TS Fuzzy Model-Based Controller Design for a Class of Nonlinear Systems Including Nonsmooth Functions

Vafamand, Navid; Asemani, Mohammad Hassan; Khayatiyan, Alireza; Khooban, Mohammad Hassan; Dragicevic, Tomislav

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Abstract—This paper proposes a novel robust controller design for a class of nonlinear systems including hard nonlinearity functions. The proposed approach is based on Takagi–Sugeno (TS) fuzzy modeling, nonquadratic Lyapunov function, and nonparallel distributed compensation scheme. In this paper, a novel TS modeling of the nonlinear dynamics with signum functions is proposed. This model can exactly represent the original nonlinear system with hard nonlinearity while the discontinuous signum functions are not approximated. Based on the bounded-input-bounded-output stability scheme and $L_1$ performance criterion, new robust controller design conditions in terms of linear matrix inequalities are derived. Three practical case studies, electric power steering system, a helicopter model and servo-mechanical system, are presented to demonstrate the importance of such class of nonlinear systems comprising signum function. Furthermore, to show the superiorities of the proposed approach, it is applied to these systems; and, the experimental real-time hardware-in-the-loop results are compared with the published literature with the same topic.

Index Terms—Electric power steering (EPS) system, hardware-in-the-loop (HIL), helicopter system, nonsmooth dynamical equations, robust $L_1$ controller, servo-system, signum function, Takagi–Sugeno (TS) fuzzy model.

I. INTRODUCTION

SIGNUM function is a discontinuous hard nonlinearity term that exists in the dynamics of many physical systems such as electrical circuits, mechanical systems and robots [1]–[3]. Signum functions can model different physical and practical phenomena such as friction force that is modeled as a function of the velocity [1], spring force that is a function of the position [4] and backlash which is formulated by linear and signum functions. The signum function has appeared in chaotic systems [5], electromechanical relays and thyristor circuits [6]. The occurrence of the signum function can lead to chattering in the physical systems, due to its discontinuous nature [1], [6].

Controlling the class of nonlinear systems comprising the signum function is difficult. Different nonlinear controllers are proposed for this class of nonlinear systems [7]–[11]. In [7], a motion control is presented for robots with nonlinear friction force modeled by signum function. The controller is designed such that the soft nonlinearities are compensated and then, the hard nonlinearities are controlled. In [12], an adaptive control of a joint robot, that the signum function appears in its dynamic, is studied. This method does not need the exact representation of the model and the system parameters including the coefficients of the signum functions can be unknown. In [13], an adaptive controller is proposed for servo actuator comprising the signum function modeling friction forces. However, introducing update law for parameter estimation of the signum parameters complicates the real implementation of the adaptive controllers [12], [13]. In [9], a robust time delay controller is designed for a smart unmanned aerial vehicle (SUAV). Dry friction is considered in nonlinear dynamic of the SUAV. Based on the error vector, the controller stabilizes the error dynamics to achieve a tracking control problem. However, the approach of [9] suffers from the highly oscillating control input signal. In [10], a local $H_{\infty}$ controller is designed for a servo-system with a flexible shaft involving backlash, coulomb and viscous friction forces. Two dynamical Riccati equations should be solved for a linearized nonsmooth system and based on the simplified linearized model; a linear controller is designed to guarantee the local stabilization of the original system. In [14], a robust switching controller for a two mass system with backlash is proposed. The system dynamic is divided into three modes. For each of these modes a linear controller and backlash compensator is designed. However, the control law must frequently and rapidly switches among these modes with the backlash amplitude decreases. In [11], a sum of square approach is proposed for pitch control of helicopter system. There exists a signum function in the nonlinear dynamic of the helicopter. First, the nonlinear terms are approximated by polynomial functions in several regions. Then, for each region, a polynomial controller is designed based on the semi-definite techniques [15]. However, to achieve a desirable closed-loop performance, higher number of region should be considered which complicates the controller design procedure.
Recently, Takagi–Sugeno (TS) model-based control has been widely used for complex nonlinear systems [16]–[20]. Parallel distributed compensation (PDC) and non-PDC control structures provide systematic approaches to accomplish the stabilization and performance issues for such TS models in terms of linear matrix inequalities (LMIs) [21]–[23]. Most of the existing TS-based control designs are presented for continuous dynamics. To the best knowledge of the authors, few TS-based literatures consider the discontinuous nonlinear dynamics such as signum functions and derive controller design conditions [24], [25].

Since, the signum function is discontinuous, it cannot be exactly represented by a TS fuzzy model. Different methods have been presented for TS modeling of this function. These approaches can be classified in two categories: first, is to approximate the nonsmooth signum function with a smooth one and deriving the corresponding TS model. Second, is to directly derive a TS model based on linearization technique [24], [25]. In the first method, different smooth functions can be considered, such as sigmoid function [26], hyperbolic tangent function [27], [28], and polynomial function [29]. Then, an approximated TS model of these smooth functions can be derived via sector nonlinearity approach. In [30], by approximating a signum function by hyperbolic function, a quasilinear parameter varying model with four vertices is derived. Then, the simplified model is discretized and a controller is designed via LMI techniques. In [25], an approximated TS model of electric power steering (EPS) system is derived. Although, the system comprises three signum functions, by considering some simplifications, a two-rule TS system with time delay state-space representation is obtained. In [24], based on the simplification method given in [25], a new TS modeling of the nonlinear EPS system is presented and PDC controller with saturation constraint is developed. In the first category, the membership functions are smooth and derivable; meanwhile the membership functions obtained based on the second category is nonderivable in general. However, both approaches provide an approximated TS model for the signum function.

In this paper, a new systematic approach for controlling a class of nonlinear systems with the signum function is proposed. This approach is based on the TS fuzzy model, the non-PDC controller and the NQLF. In this approach, the signum functions are not modeled by a TS representation; and, instead of asymptotic stabilization, bounded-input-bounded-output (BIBO) stability criterion is considered. Consequently, based on the $L_1$ performance criterion, a robust non-PDC controller is designed such that the ratio of upper bound of the signum functions with respect to the $L_\infty$ norm of the output is attenuated to be lower than a pregiven threshold. The proposed approach has some main advantages over the existing works. Since we take the signum functions as a persistent bounded disturbance, the obtained TS model exactly represents the nonlinear dynamics comprising the signum functions. In other words, signum functions are not modeled by TS fuzzy system. Consequently, the overall TS fuzzy system with less number of fuzzy rules is obtained which relaxes the controller design conditions. As a result, a simpler TS-based controller is constructed which comprises less number of fuzzy rules. To show the merits of the proposed approach, it is applied to EPS, helicopter, and servo systems and the results are compared with other recently published works concerning the same topic. Simulation results show that not only the BIBO stability is achieved but also the outputs of the aforementioned systems roughly converge to their equilibrium points. In addition, no chattering and oscillatory phenomena exhibit in the closed-looped nonlinear systems.

This paper is organized as follows. In Section II, the class of nonlinear systems with nonsmooth functions is described and three motivating practical EPS system, helicopter, and servo-motor systems are presented. In Section III, the $L_1$ performance criterion, non-PDC scheme and NQLF are studied and the main results of this paper are discussed in Section IV. In Section V, the simulation results are illustrated and compared with recent works in hand. Finally, in Section VI, the conclusion remarks are presented.

II. NONLINEAR SYSTEMS WITH NONSMOOTH FUNCTIONS

Consider the following nonlinear dynamic equation that comprises nonsmooth function:

$$\dot{x}(t) = A(x(t)) + B(x(t))u(t) + E(x(t))\text{nsf}(x(t))$$

$$y(t) = C(x(t))$$

(1)

where $A(\cdot), B(\cdot), E(\cdot)$, and $C(\cdot)$ are nonlinear functions, $x(t) \in \mathbb{R}^{n \times 1}$ and $y(t) \in \mathbb{R}^{m \times 1}$ are the state and controlled output vectors, respectively. Furthermore, $\text{nsf}(x(t)) \in \mathbb{R}^{k \times 1}$ is a vector whose arrays are bounded nonsmooth functions (such as signum and saturation) of the system’s states. We are interested in deriving TS model of the nonlinear dynamic equation (1). However, due to the discontinuity of the hard nonlinear functions such as signum function, their equivalent TS representations are not available [6]. To solve this TS modeling difficulty, we consider the following TS representation:

$$\dot{z}(t) = \sum_{i=1}^{r} h_i(z(t))[A_i x(t) + B_i u(t) + E_i \text{nsf}(x(t))$$

$$y(t) = \sum_{i=1}^{r} h_i(z(t))C_i x(t)$$

(2)

where $z(t) \in \mathbb{R}^{p \times 1}$ is a vector whose elements are bounded, smooth, and functions of the states and $h_i(z(t))$ are the normalized membership functions which satisfy the convex sum property

$$\sum_{i=1}^{r} h_i(z(t)) = 1.$$  

(3)

As it can be seen in (2), the signum functions are explicitly appeared in the TS representation. To show the necessity and applicability of the given structure in (1), in the following, we will present some motivating practical examples that can be restated as the nonlinear dynamic equation (1) and the TS model (2). The two first examples include dry and coulomb friction forces. Meanwhile, the third system involves backlash property.
A. EPS System

The conventional hydraulic steering systems have been replaced by the EPS Systems due to their modularity, tenability of steering feel, and environmental friendliness [24]. The nonlinear dynamic model of the EPS system drawn in Fig. 1 is given by the relation between mechanical steering system, a brush-type direct current motor and tire/road contact forces [25]

\[
\begin{align*}
\dot{\theta}_c &= w_c \\
\dot{w}_c &= \frac{1}{Jc}(T_d - T_c - B_c w_c - F_c \text{sgn}(w_c)) \\
\dot{\theta}_m &= \omega_m \\
\dot{\omega}_m &= \frac{1}{Jm}(T_a - T_m - B_m \omega_m - F_m \text{sgn}(\omega_m)) \\
\dot{x}_r &= \nu_r \\
\dot{\nu}_r &= \frac{T_r + G_c T_m}{\eta_p} - K_m \nu_r - B_r \nu_r - F_r \text{sgn}(\nu_r) \\
\end{align*}
\]

where \(\theta_c\) and \(w_c\) are the steering hand wheel angle and velocity, respectively, \(\theta_m\) and \(\omega_m\) are the motor angular position and velocity, respectively, and \(x_r\) and \(\nu_r\) are the rack position and velocity, respectively. In addition, \(Jc\) is the steering column moment of inertia, \(T_d\) is torque on the steering wheel enforced by the driver, \(T_c\) is the steering torque, \(B_c\) is the steering column viscous damping, and \(F_c\) is the steering column friction. Also, \(Jm\) is the motor moment of inertia, \(T_a\) is the effective assisting torque, \(T_m\) is the servo force, \(B_m\) is the motor damping, and \(F_m\) is the motor friction. Furthermore, \(M_r\) is the rack and wheel assembly mass, \(G_G\) is the motor gear ratio, \(r_p\) is the radius of the pinion, \(k_i\) is the tire spring rate, \(B_r\) is the rack damping, and \(F_r\) is the rack force. In the dynamic of the EPS system (4), the effective assisting torque, steering torque force, and servo force are derived as [24]

\[
\begin{align*}
T_a &= K_a I_m \\
T_c &= K_c (\theta_c - \frac{\nu_r}{r_p}) \\
T_m &= K_m (\theta_m - \frac{G_c T_m}{\eta_p}) \\
\end{align*}
\]

where \(K_a\) is the motor torque constant, \(I_m\) is the armature current, \(K_c\) is the steering column stiffness, and \(K_m\) is the motor torsional stiffness. The rack force \(f_r\) can be derived based on the vehicle model with single track, so-called bicycle model

\[
f_r = \frac{T_p C_f}{r_p} \left\{ \delta_f - \beta - \frac{l_r}{V} \gamma \right\} \tag{6}
\]

where \(\delta_f = \theta_c/G_{sc}\) is the front steer angle, \(G_{sc}\) is the steering system ratio, \(T_p\) is the caster trail, \(C_f\) is the cornering stiffness coefficient, \(l_r\) is the chassis length front of, and \(V\) is speed of the vehicle which is assumed to be constant [25]. In addition, \(\beta\) and \(\gamma\) denote the side slip angle and the yaw rate, respectively. By considering small angle approximations, one has [25]

\[
\begin{align*}
\dot{\beta} &= -\left(\frac{C_f + C_r}{C_m} \right) \beta - \left(\frac{C_f + C_r}{C_m} \right) \gamma + C_f \delta_f \\
\dot{\gamma} &= \left(\frac{C_f + C_r}{C_m} \right) \beta + \left(\frac{C_f + C_r}{C_m} \right) \nu_r + C_f \nu_r \\
\end{align*}
\]

\[
\tag{7}
\]

where \(M\) is the vehicle mass, \(C_r\) is the cornering rear stiffness coefficient, \(l_r\) is the chassis length rear and \(I_c\) is moment of vehicle inertia. The nonlinear dynamic of EPS (4) has three nonlinear sign functions, which are related to coulomb friction [25]. Assume that \(T_d = 0\), therefore, the EPS system (4) together with the dynamic (7) can be reformulated in the following state space representation:

\[
\dot{x} = Ax + Bu + Ensf(x) \tag{8}
\]

where \(x = [\theta_c \ w_c \ \theta_m \ w_m \ \nu_r \ \nu_r \ \beta \ \gamma]^T\) is the state vector, \(u = I_m\) is the control input, \(nsf(x) = [\text{sgn}(w_c) \ \text{sgn}(w_m) \ \text{sgn}(\nu_r)]^T\) is the signum function vector and
B. Helicopter System

Another practical example, whose dynamics can be formulated as in (1), is a rotorcraft system. A state space model was built for an experimental setup of a two-degrees-of-freedom helicopter in [20]. Consider a simplified version of the nonlinear pitch model of the helicopter as follows [11]:

$$\begin{align*}
\dot{x} &= x_2 \\
\dot{x}_2 &= \frac{1}{I_{yy}} \left(-mL_x g \cos(\theta) - mL_z g \sin(\theta) - F_{km} \text{sgn}(w) - F_{v,wm} + u\right)
\end{align*}$$

where $\theta$ and $w$ denote the pitch angle and pitch rate of the helicopter and the definitions of the system parameters are given in [11]. The schematic of the helicopter system is drawn in Fig. 2.

The equilibrium point is $x_1 = -\arctan(L_0/L_2)$ and $x_2 = 0$. The goal is to stabilize the system (9) at the origin [11]. Therefore, an offset input is needed. The dynamic equations (9) can be transformed in such way that the equilibrium point will be at the origin

$$\begin{align*}
\dot{x}_1 &= x_2 \\
\dot{x}_2 &= \frac{1}{I_{yy}} \left(-mL_x g - mL_z g \cos(x_1) - F_{km} \text{sgn}(x_2) - F_{v,wm} x_2\right)
\end{align*}$$

where $x = [x_1 \ x_2]^T = [\theta \ w]^T$ is the state vector. In the following, we are interested in deriving the equivalent TS model of (10). There exist two nonlinear terms $z_1 = mL_x g - mL_z g \cos(x_1) - mL_z g \sin(x_1)$ and $z_2 = sgn(x_2)$. The term $z_1$ satisfies the sector nonlinearity condition [18]

$$a_1 x_1 \leq z_1 \leq a_2 x_2$$

with $a_1 = -0.3034$ and $a_2 = 0.1076$. Based on the (11), one has

$$\begin{align*}
z_1 &= h_1(x_1) a_1 x_1 + h_2(x_1) a_2 x_2 \\
h_1(x_1) &= \frac{a_2 x_1 - z_1}{a_2 - a_1}, h_2(x_1) = \frac{z_1 - a_1 x_1}{a_2 - a_1 x_1}.
\end{align*}$$

The term $z_2$ comprises the signum function that cannot be exactly represented by a TS model. Substituting (12) into (10) provides an equivalent two-rule TS model as

$$\dot{x} = \sum_{i=1}^{2} h_i(x) (A_i x + B_i u^e + E_i \text{sgn}(x))$$

where the normalized membership functions are defined by (12), $u^e = u - mL_x g$ is the new control input, $nsf(x) = -\text{sgn}(x_2)$ is the sign function vector and

$$\begin{align*}
A_1 &= \begin{bmatrix} 0 & 1/F_{v,wm} \\ -F_{km}/I_{yy} & 0 \end{bmatrix}; A_2 &= \begin{bmatrix} 0 & 1/F_{v,wm} \\ F_{km}/I_{yy} & 0 \end{bmatrix} \\
B_1 &= B_2 = \begin{bmatrix} 0 \\ 1/I_{yy} \end{bmatrix}; E_1 = E_2 = \begin{bmatrix} 0 \\ F_{km}/I_{yy} \end{bmatrix}.
\end{align*}$$

C. Servo-System

Consider the dynamics of the servo-system with flexible shaft [10]

$$\begin{align*}
J_m \dot{\theta}_m + c_m \dot{\theta}_m + f_m \text{sgn} (\dot{\theta}_m) + T(\Delta_\theta) &= \tau \\
J_l \ddot{\theta}_l + c_l \ddot{\theta}_l + f_l \text{sgn} (\dot{\theta}_l) &= n T(\Delta_\theta)
\end{align*}$$

where $\theta_m$ and $\dot{\theta}_l$ are the angular position of the motor and the load, and $\tau$ is the input torque. In addition, $J_m$ and $J_l$ are the inertia of the motor and the load, $f_m$ and $f_l$ are the dry friction of the motor and the load, $c_m$ and $c_l$ are the viscous friction of the motor and the load, and $n_k$ is the gear reduction ratio. Furthermore, $T(\Delta_\theta)$ represents the transmitted torque from the motor to the load. The $T(\Delta_\theta)$ is formulated by the following dead zone model:

$$T(\Delta_\theta) = \begin{cases} 0, & \text{if } |\Delta_\theta| \leq j \\ kn_q^{-2} (\Delta_\theta - j \text{sgn}(\Delta_\theta)), & \text{otherwise} \end{cases}$$

where $\Delta_\theta = \theta_m - n_q \theta_l$, $k$ is the torsional spring stiffness and $j$ is the backlash amplitude. The diagram of the servo-system is presented in Fig. 3.

Equation (15) is equivalent to

$$T(\Delta_\theta) = T_1(\Delta_\theta) + T_2(\Delta_\theta)$$

where $T_1(\Delta_\theta) = kn_q^{-2} \Delta_\theta$ is the linear term and $T_2(\Delta_\theta)$ is the nonsmooth saturation term

$$T_2(\Delta_\theta) = \begin{cases} -kn_q^{-2} \Delta_\theta, & \text{if } |\Delta_\theta| \leq j \\ -kn_q^{-2} j \text{sgn}(\Delta_\theta), & \text{otherwise} \end{cases}$$

The overall dead zone characteristics $T(\Delta_\theta)$ and the linear and nonlinear parts $T_1(\Delta_\theta)$ and $T_2(\Delta_\theta)$, are drawn in Fig. 4. As it can be seen in Fig. 4, the saturation function $T_2(\Delta_\theta)$ is bounded.

By substituting (16) and (17) into (14) and applying some simplifications, the following state space representation of the
servo-system will be achieved:

\[
\begin{bmatrix}
\dot{x}_1 \\
\dot{x}_2 \\
\dot{x}_3 \\
\dot{x}_4
\end{bmatrix} =
\begin{bmatrix}
0 & 0 & 1 & 0 \\
-J^{-1}_{m}k_1^{-1}n^{-1} & -J^{-1}_{m}c_1 & J^{-1}_{m}k_1 & 0 \\
0 & 0 & 1 & 0 \\
J^{-1}_{I}k_1 & J^{-1}_{I}c_1 & 0 & 0
\end{bmatrix}
\begin{bmatrix}
x_1 \\
x_2 \\
x_3 \\
x_4
\end{bmatrix} +
\begin{bmatrix}
J^{-1}_{m} \\
0 \\
0 \\
0
\end{bmatrix} u +
\begin{bmatrix}
0 \\
0 \\
0 \\
0
\end{bmatrix} nsf(x). \quad (18)
\]

where \( x = [x_1 \ x_2 \ x_3 \ x_4]^T = [\theta \ x_0 \ \dot{x}_1 \ \dot{x}_2]^T \) is the state vector, \( u = \tau \) is the control input and \( nsf(x) = [-J^{-1}_{m}f_m \text{sgn}(x_2) - J^{-1}_{m}T_2(\Delta_0) - J^{-1}_{I}f_l \text{sgn}(x_4) + J^{-1}_{I}T_2(\Delta_0)]^T \) is the nonsmooth function vector. Equivalently, by defining a new control input

\[
u' = u - f_m \text{sgn}(x_2) - T_2(\Delta_0)
\]

and substituting into (18), one has

\[
\begin{bmatrix}
\dot{x}_1 \\
\dot{x}_2 \\
\dot{x}_3 \\
\dot{x}_4
\end{bmatrix} =
\begin{bmatrix}
0 & 0 & 1 & 0 \\
-J^{-1}_{m}k_1^{-1}n^{-1} & -J^{-1}_{m}c_1 & J^{-1}_{m}k_1 & 0 \\
0 & 0 & 1 & 0 \\
J^{-1}_{I}k_1 & J^{-1}_{I}c_1 & 0 & 0
\end{bmatrix}
\begin{bmatrix}
x_1 \\
x_2 \\
x_3 \\
x_4
\end{bmatrix} +
\begin{bmatrix}
J^{-1}_{m} \\
0 \\
0 \\
0
\end{bmatrix} u' +
\begin{bmatrix}
0 \\
0 \\
0 \\
0
\end{bmatrix} nsf'(x). \quad (20)
\]

where \( nsf'(x) = -J^{-1}_{I}f_l \text{sgn}(x_4) + J^{-1}_{I}T_2(\Delta_0) \). The dynamic equations (18) and (20) are in the form (2).

In the following, we will discuss our TS-based fuzzy approach in tackling with the class of nonsmooth state space representation (2).

### III. New Approach for Handling Signum Functions

In this section, first, BIBO stability analysis and L₁ performance criterion will be defined. Then, nonquadratic Lyapunov function (NQLF) and non-PDC controller will be presented. Finally, the main results of this paper will be proposed.

**A. Nonsmooth Function As Persistent Bounded Disturbance**

As discussed before, the nonsmooth functions cannot be exactly represented by a TS fuzzy model [19], [21], [22]. In this paper, a novel method is proposed for TS fuzzy handling nonsmooth terms in the model. In this approach, the nonsmooth functions are considered as a disturbance vector. The main property of the signum and saturation functions is the inherit boundedness that characterizes the disturbance as a persistent bounded signal. The persistent bounded disturbance belongs to \( L_\infty \) space and generally does not converge to zero when \( t \to \infty \) [31]. It should be noted that, considering signum function as disturbance, does not change the behavior of the original system and the system equilibrium point is maintained at the origin. This is achieved by the fact that \( nsf(0) = 0 \).

By defining the persistent bounded disturbance, the goal is to design a robust controller such that the effect of the disturbance on the closed-loop system output is minimized. For this type of disturbances, one can guarantee the boundedness of the system state vector as \( t \to \infty \). Therefore, the stabilization conditions are derived through bounded input-bounded output (BIBO) stability criterion and the robust controller design conditions are obtained based on L₁ performance index.

**B. L₁ Performance Criterion**

The optimal L₁ performance problem is to design a robust controller, such that the following problem is guaranteed for the closed-loop TS system:

\[
\min_{u} \sup_{x \neq 0} \| y \|_{\infty} = \min_{u} \sqrt{\sup_{x \neq 0} \| nsf(x) \|_{\infty}^2} \leq \Omega \quad (21)
\]

where the infinity norm of a vector signal is defined as [32]

\[
\| v \|_{\infty}^2 = \sup_{t \geq 0} v(t)^2 \quad \text{for} \quad (22)
\]

In the definition (21), the ratio of the infinity norm of the output with respect to the infinity norm of the disturbance is considered which is different from those presented in [31] and [33]. In [31] and [33], the ratio of the infinity norm of the state with respect to the infinity norm of the disturbance is utilized. Therefore, the peak of the output vector \( y \) can be reduced by the level \( \Omega \) under the effect of the peak of the persistent bounded disturbance \( nsf(x) \). The L₁ performance (21) is modified as

\[
\| y \|_{\infty} < \Phi y(0) + \Omega \| nsf(x) \|_{\infty} \quad (23)
\]

where \( \Phi \) is a positive scalar.

**C. Slack Matrices and Fruitful Lemmas**

In this section, we will present some fruitful slack matrices and lemmas that will be used in the proof procedure of the main results. Based on the convex sum property of membership functions (3), the following null term is defined [34]:

\[
\sum_{\rho=1}^{r} \bar{h}_\rho \left[ \sum_{i=1}^{r} \frac{P_i}{r} - \frac{M}{r} \right] = 0 \quad (24)
\]
where $M$ is a symmetric matrix with appropriate dimensions. The null term (24) will be added to the time derivative of NQLF to obtain more relaxed conditions.

**Lemma 1 [31]:** If a real scalar function $S(t)$ satisfies the differential inequality

$$
\dot{S}(t) \leq -\alpha S(t) + \beta v(t)
$$

where $\alpha$ and $\beta$ are positive scalar, then

$$
S(t) \leq e^{-\alpha t} S(0) + \beta \int_0^t e^{-\alpha \tau} v(\tau) d\tau.
$$

**Lemma 2 [35]:** Inequality $\sum_{i=1}^n \sum_{j=1}^n h_{ij} \gamma_{ij} < 0$ is satisfied if

$$
y_{ii} < 0, \quad \frac{1}{2} y_{ii} + y_{jj} + y_{ij} < 0, \quad \text{for } i \neq j, 1, \ldots, r.
$$

**Lemma 3 [18]:** The nonperturbed open-loop TS system (2) (i.e., $E_z = 0$) is asymptotically stabilizable with PDC controller $u = \sum_{i=1}^r h_i(z(t))K_i x(t)$ with the decay rate $\alpha$, if there exists matrices $X = X^T > 0$ and $S_i$ such that

$$
\begin{align*}
X A_i^T + A_i X + S_i^T B_i^T + B_i S_i + \alpha X < 0, \\
X (A_i^T + A_j^T) + (A_i + A_j) X + S_i^T B_j^T + B_j S_i + 2 \alpha X < 0.
\end{align*}
$$

Furthermore, the controller gains are obtained as $K_i = S_i X^{-1}$.

Lemmas 1 and 2 will be used in the main results of this paper in Section IV. In addition, Lemma 3 will be used in Section V.

### IV. MAIN RESULTS

In this section, first, the NQLF and non-PDC controller are presented. Then, through the BIBO stability, the sufficient conditions that assure the boundedness of the closed-loop TS systems including bounded nonsmooth functions are derived.

#### A. Lyapunov Function and State Feedback Controller

To obtain BIBO conditions, the NQLF and the non-PDC controller are considered as follows:

$$
V(x(t)) = x(t)^T P^{-1} x(t)
$$

$$
u(t) = \sum_{i=1}^r h_i(z(t))F_i P^{-1} x(t)
$$

where $P_z = \sum_{i=1}^r h_i(z(t))P_i$ with $P_i = P_i^T > 0$ and $F_i$ are the local feedback gains with appropriate dimensions. By substituting the non-PDC (30) into the open-loop TS system (2), closed-loop system will be obtained as

$$
\dot{x}(t) = \sum_{i=1}^r \sum_{j=1}^r h_i(z(t))h_j(z(t))\{A_i F_j + B_i F_j\} P^{-1} z x(t) + \sum_{i=1}^r h_i(z(t))E_i \text{nsf} (t).
$$

In the following, sufficient conditions will be derived in terms of LMIs such that the closed-loop TS system (31) guarantees the $L_1$ performance criterion (21).

#### B. Controller Design Conditions

The next theorem provides sufficient conditions to minimize the peak-to-peak level performance of the closed-loop TS system.

**Theorem 1:** Suppose that the time derivatives of the membership functions in the closed-loop system (31) have known lower bounds $-\varphi, \rho$ such that

$$
\dot{\varphi} \leq -\varphi, \rho \leq \dot{\rho}
$$

For a given positive scalar $\alpha$, if there exist symmetric matrices $M$ and $P_i$ and matrices $F_i$ for $i = 1, \ldots, r$ such that the following LMIs are satisfied:

$$
P_i > 0 \quad \text{for } i \leq r, \quad P_{\rho} - \sum_{i=1}^r \frac{P_i}{\rho} + \frac{M}{\rho} > 0, \quad \text{for } \rho \leq r
$$

$$
\begin{align*}
Q_{ii} < 0, \quad \frac{1}{2} Q_{ii} + Q_{jj} + Q_{ij} < 0, \quad \text{for } i \neq j, 1, \ldots, r, \\
W_{ii} > 0, \quad \frac{1}{2} W_{ii} + W_{jj} + W_{ij} > 0, \quad \text{for } i \neq j, 1, \ldots, r
\end{align*}
$$

where

$$
Q_{ij} = \left[ \begin{array}{cc} \text{sym}(A_i F_j + B_i F_j) + \alpha P_i + \varphi M & E_i \\
E_i^T & -\beta I \end{array} \right],
$$

$$
W_{ij} = \left[ \begin{array}{cc} \alpha P_i & 0 \\
0 & (\gamma - \beta) I
\end{array} \right].
$$

Then, the peak to peak ($L_1$) performance in (23) is guaranteed with an attenuated level $\Omega = \gamma$. Moreover, and the local feedback matrices of non-PDC controller (30) are derived.

**Proof:** Substituting the closed-loop system (31) into the time derivative of NQLF (29) yields

$$
\dot{V}_Q = \left[ \begin{array}{c} P_z^{-1} x \nsf (x) \\
E_z \\
E_z^T \end{array} \right] \left[ \begin{array}{c} \text{sym}(A_z F_z + B_z F_z) - P_z \\
E_z \\
E_z^T \end{array} \right] \left[ \begin{array}{c} P_z^{-1} x \nsf (x) \\
E_z \\
E_z^T \end{array} \right] - \alpha x^T P_z^{-1} x + \beta \text{nsf}(x) x^T \text{nsf}(x).
$$

By defining $\Gamma_1$ as

$$
\Gamma_1 = \left[ \begin{array}{c} \text{sym}(A_z F_z + B_z F_z) - P_z + \alpha P_z \\
E_z \\
E_z^T \end{array} \right] \left[ \begin{array}{c} P_z^{-1} x \nsf (x) \\
E_z \\
E_z^T \end{array} \right] - \beta I
$$

where $I$ is the identity matrix of appropriate dimensions. In the following, sufficient conditions will be proposed in terms of LMIs to guarantee the negative definiteness of $\Gamma_1$. By adding the null term (24) to (39), $\Gamma_1$ will be continued as

$$
\Gamma_1 = \sum_{i=1}^r \sum_{j=1}^r \left[ \begin{array}{c} \text{sym}(A_i F_j + B_i F_j) + \alpha P_j \\
E_i \\
E_i^T \end{array} \right] \left[ \begin{array}{c} P_i - \sum_{i=1}^r \frac{P_i}{\rho} + \frac{M}{\rho} \\
E_i \\
E_i^T \end{array} \right] - \beta I.
$$
Using (32) and (34), one concludes that
\[-\sum_{\rho=1}^{r} \hat{h}_\rho \left( P_\rho - \sum_{i=1}^{r} \frac{P_i}{r} + \frac{M}{r} \right) < \phi M. \tag{41}\]

Therefore, \( \Gamma_1 \leq \Gamma_2 \), where \( \Gamma_2 \) equals to
\[
\Gamma_2 = \sum_{j=1}^{r} \left[ \left\{ \begin{array}{c}
\text{sym}(A_i P_i + B_i F_j) \\
\alpha + \phi M \end{array} \right\} E_i F_i^T \right]. \tag{42}\]

Based on Lemma 3, the negative definiteness of (42) is enforced if the LMIs (35) are satisfied. Therefore, from (38), one has
\[
\dot{V}_Q \leq -\alpha x^T P_{\infty}^{-1} x + \beta \text{nsf}(x)^T \text{nsf}(x) \\
= -\alpha V_Q + \beta \text{nsf}(x)^T \text{nsf}(x). \tag{43}\]

Considering Lemma 1, we obtain
\[
x^T P_{\infty}^{-1} x \leq e^{-\alpha t} x(0)^T P_{\infty}^{-1} x(t) + \beta \int_0^t e^{-\alpha \tau} \text{nsf}(x(t - \tau))^T \text{nsf}(x(t - \tau)) d\tau \\
\leq \sup_{\tau \in [0,t]} e^{-\alpha \tau} x(0)^T P_{\infty}^{-1} x(t) + \beta \int_0^t e^{-\alpha \tau} \text{nsf}(x(t - \tau))^T \text{nsf}(x(t - \tau)) d\tau \\
\leq x(0)^T P_{\infty}^{-1} x(t) + \frac{\beta}{\alpha} \|\text{nsf}(x(t))\|^T \|\text{nsf}(x(t))\| \tag{44}\]

where \( P_{\infty} = \sum_{i=1}^{r} h_i(z(0)) P_i \). Utilizing (2), one can write the following equality for the system output:
\[
y^T y = [x^T \text{nsf}(x)^T] \begin{bmatrix}
C_z^T & 0 \\
0 & C_z & 0
\end{bmatrix} \begin{bmatrix}
x \\
\text{nsf}(x)
\end{bmatrix}. \tag{45}\]

Now, suppose that the following inequality holds:
\[
y^T y \leq \gamma \left[ x^T P_{\infty}^{-1} \text{nsf}(x)^T \right] \begin{bmatrix}
\alpha P_z & 0 \\
0 & (\gamma - \beta) I
\end{bmatrix} \begin{bmatrix}
P_{\infty}^{-1} x \\
\text{nsf}(x)
\end{bmatrix}. \tag{46}\]

Then, we have
\[
y^T y \leq \gamma \left( \alpha x^T P_{\infty}^{-1} x + (\gamma - \beta) \text{nsf}(x)^T \text{nsf}(x) \right). \tag{47}\]

Substituting (44), results in
\[
y^T y \leq x(0)^T P_{\infty}^{-1} x(0) + \gamma^2 \text{nsf}(x)^T \text{nsf}(x). \tag{48}\]

Therefore, the \( L_1 \) performance will be
\[
\Omega = \sup_{x \neq 0} \frac{\|y\|_\infty}{\|\text{nsf}(x)\|_\infty} = \sqrt{\sup_{x \neq 0} \frac{y^T y}{\|\text{nsf}(x)\|^T \|\text{nsf}(x)\|}} = \gamma. \tag{49}\]

The remaining issue is to derive sufficient conditions for satisfying (46). By applying Schur complement and Lemma 2, (46) is implied by the LMIs (35). The proof is completed.

Remark 1: In Theorem 1, the sufficient conditions of robust controller design is derived such that the effect of nonsmooth functions on the system output is minimized. However, by decreasing this effect, a higher amplitude control input is achieved. In practice, we encounter physical restrictions on employing amplitude-bounded inputs. Therefore, instead of minimizing the \( L_1 \) performance gain in Theorem 1, we may choose this gain in prior. It should be noted that although our approach assures the BIBO stability with a threshold value of \( \gamma \), in most cases, the ultimate bound of the system outputs affected by the disturbance are very less than the prechosen parameter. This property can be observed in the simulation section.

Remark 2: In comparison with the existing methods based on the NQLF and the non-PDC scheme [18]–[23], the proposed control method considers the BIBO stability analysis. Deriving the LMI formulations based on the BIBO scheme has some difficulties over the conventional Lyapunov stability theory: 1) negative definiteness of matrices cannot be directly assured due to the existence of zero diagonal elements and 2) establishing a proper relation between the system output and the nonsmooth functions through the BIBO scheme.

Remark 3: In several practical systems, the source of the nonlinearity is the signum function appearing in the dynamic equation of the system [10], [24], [28], [36], [37]. Therefore, by considering the nonsmooth signum functions as persistent bounded disturbance, a linear system will be remained. Consequently, the nonlinear TS fuzzy (2) alters to a linear system
\[
\dot{x} = Ax + Bu + \text{nsf}(x) \\
y = Cx. \tag{50}\]

Furthermore, the nonlinear non-PDC controller (30) turns to a linear one
\[
u = FP_{\infty}^{-1} x \tag{51}\]

where \( P = P^T > 0 \) is the common Lyapunov matrix and \( F \) is the controller feedback gains. For this special case, we will present following corollary.

Corollary 1: For the system (50) and controller (51), the peak to peak (\( L_1 \)) performance in (23) is guaranteed with an attenuated level \( \Omega = \gamma \), if there exist symmetric matrix \( P \) and matrix \( F \) such that the following LMIs are satisfied:
\[
P > 0 \tag{52}\]
\[
Q = \begin{bmatrix}
\text{sym}(AP + BF) + \alpha P & E \\
E^T & -\beta I
\end{bmatrix} < 0. \tag{53}\]
\[
W = \begin{bmatrix}
\alpha P & \beta E \\
E^T & -\gamma I
\end{bmatrix} > 0. \tag{54}\]

Proof: Let \( A_i = A, B_i = B, C_i = C, E_i = E, \) and \( P_i = P \) for \( i \leq r \) and \( M = 0 \) in the conditions of Theorem 1. Consequently, the conditions of Corollary 1 are obtained. The proof is completed.

Remark 4: Fig. 5 illustrates the general schematic of the proposed robust controller design procedure for nonlinear systems with nonsmooth function.

V. Simulation

In this section, to show the advantages of the proposed approach in controlling the nonlinear dynamic systems including signum functions, we consider the EPS, helicopter, and
A. EPS System

Consider the nonlinear dynamic equation (8) with the parameters given in Table I.

By considering the nonlinear signum functions as persistent bounded disturbance, the dynamic equation (8) turns into a linear system with disturbance. Consequently, a simple linear feedback controller of the following form is obtained:

\[ I_m = FP^{-1}x = Sx. \]  

(55)

By letting values 40, 50, and 117.3 for \( \alpha \), respectively, in Corollary 1, feedback controller (55) with different matrix gains will be achieved. Each of these feedback matrices is obtained based on the corresponding value of \( \alpha \) and specified by a superscript

\[ S^1 = [-0.004 -0.0002 0.0022 0 -5.89 0.0143 -0.028] \]

\[ S^2 = [-0.011 -0.0003 0.0021 0 -5.89 0.0206 -0.131] \]

\[ S^3 = [-1.191 -0.0147 0.0031 0 -5.95 0.3194 -70.001] \]

Furthermore, based on the procedure presented in [24] and [25], the following approximated 2-rule TS model of (8) is achieved:

\[ \dot{x} = \sum_{i=1}^{2} h_i(x)(A_i x + B_i u) \]  

(56)

where the matrices \( B_1 = B_2 \) are equal to \( B \) in (8). Also, the most arrays of \( A_1 = [a_{ij}] \) and \( A_2 = [a_{2ij}] \) are the same as the \( A = [a_{ij}] \) presented in (8) (i.e., \( a_{1ij} = a_{2ij} = a_{ij} \)), except

\[ a_{122} = -\frac{B_c + F_c}{j_c}; \quad a_{144} = -\frac{B_m + F_m}{j_m}; \quad a_{166} = -\frac{C_f + C_r}{MV} \]

\[ a_{222} = -\frac{F_c - B_c}{j_c}; \quad a_{244} = -\frac{F_m - B_m}{j_m}; \quad a_{266} = -\frac{F_r - B_f}{Mr}. \]

Furthermore, the normalized membership functions are obtained as

\[ h_1(x) = \frac{x - x_{2\min}}{x_{2\max} - x_{2\min}}; \quad h_2(x) = \frac{x_{2\max} - x}{x_{2\max} - x_{2\min}}. \]

By applying the proposed method in [18] (Lemma 3) on (56) and letting \( \alpha = 5 \), the following feedback gains are achieved:

\[ K_1 \cong K_2 \cong [-0.175 -0.002 0.0015 0 -5.90 0 -0.003 -0.0004]. \]

Fig. 6(a)–(h) illustrates the closed-loop EPS system derived by Corollary 1 with persistent bounded disturbance (C1-PBD) and the PDC controller [18] with approximated TS model [24] (PDC+App. TS). In Corollary 1, a simple linear controller is designed, meanwhile using the approximated TS model [24] with the controller design [18] provides a nonlinear controller. The control input signal and the error of the closed-loop EPS system output are drawn in Figs. 7(a)–(h) and 8(a) and (b).
Fig. 7 demonstrates that the four controllers can effectively stabilize the EPS system. However, the first three controllers designed based on the proposed approach in this paper, provide higher steady state performance than the controller design based on the approximated TS system. The reason is that due to the discontinuity of the signum functions, the TS model cannot effectively represent them in the neighborhoods of the origin. Therefore, the behavior of these functions cannot be exactly captured by TS models derived based on linearization method. In some cases (such as the EPS system), the origin is the equilibrium point of system, and the system must be stabilized at origin. Since, the TS model is only an approximation of the original nonlinear including signum function system near the equilibrium point, the controller designed by this TS model does not provide a high performance near the equilibrium point. Thus, the closed-loop system evolution has an oscillatory nature near the equilibrium point that is evident in the Fig. 8(b). This disadvantage in modeling signum functions with TS model is eliminated in the proposed approach of this paper. The robust $L_1$ controller design based on the TS modeling with persistent bounded disturbance, can force the states to converge to their equilibrium point without any oscillatory behavior, in this paper.

Furthermore, Table II, demonstrates the $L_2$ and $L_\infty$ norms of the control input, settling time (0.02%), and $L_2$ norm of the EPS output.
Section IV-A, each of these feedback matrices is obtained for different controller matrices gains will be achieved. Similar to the polynomial controller is designed in several ranges and a polynomial controller is designed in [11]. In [11], a piecewise affine (PWA) controller is presented that is applicable for systems with discontinuous vector fields. In both [11] and [24], it is evident that the energy of the control input norms compared to approaches based on the approximated TS models [24]. For instance, by comparing the case $C1 (\alpha = 40)$ and [24], it is evident that the energy of the control input of the proposed approach is extensively reduces the energy (more than 11 times smaller than [24]). In addition, in the viewpoint of EPS system output, the proposed controllers for $\alpha = 50$ and 117.3, outperform the controller designed based on approximated TS model [24].

**B. Helicopter System**

Consider the nonlinear dynamic equation (9) with the parameters given in Table III.

By applying Theorem 1 to the TS fuzzy system (13) and choosing $\varphi = 1$ and $\alpha = 1, 2, \text{ and } 3$, respectively, different controller matrices gains will be achieved. Similar to Section IV-A, each of these feedback matrices is obtained based on the corresponding value of $\alpha$ and specified by a superscript

$$F_1^\alpha = \begin{bmatrix} 0.348 & -0.788 \\ -1.837 & 6.038 \end{bmatrix}; \quad F_2^\alpha = \begin{bmatrix} -0.289 & -0.011 \\ -2.038 & 6.038 \end{bmatrix}$$

$$P_1^\alpha = \begin{bmatrix} 1.566 \\ -1.837 \end{bmatrix}; \quad P_2^\alpha = \begin{bmatrix} 1.566 \\ -2.038 \end{bmatrix}$$

$$F_1^\alpha = \begin{bmatrix} 1.028 \\ -3.169 \end{bmatrix}; \quad F_2^\alpha = \begin{bmatrix} -0.628 \\ -7.062 \end{bmatrix}$$

$$P_1^\alpha = \begin{bmatrix} 3.993 \\ -6.660 \\ 19.013 \end{bmatrix}; \quad P_2^\alpha = \begin{bmatrix} 4.030 \\ -7.062 \\ 20.177 \end{bmatrix}$$

$$F_1^\alpha = \begin{bmatrix} 1.389 \\ -8.421 \end{bmatrix}; \quad F_2^\alpha = \begin{bmatrix} -1.008 \\ -2.617 \end{bmatrix}$$

$$P_1^\alpha = \begin{bmatrix} 5.860 \\ -14.072 \\ 54.229 \end{bmatrix}; \quad P_2^\alpha = \begin{bmatrix} 5.924 \\ -14.70 \end{bmatrix}$$

In [11], a piecewise affine (PWA) controller is presented that is applicable for systems with discontinuous vector fields. In that paper, the nonlinearities are approximated by PWA functions in several ranges and a polynomial controller is designed via sum-of-squares technique for each area [11].

Figs. 9(a) and (b) and 10(a) and (b) demonstrate the helicopter states evolution and control input efforts for the proposed approach based on Theorem 1 ($T1$) and the PWA controller in [11]. Both approaches can stabilize the helicopter states and force the pitch angle to converge to its desired reference. However, as it can be seen in Fig. 9(b), the steady state error of the closed-loop system derived based on $T1$ is close to zero, while the one based on [11] is about 0.2. Therefore, the proposed approach has a better performance compared to [11]. Moreover, it can be seen that by approximate selection of the parameter $\alpha$ in Theorem 1, the steady state error converges sufficiently close to zero.
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Fig. 11. States (a) $x_1, \theta_m$; (b) $x_2, \theta_m$; (c) $x_3, \theta_l$; and (d) $x_4, \theta_l$ (C1 [based on the model (18)] red, C1 [based on the model (20)] by blue).

Table IV demonstrates the metrics of the control input, settling time ($0.02\%$) and the error of the EPS system output at time $t = 4.9$ s.

From Table IV, it is inferred that the proposed approach provides significantly less steady state output error compared to the PWA controller [11]. From the quantitative viewpoint, by comparing the controllers based on the cases $T1 (\alpha = 1)$ and [11], one concludes that the norm-2 and $-\infty$ of the proposed approach is about 1.03 and 1.05 time higher, respectively. However, the steady-state error is reduced about 109 time less than [11]. Consequently, the steady-state performance is greatly improved by the expense of a litter increase in the control input amplitude and energy. Furthermore, by increasing the value of the parameter $\alpha$, the closed-loop convergence speed of the system and the amplitude and the energy of control effort will be increased.

C. Servo-System

Consider the nonlinear dynamic equation of the servo-system (9) with the parameters given in Table V.

Applying Corollary 1 (C1) with $\alpha = 1$ and $\beta = 5$ to the representations (18) and (20), the following control signals are provided (each of the feedback matrices is obtained based on the corresponding system matrices and specified by a superscript):

$$ u = S^1 x $$
$$ S^1 = \begin{bmatrix} 0.8001 & -0.0676 & -9.3101 & -0.0733 \end{bmatrix} $$
$$ u' = u - f_m \text{sgn}(x_2) + T_2(\Delta \theta) = S^2 x $$
$$ S^2 = \begin{bmatrix} 1.1252 & -0.0435 & -8.0555 & -0.0663 \end{bmatrix} $$

Fig. 11 (a)–(d) shows the closed-loop servo-system derived by Corollary 1 based on the (18) and (20). Also, the control input signal and the error of the closed-loop servo-system output are drawn in Fig. 12.

As it can be seen in Fig. 11, the controller designed based on (20) provides a better performance compared to the controller based on (18). In the model (18), the persistent bounded disturbance input is the vector $[-J^{-1}_m f_m \text{sgn}(x_2) - J^{-1}_m T_2(\Delta \theta) - J^{-1}_l f_l \text{sgn}(x_4) + J^{-1}_l T_2(\Delta \theta)]$. Meanwhile, in the model (20), a new controller is defined to completely eliminate the effect of the nonsmooth term $-J^{-1}_m f_m \text{sgn}(x_2) - J^{-1}_m T_2(\Delta \theta)$. Consequently, only the term $-J^{-1}_l f_l \text{sgn}(x_4) + J^{-1}_l T_2(\Delta \theta)$ is considered as the persistent bounded disturbance input and alleviated by the controller (19). Therefore, the designed controller based on the model (20) improves the closed-loop system performance.

VI. CONCLUSION

This paper proposes a novel robust controller design for a class of nonlinear systems including hard nonlinearity signum and saturation functions. The proposed approach is based on TS fuzzy modeling, NQLF and non-PDC scheme. By employing the sector nonlinearity approach, this class of systems is represented by TS fuzzy models. The main advantage of this modeling is that no approximation due to existence of the discontinuous hard nonlinearities is accomplished. Therefore, the obtained TS fuzzy model can exactly represent the original nonlinear system. Based on the inherit properties of signum functions, the BIBO stability scheme,
and $L_1$ performance criterion are considered. Consequently, new robust controller design conditions are derived in terms of LMIs. EPS system, helicopter model, and servo-system are studied as three case studies. In these systems, the friction forces are modeled by sigmoid functions. Experimental simulation results illustrate that the proposed controller can force the states to converge to their equilibrium point without any oscillation behavior. For the future work, deriving the local stability analysis for the class of nonlinear systems with nonsmooth function can be a good research area.

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Authors’ photographs and biographies not available at the time of publication.