Global Sampled-Data Output Feedback Stabilization for Nonlinear Systems via Intermittent Hold

Le Chang, Cheng Fu, *Student Member, IEEE*, and Huanjun Zhang

***Abstract*—This paper introduces a sampled-data and intermittent-hold controller for nonlinear feedforward systems. The intermittent hold allows the control signal to be held in a portion of each sampled period, which does not require the control to be persistently implemented, and thus has less control time. But, less control time degrades the performance of a continuous-time control system or even destabilizes it, especially when the holding portion is sufficiently small. To tackle this obstacle, we first introduce the notion of activating rate to describe the intermittent hold, and give the sampled-data and intermittent-hold controller based on some tuning parameters. Then it is proved that for any activating rate, these parameters can be designed to achieve the stability of the considered systems under appropriately choosing the sampling size. Finally, simulation examples are given to illustrate the effectiveness of the proposed method.**

***Index Terms*—Adaptive control, intermittent problem, nonlinear continuous-discrete system, nonlinear uncertain system, sampled-data control.**

I. Introduction

THE sampled-data control problem is an important research topic [1], and it aims to analyze the behavior of a continuous-time system that is controlled by a digital device. A digital-to-analog converter, such as the zero-order hold, is generally required to permit the system to possess a discontinuous input signal. However, in some scenarios, this converter works intermittently, and the control is not persistently implemented. For example, spacecrafts in [2] can not leave the engine on with a limited fuel supply and the engine only operate for a short period of an orbit period time. Other examples were the network attack [3], [4], and the communication network was intermittently working [5]. This mechanism, possibly due to constraints, failures or requirements, does not operate continuously but rather intermittent. The challenge to come is how to ensure system stability, especially for the nonlinear system which is too complex to get the solution.

The intermittent hold is introduced to describe the case when the control input is only holding during a portion of the sampling periods (see Fig. 1). The control signal with intermittent hold is a piece-wise constant, but periods of open-loop control are combined with feedback control. Such a control was considered in [6] when the control input is missing in the sampled-data system. It is also related to the intermittent control [7]–[10], where the continuous-time control signal is combined with intermittent feedback. The intermittent controller can also be studied when the control direction is unknown [11] or the fault controller [12]. Since more and more controllers are implemented on digital computers in practice, developing the sampling control with intermittent hold is a better choice.

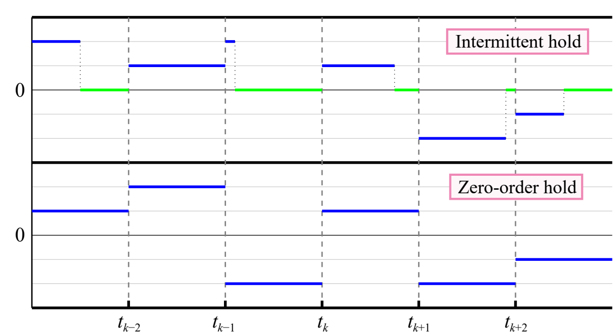

Fig. 1. The comparison between intermittent hold and zero-order hold. The blue line means holding the signal, while the green line means rest.

For a nonlinear system, designing a sampled-data feedback controller leads to challenging control problems. As summarized in [13], there are usually two approaches to solve the sampled-data control problem for nonlinear systems. One approach is first to derive a discrete-time model of the nonlinear system by integration, and then to design the controller using the obtained discrete-time model. Through representing the evolution of the system state at sampling times, [14] designed the sampled-data output feedback stabilizing controller for strict-feedback nonlinear systems. This approach was also developed in [15], [16] to design the sampled-data observer for nonlinear systems. However, for the sampled-data control system with intermittent hold, even if a discrete-time approximate model is obtained by the Euler discretiza-

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L. Chang is with the College of Electric Power Engineering, Shanghai University of Electric Power, Shanghai 200090, China (e-mail: lchang@shiep.edu.cn).

C. Fu is with the School of Automation, Qingdao University, Qingdao 266071, China (e-mail: cfu@qdu.edu.cn).

H. Zhang is with the School of Mathematics and Statistics, Shandong Normal University, Jinan 250014, China (e-mail: zhhuanjun@sdnu.edu.cn).

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tion for instance, how to design a discrete-time stabilizer is still an open problem since the Euler model is dependent on the holding length. Another approach is first designing a continuous-time controller using classical methods and discretizing it. It is called the emulation method, which has been employed in many results, e.g., [17]–[19], to solve the stabilizing control design problem for nonlinear systems. For a zero-order hold with sufficiently small sampling intervals, the emulation method gives an approximation of the continuous-time control problem, which guarantees the system performance under the designed controller. Different from the zero-order hold, the intermittent hold may still make the control approximating the pulse control, which means that the system stability is difficult to be guaranteed even for a small sampling size. Therefore, these existing approaches require the data to be the zero-order hold during the sampling cycle, which is no longer possible for the considered intermittent hold in this paper.

The feedforward nonlinear system belongs to an important class of nonlinear systems, and many excellent results have been obtained on stabilization problem of feedforward nonlinear systems, for instance [20]–[24]. Two main approaches are developed to design of the stabilizing controllers. One is based on the saturated method. For example, the saturation control design method was introduced to design the stabilizing controller for the feedforward nonlinear system in [20], and recently a dynamic gain-based saturation control method was developed in [25]. The low gain feedback control method is another approach to design the stabilizing controller. It was introduced in [23], [26] to study a feedforward nonlinear system, and developed in [27]–[30] for complex or uncertain cases. For the sampled-data control problem, [18], [31]–[33] developed the stabilizing controllers by employing the zero-order hold for feedforward nonlinear systems. Since the feedforward nonlinear system has a tendency to be controlled by a bounded control [34] or a saturated control [25], the study of sampled-data feedback control on feedforward nonlinear systems has some advantages, such as the arbitrary sampling size. Thus, it is necessary to study the intermittent hold for a class of feedforward nonlinear systems, which will be studied in current paper.

In this paper, we will address the issue of global stabilization when the system input is intermittent hold. Since the hold length can be varying, the emulation method for zero-order-hold problem becomes invalid even for a small sampling size. Moreover, the discrete-time model of a continuous-time system includes an extra parameter. The difficulties come from both getting discrete-time model for a nonlinear system and building the controller for a parameter-depended system. The contributions of this work can be characterized by the following novel features: 1) We do not require the control to be continuously implemented, which is widely considered in the sampled-data control systems such as [18], [32], [33], and thus reduce the control time; 2) It is proved that for any non-zero holding length, the stability of the considered systems can be guaranteed by choosing appropriate sampling size and control parameters; and 3) Since the holding length can be sufficient small, the controller is approximating the impulsive control. Hence, our method builds a relationship between the sampled zero-order controller and the impulsive control.

This paper is organized as below: Section II describes the stabilizing problem. Section III introduces the control design method, and analyzes the system stability. Simulation examples are presented in Section IV, and the concluding remarks are provided in Section V. A reference list ends this paper.

Notations: $\mathbb{R}$ is the set of real numbers, and $\mathbb{R}^n$ denotes the n-dimensional real number space. I is the identity matrix of appropriate dimension. $\|\cdot\|$ denotes the Euclidean norm for vectors, or the induced Euclidean norm for a matrix. For any matrix x, x^T is its transpose. We use $x_i(t)$ to represent the ith element of state $x(t)$, and x_i to represent the value of state $x(t)$ at the instant t_i.

II. Problem Formulation

The framework in this paper is shown as Fig. 2. Consider the nonlinear system

$$\dot{x}(t) = Ax(t) + Bu(t) + F(u(t), x(t)) \tag{1}$$

where $x \in \mathbb{R}^n$ is the system state, $u \in \mathbb{R}$ is the system input. The initial instant t_0 is assumed as 0, and the initial state is $x(0) \in \mathbb{R}^n$. Matrices $A \in \mathbb{R}^{n\times n}$, $B \in \mathbb{R}^{n\times 1}$ are in the form

$$A = \begin{pmatrix} 0 & 1 & 0 & \cdots & 0 \\ 0 & 0 & 1 & \cdots & 0 \\ \vdots & \vdots & \vdots & & \vdots \\ 0 & 0 & 0 & \cdots & 1 \\ 0 & 0 & 0 & \cdots & 0 \end{pmatrix}, \quad B = \begin{pmatrix} 0 \\ 0 \\ \vdots \\ 0 \\ 1 \end{pmatrix}.$$

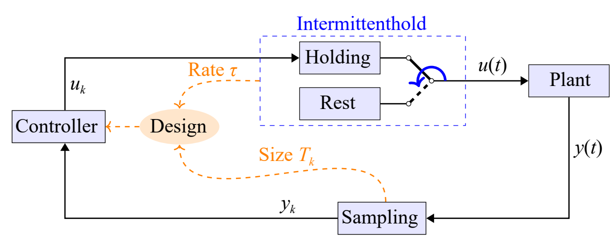

Fig. 2. The system framework in this paper.

The vector function $F(u, x) = (f_1(u, x), \ldots, f_n(u, x))^T$ with $f_i(u, x) : \mathbb{R} \times \mathbb{R}^n \to \mathbb{R}$ being a continuous function with respect to its variables for any $i \in \{1, 2, \ldots, n\}$.

In this paper, we assume that the sampling instants are a given sequence of discrete instants t_k, with $k = 0, 1, 2, \ldots$ being the index. Then, the measured output y_k is described as

$$y_k = Cx(t_k) \tag{2}$$

where $C = (1, 0, 0, \ldots, 0) \in \mathbb{R}^{1\times n}$. Each sampling interval $[t_k, t_{k+1})$ is partitioned into two parts: $[t_k, t_k + d_k)$ and $[t_k + d_k, t_{k+1})$, where d_k is the holding length during kth sampling cycle. During the first part, the controller $u(t)$ is holding the signal u_k which is to be designed. The second part is rest, and it is denoted $u(t) = 0$. Mathematically, it is described as

$$u(t) = \begin{cases} u_k, & t \in [t_k, t_k + d_k) \\ 0, & t \in [t_k + d_k, t_{k+1}). \end{cases} \tag{3}$$

It should satisfy $d_k \in (0, t_{k+1} - t_k]$ to make the signal well-defined. When $d_k \to 0$, the control is characterized by impulsivity. When $d_k = t_{k+1} - t_k$, the control is the conventional zero-order hold. For each period, the control is only activating on the first part $[t_k, t_k + d_k)$. We introduce the activating rate τ_k as

$$\tau_k = \frac{d_k}{t_{k+1} - t_k} \in (0, 1].$$

Remark 1: It is noted that when a limited number of d_k are zero, the problem can be converted into the case we considered. For example, if $d_k \neq 0, d_{k+1} = 0$, we can drop the sampling measurement at the instant t_{k+1}, and re-write the period $[t_k, t_{k+1})$, $[t_{k+1}, t_{k+2})$ as a new sampling period $[t_k, t_{k+2})$. We repeat this process, and the holding length d_k is not zero.

Before giving our objective, we recall the definition of globally asymptotically stable from [35].

Definition 1 (Globally asymptotically stable): Let $x(t, x_0)$ be a solution of $\dot{x} = \phi(x)$ with initial condition $x(0, x_0) = x_0 \in \mathbb{R}^n$. The system $\dot{x} = \phi(x)$ is globally asymptotically stable at equilibrium $x = 0$ if

1) For each $\epsilon > 0$ there is $\delta > 0$ such that

$$\|x_0\| \leq \delta \Rightarrow \|x(t, x_0)\| \leq \epsilon, \ \forall t \geq 0.$$

2) For any initial condition $x_0 \in \mathbb{R}^n$, it holds

$$\lim_{t \to +\infty} x(t, x_0) = 0.$$

Our problem in this paper is that by considering the activating rate τ_k and the sampled size $T_k = t_{k+1} - t_k$, design the variable u_k in (3) to regulate system (1), (2) to be globally asymptotically stable at the equilibrium $x = 0$.

Remark 2: When $\tau_k = 1$, the objective is to design the sampled-data control with zero-order hold. The controller becomes

$$u(t) = u_k, \quad t \in [t_k, t_{k+1})$$

which was considered in [31] for feedforward nonlinear systems. Allowing $\tau_k \in (0, 1]$ is a more general problem. The existing methods are no longer possible to provide the sampled-data controller when τ_k is sufficient small.

III. Feedback Stabilization for Feedforward Nonlinear Systems

For a feedforward nonlinear, the nonlinearities satisfy the following assumption.

Assumption 1: There exists a continuous function $\theta(u) \geq 0$ such that for $i = 1, 2, \ldots, n-1$, it holds

$$|f_i(u, x)| \leq \theta(u)(|x_{i+2}| + |x_{i+3}| + \cdots + |x_n| + |x_{n+1}|)$$

for any $x = (x_1, x_2, \ldots, x_n)^T \in \mathbb{R}^n$ and $u = x_{n+1} \in \mathbb{R}$. Moreover, it holds $f_n(u, x) = 0$.

Remark 3: Assumption 1 is widely considered when studying the feedforward nonlinear system, such as the results in [27]−[29], [32]. Some physical systems such as the rectors chemical system [18], the nonlinear LLC resonant circuit system [21] can be modeled as the feedforward nonlinear system satisfying Assumption 1. The sampled-data control for such a nonlinear system was also designed in [31] with the zero-order hold. But, the sampled-data control problem with intermittent hold has not been reported in the existing literatures, which will be solved in this paper.

A. Controller Design

We design the variable u_k in control (3) as

$$\begin{aligned} u_k &= \mathcal{K}(\rho_k, \tau_k)\hat{x}_k \\ \hat{x}_{k+1} &= (I + \mathcal{L}(T_k, \rho_k)C)\mathcal{M}_1(T_k)\hat{x}_k - \mathcal{L}(T_k, \rho_k)y_{k+1} \\ &\quad + (I + \mathcal{L}(T_k, \rho_k)C)\mathcal{M}_2(T_k, \tau_k)u_k \end{aligned} \tag{4}$$

where $\hat{x}_k$ is updating from an initial value $\hat{x}_0 \in \mathbb{R}^n$, $T_k = t_{k+1} - t_k$, and ρ_k is a dynamic parameter to be designed. The gain matrices are

$$\mathcal{M}_1(T_k) = e^{AT_k}, \ \mathcal{M}_2(T_k, \tau_k) = \int_0^{T_k\tau_k} e^{A(T_k - s)}Bds$$

and

$$\mathcal{K}(\rho_k, \tau_k) = \frac{1}{\tau_k}K\Gamma(\rho_k), \quad \mathcal{L}(T_k, \rho_k) = \frac{T_k}{\rho_k^{n+1}}\Gamma^{-1}(\rho_k)L$$

where $\Gamma(\rho_k) = \mathrm{diag}\left\{\frac{1}{\rho_k^n}, \frac{1}{\rho_k^{n-1}}, \ldots, \frac{1}{\rho_k}\right\}$. Matrices K, L are chosen such that there exist a positive constant γ and a positive definite matrix P satisfying

$$P\mathcal{A}_0 + \mathcal{A}_0^T P \leq -I, \quad P\mathcal{D} + \mathcal{D}P \geq \gamma P \tag{5}$$

where

$$\mathcal{A}_0 = \begin{pmatrix} A + BK & -LC \\ 0 & A + LC \end{pmatrix}, \quad \mathcal{D} = \begin{pmatrix} D & 0 \\ 0 & D \end{pmatrix}$$

with $D = \mathrm{diag}\{n, n-1, \ldots, 1\}$. Although (5) is not linear, the matrices K, L, P can always be achieved, see [23] for more details.

Remark 4: The variable $\hat{x}_k$ is actually estimating the system state $x(t)$ at the sampling instants $\{t_k\}_{k\geq 0}$. Its dynamic in (4) is the discrete-time form of the continuous-discrete observer and gives

$$\begin{aligned} \dot{\hat{x}}(t) &= A\hat{x}(t) + Bu(t), \quad t \in [t_k, t_{k+1}) \\ \hat{x}(t_{k+1}) &= \hat{x}(t_{k+1}^-) + \mathcal{L}(T_k, \rho_k)(C\hat{x}(t_{k+1}^-) - y_{k+1}) \end{aligned}$$

where $\hat{x}(t_{k+1}^-) = \lim_{s>0, s\to 0}\hat{x}(t_{k+1} - s)$. The continuous-discrete observer was studied in [15], [16] for sampled-data control problems. Due to its impulsive character, we extend it by including the activating rate τ_k to solve our problem.

B. Stability Analysis

In the control (4), the only left parameter to be determined is ρ_k. In the following, we show that ρ_k can be determined under a condition on τ_k, T_k.

Theorem 1: Under Assumption 1, for any constant $\tau_{\min} \in (0, 1]$ and $T_{\max} > 0$, if the activating rate τ_k and the sampling size T_k are satisfying $\tau_k \in [\tau_{\min}, 1]$, $T_k \in (0, T_{\max}]$, the dynamic parameter ρ_k can be designed such that system (1) can be globally stabilized through the control (3), (4).

Proof: To analyze the system stability, we consider the closed-loop system (1)−(4). Denote $x_k = x(t_k)$ for $k = 0, 1, 2, \ldots$, and from (1), we get

$$x_{k+1} = \mathcal{M}_1(T_k)x_k + \mathcal{M}_2(T_k,\tau_k)u_k + \mathcal{F}_k \tag{6}$$

where $\mathcal{F}_k = \int_{t_k}^{t_{k+1}} e^{A(t_{k+1}-s)}F(u(s),x(s))ds$. According to (2), the output y_{k+1} is expressed as

$$y_{k+1} = C\mathcal{M}_1(T_k)x_k + C\mathcal{M}_2(T_k,\tau_k)u_k + C\mathcal{F}_k.$$

Let $\tilde{x}_k = x_k - \hat{x}_k$. We substitute (3), (4) into (6) to get the closed-loop system as

$$\begin{cases} \tilde{x}_{k+1} = (I+\mathcal{L}(T_k,\rho_k)C)\mathcal{M}_1(T_k)\tilde{x}_k \\ \qquad\quad + (I+\mathcal{L}(T_k,\rho_k)C)\mathcal{F}_k \\ \hat{x}_{k+1} = \left(\mathcal{M}_1(T_k) + \frac{1}{\tau_k}\mathcal{M}_2(T_k,\tau_k)K\Gamma(\rho_k)\right)\hat{x}_k \\ \qquad\quad - \mathcal{L}(T_k,\rho_k)C\mathcal{M}_1(T_k)\tilde{x}_k - \mathcal{L}(T_k,\rho_k)C\mathcal{F}_k. \end{cases}$$

To continue, we introduce two auxiliary variables

$$z_k = \Gamma(\rho_k)\hat{x}_k, \quad e_k = \Gamma(\rho_k)\tilde{x}_k.$$

Then, we get

$$\begin{aligned} z_{k+1} &= \Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)\Gamma(\rho_k)\hat{x}_{k+1} \\ &= \Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)\left(\mathcal{M}_1(\frac{T_k}{\rho_k}) + \frac{1}{\tau_k}\mathcal{M}_2(\frac{T_k}{\rho_k},\tau_k)K\right)z_k \\ &\quad - \frac{T_k}{\rho_k}\Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)LC\mathcal{M}_1(\frac{T_k}{\rho_k})e_k \\ &\quad - \frac{T_k}{\rho_k}\Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)LC\Gamma(\rho_k)\mathcal{F}_k \end{aligned} \tag{7}$$

and

$$\begin{aligned} e_{k+1} &= \Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)\Gamma(\rho_k)\tilde{x}_{k+1} \\ &= \Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)\left(\Gamma(\rho_k) + \frac{T_k}{\rho_k^{n+1}}LC\right)e^{AT_k}\tilde{x}_k \\ &\quad + \Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)\left(\Gamma(\rho_k) + \frac{T_k}{\rho_k^{n+1}}LC\right)\mathcal{F}_k \\ &= \Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)\left(I + \frac{T_k}{\rho_k}LC\right)\mathcal{M}_1\left(\frac{T_k}{\rho_k}\right)e_k \\ &\quad + \Gamma^{-1}\left(\frac{\rho_k}{\rho_{k+1}}\right)\left(I + \frac{T_k}{\rho_k}LC\right)\Gamma(\rho_k)\mathcal{F}_k \end{aligned} \tag{8}$$

where $C\Gamma(\rho_k) = \frac{1}{\rho_k^n}C$, $\Gamma(\rho_k)e^{AT_k} = e^{A\frac{T_k}{\rho_k}}\Gamma(\rho_k)$ are employed.

Denote $Z_k = (z_k^T, e_k^T)^T$, $\mathcal{H}(\alpha) = \text{diag}\{\Gamma^{-1}(\alpha), \Gamma^{-1}(\alpha)\}$, and the matrix function

$$\mathcal{A}(\alpha,\tau) = \begin{pmatrix} \mathcal{M}_1(\alpha) + \frac{1}{\tau}\mathcal{M}_2(\alpha,\tau)K & -\alpha LC\mathcal{M}_1(\alpha) \\ 0 & (I+\alpha LC)\mathcal{M}_1(\alpha) \end{pmatrix}$$

for the variables α and τ. We arrange the dynamics (7) and (8) into the matrix form

$$Z_{k+1} = \mathcal{H}\left(\frac{\rho_k}{\rho_{k+1}}\right)\left(\mathcal{A}\left(\frac{T_k}{\rho_k},\tau_k\right)Z_k + \tilde{F}_k(\rho_k)\right) \tag{9}$$

where

$$\tilde{F}_k(\rho_k) = \begin{pmatrix} -\frac{T_k}{\rho_k}LC\Gamma(\rho_k)\mathcal{F}_k \\ \left(I + \frac{T_k}{\rho_k}LC\right)\Gamma(\rho_k)\mathcal{F}_k \end{pmatrix}.$$

In the next, we respectively consider the matrix $\mathcal{A}(\alpha,\tau)$ and the nonlinear term $\tilde{F}_k(\rho_k)$.

Since

$$\frac{1}{\tau}M_2(\alpha,\tau) = \int_0^{\alpha} e^{A(\alpha-\tau s)}Bds$$

each element in the matrix $\mathcal{A}(\alpha,\tau)$ is of class C^∞ on $\mathbb{R}\times\mathbb{R}$. Thus, the norms of matrices $\mathcal{A}(\alpha,\tau)$, $\partial\mathcal{A}(\alpha,\tau)/\partial\alpha$ and $\partial^2\mathcal{A}(\alpha,\tau)/\partial\alpha^2$ are bounded on the area $(0,1]\times(0,1]$.

We consider the function

$$\omega(\alpha,\tau,v) = v^T\mathcal{A}^T(\alpha,\tau)P\mathcal{A}(\alpha,\tau)v$$

on the area $(0,1]\times(0,1]\times\Omega$ with $\Omega = \{v \mid \|v\| = 1\}$. Using the Taylor formula, one obtains

$$\omega(\alpha,\tau,v) = v^TPv - \alpha v^Tv + \left.\frac{\partial^2\omega(\alpha,\tau,v)}{\partial\alpha^2}\right|_{\alpha=\epsilon}\alpha^2$$

for any $\alpha\in(0,1]$, $\tau\in(0,1]$ and some $\epsilon\in(0,1]$. Here, $\omega(0,\tau,v) = v^TPv$ and $\partial\omega(\alpha,\tau,v)/\partial\alpha|_{\alpha=0} = -v^Tv$ are employed. It is noted that $\omega(\alpha,\tau,v)$ is also of class C^∞, and thus $\partial^2\omega(\alpha,\tau,v)/\partial\alpha^2$ is bounded on the area $(0,1]\times(0,1]\times\Omega$. We can find a constant α_m such that

$$\omega(\alpha,\tau,v) \le v^TPv - \frac{1}{2}\alpha v^TPv$$

holds for any $\alpha\in(0,\alpha_m]$, $\tau\in(0,1]$, $v\in\Omega$. Because v can be any vector in Ω, we achieve that for any $\alpha\in(0,\alpha_m]$, $\tau\in(0,1]$, it holds

$$\mathcal{A}^T(\alpha,\tau)P\mathcal{A}(\alpha,\tau) \le P - \frac{\alpha}{2}P. \tag{10}$$

Then, we estimate the term $\tilde{F}_k(\rho_k)$. Before doing this, we first consider $\Gamma(\rho_k)F(u,x)$. Under Assumption 1, we get

$$\begin{aligned} &\left|\frac{1}{\rho_k^{n+1-i}}f_i(u(t),x(t))\right| \\ &\quad \le \frac{\theta(u)}{\rho_k^2}\left(\frac{1}{\rho_k^{n-1-i}}|x_{i+2}(t)| + \cdots + \frac{1}{\rho_k}|x_n(t)| + |u(t)|\right) \\ &\quad \le \frac{\theta(u)}{\rho_k^2}\left(\sqrt{n}\|\Gamma(\rho_k)x(t)\| + |u(t)|\right) \end{aligned}$$

for $i = 1,2,\ldots,n-2$, and

$$\left|\frac{1}{\rho_k^2}f_i(u(t),x(t))\right| \le \frac{\theta(u)}{\rho_k^2}|u(t)|$$

where $\rho_k \ge 1$ is utilized. During $t\in[t_k,t_{k+1})$, the input $u(t)$ is u_k or 0. Denoting $\theta_k = \max\{\theta(u_k),\theta(0)\}$, one obtains

$$\begin{aligned} &\|\Gamma(\rho_k)F(u(t),x(t))\| \\ &\quad \le \frac{\theta_k}{\rho_k^2}\left(n\|\Gamma(\rho_k)x(t)\| + \sqrt{n}|u(t)|\right), \quad t\in[t_k,t_{k+1}). \end{aligned} \tag{11}$$

On the other hand, considering system (1), we get

$$\Gamma(\rho_k)\dot{x}(t) = \frac{1}{\rho_k}A\Gamma(\rho_k)x(t) + \frac{1}{\rho_k}Bu(t) + \Gamma(\rho_k)F(u(t),x(t))$$

when $t \in [t_k, t_{k+1})$. Then, the norm of $\Gamma(\rho_k)x(t)$ on $[t_k, t_{k+1})$ satisfies

$$\begin{aligned}\frac{d}{dt}\|\Gamma(\rho_k)x(t)\| &\le \|\Gamma(\rho_k)\dot{x}(t)\| \\ &\le (\|A\| + n\theta_k)\|\Gamma(\rho_k)x(t)\| + \left(\|B\| + \theta_k\sqrt{n}\right)|u(t)|.\end{aligned}$$

Using the Grönwall's inequality, we get

$$\begin{aligned}\|\Gamma(\rho_k)x(t)\| \le\; & e^{(t-t_k)(\|A\|+n\theta_k)}\|\Gamma(\rho_k)x_k\| \\ & + e^{(t-t_k)(\|A\|+n\theta_k)}\left(\|B\| + \theta_k\sqrt{n}\right)\int_{t_k}^{t}|u(s)|ds \\ \le\; & e^{(t-t_k)(\|A\|+n\theta_k)}\|e_k + z_k\| \\ & + e^{(\|A\|+n\theta_k)T_k}\left(\|B\| + \theta_k\sqrt{n}\right)\frac{d_k}{\tau_k}|Kz_k| \\ \le\; & e^{(\|A\|+n\theta_k)T_k}\|e_k\| \\ & + e^{(\|A\|+n\theta_k)T_k}\left(\left(\|B\| + \theta_k\sqrt{n}\right)T_k\|K\| + 1\right)\|z_k\|\end{aligned} \tag{12}$$

for any $t \in [t_k, t_{k+1})$, where the relation $T_k = d_k/\tau_k$ is utilized. Back to (11), we achieve the estimation

$$\begin{aligned}&\|\Gamma(\rho_k)F(u(t),x(t))\| \\ &\le \frac{n\theta_k}{\rho_k^2}e^{(\|A\|+n\theta_k)T_k}\left(\left(\|B\| + \theta_k\sqrt{n}\right)T_k\|K\| + 1\right)\|z_k\| \\ &\quad + \frac{\theta_k}{\rho_k^2}\sqrt{n}\frac{1}{\tau_k}\|K\|\|z_k\| + \frac{n\theta_k}{\rho_k^2}e^{(\|A\|+n\theta_k)T_k}\|e_k\|\end{aligned}$$

during $t \in [t_k, t_k + d_k)$, and

$$\begin{aligned}&\|\Gamma(\rho_k)F(u(t),x(t))\| \\ &\le \frac{n\theta_k}{\rho_k^2}e^{(\|A\|+n\theta_k)T_k}\left(\left(\|B\| + \theta_k\sqrt{n}\right)T_k\|K\| + 1\right)\|z_k\| \\ &\quad + \frac{n\theta_k}{\rho_k^2}e^{(\|A\|+n\theta_k)T_k}\|e_k\|\end{aligned}$$

during $t \in [t_k + d_k, t_{k+1})$.

Then, since $\|z_k\| \le \|Z_k\|$ and $\|z_k\| + \|e_k\| \le \sqrt{2}\|Z_k\|$, we get

$$\begin{aligned}\|\Gamma(\rho_k)\mathcal{F}_k\| \le\; & \int_{t_k}^{t_{k+1}} e^{\|A\|T_k}\|\Gamma(\rho_k)F(u(s),x(s))\|ds \\ \le\; & \frac{\theta_k n}{\rho_k^2}T_k e^{(2\|A\|+n\theta_k)T_k}\left(\left(\|B\| + \theta_k\sqrt{n}\right)T_k\|K\| + 1\right)\|z_k\| \\ & + \frac{\theta_k}{\rho_k^2}\frac{d_k}{\tau_k}\sqrt{n}e^{\|A\|T_k}\|K\|\|z_k\| + \frac{\theta_k n}{\rho_k^2}T_k e^{(2\|A\|+n\theta_k)T_k}\|e_k\| \\ \le\; & \frac{\theta_k n}{\rho_k^2}T_k e^{(2\|A\|+n\theta_k)T_k}\left(\left(\|B\| + \theta_k\sqrt{n}\right)T_k\|K\| + \sqrt{2}\right)\|Z_k\| \\ & + \frac{\theta_k}{\rho_k^2}\frac{d_k}{\tau_k}\sqrt{n}e^{\|A\|T_k}\|K\|\|Z_k\|.\end{aligned} \tag{13}$$

Because

$$\begin{aligned}\left\|\tilde{F}_k(\rho_k)\right\| &\le \left\|\frac{T_k}{\rho_k}LC\Gamma(\rho_k)\mathcal{F}_k\right\| + \left\|\left(I + \frac{T_k}{\rho_k}LC\right)\Gamma(\rho_k)\mathcal{F}_k\right\| \\ &\le (2\|LC\| + n)\|\Gamma(\rho_k)\mathcal{F}_k\|\end{aligned}$$

one achieves the estimation

$$\left\|\tilde{F}_k(\rho_k)\right\| \le \frac{T_k}{\rho_k^2}\sigma_k\|Z_k\| \tag{14}$$

where σ_k is the constant depended on θ_k and T_k. It can be chosen as $\sigma_k = \theta_k((2\|LC\| + n)\sqrt{n}e^{\|A\|T_k}\|K\| + (2\|LC\| + n)ne^{(2\|A\|+n\theta_k)T_k}((\|B\| + \theta_k\sqrt{n})T_k\|K\| + \sqrt{2}))$.

Now, we can analyze the stability of system (9) by employing the Lyapunov method. Let the Lyapunov function candidate be

$$V_k = Z_k^T P Z_k, \quad k = 0, 1, \ldots$$

Since

$$\begin{aligned}&\frac{1}{\alpha^\gamma}v^T\mathcal{H}(\alpha)P\mathcal{H}(\alpha)v - v^T P v \\ &= \left.\frac{d(\frac{1}{s^\gamma}v^T\mathcal{H}(s)P\mathcal{H}(s)v)}{ds}\right|_{s=\xi\in(\alpha,1)}(\alpha - 1) \\ &= (\alpha - 1)\frac{1}{s^{1+\gamma}}v^T\mathcal{H}(s)(PD + DP)\mathcal{H}(s)v \\ &\quad - \gamma(\alpha - 1)\frac{1}{s^{1+\gamma}}v^T\mathcal{H}(s)P\mathcal{H}(s)v \\ &\le -(1 - \alpha)\frac{1}{s^{1+\gamma}}v^T\mathcal{H}(s)(PD + DP - \gamma P)\mathcal{H}(s)v \\ &\le 0\end{aligned}$$

holds for any non-zero vector $v \in \Omega = \{v|\ \|v\| = 1\}$ and constant $\alpha \in (0,1]$, we get

$$\mathcal{H}\left(\frac{\rho_k}{\rho_{k+1}}\right)P\mathcal{H}\left(\frac{\rho_k}{\rho_{k+1}}\right) \le \left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma P.$$

Thus, using the estimation (10) and (14), it is calculated as

$$\begin{aligned}V_{k+1} &= Z_{k+1}^T P Z_{k+1} + \left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma \tilde{F}_k^T(\rho_k)P\tilde{F}_k(\rho_k) \\ &\le \left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma V_k - \frac{1}{2}\frac{T_k}{\rho_k}\left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma V_k \\ &\quad + 2\varpi\|P\|\sigma_k\frac{T_k}{\rho_k^2}\left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma\|Z\|_k^2 \\ &\quad + \|P\|\sigma_k^2\frac{T_k^2}{\rho_k^4}\left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma\|Z\|_k^2 \\ &\le \left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma V_k - \frac{1}{2}\frac{T_k}{\rho_k}\left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma V_k \\ &\quad + 2\varpi\|P\|\sigma_k\frac{T_k}{\rho_k^2}\left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma\frac{1}{\lambda_{\min}(P)}V_k \\ &\quad + \|P\|\sigma_k^2\frac{T_k^2}{\rho_k^4}\left(\frac{\rho_k}{\rho_{k+1}}\right)^\gamma\frac{1}{\lambda_{\min}(P)}V_k\end{aligned}$$

where ϖ is employed to denote the upper bounded of $\|\mathcal{A}(\alpha,\tau)\|$ in the area $(0,1]\times(0,1]$, and $\lambda_{\min}(P)$ is the minimal eigenvalue of matrix P.

Through choosing

$$\rho_{k+1} = \max\left\{\frac{\rho_k}{2}\Big(1-\frac{T_k}{2\rho_k}+2\varpi\|P\|\sigma_k\frac{T_k}{\rho_k^2}\frac{1}{\lambda_{\min}(P)} + \|P\|\sigma_k^2\frac{T_k^2}{\rho_k^4}\frac{1}{\lambda_{\min}(P)}\Big)^{\frac{1}{\gamma}},\ \frac{T_k}{\alpha_m},\ \rho_k\right\} \tag{15}$$

with initial value $\rho_0\geq 1$, we obtain

$$V_{k+1}\leq\left(\frac{1}{2}\right)^{\gamma}V_k.$$

Therefore, V_k is decreasing, and Z_k is bounded and converging to 0. Since $|u_k| = |\frac{1}{\tau_k}Kz_k|\leq\frac{1}{\tau_{\min}}\|K\|\|Z_k\|$, we conclude that the input u_k is bounded, which can ensure θ_k and σ_k to be bounded. That is, a constant $\bar{\sigma}$ can be found such that $\sigma_k\leq\overline{\sigma}$. If

$$\rho_k \geq 4\varpi\|P\|\bar{\sigma}\frac{1}{\lambda_{\min}(P)}+2\|P\|\overline{\sigma}^2 T_{\max}\frac{1}{\lambda_{\min}(P)}$$

we get

$$\frac{T_k}{2\rho_k}\geq 2\varpi\|P\|\sigma_k\frac{T_k}{\rho_k^2}\frac{1}{\lambda_{\min}(P)}+\|P\|\sigma_k^2\frac{T_k^2}{\rho_k^4}\frac{1}{\lambda_{\min}(P)}.$$

Then, (15) is turned into $\rho_{k+1} = \max\{\frac{T_k}{\alpha_m},\rho_k\}$. Because $T_k\leq T_{\max}$, we ensure ρ_k is a bounded parameter. Since Z_k is converging to 0, we conclude that the states $\hat{x}_k$, x_k are asymptotic converging to 0. From (12), we also get the upper bound of $\|x(t)\|$ on $[t_k,t_{k+1})$ is proportional to $\|x_k\|$. Thus, we get $\lim_{t\to+\infty}x(t) = 0$, and conclude that system (1)–(4) is globally asymptotically stable at equilibrium $x = 0$. ■

Remark 5: The contributions of Theorem 1 are twofold: 1) A new design approach is developed to dominate the intermittent hold for sampled-data control problem. This approach is inspired by the design principle of continuous-discrete observer [14]–[16]. Different from their results, our result considers the parameter-depended coefficient matrix $\mathcal{A}(\alpha,\tau)$, and the second variable τ is chosen in an open interval $(0,1]$. Thus, additional analysis is processed to ensure the bound of $\mathcal{A}(\alpha,\tau)$ on the open interval $(0,1]\times(0,1]$. 2) It is noted that a main design principle in this paper is that $\int_{t_k}^{t_{k+1}}u_k = K\Gamma(\rho)\hat{x}_k$ which is utilized to achieve the estimation (12) and (13). This means that even for a sufficiently small τ_k, the value of u_k maybe large, but the integrating term $\int_{t_k}^{t_{k+1}}u_k$ is bounded.

Remark 6: Theorem 1 presents a robust result that the control period $[t_k,t_k+d_k)$ is pre-given. We can extend this by design the holding length d_k. For example, the control length is determined through a self/event-triggered mechanism.

Remark 7: The semi-global asymptotical stability can be achieved through designing the dynamic parameter ρ_k as a constant parameter ρ. From the design (15), the dynamic parameter ρ_k is depended on u_k. Since $u_k\leq\frac{1}{\tau_{\min}}\|K\|\|Z_k\|\leq\frac{1}{\tau_{\min}\lambda_{\min}(P)}\|K\|\sqrt{V_k}\leq\frac{1}{\tau_{\min}\lambda_{\min}(P)}\|K\|\sqrt{V_0}$, we can find $\bar{\sigma}_\Omega$ such that $\sigma_k\leq\bar{\sigma}_\Omega$ when $x(0)\in\Omega$ with $\Omega\subset\mathbb{R}^n$ being any closed set. Then, constant ρ can be chosen with the set Ω, and we can design a constant parameter ρ to achieve the semi-global asymptotical stability.

Remark 8: The only design parameter in our control (4) is ρ_k. This parameter ρ_k increases with the sampling size T_k, the activating rate τ_k, the system order n, and the nonlinear growth rate $\theta(u)$. From the definition of V_k, a high parameter ρ_k may result in an overshoot or a slow converging rate. But, to get a sufficient condition, we employ many inequalities to estimate the upper bound of the system states. In this case, a smaller parameter ρ_k may also ensure the system stability.

IV. Simulation

We consider two examples to illustrate the effectiveness of our method.

Example 1: Consider a three-stage rectors chemical system. Following the description in [18], it can be described as

$$\begin{aligned}\dot{x}_1 &= \frac{1-R_1}{V_1}x_2+(k_1+k_2k_3)x_3\\ \dot{x}_2 &= \frac{1-R_2}{V_2}x_3\\ \dot{x}_3 &= \frac{F}{V_3}u\end{aligned} \tag{16}$$

where x_1, x_2 and x_3 are respectively the compositions of the produce streams, R_1 and R_2 denote the recycle flow rates, V_1, V_2 and V_3 denote the rector volumes, k_1, k_2 and k_3 denote the reaction constants, and u, F are the fresh feed rates.

The sampled-data output is measured as

$$y_k = x_1(t_k),\quad k = 0,1,2,\ldots$$

where $t_k = kT$ with T being the sampling size.

Consider the state transformation $z_1 = x_1$, $z_2 = \frac{1-R_1}{V_1}x_2$, $z_3 = \frac{(1-R_1)(1-R_2)}{V_1V_2}x_3$, $v = \frac{(1-R_1)(1-R_2)F}{V_1V_2V_3}u$. Then, we obtain

$$\dot{z}_1 = z_2+\kappa z_3,\quad \dot{z}_2 = z_3,\quad \dot{z}_3 = v$$

where $\kappa = (k_1+k_2k_3)\frac{V_1V_2}{(1-R_1)(1-R_2)}$.

For the three-stage rectors chemical system (16), we choose the parameters as below: The recycle flow rates of the rector chemical system are chosen as $R_1 = 0.4$ m/s and $R_2 = 0.3$ m/s; the rector volumes are chosen as $V_1 = 0.4$ L, $V_2 = 0.5$ L and $V_3 = 0.3$ L; the reaction constants are $k_1 = 0.2$, $k_2 = 0.3$, and $k_3 = 0.1$; the fresh feed rate is $F = 0.1$ m/s.

It can be calculated that $\kappa = 0.110$. It is verified that Assumption 1 is satisfied with $\theta = 0.110$. In this simulation, we consider the periodic sampling, and the sampling size T is $T = 1$. Then, the control is designed as

$$v = \begin{cases}-\dfrac{5\hat{z}_{1,k}}{\tau\rho^3}-\dfrac{8\hat{z}_{2,k}}{\tau\rho^2}-\dfrac{6\hat{z}_{3,k}}{\tau\rho}, & t\in[k,k+\tau)\\ 0, & t\in[k+\tau,k+1)\end{cases} \tag{17}$$

where $\hat{z}_k = (\hat{z}_{1,k},\hat{z}_{2,k},\hat{z}_{3,k})^T$ satisfies

$$\hat{z}_{k+1} = M_1\hat{z}_k+M_2u-M_3y_k.$$

Here, $M_1 = (I + M_3C)e^A$, $M_2 = (I + M_3C)\int_0^{\tau} e^{A(1-s)}ds$, $M_3 = \left(-\frac{6}{\rho}, -\frac{12}{\rho^2}, -\frac{8}{\rho^3}\right)^T$, $A = \begin{pmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 0 \end{pmatrix}$, $C = (1,0,0)$.

To show the effectiveness of our compensating design, we consider the case of the semi-global stabilization. We let $\rho = 15$, and respectively consider the activating rates $\tau = 0.2$, $\tau = 0.4$, $\tau = 0.6$, $\tau = 0.8$, $\tau = 1$. The state trajectory is shown in Fig. 3 with the initial condition $x_0 = (0.02, -0.02, 0.05)^T$, $z_0 = (0,0,0)^T$. We observe that all system states converge to the equilibrium $z_1, z_2, z_3 = 0$. Meanwhile, the trajectories are almost same. Thus, through our compensating design for the intermittent, the system performance is guaranteed, and the system stability can be achieved, even for a small activating rate τ. This makes our result significant in the study of nonlinear systems.

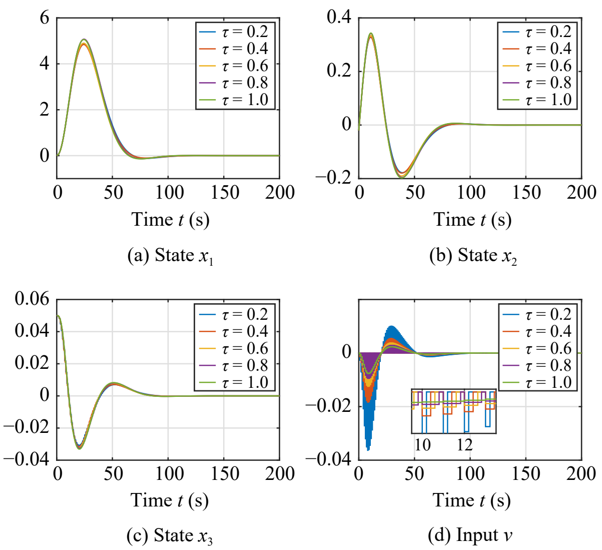

Fig. 3. Trajectory of system (16) under control (17) with different activating rate τ.

Example 2: Consider the nonlinear system

$$\begin{aligned} \dot{x}_1 &= x_2 + 2ux_3 \\ \dot{x}_2 &= x_3 + u^2 \\ \dot{x}_3 &= u \end{aligned} \tag{18}$$

with the output measurement $y_k = x_1(t_k)$, $k = 0,1,2,\ldots$ The instants $\{t_k\}_{k\geq 0}$ are given as $t_0 = 0$, and $t_{i+1} = t_i + T_i, i = 0,1,\ldots$ with T_i being randomly chosen in $[10^{-3}, 10^{-2}]$. We also randomly choose the activating rate τ_k in $[0.1, 1]$. The control $u(t)$ is given as

$$u = \begin{cases} -\dfrac{5\hat{z}_{1,k}}{\tau\rho_k^3} - \dfrac{8\hat{z}_{2,k}}{\tau\rho_k^2} - \dfrac{6\hat{z}_{3,k}}{\tau\rho_k}, & t \in [t_k, t_k + d_k) \\ 0, & t \in [t_k + d_k, t_{k+1}) \end{cases} \tag{19}$$

where $d_k = \tau_k T_k$, and $\hat{z}_k = (\hat{z}_{1,k}, \hat{z}_{2,k}, \hat{z}_{3,k})^T$ satisfies

$$\hat{z}_{k+1} = M_1\hat{z}_k + M_2 u - M_3 y_k.$$

Here,

$$M_1 = (I + M_3C)e^{AT_k}, \quad M_2 = (I + M_3C)\int_0^{T_k\tau_k} e^{A(T_k - s)}ds$$

$$M_3 = T_k\left(-\frac{6}{\rho_k}, -\frac{12}{\rho_k^2}, -\frac{8}{\rho_k^3}\right)^T, \quad A = \begin{pmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 0 \end{pmatrix}$$

and $C = (1,0,0)$. Following Theorem 1, the dynamic parameter ρ_k can be updated as

$$\begin{aligned} \rho_{k+1} = \max\Bigg\{&\frac{\rho_k}{2}\Big(1 - \frac{T_k}{2\rho_k} + 24(|u_k| + e^{|u_k|})\frac{T_k}{\rho_k^2} \\ &+ 100(|u_k| + e^{|u_k|})^2\frac{T_k^2}{\rho_k^4}\Big)^{10}, \ 10^2 T_k, \ \rho_k\Bigg\}. \end{aligned} \tag{20}$$

The simulation results are shown in Fig. 4. One can see that all the system states x_1, x_2, x_3 are converging to 0. Thus, the system is asymptotically stable at the point $x = 0$. Meanwhile, the control input is piece-wise constant, which is the intermittent-hold mechanism as we described. Therefore, the simulation verified the effectiveness of our designed controller.

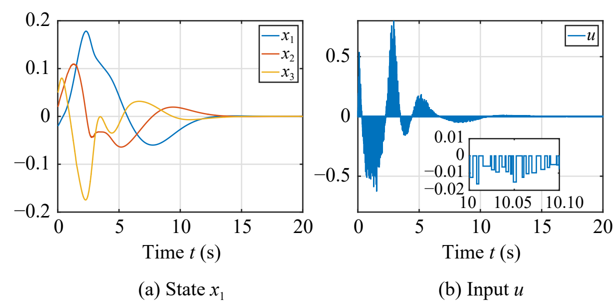

Fig. 4. Trajectory of system (18) under control (19) and (20).

V. Conclusion

This paper considered the sampled-data control with intermittent hold for feedforward nonlinear systems. We assumed the control signal to be holded during a given activating period $[t_k, t_k + d_k)$, and to be zero during the other period $[t_k + d_k, t_{k+1})$. It is proved that the stabilizing controller can be designed for the feedforward nonlinear system (1) under Assumption 1 if $\frac{d_k}{t_{k+1} - t_k} \in [\tau_{\min}, 1]$ with $\tau_{\min} > 0$. The introduced method successfully built a relationship between the impulsive controller and the sampled-data controller. We think the possible future works may consider the more complex systems. For example, how to design the intermittent-hold controller for the uncontrollable system $\dot{x}_1 = x_2$, $\dot{x}_2 = x_3^3$, $\dot{x}_3 = u$?

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Le Chang received the B.S. degree in mathematics and the Ph.D. degree in control science and engineering from Shandong University in 2012 and 2020, respectively. From 2017 to 2019, he was a visiting Ph.D. student with the Swinburne University of Technology, Australia. From 2020 to 2023, he was a Postdoctoral Researcher at the Shanghai Jiao Tong University. Since 2023, he has been a Researcher in the College of Electric Power Engineering, Shanghai University of Electric Power. His research interests include distributed estimation, control of nonlinear systems, and advanced control for power systems.

Cheng Fu (Student Member, IEEE) received the B.S. degree in automation and the M.S. degree in the control science and engineering, Qingdao University, in 2016 and 2019, respectively, the Ph.D. degree in power electronics from Shandong University in 2024. He is currently a Distinguished Professor at the School of Automation, Qingdao University. His research interests include advanced control of power converters, applied nonlinear control and intelligent systems.

Huanjun Zhang received the B.S. degree in mathematics form Shandong Normal University in 2011, and the Ph.D. degree in control science and engineering from Shandong University in 2020. Since 2020, she has been as a Lecturer in the School of Mathematical and Statistics, Shandong Normal University. Her current research interests include stochastic optimal control, and mathematical finance.