, a larger rate is regulated for the better condition and a smaller rate is for the worse one. Here, SNR is used as a useful indicator for communication condition. As it is always time-varying due to the mobility of node, the signal-attenuation, and interference channels, we introduce the ACM scheme to adapt its varying. The ACM scheme is a promising technique to change the data rate that can be reliably transmitted over fading channels according to SNR, and it may utilize a set of classical codes, each of which is designed for good performance on an additive white Gaussian noise (AWGN) channel.

The ACM was proposed over a fading channel in [22–29]. In [22], the authors applied coset codes to an adaptive modulation in fading channels. A general technique was developed to determine the average spectral efficiency of the coding scheme for any set of 2-D trellis codes in [27], and all codes could be generated by the same encoder and decoded. In [29], the authors examined adaptive modulation schemes that were subjected to an average power and BER constraint for flat-fading channels, where the data rate, transmit power, and instantaneous BER were varied to maximize spectral efficiency. Furthermore, in [23], new

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upper bounds on the average spectral efficiency (ASE) of ACM were presented with maximum likelihood decoding and sequential decoding. These works showed that the ACM scheme could achieve a high transmission rate, improve the BER performance, and need not increase the channel bandwidth. In this paper, such ACM scheme is applied to wireless networked control system for the first time. When the SNR is increasing, the transmission rate will be increased. Similarly, the transmission rate will fall off according to the decreased SNR, which results in decreasing the occurrence of network-induced delay and packet loss, and improving the overall closed-loop control performance ultimately.

To jointly design control and communication with ACM scheme, all communication faults (such as time-varying routing and interference) need to be dealt with in an integrated fashion and under a more realistic scenario. But it will become even more difficult to quantify the impact of a wireless network design on the control performance analytically when the communication faults above occur. Such as how these communication faults are incarnated in the control system model, how to use ACM to decrease the occurrence of delay and packet loss, and how to design the controller in this condition will be a very challenging task as well.

This paper casts the joint control and communication design problem with ACM scheme, and focuses on solving the problem above. The characteristics of varying interference, routing, and relative rate in wireless transmission channel are analyzed, and transformed into a new concept − equivalent delay (ED). The analytic computation formula, the lower and upper bounds of ED are given. Moreover, we model the WNCS as a discrete-time system with time-varying input delay under consideration of varying interference, routing, and rate in the wireless transmission. Then, communication faults can be dealt with by the deriving of sufficient conditions on the stability of the closed-loop WNCS. Finally, the numerical result shows the validity of our proposed control strategies.

The paper is organized as follows. Section 1 describes the system model, where the ACM scheme is utilized in the feedback loop. The main analytical properties of this WNCS and the concept of ED are framed. How different communication factors affect the nature of ED is also discussed in this section. Section2 models the WNCS utilizing ACM as a discrete-time system with time-varying input delay. In terms of the given model, a controller design method is proposed in Section 3. Numerical result is presented in Section 4 to show the feasibility of our method. Conclusions will be given in Section 5.

1 Wireless channel model analysis

In this section, we introduce the WNCS and describe the architecture and parameters of the wireless network, through which the control system will communicate. Our system model is outlined in Fig. 1. Multiple control systems coexist and their feedback loops are closed over a shared wireless network. After the data (sensor measurements) are converted into a binary bit stream, they (or the control command) go through the network.

Due to limited bandwidth and large amount of data packets transmitted over one shared medium, some interesting phenomena are observed. For example, a packet, which fails to reach its destination or an intermediate node, will be dropped. The retroactivity phenomenon observed as a consequence of the unreliable services, is becoming even more dominant as a delay factor in network traffic conditions. We will next discuss the factors that contribute to this behavior. One is the distance among sensors, controllers, and actuators. As separating distance decreases, the electromagnetic communication signal strength is attenuated and the SNR will be increased. High SNR values indicate a strong connection and allow communication at high data transfer rates. The other factor is the routing. The number of hops during the packet transmission will change as the nodes move from one position to another. This case is described in Fig. 2. In Fig. 2 (a), for the i-th plant, the packet of sensor i would travel over a set of U_1^i links from source (sensor i) to destination (actuator i) at time t. But it travels over U_2^i links at time t_1, see Fig. 2 (b). For convenience, we drop the superscript "i" except when needed for clarification.

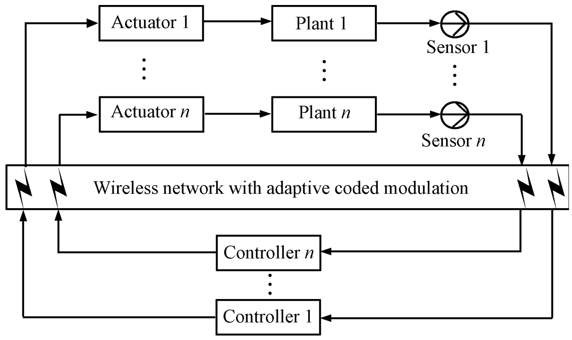

Fig. 1 The wireless networked control system

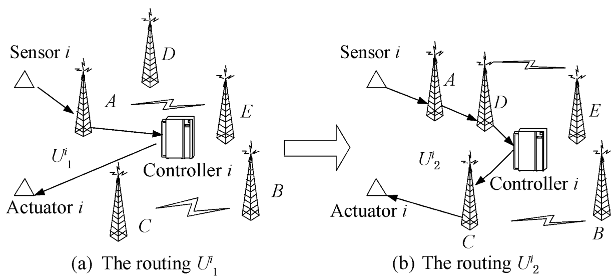

Fig. 2 The change of routing during the transmission for a packet

We denote the l-th link of U_1 as L in this paper. For every link of the wireless network, the ACM scheme is employed to exhibit the high degree of adaptability on the time-varying channel conditions. It will adjust the coding and modulation mode according to the quality of channel. The discrete-time system model of link L, described in Fig. 3, consists of a transmitter and a receiver communicating over a wireless channel degraded by multipath fading.

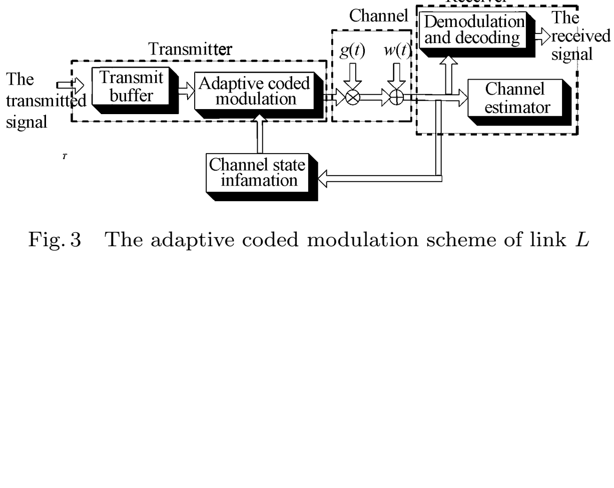

Fig. 3 The adaptive coded modulation scheme of link L

We assume that both the transmitter and receiver have perfect knowledge of the instantaneous received SNR all the time. Note that this feedback channel from the receiver to the transmitter is optional and only required by some media access control (MAC) protocols. This technique provides many parameters that can be adjusted relative to the channel fading, including data rate, instantaneous SNR, symbol rate, and channel code rate or any combination of these parameters. Here, we only consider data rate and instantaneous SNR, and assume others are constant.

Furthermore, we will obtain the instantaneous data rate for link L, and denote the instantaneous received SNR as γ. Note that all of the links have the same SNR region boundaries, and are defined as the ranges of γ values, over which the different constellations are used by the transmitter and are denoted by $\{\gamma_i\}_{i=0}^{N}$. One trellis code designed to combat AWGN is assigned to each fading region $C_n = [\gamma_{n-1}, \gamma_n)$, $1 \le n \le N$[26−29]. The different constellations are transmitted over the range of γ values. Specifically, we assign one signal constellation and a corresponding data rate of k_L^n bits/symbol to each SNR region C_n for link L. When we send $k_L^n = \log_2 M_L^n$ (bits/symbol), the instantaneous data rate is $R_L^n = k_L^n/T_s$ (bps), where T_s is the symbol time, and it is constant in fact.

To capture all the communication faults above, we introduce the concept of ED τ_k, and make the following assumptions.

Assumption 1. The ACM scheme is designed such that the BER never exceeds a given target maximum BER_0 for any SNR γ.

Assumption 2. There are retransmissions, and for simplicity, all packets are assumed to have the same size P.

Assumption 3. The size of the link storage buffer is configurable (β_L data packets for Link L).

Some studies[24−25] have indicated that the network-induced delay is related to the fixed link delay, link speed, and link buffer size per link. Therefore, considering that the instantaneous received SNR γ falls in region C_n, the ED possible of a WNCS packet Δ (traveling over the set of U_1 links from source to destination) will be (1), where δ_L is the fixed link delay of link L, P is the packet size in bytes, $R_L^n \left(R_L^f\right)$ is the link speed of link L in bps (data rate when the instantaneous received SNR γ falls in a region C_n (C_f)). F_q represents a set that contains q values in $\{1, 2, \cdots, N\}$, and β_L is the buffer size of link L in a number of packets. Ω_n and Ω_f represent the retransmission numbers, which belong to $\{1, 2, \cdots, m\}$. m is the maximum of retransmission numbers. t_L^{ni} is the instant acknowledgement timeout, that is, the time of i-th retransmission for packet Δ is t_L^{ni}. The maximum of them is t_{out}. In (1), $8P/R_L^n$ represents the transmission delay of packet Δ for link L. For simplification, take the second formula in (1) for example. Suppose there is one packet before Δ in the buffer, that is, f is one of the values in $\{1, 2, \cdots, m\}$. Suppose $f = 2$, and the retransmission numbers are 1 and 2 for packet Δ and packet before Δ, respectively. Then, packet Δ needs $\left(\dfrac{8P}{R_L^n} + t_L^{n1}\right)$ s to arrive at the receiver of link L, and the ED of Δ for link L will be $\left(\delta_L + \left(\dfrac{8P}{R_L^n} + t_L^{n1}\right) + \left(\dfrac{8P}{R_L^2} + \sum_{i=1}^{2} t_L^{2i}\right)\right)$. Similarly, we can get other formulas in (1).

When the instantaneously received SNR γ belongs to C_1 (the link speed is minimum R_L^1), $U_{\max}$ contains the maximum link number, and we have the maximum ED

$$\tau_{\max} = \sum_{L\in U_{\max}} \left(\delta_L + \left(\frac{8P}{R_L^1} + m t_{out}\right)(\beta_L + 1)\right) \tag{2}$$

When the instantaneously received SNR γ belongs to C_N (the link speed is maximum R_L^N),the minimum ED for a network path $U_{\min}$ that contains the minimum link number is

$$\tau_{\min} = \sum_{L\in U_{\min}} \left(\delta_L + \frac{8P}{R_L^N}\right) \tag{3}$$

Remark 1. From the analysis above, we can see that ACM scheme can change the data rate that can be reliably transmitted over fading channels according to the temporal SNR. With (1), we translate the characteristics of wireless network utilizing ACM schemes into ED. When the network environment is varying, the corresponding ED τ_k will change, and the variational region will be $[\tau_{\min}, \tau_{\max}]$, which are given to sustain the modeling of WNCS. On the other hand, if the ACM scheme is not applied to WNCS, the data rate must be constant, and the number and time interval of retransmission must increase. This will make the ED ampliative. In the following section, based on (2) and (3), we will give the general form of WNCS.

2 Model description of WNCS

Assume that the model of the i-th plant is a linear time-invariant system as

$$\dot{\boldsymbol{x}}(t) = A\boldsymbol{x}(t) + B\boldsymbol{u}(t) \tag{4}$$

$$\boldsymbol{u}(t) = K\boldsymbol{x}(t) \tag{5}$$

$$\tau_k = \begin{cases} \displaystyle\sum_{L\in U_1}\left(\delta_L + \left(\frac{8P}{R_L^n} + \sum_{i=1}^{\Omega_n} t_L^{ni}\right)\right), \text{ if there is no packet before } \Delta \text{ in the buffer} \\ \displaystyle\sum_{L\in U_1}\left(\delta_L + \left(\frac{8P}{R_L^n} + \sum_{i=1}^{\Omega_n} t_L^{ni}\right) + \sum_{f\in F_1}\left(\frac{8P}{R_L^f} + \sum_{i=1}^{\Omega_f} t_L^{fi}\right)\right), \text{ if there is one packet before } \Delta \text{ in the buffer} \\ \qquad\vdots \qquad\qquad \vdots \qquad\qquad \vdots \\ \displaystyle\sum_{L\in U_1}\left(\delta_L + \left(\frac{8P}{R_L^n} + \sum_{i=1}^{\Omega_n} t_L^{ni}\right) + \sum_{f\in F_q}\left(\frac{8P}{R_L^f} + \sum_{i=1}^{\Omega_f} t_L^{fi}\right)\right), \text{ if there is } q \text{ packets before } \Delta \text{ in the buffer} \\ \qquad\vdots \qquad\qquad \vdots \qquad\qquad \vdots \\ \displaystyle\sum_{L\in U_1}\left(\delta_L + \left(\frac{8P}{R_L^n} + \sum_{i=1}^{\Omega_n} t_L^{ni}\right) + \sum_{f\in F_{\beta_L}}\left(\frac{8P}{R_L^f} + \sum_{i=1}^{\Omega_f} t_L^{fi}\right)\right), \text{ if there is } \beta_L \text{ packets before } \Delta \text{ in the buffer} \end{cases} \tag{1}$$

where $\boldsymbol{x}(t) \in \mathbf{R}^n$ is the state, $\boldsymbol{u}(t) \in \mathbf{R}^m$ is the control input, A and B are system matrices of appropriate dimensions, and K is the state feedback gain to be determined. Discretizing the system at the sampling instants with sampling period h, we have

$$\boldsymbol{x}(k+1) = \Phi\boldsymbol{x}(k) + \Gamma\boldsymbol{u}(k) \tag{6}$$

$$\boldsymbol{u}(k) = K\boldsymbol{x}(k)$$

where

$$\Phi = \mathrm{e}^{Ah}, \quad \Gamma = \int_0^h \mathrm{e}^{As}\mathrm{d}sB$$

Suppose that the sensor and the actuator are clock-driven, the controller is event driven, and the data are transmitted with a single packet. At the same time, we consider the effect of varying interference and routing on the NCS with ACM scheme, and we can obtain (2) and (3). Thus, there are at most $\left\lfloor \frac{\tau_{\max}}{h} \right\rfloor - \left\lfloor \frac{\tau_{\min}}{h} \right\rfloor + 1$ packets arriving at the actuator during a sampling period. If we assume that there are q_k packets arriving at the actuator during $[kh, (k+1)h)$, the network-induced delay are τ_k^1, τ_k^2, $\cdots$, $\tau_k^{q_k-1}$, and $\tau_k^{q_k}$, orderly. The real control system (6) can be modeled as a discrete-time system with time-varying input delay as

$$\boldsymbol{x}(k+1) = \Phi\boldsymbol{x}(k) + \Gamma\left(\delta\boldsymbol{u}\left(k - \left\lfloor \frac{\tau_{k-1}^{q_{k-1}}}{h} \right\rfloor - 1\right) + (1-\delta)\boldsymbol{u}\left(k - \left\lfloor \frac{\tau_k^1}{h} \right\rfloor\right)\right) \tag{7}$$

$$\boldsymbol{u}(k-d) = K\boldsymbol{x}(k-d),$$

$$\tau_{k-1}^{q_{k-1}}, \tau_k^1, d \in [\tau_{\min}, \tau_{\max}] \tag{8}$$

where $\tau_{k-1}^{q_{k-1}}$ is the time delay of the last packet that reaches the actuator during the instant $[(k-1)h, kh)$, and it is easy to see that $\tau_{k-1}^{q_{k-1}}$ is time-varying. τ_1^{k-1} represents the network-induced delay of the last packet during a sampling interval $[(k-1)h, kh)$. $\lfloor\cdot\rfloor$ denotes the lower integer bound. $\tau_{\min}$ and $\tau_{\max}$ are obtained from (1) and (2), respectively.

Remark 2. In (7), if $\delta = 0$, it is meant that the packet with time-delay τ_k^1 arrives at the actuator at instant kh. Otherwise, if $\delta = 1$, this packet arrives after the moment kh. $\left\lfloor \frac{\tau_{k-1}^{q_{k-1}}}{h} \right\rfloor$ is a subset of $\{\left\lfloor \frac{\tau_{\min}}{h} \right\rfloor, \left\lfloor \frac{\tau_{\min}}{h} \right\rfloor + 1, \cdots, \left\lfloor \frac{\tau_{\max}}{h} \right\rfloor\}$. If $\left\lfloor \frac{\tau_{k-1}^{q_{k-1}-1}}{h} \right\rfloor = \left\lfloor \frac{\tau_{k-1}^{q_{k-1}}}{h} \right\rfloor + 1$, it is meant that no packet dropout occurs in the transmission. When $\left\lfloor \frac{\tau_{k-1}^{q_{k-1}-1}}{h} \right\rfloor > \left\lfloor \frac{\tau_{k-1}^{q_{k-1}}}{h} \right\rfloor + 1$, packets $\boldsymbol{x}\left(k - \left\lfloor \frac{\tau_{k-1}^{q_{k-1}-1}}{h} \right\rfloor + 1\right)$, $\boldsymbol{x}\left(k - \left\lfloor \frac{\tau_{k-1}^{q_{k-1}-1}}{h} \right\rfloor + 2\right), \cdots, \boldsymbol{x}\left(k - \left\lfloor \frac{\tau_{k-1}^{q_{k-1}}}{h} \right\rfloor - 1\right)$ must be dropped, or there may be disorder. If $\left\lfloor \frac{\tau_{k-1}^{q_{k-1}-1}}{h} \right\rfloor < \left\lfloor \frac{\tau_{k-1}^{q_{k-1}}}{h} \right\rfloor$, it is meant that the new data packet may reach the plant before the old one. Thus, system (7) is a general model for the WNCS utilizing ACM schemes, since the effects of the characteristics of WNCS utilizing ACM schemes are simultaneously considered by the ED.

Remark 3. This model is similar to the one considered in [1]. However, there are at least two differences between them. The first is that our framework is more general. Our work accommodates both continuous and discrete plants, and we will give the stabilization method of system using a recent result on stability of discrete-time systems, while in [1] the model failed for the discrete plant, and simple Lyapunov functional for continuous systems was used. The second is that our model gives the lower and upper EDs through analysis of the characteristics of WNCS utilizing ACM schemes, but [1] gave the maximum allowable value through the method of time-delay.

Next, conditions are sought for the stabilization of system (7) to ensure the stability of the closed loop system.

3 Feedback gain design

In this section, we use the Lyapunov functions and some inequality techniques for the stabilization of WNCS (7) based on [30], where the authors only analyzed the asymptotic stability of the system with time-varying state delay and did not consider the controller design. Here, we use some new techniques to synthesize the controller that stabilizes WNCS (7) with input delay.

Theorem 1. Suppose that for some prescribed scalars ε, there exist matrices

$$P = \begin{bmatrix} P_{11} & P_{12} & P_{13} \\ * & P_{22} & P_{23} \\ * & * & P_{33} \end{bmatrix} > 0, \quad Q = \begin{bmatrix} Q_{11} & Q_{12} \\ * & Q_{22} \end{bmatrix} \geq 0$$

$$R = \begin{bmatrix} R_{11} & R_{12} \\ * & R_{22} \end{bmatrix} \geq 0 \tag{9}$$

$Z_i > 0, i = \{1, 2\}, M, S, N$, and T such that the following linear matrix inequality is feasible

$$\begin{bmatrix} \Theta_1 + \Theta_2 + \Theta_2^{\mathrm{T}} + \Theta_3 + \Theta_3^{\mathrm{T}} & \Phi_4 \\ * & \Phi_5 \end{bmatrix} < 0 \tag{10}$$

where

$$\Theta_1 = \begin{bmatrix} \Theta_4 & 0 & 0 & \Theta_5 & P_{12} & P_{13} \\ * & -Q_{11} & 0 & P_{12}^{\mathrm{T}} & P_{22}-Q_{12} & P_{23} \\ * & * & -R_{11} & P_{13}^{\mathrm{T}} & P_{23}^{\mathrm{T}} & P_{33}-R_{12} \\ * & * & * & \Theta_6 & P_{12} & P_{13} \\ * & * & * & * & P_{22}-Q_{22} & P_{23} \\ * & * & * & * & * & P_{33}-R_{22} \end{bmatrix}, \quad \Theta_2 = \begin{bmatrix} M+N & S-M & -S-N & 0 & 0 & 0 \end{bmatrix}$$

$$\Theta_3 = T\begin{bmatrix} -(\Phi-I) & 0 & 0 & I & 0 & 0 \end{bmatrix}, \quad \Theta_4 = \alpha Q_{11}+R_{11}, \quad \Theta_5 = P_{11}+\alpha Q_{12}+R_{12}$$

$$\Theta_6 = P_{11}+\alpha Q_{22}+R_{22}+\left(\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1\right)(Z_1+Z_2), \quad \alpha = \left\lfloor\frac{\tau_{\max}}{h}\right\rfloor - \left\lfloor\frac{\tau_{\min}}{h}\right\rfloor+1$$

$$\Phi_4 = \begin{bmatrix} \sqrt{\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1}M & \sqrt{\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-\left\lfloor\frac{\tau_{\min}}{h}\right\rfloor}S & \sqrt{\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1}N & -\varepsilon T & \Phi \end{bmatrix}$$

$$\Phi_5 = \mathrm{diag}\left\{ -Z_1 \quad -Z_1 \quad -Z_2 \quad -\varepsilon I \quad -\varepsilon I \right\}, \quad \Phi = \begin{bmatrix} 0 & \Gamma K & 0 & 0 & 0 & 0 \end{bmatrix}^{\mathrm{T}}$$

then WNCS (7) can be asymptotically stabilized. Moreover, the state feedback controller in (8) can be given by (10).

Proof. Consider the state feedback control law (8). The resulting closed-loop system can be rewritten as

$$\boldsymbol{x}(k+1) = \Phi\boldsymbol{x}(k)+\Gamma K\boldsymbol{x}(k-\tau_\sigma) \tag{11}$$

$$\left\lfloor\frac{\tau_{\min}}{h}\right\rfloor+1 \le \tau_\sigma \le \left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1$$

Let the Lyapunov functional for system (11) be

$$V(k) = V_1(k)+V_2(k)+V_3(k)+V_4(k)+V_5(k) \tag{12}$$

$$\boldsymbol{\gamma}(k) = \begin{bmatrix} \boldsymbol{x}(k) \\ \boldsymbol{\lambda}(k) \end{bmatrix}$$

$$\boldsymbol{\lambda}(k) = \boldsymbol{x}(k+1)-\boldsymbol{x}(k)$$

$$V_1(k) = \boldsymbol{\zeta}^{\mathrm{T}}(k)P\boldsymbol{\zeta}(k)\boldsymbol{\zeta}^{\mathrm{T}}(k) = \begin{bmatrix} \boldsymbol{x}^{\mathrm{T}}(k) & \boldsymbol{x}^{\mathrm{T}}(k-\tau_\sigma)\,\boldsymbol{x}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right) \end{bmatrix}$$

$$V_2(k) = \sum_{n=k-\tau_\sigma}^{k-1}\boldsymbol{\gamma}^{\mathrm{T}}(n)Q\boldsymbol{\gamma}(n)$$

$$V_3(k) = \sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-1}\boldsymbol{\gamma}^{\mathrm{T}}(n)R\boldsymbol{\gamma}(n)$$

$$V_4(k) = \sum_{\theta=-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor}^{-\left\lfloor\frac{\tau_{\min}}{h}\right\rfloor-1}\sum_{n=k+\theta}^{k-1}\boldsymbol{\gamma}^{\mathrm{T}}(n)Q\boldsymbol{\gamma}(n)$$

$$V_5(k) = \sum_{\theta=-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{-1}\sum_{n=k+\theta}^{k-1}\boldsymbol{\lambda}^{\mathrm{T}}(n)(Z_1+Z_2)\boldsymbol{\lambda}(n)$$

where P, Q, and R with structures given in (9), and $Z_i > 0$, $i=\{1,2\}$ are matrices to be determined. Taking the forward difference for Lyapunov functional (9) obtains

$$\Delta V_1(k) = 2\boldsymbol{\zeta}^{\mathrm{T}}(k)P\boldsymbol{\Delta}(k)+\boldsymbol{\Delta}^{\mathrm{T}}(k)P\boldsymbol{\Delta}(k)$$

$$\boldsymbol{\Delta}^{\mathrm{T}}(k) = \left[\boldsymbol{\lambda}^{\mathrm{T}}(k)\ \boldsymbol{\lambda}^{\mathrm{T}}(k-\tau_\sigma)\ \boldsymbol{\lambda}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right)\right]$$

$$\Delta V_2(k) \le \boldsymbol{\gamma}^{\mathrm{T}}(k)Q\boldsymbol{\gamma}(k)-\boldsymbol{\gamma}^{\mathrm{T}}(k-\tau_\sigma)Q\boldsymbol{\gamma}(k-\tau_\sigma)+\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor}^{k-\left\lfloor\frac{\tau_{\min}}{h}\right\rfloor-1}\boldsymbol{\gamma}^{\mathrm{T}}(n)Q\boldsymbol{\gamma}(n)$$

$$\Delta V_3(k) \le \boldsymbol{\gamma}^{\mathrm{T}}(k)R\boldsymbol{\gamma}(k)-\boldsymbol{\gamma}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right)\times R\boldsymbol{\gamma}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right)$$

$$\Delta V_4(k) \le (\alpha-1)\boldsymbol{\gamma}^{\mathrm{T}}(k)Q\boldsymbol{\gamma}(k)-\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor}^{k-\left\lfloor\frac{\tau_{\min}}{h}\right\rfloor-1}\boldsymbol{\gamma}^{\mathrm{T}}(n)Q\boldsymbol{\gamma}(n)$$

$$\Delta V_5(k) \le \left(\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1\right)\boldsymbol{\lambda}^{\mathrm{T}}(k)(Z_1+Z_2)\boldsymbol{\lambda}(k)-\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-\tau_\sigma-1}\boldsymbol{\lambda}^{\mathrm{T}}(n)Z_1\boldsymbol{\lambda}(n)-\sum_{n=k-\tau_\sigma}^{k-1}\boldsymbol{\lambda}^{\mathrm{T}}(n)Z_1\boldsymbol{\lambda}(n)-\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-1}\boldsymbol{\lambda}^{\mathrm{T}}(n)Z_2\boldsymbol{\lambda}(n)$$

In addition, based on (11), we have

$$\boldsymbol{\lambda}(k) = \boldsymbol{x}(k+1)-\boldsymbol{x}(k) = (\Phi-I)\boldsymbol{x}(k)+\Gamma K\boldsymbol{x}(k-\tau_\sigma)$$

Then, for any matrix T, we have

$$\boldsymbol{v}^{\mathrm{T}}(k)T\left[\boldsymbol{\lambda}(k)-(\Phi-I)\boldsymbol{x}(k)-\Gamma K\boldsymbol{x}(k-\tau_\sigma)\right] = 0$$

Define

$$\boldsymbol{v}(k) = \begin{bmatrix} \boldsymbol{x}^{\mathrm{T}}(k) & \boldsymbol{x}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_\sigma}{h}\right\rfloor\right) & \boldsymbol{x}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor\right) & \boldsymbol{\lambda}^{\mathrm{T}}(k) & \boldsymbol{\lambda}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_\sigma}{\mathrm{h}}\right\rfloor\right) & \boldsymbol{\lambda}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_{\max}}{\mathrm{h}}\right\rfloor\right) \end{bmatrix}^{\mathrm{T}}$$

As we all know that

$$\begin{aligned} &2\boldsymbol{v}^{\mathrm{T}}(k)T\left[\boldsymbol{\lambda}(k)-(\Phi-I)\boldsymbol{x}(k)-\Gamma K\boldsymbol{x}(k-\tau_\sigma)\right] \le \\ &2\boldsymbol{v}^{\mathrm{T}}(k)T\left[\boldsymbol{\lambda}(k)-(\Phi-I)\boldsymbol{x}(k)\right]+ \\ &\varepsilon\boldsymbol{v}^{\mathrm{T}}(k)(-T)(-T)^{\mathrm{T}}\boldsymbol{v}(k)+ \\ &\varepsilon^{-1}\boldsymbol{x}^{\mathrm{T}}(k-\tau_\sigma)K^{\mathrm{T}}\Gamma^{\mathrm{T}}\Gamma K\boldsymbol{x}(k-\tau_\sigma) \end{aligned} \tag{13}$$

we have (14).

Since $Z_i > 0$, $i=\{1,2\}$, the last three terms in the previous equation are all nonpositive. By the Schur complement, (10) guarantees

$$\begin{aligned}\Delta V(k) \leq\ & 2\boldsymbol{\zeta}^{\mathrm{T}}(k) P\boldsymbol{\Delta}(k)+\boldsymbol{\Delta}^{\mathrm{T}}(k) P\boldsymbol{\Delta}(k)+\boldsymbol{\gamma}^{\mathrm{T}}(k) Q\boldsymbol{\gamma}(k)-\boldsymbol{\gamma}^{\mathrm{T}}(k-\tau_\sigma) Q\boldsymbol{\gamma}(k-\tau_\sigma)+\boldsymbol{\gamma}^{\mathrm{T}}(k) R\boldsymbol{\gamma}(k)- \\ & \boldsymbol{\gamma}^{\mathrm{T}}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right) R\boldsymbol{\gamma}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right)+(\alpha-1)\boldsymbol{\gamma}^{\mathrm{T}}(k) Q\boldsymbol{\gamma}(k)+ \\ & \left(\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1\right)\boldsymbol{\lambda}^{\mathrm{T}}(k)(Z_1+Z_2)\boldsymbol{\lambda}(k)-\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-\tau_\sigma-1}\boldsymbol{\lambda}^{\mathrm{T}}(n) Z_1\boldsymbol{\lambda}(n)-\sum_{n=k-\tau_\sigma}^{k-1}\boldsymbol{\lambda}^{\mathrm{T}}(n) Z_1\boldsymbol{\lambda}(n)- \\ & \sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-1}\boldsymbol{\lambda}^{\mathrm{T}}(n) Z_2\boldsymbol{\lambda}(n)+2\boldsymbol{v}^{\mathrm{T}}(k)T\left[\boldsymbol{\lambda}(k)-(\Phi-I)\boldsymbol{x}(k)\right]+\varepsilon\boldsymbol{v}^{\mathrm{T}}(k)(-T)(-T)^{\mathrm{T}}\boldsymbol{v}(k)+ \\ & \varepsilon^{-1}\boldsymbol{x}^{\mathrm{T}}(k-\tau_\sigma) K^{\mathrm{T}}\Gamma^{\mathrm{T}}\Gamma K\boldsymbol{x}(k-\tau_\sigma)+2\boldsymbol{v}^{\mathrm{T}}(k) M\left[\boldsymbol{x}(k)-\boldsymbol{x}(k-\tau_\sigma)-\sum_{n=k-\tau_\sigma}^{k-1}\boldsymbol{\lambda}(n)\right]+ \\ & 2\boldsymbol{v}^{\mathrm{T}}(k) S\left[\boldsymbol{x}(k-\tau_\sigma)-\boldsymbol{x}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right)-\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-\tau_\sigma-1}\boldsymbol{\lambda}(n)\right]+ \\ & 2\boldsymbol{v}^{\mathrm{T}}(k) N\left[\boldsymbol{x}(k)-\boldsymbol{x}\left(k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1\right)-\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-1}\boldsymbol{\lambda}(n)\right]\leq \\ & \boldsymbol{v}^{\mathrm{T}}(k)\left[\Theta_1+\Theta_2+\Theta_2^{\mathrm{T}}+\Theta_3+\Theta_3^{\mathrm{T}}+\left(\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1\right) MZ_1^{-1}M^{\mathrm{T}}+(\alpha-1) SZ_1^{-1}S^{\mathrm{T}}+\right. \\ & \left.\left(\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1\right) NZ_2^{-1}N^{\mathrm{T}}+\varepsilon^{-1}(-\varepsilon T)(-\varepsilon T)^{\mathrm{T}}+\varepsilon^{-1}\Phi\Phi^{\mathrm{T}}\right]\boldsymbol{v}(k)-\sum_{n=k-\tau_\sigma}^{k-1}\left[\boldsymbol{v}^{\mathrm{T}}(k) M+\boldsymbol{\lambda}^{\mathrm{T}}(n) Z_1\right]\times \\ & Z_1^{-1}\left[M^{\mathrm{T}}\boldsymbol{v}(k)+Z_1\boldsymbol{\lambda}(n)\right]-\sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-\tau_\sigma-1}\left[\boldsymbol{v}^{\mathrm{T}}(k) S+\boldsymbol{\lambda}^{\mathrm{T}}(n) Z_1\right] Z_1^{-1}\left[S^{\mathrm{T}}\boldsymbol{v}(k)+Z_1\boldsymbol{\lambda}(n)\right]- \\ & \sum_{n=k-\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor-1}^{k-1}\left[\boldsymbol{v}^{\mathrm{T}}(k) N+\boldsymbol{\lambda}^{\mathrm{T}}(n) Z_2\right] Z_2^{-1}\left[N^{\mathrm{T}}\boldsymbol{v}(k)+Z_2\boldsymbol{\lambda}(n)\right] \qquad (14)\end{aligned}$$

$$\begin{aligned}&\Theta_1+\Theta_2+\Theta_2^{\mathrm{T}}+\Theta_3+\Theta_3^{\mathrm{T}}+\left(\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1\right)\times \\ &MZ_1^{-1}M^{\mathrm{T}}+(\alpha-1) SZ_1^{-1}S^{\mathrm{T}}+\left(\left\lfloor\frac{\tau_{\max}}{h}\right\rfloor+1\right)\times \\ &NZ_2^{-1}N^{\mathrm{T}}+\varepsilon^{-1}(-\varepsilon T)(-\varepsilon T)^{\mathrm{T}}+\varepsilon^{-1}\Phi\Phi^{\mathrm{T}}<0 \qquad (15)\end{aligned}$$

Therefore, (10) yields $\Delta V(k) < -\epsilon\|\boldsymbol{x}(k)\|^2$ for a sufficiently small $\epsilon > 0$, and the WNCS (7) is asymptotically stabilized. □

Remark 4. By defining new Lyapunov functions and making use of more matrix variables to achieve delay dependence, a new condition is obtained for the controller design of WNCS (7). The merit of the proposed conditions is that they can deal with larger ED and get better performance for system (7).

4 Simulation

In this section, we use the Matlab linear matrix inequality toolbox and Simulink to illustrate the effectiveness and advantage of our proposed method. All simulations use the network settings given in Table 1.

Table 1 System parameters values

Parameter	Value
Frame size	1 024 bits
ACK timeout t_{out}	0.864 ms
Retry limit m	3
Error coding threshold	0.03
Fixed link delay	1ms
Buffer size	50 packets

A DC motor speed control system[31] is introduced based on WNCS utilizing adaptive coded modulation schemes. The loop equation for the electrical circuit is

$$u(t)=e_a=L\frac{\mathrm{d}i_a}{\mathrm{d}t}+Ri_a+e_b \qquad (16)$$

The mechanical torque balance based on Newton′s law is

$$J\frac{\mathrm{d}\omega}{\mathrm{d}t}+B\omega+T_l=T_e=Ki_a \qquad (17)$$

By letting $x_1=i_a$ and $x_2=\omega$, the electromechanical dynamics of the DC motor can be described by the following state-space descriptions

$$\begin{aligned}\dot{x}_1 &= -\frac{R}{L}x_1-\frac{K_b}{L}x_2+\frac{1}{L}u \\ \dot{x}_2 &= \frac{K}{J}x_1-\frac{B}{J}x_2 \qquad (1)\end{aligned}$$

where the parameters of the motor are shown in Table 2.

Table 2 DC motor parameters

J	Inertia	42.6E-6 Kg·m^2
L	Inductance	170E-3 H
R	Resistance	4.67 Ω
B	Damping coefficient	47.3E-6 N·m·s/rad
K	Torque constant	14.7E-3 N·m/A
K_b	Back-EMF constant	14.7E-3 V·s/rad

First, we give the controller for WNCS applying ACM schemes by Theorem 1. Here, the result in [27] is used, and it is an adaptive code with eight 4-D trellis codes utilizing the eight nested 2-D signal constellations. The overall encoder is applied in the International Telecommunications

Union's ITU-T V.34 modem standard. Using the curve fitting with the least squares method, we get the thresholds $\{\gamma_n\}$, which are listed in Table 3.

Table 3 Codes and calculated γ_n (dB) for target $\text{BER}_0 = 10^{-3}$

n	M_n	a_n	b_n	γ_n
1	4	896.0704	10.7367	7.1
2	8	404.4353	6.8043	11.8
3	16	996.5492	8.7345	14.0
4	32	443.1272	8.2282	17.0
5	64	296.6007	7.9270	20.1
6	128	327.4874	8.2036	23.0
7	256	404.2837	7.8824	26.2
8	512	310.5283	8.2425	29.0

If we choose $\text{T}_\text{s} = 0.0083\,\text{ms}$, and $\text{h} = 0.7\,\text{s}$, the transmission rate will be 0.24E+6 bps, 0.36E+6 bps, 0.48E+6 bps, 0.60E+6 bps, 0.72E+6 bps, 0.84E+6 bps, and 0.96E+6 bps. According to (1) and (2), the total region of ED will be $[0.0095\,\text{s}, 1.7419\,\text{s}]$. Based on Theorem 1, we can get the following controller gain

$$\boldsymbol{K} = \left[\begin{array}{cc} -0.0446 & -0.0231 \end{array}\right]$$

On the other hand, if there is no ACM used in WNCS, the total region of ED will be $[0.0465\,\text{s}, 2.3222\,\text{s}]$. The following controller can be obtained

$$\boldsymbol{K} = \left[\begin{array}{cc} -0.5568 & -0.5704 \end{array}\right]$$

The responses of the resulting closed-loop system with and without ACM scheme are shown in Figs. 4 and 5, where the initial condition is $\boldsymbol{x}(0) = [0.08; 0.1]$. It is easy to see that the system performance is improved when there exists ACM scheme in WNCS. The settling time, when ACM is applied (5 s for x_1 and 10 s for x_2), is less than that without ACM (17 s for x_1 and 40 s for x_2). Moreover, the overshoot of the states with ACM schemes are less than the ones without ACM.

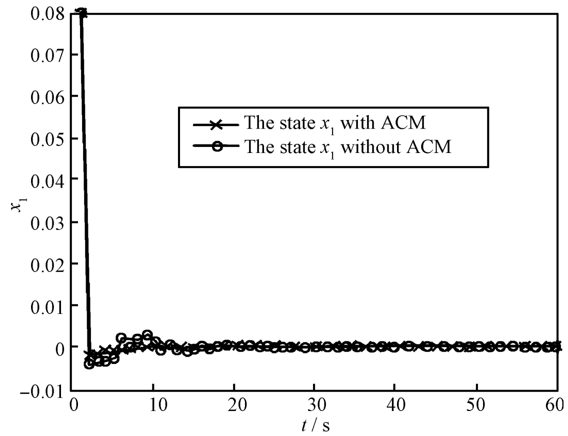

Fig. 4 Responses of state x_1

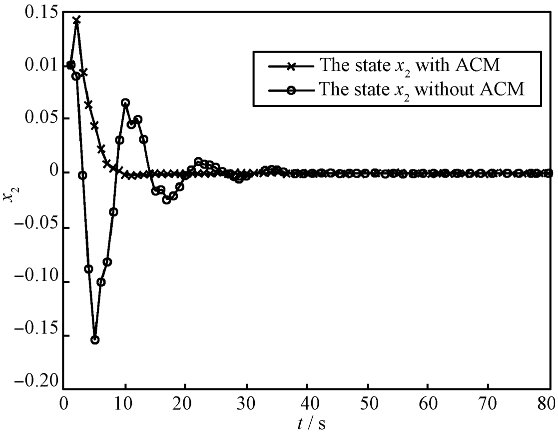

Fig. 5 Responses of state x_2

5 Conclusion

In this paper, we have modelled the WNCS with ACM scheme as a discrete-time system with time-varying input delay. The concept of ED is presented to capture the characteristics of varying rate, interference, and routing in wireless transmission channel. Delay-dependent sufficient conditions on the stability of the closed-loop WNCS and the state feedback controller are derived in terms of linear matrix inequalities by using a recent result on the discrete time delay. Simulation result shows the effectiveness and advantage of the proposed method.

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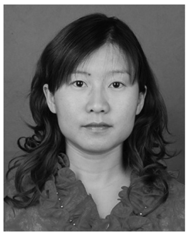

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