

# Data-Driven Control for Linear Discrete-Time Delay Systems

Juan G. Rueda-Escobedo, Emilia Fridman and Johannes Schiffer

**Abstract**—The increasing ease of obtaining and processing data together with the growth in system complexity has sparked the interest in moving from conventional model-based control design towards data-driven concepts. Since in many engineering applications time delays naturally arise and are often a source of instability, we contribute to the data-driven control field by introducing data-based formulas for state feedback control design in linear discrete-time time-delay systems with uncertain delays. With the proposed approach, the problems of system stabilization as well as of guaranteed cost and  $H_\infty$  control design are treated in a unified manner. Extensions to determine the system delays and to ensure robustness in the event of noisy data are also provided.

**Index Terms**—Data-driven control, sampled data control, delay systems, robust control.

## I. INTRODUCTION

There is a growing stream of efforts for developing novel control design methods that only rely on data, enabling a direct control synthesis while avoiding intermediate steps, such as system modeling or system identification [1], [2]. This trend is driven by several factors. These comprise the increasing ease of obtaining and processing data, which is facilitated by modern computers and communication networks, the growth in system complexity in many modern applications and the desire of systematizing the control design.

Although this movement has its roots in computer science [2], where, among other techniques, neural networks, fuzzy systems, online optimization, learning methods, etc., are used for system control, the area of data-driven control has recently shifted towards the development of controller synthesis approaches, which are based on more conventional control strategies. A main reason for this is the need of rigorous guarantees on the system operation, in other words, the need of robust controllers. With this premise in mind, there has been a number of recent contributions in the area of linear system control. The main idea is to assume an underlying linear system to interpret the data and to develop data-driven control formulas which leads to robust controllers by accounting for model mismatches, noise and disturbances. Recent contributions comprise works on linear quadratic tracking [3],

[4], dynamical feedback [5], predictive control [6]–[8], state-feedback and optimal control [9]–[17] as well as extensions to nonlinear discrete Volterra systems and subclasses thereof [18], [19].

Among the most relevant robust control problems is the stabilization of time-delay systems (TDSs). Time delays are an ubiquitous phenomenon in many engineering applications, such as biological and chemical systems as well as networked control and sampled-data systems [20]–[22]. Yet in this important direction, to date there are only few contributions from a data-driven control perspective. One of these is [23], where the authors extend the Virtual Reference Feedback Tuning (VRFT) method to SISO linear discrete-time (LDT) TDS with known input delay. The VRFT is combined with a data-based Smith predictor to account for the effect of the delay. A similar approach is presented in [24] for a SISO linear continuous-time TDSs with unknown input delay. In the field of optimal control for TDSs, a data-driven quadratic guaranteed cost control for continuous time TDSs with known delay, but unknown system matrices is proposed in [25]. Therein, the system data is used to characterize the cost and to update the control gains. In a similar direction, in [26], [27], the authors propose a data-based adaptive dynamic programming method for optimal and  $H_\infty$  control design.

These recent advances motivate the work in the present paper, which is focused on data-driven control design for LDT-TDSs with state and input delays. Inspired by [13] and [21], we provide data-driven formulas for the computation of state feedback gains to achieve system stabilization as well as for guaranteed cost and  $H_\infty$  control design in a unified manner. In contrast to the approaches in [23], [26], [27], the proposed method addresses the case of uncertain and time-varying delays. Furthermore, the impact of noise in the data is analyzed and taken into account for the feedback design, resulting in robust stability guarantees for the closed-loop system. More precisely, the following contributions are made:

- 1) From input-state data and for *known* delays, we provide data-based formulas to replace the system model by the data itself. These formulas can be used for system representation or for control design.
- 2) By using these data-based formulas, we provide data-driven formulas for the design of state-feedback gains. These formulas are given for three control problems: stabilization, control with guaranteed cost, and for  $H_\infty$  control. In all these cases, uncertain delays are considered.
- 3) The proposed approach is extended to the cases of *un-*

J.G. Rueda-Escobedo and J. Schiffer are with the Control Systems and Network Control Technology Group, Brandenburg University of Technology Cottbus-Senftenberg (BTU C-S), 03046 Cottbus, Germany (e-mail: {ruedaesc,schiffer}@b-tu.de).

E. Fridman is with the Department of Electrical Engineering-Systems, Tel-Aviv University, Ramat-Aviv, Tel-Aviv 69978, Israel (e-mail: emilia@eng.tau.ac.il)

known delays and data corrupted by noise. In particular, we provide an algorithm to determine the system delays from disturbed data, and we robustify the data-driven formulas to account for the impact of noise.

The organization of the paper is as follows. In Section II the addressed system is described and the main goals of the paper are outlined. In Section III, a data-based representation for linear discrete-time time-delay systems is introduced. By using the results of Section III, in Section IV data-based formulas for control design are given. The formulas address three basic control problems: stabilization, control with guaranteed cost, and  $H_\infty$  control. In Section V we investigate the effect of uncertainties in the data. The application of the control formulas is illustrated with a numerical example in Section VI. In Section VII some concluding remarks are given. Finally, the proofs of all the claims are given in the Appendix.

### A. Notation

The set of integer numbers is denoted by  $\mathbb{Z}$  and  $\mathbb{R}$  represents the set of real numbers. Let  $\mathbb{F}$  be either  $\mathbb{Z}$  or  $\mathbb{R}$ . Then  $\mathbb{F}_{>0}$  ( $\mathbb{F}_{\geq 0}$ ) denotes the set of all elements of  $\mathbb{F}$  greater than (or equal to) zero. The identity matrix of order  $n \in \mathbb{Z}_{>0}$  is denoted by  $I_n$ . For  $A \in \mathbb{R}^{n \times n}$ ,  $A > 0$  means that  $A$  is symmetric positive definite. The elements below the diagonal of a symmetric matrix are denoted by  $\star$ . Given a matrix  $A \in \mathbb{R}^{n \times m}$ ,  $A^\dagger$  denotes its Moore-Penrose inverse. If  $A$  has full-row rank, we have  $AA^\dagger = I_n$  with

$$A^\dagger = A^\top (AA^\top)^{-1}.$$

For  $v \in \mathbb{R}^n$ ,  $\|v\|_2 = \sqrt{v^\top v}$  denotes the Euclidean norm of  $v$ . For  $A \in \mathbb{R}^{m \times n}$ ,  $\|A\|_2 = \max_{\|v\|_2=1} \|Av\|_2$  with  $v \in \mathbb{R}^n$  denotes the induced Euclidean norm of  $A$ .

Given a signal  $z : \mathbb{Z} \rightarrow \mathbb{R}^n$  and two integers  $k$  and  $r$ , where  $r \geq k$ , we define  $z_{[k,r]} := \{z(k), z(k+1), \dots, z(r)\}$ . Given a signal  $z$  and a positive integer  $T$ , we define

$$Z_{\{i\}} = Z_{\{i,T\}} := [z(i) \quad z(i+1) \cdots z(T+i-1)]. \quad (\text{I.1})$$

Finally, given signals  $x(k) \in \mathbb{R}^n$  and  $h(k) \in \mathbb{Z}_{\geq 0}$  for  $k \in \mathbb{Z}$ , we introduce the short hand  $x_{h(k)}(k) := x(k - h(k))$ . If  $h$  is independent of  $k$ , then  $x_h(k) := x(k - h)$ .

## II. CONSIDERED CLASS OF SYSTEMS AND OBJECTIVES

The following LDT-TDS is considered in this paper:

$$x(k+1) = A_0 x(k) + A_1 x_{h_1(k)}(k) + B u_{h_2(k)}(k), \quad (\text{II.1})$$

with  $k \in \mathbb{Z}_{\geq 0}$ , state vector  $x(k) \in \mathbb{R}^n$  and input  $u(k) \in \mathbb{R}^m$ . Furthermore,  $h_1(k)$  and  $h_2(k)$ , where  $h_1(k) \in \mathbb{Z}_{\geq 0}$  and  $h_2(k) \in \mathbb{Z}_{\geq 0}$ , represent uncertain, bounded delays with upper bound  $\bar{h} \geq h_i(k)$  for  $i = \{1, 2\}$  and all  $k \in \mathbb{Z}_{\geq 0}$ . With respect to the system's initial condition and past inputs we assume  $x(j) = \phi_j \in \mathbb{R}^n$  and  $u(j) = 0$  for  $j \in \mathbb{Z} \cap [-\bar{h}, 0]$ .

The main objective of this paper is to design state feedback controllers *directly from input-state data* that stabilize the system (II.1) in the presence of - possibly uncertain - delays  $h_1(k)$  and  $h_2(k)$ . The analysis is conducted under the following assumptions on the system (II.1).

*Assumption II.1.*

- 1) The system matrices  $A_0$ ,  $A_1$  and  $B$  are constant but unknown.
- 2) An upper bound  $\bar{h} \in \mathbb{Z}_{>0}$  for the input and state delay  $h_1(k)$  and  $h_2(k)$ , respectively, is known.
- 3) Input and state sequences  $u_{[-\bar{h},T]}$  and  $x_{[-\bar{h},T]}$  are available, where  $T \in \mathbb{Z}_{>0}$  with  $T > \bar{h}$  is the number of recorded samples and the delays  $h_1$  and  $h_2$  were constant during the time window in which the data was recorded.

Assumption II.1.1 and Assumption II.1.2 are standard. Assumption II.1.3 can be contextualized as follows. Consider a scenario in which the recorded data is produced in a controlled experiment where the state and input delays are constant. However, during the system operation, the delays might change. Another scenario in which Assumption II.1.3 is reasonable is in networked control. Suppose that for generating the data the system is operated in open-loop and that the input delay remains constant. Then the system state can be recorded locally and transmitted later for its processing. Hence, Assumption II.1.3 follows. However, when the system is operated in closed-loop, the input delay becomes uncertain (but bounded) due to the transmission of the state measurement through the network [21, Sec. 7.8.1], [22, Sec. 3].

In the following, a data-based representation framework is introduced for the system (II.1) under Assumption II.1. At first, this is done for the case of *known* delays. By using the resulting framework, three control designs are derived in Section IV. These are stabilizing control, guaranteed cost control, and  $H_\infty$  control. In addition, in Section V the proposed approach is extended to the case of *unknown* delays and noisy data.

## III. DATA-BASED SYSTEM REPRESENTATION OF LINEAR DISCRETE-TIME TIME-DELAY SYSTEMS

In this section, we derive a data-based representation of the system (II.1) using the data provided by the sequences  $u_{[-\bar{h},T]}$  and  $x_{[-\bar{h},T]}$ , see Assumption II.1.3. To this end, we at first assume that the delays  $h_1$  and  $h_2$  are known and constant. The case of unknown delays is then treated in Section V. Under these considerations, the system (II.1) can be rewritten as

$$x(k+1) = [B \quad A_1 \quad A_0] \begin{bmatrix} u_{h_2(k)} \\ x_{h_1(k)} \\ x(k) \end{bmatrix}. \quad (\text{III.1})$$

To represent the system (III.1) solely by data, consider also the matrices  $U_{h_2,\{0\}} \in \mathbb{R}^{m \times T}$ ,  $X_{h_1,\{0\}} \in \mathbb{R}^{n \times T}$ ,  $X_{\{0\}} \in \mathbb{R}^{n \times T}$ ,  $X_{\{1\}} \in \mathbb{R}^{n \times T}$ , and

$$W_0 := \begin{bmatrix} U_{h_2,\{0\}} \\ X_{h_1,\{0\}} \\ X_{\{0\}} \end{bmatrix} \in \mathbb{R}^{(m+2n) \times T}. \quad (\text{III.2})$$

Here, the matrices  $U_{h_2,\{0\}}$ ,  $X_{h_1,\{0\}}$ ,  $X_{\{0\}}$  and  $X_{\{1\}}$  are built using the sequences corresponding to  $u_{h_2(k)}$ ,  $x_{h_1(k)}$ ,  $x(k)$  and  $x(k+1)$ , respectively, in accordance with the definition given in (I.1). We have the following result.

**Proposition III.1** (Open-Loop Data-Based Representation). *The system trajectories of (III.1) and the ones of the system*

$$x(k+1) = X_{\{1\}} W_0^\dagger \begin{bmatrix} u_{h_2}(k) \\ x_{h_1}(k) \\ x(k) \end{bmatrix} \quad (\text{III.3})$$

are equivalent if and only if  $W_0$  given in (III.2) satisfies  $\text{rank}(W_0) = m + 2n$ . Furthermore, it holds that  $[B \ A_1 \ A_0] = X_{\{1\}} W_0^\dagger$ .

The condition on the rank of  $W_0$  in Proposition III.1 is equivalent to the requirement that the recorded data is rich enough. Since this rank condition is necessary and sufficient, it is analogous to [13, Eq. 6], but for LDT-TDSs of the form (III.1). The rank condition  $\text{rank}(W_0) = m + 2n$  will appear repeatedly along this note. From this rank condition it also follows that a minimal requirement on the data length  $\ell$  is that  $\ell \geq m + 2n + \bar{h}$  (since the sequences  $u_{h_2}(k)$ ,  $x_{h_1}(k)$ ,  $x(k)$  and  $x(k+1)$  are used to build  $W_0$  in (III.2)).

In a similar way, one can find a system representation in closed-loop by using the recorded data. While Proposition III.1 represents an identification-like result, the following lemma provides a system representation that can be used for control design while avoiding the identification of the system matrices.

**Lemma III.2** (Closed-Loop Data-Based Representation). *Consider the system (III.1) and assume a feedback control of the form  $u(k) = Kx(k)$  with  $K \in \mathbb{R}^{m \times n}$ . The trajectories of the closed-loop system*

$$x(k+1) = [BK \ A_1 \ A_0] \begin{bmatrix} x_{h_2}(k) \\ x_{h_1}(k) \\ x(k) \end{bmatrix} \quad (\text{III.4})$$

and the ones of the system

$$x(k+1) = X_{\{1\}} G_K \begin{bmatrix} x_{h_2}(k) \\ x_{h_1}(k) \\ x(k) \end{bmatrix}, \quad (\text{III.5})$$

where  $G_K$  is a  $T \times 3n$  matrix satisfying

$$\begin{bmatrix} K & 0 & 0 \\ 0 & I_n & 0 \\ 0 & 0 & I_n \end{bmatrix} = W_0 G_K, \quad (\text{III.6})$$

are equivalent for every  $K$  if and only if  $\text{rank}(W_0) = m + 2n$  with  $W_0$  given in (III.2). In particular, one has

$$u(k) = U_{h_2, \{0\}} G_K [I_n \ 0 \ 0]^\top x(k). \quad (\text{III.7})$$

Note that Lemma III.2 not only provides a purely data-based representation for the closed-loop system, but also for the control input, and more importantly, for the feedback gain. These characteristics are exploited in the next section for controller synthesis.

**Remark III.3.** By setting  $h_1$  or  $h_2$  in (III.1) to zero, respectively, the following standard LDT-TDS can be described using the same approach as in Proposition III.1:

- $x(k+1) = A_0 x(k) + Bu(k - h_2)$ ,
- $x(k+1) = A_0 x(k) + A_1 x(k - h_1) + Bu(k)$ .

Similarly, the case  $h = h_1 = h_2 > 0$  can be addressed in this manner.

**Remark III.4.** By introducing the augmented state vector, see [21],

$$x_{\text{aug}}(k) = [x^\top(k) \ x^\top(k-1) \ \dots \ x^\top(k-h_1)]^\top,$$

it is possible to obtain an augmented non-delayed system dynamics corresponding to (III.1), namely

$$x_{\text{aug}}(k+1) = A_{\text{aug}} x_{\text{aug}}(k) + B_{\text{aug}} u(k - h_2), \quad (\text{III.8})$$

$$k \in \mathbb{Z}_{\geq 0}, \quad x_{\text{aug}}(k) \in \mathbb{R}^{(h_1+1)n}, \quad u(k) \in \mathbb{R}^m,$$

with

$$A_{\text{aug}} = \begin{bmatrix} A_0 & 0 & \dots & A_1 \\ I_n & 0 & \dots & 0 \\ \vdots & \ddots & \ddots & \vdots \\ 0 & \dots & I_n & 0 \end{bmatrix}, \quad B_{\text{aug}} = \begin{bmatrix} B \\ 0 \\ \vdots \\ 0 \end{bmatrix}. \quad (\text{III.9})$$

In principle, this augmented dynamics could also be used to derive a data-driven formula suitable for control design, e.g. by using the results from [13]. However, this would require that

$$\text{rank} \begin{bmatrix} U_{h_2, \{0\}} \\ X_{\text{aug}, \{0\}} \end{bmatrix} = (h_1 + 1)n + m.$$

Clearly, this can be very demanding in the reasonable scenario that  $h_1 \gg 0$  and also seems unnecessary since only  $A_0$  and  $A_1$  in  $A_{\text{aug}}$  and  $B$  in  $B_{\text{aug}}$  are unknown, see (III.9). In addition, practically meaningful scenarios in which the delay becomes uncertain (and possibly time-varying) in closed-loop cannot be addressed with the augmented dynamics, see also the discussion below Assumption II.1.

**Remark III.5.** In view of the representation (III.9), Willems' Lemma [28, Cor. 2] provides sufficient conditions for guaranteeing  $\text{rank}(W_0) = m + 2n$  with  $W_0$  as in (III.2), i.e., the controllability of the pair  $(A_{\text{aug}}, B_{\text{aug}})$  and the persistence of excitation of order  $(h_1 + 1)n + 1$  of the input sequence.

#### IV. DATA-DRIVEN FORMULAS FOR CONTROLLING LINEAR DISCRETE-TIME TIME-DELAY SYSTEMS

This section is dedicated to the derivation of data-based controller synthesis formulas for the system (II.1) in the presence of uncertain, time-varying, bounded input and state delays  $h_1(k) < \bar{h}$  and  $h_2(k) < \bar{h}$ , respectively. The main tool to achieve this goal is Lemma III.2 together with the recorded data sequences  $u_{[-\bar{h}, T]}$  and  $x_{[-\bar{h}, T]}$ . More precisely, three goals are pursued in this section for the system (II.1):

- 1) Design of a feedback gain for system stabilization.
- 2) Design of a feedback gain, which ensures a prescribed cost for the input and state trajectory.
- 3) Design of a feedback gain, which ensures a prescribed  $L_2$ -gain of the system with respect to additive disturbances.

With regard to item 3), we note that in the present setting the  $H_\infty$  control design is performed in the time domain by using the  $L_2$ -gain, which we recall is defined as the

maximum energy amplification ratio of the system [21]. Also, as discussed in Section II, we account for the event that the delays  $h_1(k)$  and  $h_2(k)$  may become uncertain during the operation of the closed-loop system.

We start with the first item, i.e., system stabilization, which not only is the simplest scenario, but also paves the path for finding solutions to the other two items. Hence, formally the first problem we address is the following.

**Problem IV.1.** Consider the system (II.1) with given  $\bar{h} > 0$  an upper bound for the input and state delays. Find a feedback gain  $K$ , such that the origin of system (II.1) with feedback  $u(k) = Kx(k)$  is an asymptotically equilibrium point for all uncertain delays  $h_1(k) \in [0, \bar{h}]$  and  $h_2(k) \in [0, \bar{h}]$ .  $\nabla\nabla\nabla$

In an analogous fashion to the non-delayed case [13], also in the present delayed setting the matrix  $G_K$  in (III.5) plays the role of a decision variable in a direct data-driven controller synthesis. By exploiting this fact together with the closed-loop data representation given in Lemma III.2, we provide the following solution to Problem IV.1.

**Theorem IV.2** (Stabilization with Static State Feedback). Consider the system (II.1) and suppose that  $\text{rank}(W_0) = m + 2n$  with  $W_0$  as in (III.2). Given a positive delay bound  $\bar{h}$  and a tuning parameter  $\varepsilon > 0$ , let there exist  $n \times n$  matrices  $\bar{P} > 0$ ,  $\bar{S} > 0$ ,  $\bar{R}_i > 0$ ,  $\bar{S}_{12,i}$ , with  $i = \{1, 2\}$ , and  $T \times n$  matrices  $Q_1$ ,  $Q_2$  and  $Q_3$  such that

$$\bar{\Phi} > 0, \quad (IV.1)$$

$$\begin{bmatrix} \bar{R}_i & \bar{S}_{12,i} \\ \star & \bar{R}_i \end{bmatrix} \geq 0, \quad (IV.2)$$

with  $\bar{\Phi}$  given in (IV.5) on p. 5 and

$$\begin{aligned} U_{h_2, \{0\}} Q_2 &= U_{h_2, \{0\}} Q_3 = 0, \\ X_{h_1, \{0\}} Q_1 &= X_{h_1, \{0\}} Q_3 = 0, \\ X_{\{0\}} Q_1 &= X_{\{0\}} Q_2 = 0, \\ X_{h_1, \{0\}} Q_2 &= X_{\{0\}} Q_3. \end{aligned} \quad (IV.3)$$

Choose the feedback gain as

$$K = U_{h_2, \{0\}} Q_1 \left( X_{\{0\}} Q_3 \right)^{-1}. \quad (IV.4)$$

Then for all delays  $h_1(k) \in [0, \bar{h}]$  and  $h_2(k) \in [0, \bar{h}]$  for all  $k \in \mathbb{Z}_{\geq 0}$ , the origin of (II.1) in closed-loop with the control  $u(k) = Kx(k)$  is asymptotically stable.  $\nabla\nabla\nabla$

Note that  $\bar{\Phi} > 0$  in (IV.5) implies that  $\bar{\Phi}_{55} > 0$ , meaning that  $X_{\{0\}} Q_3 + Q_3^\top X_{\{0\}}^\top > 0$ , i.e.,  $X_{\{0\}} Q_3$  is nonsingular. To build the inequalities (IV.1) and (IV.2) only the recorded data from the sequences  $x_{[-\bar{h}, T]}$  and  $u_{[-\bar{h}, T]}$  is needed. Once the matrices  $Q_1$ ,  $Q_2$  and  $Q_3$  are found such that (IV.1), (IV.2) and (IV.3) hold, the feedback gain  $K$  can be computed directly from (IV.4). In this way, the process of identifying the system matrices and the a posteriori controller design is combined into a single direct data-driven synthesis step.

Differently from the non-delay case, see e.g., [13], in the present setting the LMIs (IV.1) and (IV.2) are accompanied by the equality constraints (IV.3). The reason for this lies in the closed-loop representation (III.5) and in particular (III.6).

To see this, consider the right hand-side of (III.6). Not only the matrices in the main diagonal are needed to obtain (III.5), but also the zeros, which gives rise to the equality constraints (IV.3). This does not happen in the non-delay case since there the closed-loop system is fully described by the single matrix  $A + BK$ , while in the present case three separated matrices are required.

Once a solution for Problem IV.1 is given, we can think of including performance criteria in the controller design. For linear systems, it is common to attempt the minimization of the system trajectories and the control effort. This results in a linear quadratic regulator (LQR) design. However, for systems of the form (II.1) an optimal control gain does not exist due to the uncertain delays [21, Sec. 6.2.3]. Instead, one can attempt to find a feedback gain which guarantees a certain cost. This yields the problem formulation below.

**Problem IV.3.** Consider the system (II.1) with  $x(0) = x_0$  and  $x(k) = 0$  for  $k < 0$  with cost function

$$J = \sum_{j=0}^{\infty} z^\top(j) z(j), \quad (IV.6)$$

and performance output

$$z(k) = L_1 x(k) + L_2 x_{h_1(k)}(k) + D u_{h_2(k)}(k), \quad (IV.7)$$

with  $z(k) \in \mathbb{R}^q$  and constant matrices  $L_1 \in \mathbb{R}^{q \times n}$ ,  $L_2 \in \mathbb{R}^{q \times n}$  and  $D \in \mathbb{R}^{q \times m}$ . Given a cost  $\delta > 0$ , find a feedback gain  $K$  that guarantees  $J \leq \delta$  for all uncertain delays  $h_1(k) \in [0, \bar{h}]$  and  $h_2(k) \in [0, \bar{h}]$ .  $\nabla\nabla\nabla$

The result provided in Theorem IV.2 can be extended to address Problem IV.3 by including the effect of the cost  $\delta$  and the functional  $J$  in the inequalities (IV.1) and (IV.2). By doing so, we obtain the following result.

**Corollary IV.4** (Guaranteed Cost Control). Consider the system (II.1) together with the considerations presented in Problem IV.3. Suppose that  $\text{rank}(W_0) = m + 2n$  with  $W_0$  as in (III.2). Given a positive delay bound  $\bar{h}$ , the cost  $\delta > 0$  and a tuning parameter  $\varepsilon > 0$ , let there exist  $n \times n$  matrices  $\bar{P} > 0$ ,  $\bar{S} > 0$ ,  $\bar{R}_i > 0$ ,  $\bar{S}_{12,i}$ , with  $i = \{1, 2\}$ , and  $T \times n$  matrices  $Q_1$ ,  $Q_2$  and  $Q_3$  such that

$$\bar{\Psi} = \begin{bmatrix} \bar{\Phi} & \kappa^\top \\ \star & -I_n \end{bmatrix} > 0 \quad (IV.8)$$

$$\begin{aligned} \kappa &= [L_1 X_{\{0\}} Q_3 \quad L_2 X_{\{0\}} Q_3 \quad D U_{h_2, \{0\}} Q_1 \quad 0 \quad 0], \\ \begin{bmatrix} \bar{R}_i & \bar{S}_{12,i} \\ \star & \bar{R}_i \end{bmatrix} &\geq 0, \end{aligned} \quad (IV.9)$$

together with (IV.3) are satisfied with  $\bar{\Phi}$  given in (IV.5), in addition to

$$\begin{bmatrix} \delta & -x_0^\top \\ \star & X_{\{0\}} Q_3 + Q_3^\top X_{\{0\}}^\top + \bar{P} \end{bmatrix} > 0. \quad (IV.10)$$

Choose the feedback gain

$$K = U_{h_2, \{0\}} Q_1 \left( X_{\{0\}} Q_3 \right)^{-1}. \quad (IV.11)$$

Then for all delays  $h_1(k) \in [0, \bar{h}]$  and  $h_2(k) \in [0, \bar{h}]$  for all  $k \in \mathbb{Z}_{\geq 0}$ , the origin of (II.1) in closed-loop with the

$$\bar{\Phi} = \begin{bmatrix} \bar{\Phi}_{11} & \bar{\Phi}_{12} & \bar{\Phi}_{13} & -\bar{S}_{12,1} - \bar{S}_{12,2} & \bar{\Phi}_{15} \\ \star & 2\bar{R}_1 - \bar{S}_{12,1} - \bar{S}_{12,2}^\top & 0 & -\bar{R}_1 + \bar{S}_{12,1} & -\varepsilon(X_{\{1\}}Q_2)^\top \\ \star & \star & 2\bar{R}_2 - \bar{S}_{12,2} - \bar{S}_{12,2}^\top & -\bar{R}_2 + \bar{S}_{12,2} & -\varepsilon(X_{\{1\}}Q_1)^\top \\ \star & \star & \star & \bar{R}_1 + \bar{R}_2 + \bar{S} & 0 \\ \star & \star & \star & \star & \bar{\Phi}_{55} \end{bmatrix}, \quad (\text{IV.5})$$

$$\begin{aligned} \bar{\Phi}_{11} &= \bar{P} - \bar{S} + (1 - \bar{h}^2)(\bar{R}_1 + \bar{R}_2) - X_{\{1\}}Q_3 - (X_{\{1\}}Q_3)^\top, & \bar{\Phi}_{12} &= -\bar{R}_1 + \bar{S}_{12,1} - X_{\{1\}}Q_2, \\ \bar{\Phi}_{13} &= -\bar{R}_2 + \bar{S}_{12,2} - X_{\{1\}}Q_1, & \bar{\Phi}_{15} &= \bar{h}^2(\bar{R}_1 + \bar{R}_2) + X_{\{0\}}Q_3 - \varepsilon(X_{\{1\}}Q_3)^\top, \\ \bar{\Phi}_{55} &= -\bar{P} - \bar{h}^2(\bar{R}_1 + \bar{R}_2) + \varepsilon(X_{\{0\}}Q_3 + Q_3^\top X_{\{0\}}). \end{aligned}$$

control  $u(k) = Kx(k)$  is exponentially stable. Furthermore, this control ensures a guaranteed cost  $\delta$  for  $J$  given in (IV.6), i.e.,  $J \leq \delta$ .  $\nabla\nabla\nabla$

Another way of introducing performance criteria into the control design is to consider external disturbances affecting the system and to impose restrictions to the response of the system subject to these disturbances. For linear time invariant systems, this is usually done by minimizing the  $H_\infty$  norm of the system. In the present setting, the  $H_\infty$  control design is performed in the time domain by using the  $L_2$ -gain. The resulting control problem is formalized as follows.

**Problem IV.5.** Consider the system (II.1) with an additive disturbance  $\omega(k) \in \mathbb{R}^p$  and feedback gain  $K$ , i.e.,

$$\begin{aligned} x(k+1) &= A_0x(k) + A_1x_{h_1(k)}(k) \\ &+ BKx_{h_2(k)}(k) + D_0\omega(k), \end{aligned} \quad (\text{IV.12})$$

together with the performance output

$$z(k) = L_1x(k) + L_2x_{h_1(k)}(k) + DKx_{h_2(k)}(k), \quad (\text{IV.13})$$

with  $z(k) \in \mathbb{R}^q$  and constant matrices  $L_1 \in \mathbb{R}^{q \times n}$ ,  $L_2 \in \mathbb{R}^{q \times n}$ ,  $D_0 \in \mathbb{R}^{n \times p}$  and  $D_1 \in \mathbb{R}^{q \times m}$ . Fix a constant  $\gamma > 0$ . For all uncertain delays  $h_1(k) \in [0, \bar{h}]$  and  $h_2(k) \in [0, \bar{h}]$ , find a feedback gain  $K$  such that, for  $\omega(k) = 0$ , the origin of (IV.12) is an asymptotically stable equilibrium point and for  $\omega(k) \neq 0$ , the system (IV.12) has an  $L_2$ -gain less than  $\gamma$ .  $\nabla\nabla\nabla$

As before, it is possible to give a solution to Problem IV.5 by extending Theorem IV.2 and including the required conditions in the data-based inequalities (IV.1) and (IV.2). The following results is consistent with such approach.

**Corollary IV.6** (Static  $H_\infty$  Control). Consider the system (IV.12) together with the considerations given in Problem IV.5. Suppose that  $\text{rank}(W_0) = m + 2n$  with  $W_0$  given in (III.2). Given positive constants  $\bar{h}$  and  $\gamma$  and the tuning parameter  $\varepsilon > 0$ , suppose that there exists  $n \times n$  matrices  $\bar{P} > 0$ ,  $\bar{S} > 0$ ,  $\bar{R}_i > 0$  and  $\bar{S}_{12,i} > 0$ , for  $i = \{1, 2\}$ , and  $T \times n$  matrices  $Q_1$ ,  $Q_2$  and  $Q_3$  such that the following data-based inequalities are

satisfied.

$$\bar{\Gamma} = \begin{bmatrix} \bar{\Psi} & \kappa_2^\top \\ \star & \gamma I_n \end{bmatrix} > 0, \quad (\text{IV.14})$$

$$\begin{aligned} \kappa_2 &= \begin{bmatrix} -Q_3^\top X_{\{0\}}^\top D_0^\top & 0 & 0 & 0 & -\varepsilon Q_3^\top X_{\{0\}}^\top D_0^\top & 0 \end{bmatrix}, \\ \begin{bmatrix} \bar{R}_i & \bar{S}_{12,i} \\ \star & \bar{R}_i \end{bmatrix} &\geq 0, \end{aligned} \quad (\text{IV.15})$$

together with (IV.3), and where  $\bar{\Psi}$  is given in (IV.8). Choose the feedback gain

$$K = U_{h_2, \{0\}} Q_1 (X_{\{0\}} Q_3)^{-1}. \quad (\text{IV.16})$$

Then for all delays  $h_1(k) \in [0, \bar{h}]$  and  $h_2(k) \in [0, \bar{h}]$ , the origin of (IV.12) is an asymptotically stable equilibrium point for  $\omega(k) = 0$ . Furthermore, for  $\omega(k) \neq 0$  the system (IV.12) has an  $L_2$ -gain less than  $\gamma$ .  $\nabla\nabla\nabla$

## V. HANDLING UNKNOWN CONSTANT DELAYS AND NOISY DATA

In this section we address the problems that the delays  $h_1$  and  $h_2$  of the system (II.1) are unknown and that the available data is corrupted by noise. To this end, we assume that the sequences  $x_{[-\bar{h}, T]}$  and  $u_{[-\bar{h}, T]}$  can be expressed as the sum of two sequences:

$$\begin{aligned} x_{[-\bar{h}, T]} &= x_{[-\bar{h}, T]}^{\text{nom}} + x_{[-\bar{h}, T]}^\delta, \\ u_{[-\bar{h}, T]} &= u_{[-\bar{h}, T]}^{\text{nom}} + u_{[-\bar{h}, T]}^\delta, \end{aligned} \quad (\text{V.1})$$

where the superscript ‘nom’ denotes the sequence that corresponds to the dynamics (III.1), i.e., the *nominal* part of the data, whereas the superscript ‘ $\delta$ ’ denotes the sequence corresponding to the measurement noise.

Hence, the objectives of this section are to provide formulas for determining the delays  $h_1$  and  $h_2$  from the recorded data and to robustify the design of the feedback gains from Section IV with respect to additive noise.

### A. Data-Based System Representations for Unknown Constant Delays

By using the sequences  $u_{[-\bar{h}, T]}$  and  $x_{[-\bar{h}, T]}$  as described in (V.1), we can build the matrix  $W_0$  as in (III.2). Since the construction of  $W_0$  is linear, it is possible to split it in two

parts, one corresponding to the data generated by the system (III.1), i.e. the nominal ('nom') data, and one to the noise, i.e.,

$$W_0 = W_0^{\text{nom}} + W_0^\delta. \quad (\text{V.2})$$

However,  $W_0$  in (V.2) will not result in useful data for arbitrary  $W_0^\delta$ . To study when  $W_0$  retains the system information, let  $U \in \mathbb{R}^{(m+2n) \times (m+2n)}$  and  $V \in \mathbb{R}^{T \times T}$  be orthonormal matrices such that

$$\text{Range}(U) = \text{Range}(W_0), \quad \text{Range}(V) = \text{Range}(W_0^\top).$$

Consider the factorization [29, Sec. 2]

$$\begin{aligned} W_0 &= U \begin{bmatrix} W_0^{11} & 0 \end{bmatrix} V^\top, \\ W_0^\delta &= U \begin{bmatrix} W_0^{\delta,11} & W_0^{\delta,12} \end{bmatrix} V^\top, \end{aligned} \quad (\text{V.3})$$

where  $W_0^{11} \in \mathbb{R}^{(m+2n) \times (m+2n)}$ ,  $W_0^{\delta,11} \in \mathbb{R}^{(m+2n) \times (m+2n)}$  and  $W_0^{\delta,12} \in \mathbb{R}^{(m+2n) \times T}$ . By using the factorization (V.3), we introduce the following assumptions related to the impact of the noise.

**Assumption V.1.**

- 1)  $\text{rank}(W_0) = \text{rank}(W_0^{\text{nom}}) = m + 2n$ .
- 2)  $\|W_0^{\delta,11}\|_2 \|W_0^\dagger\|_2 = \|W_0^{\delta,11}\|_2 \|(W_0^{11})^{-1}\|_2 < 1$ .

Assumption V.1.1) implies that  $W_0^\delta$  does not modify the rank of  $W_0^{\text{nom}}$ , whereas Assumption V.1.2) restricts the size of the perturbation term  $W_0^\delta$ . The premises of Assumption V.1 ensure that  $W_0$  is an *acute* perturbation of  $W_0^{\text{nom}}$  [29], [30].

Now, in order to identify the system delays, consider the matrices

$$W_{0,(i,j)} = \begin{bmatrix} U_{i,\{0\}} \\ X_{j,\{0\}} \\ X_{\{0\}} \end{bmatrix}, \quad (\text{V.4})$$

for  $i = \{0, 1, \dots, \bar{h}\}$ ,  $j = \{0, 1, \dots, \bar{h}\}$ , where  $U_{i,\{0\}}$  is built with  $u(k-i)$ ,  $X_{j,\{0\}}$  with  $x(k-j)$ , and  $\bar{h} \geq 0$  is the upper bound for the delays. In the unperturbed case, i.e., for  $x_{[-\bar{h},T]}^\delta = 0$  and  $u_{[-\bar{h},T]}^\delta = 0$ , one can verify the next rank conditions in order to determine the system delays

$$\text{rank}(W_{0,(i,j)}) = \text{rank}\left(\begin{bmatrix} W_{0,(i,j)} \\ X_{\{1\}} \end{bmatrix}\right) = m + 2n, \quad (\text{V.5})$$

for all  $i$  and  $j$  in  $\{0, 1, \dots, \bar{h}\}$ . If for some pair  $(i^*, j^*)$  the condition above holds, then one can take  $h_1 = j^*$  and  $h_2 = i^*$  since  $X_{\{1\}}$  belongs to the row space of  $W_{0,(i^*,j^*)}$ . If for two or more pairs  $(i, j)$  the condition (V.5) holds, then it is not possible to identify the delays from the recorded data. However, in the perturbed case the condition (V.5) might never hold due to the effect of the noise. Therefore, instead of (V.5), we propose to use the orthogonal distance of  $X_{\{1\}}$  to the row space of each  $W_{0,(i,j)}$  in order to determine the delays. This yields the next proposition, for the presentation of which we introduce the matrix

$$X_{\{1\}} = X_{\{1\}}^{\text{nom}} + X_{\{1\}}^\delta, \quad (\text{V.6})$$

where  $X_{\{1\}}^{\text{nom}}$  denotes the part of the data that corresponds to the dynamics of (III.1) and  $X_{\{1\}}^\delta$  denotes the part corresponding to the noise. In addition, we define the orthogonal distance

$$d_{(i,j)}(X_{\{1\}}) := \left\| X_{\{1\}} \left( I_T - (W_{0,(i,j)})^\dagger W_{0,(i,j)} \right) \right\|_2, \quad (\text{V.7})$$

and the function  $\psi : \mathbb{R}_{\geq 0} \rightarrow \mathbb{R}_{\geq 0}$

$$\psi(\sigma) = \frac{\sigma}{[1 + \sigma^2]^{1/2}}. \quad (\text{V.8})$$

In addition, upper bounds for the noisy matrices are required. These are represented by the positive, data-dependent constants  $r_{X_{\{1\}}^\delta}$ ,  $r_{W_{0,(i,j)}^{\delta,11}}$  and  $r_{W_{0,(i,j)}^{\delta,12}}$  satisfying

$$\begin{aligned} \|X_{\{1\}}^\delta\|_2 &\leq r_{X_{\{1\}}^\delta}, \quad \|W_{0,(i,j)}^{\delta,12}\|_2 \leq r_{W_{0,(i,j)}^{\delta,12}}, \\ \|W_{0,(i,j)}^{\delta,11}\|_2 \|W_{0,(i,j)}^\dagger\|_2 &\leq r_{W_{0,(i,j)}^{\delta,11}} \|W_{0,(i,j)}^\dagger\|_2 < 1, \end{aligned} \quad (\text{V.9})$$

with  $X_{\{1\}}^\delta$  and  $W_{0,(i,j)}^\delta$  as in (V.6) and (V.2), respectively, and  $W_{0,(i,j)}^{\delta,11}$  and  $W_{0,(i,j)}^{\delta,12}$  as in (V.3). Clearly,  $r_{W_{0,(i,j)}^{\delta,11}}$  exists for any data set consistent with Assumption V.1.2).

**Proposition V.2** (Data-Based System Representations for Unknown Delays). *Consider the system (II.1) with corresponding perturbed data sequences  $x_{[-\bar{h},T]}$  and  $u_{[-\bar{h},T]}$  as introduced in (V.1). Let  $\bar{h} > 0$  be the upper bound for the constant but unknown state and input delays  $h_1 \in \mathbb{Z}_{\geq 0}$  and  $h_2 \in \mathbb{Z}_{\geq 0}$ . Build the matrices  $W_{0,(i,j)}$  as in (V.4) for  $i$  and  $j$  in  $\{0, 1, \dots, \bar{h}\}$ , and suppose that, for all  $i$  and  $j$ , Assumption V.1 holds for each of the matrices  $W_{0,(i,j)}$  and the respective perturbation  $W_{0,(i,j)}^\delta$ . Furthermore, suppose that the constants  $r_{X_{\{1\}}^\delta}$ ,  $r_{W_{0,(i,j)}^{\delta,11}}$  and  $r_{W_{0,(i,j)}^{\delta,12}}$  defined in (V.9) are known.*

*Recall the orthogonal distance  $d_{(i^*,j^*)}(X_{\{1\}})$  given in (V.7) and the function  $\psi(\cdot)$  defined in (V.8). If*

$$d_{(i^*,j^*)}(X_{\{1\}}) \leq r_{X_{\{1\}}^\delta} + (\|X_{\{1\}}\|_2 + r_{X_{\{1\}}^\delta}) \cdot \psi(\sigma^*) \quad (\text{V.10})$$

where

$$\sigma^* = \frac{r_{W_{0,(i^*,j^*)}^{\delta,12}} \|W_{0,(i^*,j^*)}^\dagger\|_2}{1 - r_{W_{0,(i^*,j^*)}^{\delta,11}} \|W_{0,(i^*,j^*)}^\dagger\|_2}, \quad (\text{V.11})$$

for only one pair  $(i^*, j^*)$ , then  $h_1 = j^*$  and  $h_2 = i^*$ . Moreover, the corresponding open- and closed-loop data-based representations are obtained via Proposition III.1 and Lemma III.2, respectively, by using the matrices  $U_{i^*,\{0\}}$ ,  $X_{j^*,\{0\}}$  together with  $X_{\{0\}}$  and  $X_{\{1\}}$ .

*If condition (V.10) holds for two or more pairs  $(i, j)$ , then the delays are not decidable from the available data.  $\nabla \nabla \nabla$*

Proposition V.2 provides a tool for deriving data-based system representations for *unknown* delays, even in the presence of noise. In addition, the delays themselves are also determined.

In general,  $(\bar{h} + 1)^2$  evaluations of (V.10) are required. This number reduces to  $\bar{h} + 1$  if, for example,  $h_1 = h_2$  or if one of the delays is known. Furthermore, the same idea can be used to identify time-dependent delays. However, in such case,  $(\bar{h} + 1)^T$  evaluations are required, which might not be computationally feasible.

## B. Stabilization with Noisy Data

Now we proceed to analyze the impact of noisy data on the controller synthesis formulas derived in Section IV. The main

objective is to extend the result of Theorem IV.2 to incorporate a criterion to ensure closed-loop stability of the system (II.1) even when the feedback gain  $K$  is computed with corrupted data. In order to account for the impact of the noise in the data, consider the following matrix

$$\Delta_{[\cdot]} := [B \quad A_1 \quad A_0] W_0^\delta - X_{\{1\}}^\delta. \quad (\text{V.12})$$

The quantity  $\|\Delta_{[\cdot]}\|_2$  is a measurement of how far the noise is from being a system trajectory. If the noise would correspond to a system trajectory, then it would not affect any of the calculations; though in such case it might not be classified as noise. Therefore, it is logical that only  $\Delta_{[\cdot]}$  has an impact on the computation of  $K$ . By using this measurement of the noise, it is possible to account for it in the feedback design. This approach yields the next result.

**Theorem V.3** (Stabilization with Noisy Data). *Consider the premises of Theorem IV.2. Let the recorded data be corrupted by noise as in (V.1). Suppose that  $\Delta_{[\cdot]}$  in (V.12) is bounded as  $\|\Delta_{[\cdot]}\|_2 \leq \alpha$ , with  $\alpha > 0$  known. Given a positive delay bound  $h$  and a tuning parameter  $\varepsilon > 0$ , let there exist  $n \times n$  matrices  $\bar{P} > 0$ ,  $\bar{S} > 0$ ,  $\bar{R}_i > 0$ ,  $\bar{S}_{12,i}$ , with  $i = \{1, 2\}$ ,  $T \times n$  matrices  $Q_1, Q_2, Q_3$ , and  $\lambda > 0$ , such that*

$$\begin{bmatrix} \bar{\Phi} - \alpha^2 \lambda I_{5n} & \mathbf{Q}^\top \\ \mathbf{Q} & \lambda I_{5T} \end{bmatrix} > 0, \quad (\text{V.13})$$

$$\begin{bmatrix} \bar{R}_i & \bar{S}_{12,i} \\ \star & \bar{R}_i \end{bmatrix} \geq 0, \quad (\text{V.14})$$

together with (IV.3) hold, where  $\bar{\Phi}$  is given in (IV.5) and

$$\mathbf{Q} = \begin{bmatrix} Q_3 & Q_2 & Q_1 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 \\ \varepsilon Q_3 & \varepsilon Q_2 & \varepsilon Q_1 & 0 & 0 \end{bmatrix}. \quad (\text{V.15})$$

Choose the feedback gain

$$K = U_{h_2, \{0\}} Q_1 (X_{\{0\}} Q_3)^{-1}. \quad (\text{V.16})$$

Then for all delays  $h_1(k) \in [0, \bar{h}]$  and  $h_2(k) \in [0, \bar{h}]$  for all  $k \in \mathbb{Z}_{\geq 0}$ , the origin of (II.1) in closed-loop is asymptotically stable.  $\nabla \nabla \nabla$

For  $\alpha = 0$ , i.e., in the noise free case, the inequality (V.13) reduces to the one in (IV.1). As in (V.13), the inequalities (IV.8) and (IV.14) can be extended to account for data corrupted by noise. Therefore, from Theorem IV.2 analogous corollaries to Corollary IV.4 and Corollary IV.6 can be derived in a straightforward manner. Hence, their explicit presentation is omitted.

## VI. NUMERICAL EXAMPLE: UNSTABLE BATCH REACTOR

To exemplify the proposed method, we consider the unstable linearized batch reactor in [31, pp. 63] controlled through a

TABLE I

COMPUTED VALUES OF THE DISTANCE  $d_{(i,j)}(X_{\{1\}})$  GIVEN IN (V.7) FOR THE DIFFERENT VALUES OF  $i$ , WITH  $j = 0$  AND  $T = 10$ , IN THE ABSENCE OF NOISE WITH THE DATA RECORDED FROM THE SYSTEM (VI.1).

$i$	0	1	2
$d_{(i,0)}(X_{\{1\}})$	$6.85 \times 10^{-4}$	$2.44 \times 10^{-3}$	$3.92 \times 10^{-4}$
$i$	3	4	5
$d_{(i,0)}(X_{\{1\}})$	$2.75 \times 10^{-10}$	$2.76 \times 10^{-3}$	$6.83 \times 10^{-3}$
$i$	6	7	8
$d_{(i,0)}(X_{\{1\}})$	$3.88 \times 10^{-3}$	$1.97 \times 10^{-2}$	$1.06 \times 10^{-2}$

network, and described by the dynamics

$$\begin{aligned} \dot{x}(t) = & \begin{bmatrix} 1.38 & -0.20 & 6.71 & -5.67 \\ -0.58 & -4.29 & 0 & 0.67 \\ 1.06 & 4.27 & -6.65 & 5.89 \\ 0.04 & 4.27 & 1.34 & -2.10 \end{bmatrix} x(t) \\ & + \begin{bmatrix} 0 & 0 \\ 5.67 & 0 \\ 1.13 & -3.14 \\ 1.13 & 0 \end{bmatrix} u(t - h_2). \end{aligned} \quad (\text{VI.1})$$

The input to the system is generated using a zero-order hold (ZOH) with a sampling time of 10 [ms]. This sampling time is taken as the base time. Additionally, a constant input delay  $h_2 = 3$  (30 [ms]) is introduced in the system (VI.1). Furthermore, we assume a maximum delay length of  $\bar{h} = 8$ . We consider that the plant has been in operation for a certain time using the PI-control given in [31]. In the context of the present paper, it is assumed that this controller implementation is based on expertise rather than on a model. Furthermore, the plant operation point is assumed known and corresponds to

$$\begin{aligned} x_{\text{op}} &= [24.35 \quad 14 \quad 48.78 \quad 63.13]^\top, \\ u_{\text{op}} &= [5.565 \quad 44.36]^\top. \end{aligned} \quad (\text{VI.2})$$

The overall objective is to stabilize the system around (VI.2).

### A. Scenario 1: Noise-Free Data and Unknown Delay

For generating the input-state data, and to characterize the system (VI.1), we feed the reference  $y_{\text{ref}} = [10 \quad 14]^\top$  in combination with the excitation signal  $u_{\text{exc}}(t)$  defined below to the PI-control already available in the plant [31].

$$u_{\text{exc}}(t) = \begin{bmatrix} 10 \sin(7\pi t) - 5 \sin(11\pi t) \\ 8 \sin(9\pi t) - 6 \sin(13\pi t) \end{bmatrix}.$$

The resulting excitation signal is shown in Figure 1. For the control design, we assume  $h_2$  constant, but unknown, and given the physical background of the system (VI.1), we have  $h_1 = 0$ .

As first step, we seek to investigate the value of  $h_2$  in the range  $\{0, 1, \dots, \bar{h}\}$ . Note that for  $T = m + n$ , which yields a square  $W_0$ , the distance  $d_{(i,j)}(X_{\{1\}})$  defined in (V.7) is always zero since  $I_{n+m} - W_0 W_0^{-1} = 0$ . Therefore, we choose  $T = 10 > n + m$ . We identify the length of  $h_2$  using the result of Proposition V.2. For the noise free case

( $r_{X_{\{1\}}^\delta} = r_{W_0^{\delta,11}} = r_{W_0^{\delta,12}} = 0$ ), the criterion given in (V.10) reads as  $d_{(i,j)}(X_{\{1\}}) \leq 0$ . The resulting values for the distance  $d_{(i,j)}(X_{\{1\}})$  for the different values of  $i$ , with  $j = 0$  (since  $h_1 = 0$ ) and  $T = 10$ , are shown in Table I. From Table I, the input delay can be clearly determined as  $h_2 = 3$  since the distance  $d_{(3,0)}(X_{\{1\}})$  is practically zero and its value is due to numerical errors.

Now that  $h_2$  has been determined, the matrices  $X_{\{0\}}$ ,  $X_{\{1\}}$  and  $U_{h_2,\{0\}}$  can be built. To illustrate the application of the data-driven controller synthesis from Section IV, we consider the stabilization of the system (VI.1) at the operational point (VI.2). We assume that the network-induced delay takes values in the set  $\{0, 1, \dots, 5\}$ , whereas the input delay remains constant at  $h_2 = 3$ . This satisfies the delay upper bound  $\bar{h} = 8$ , which is used in the formulas provided in Theorem IV.2. By using the data-based matrices  $X_{\{0\}}$ ,  $X_{\{1\}}$  and  $U_{h_2,\{0\}}$ , and following Theorem IV.2, we solve (IV.1), (IV.2) and (IV.3) with  $\bar{R}_1 = \bar{S}_{12} = 0$  and  $Q_2 = 0$  using CVX<sup>1</sup> [33]. For this, we used  $\varepsilon = 3$ . This yields the following feedback gain:

$$K = \begin{bmatrix} 0.813 & -0.282 & 0.115 & -1.121 \\ 2.255 & -0.549 & 1.894 & -1.226 \end{bmatrix}. \quad (\text{VI.3})$$

To compare our result with a model-based approach, we also computed a stabilizing gain following [21, Chap. 6] by discretizing the batch reactor model in (VI.1) with the given base time of 10 [ms]. By using the given delay upper bound  $\bar{h} = 8$  and with  $\varepsilon = 3$ , we obtained the controller gain

$$K_{\text{MB}} = \begin{bmatrix} 0.338 & -0.511 & -0.081 & -0.626 \\ 2.117 & 0.034 & 1.512 & -0.914 \end{bmatrix}. \quad (\text{VI.4})$$

We simulate the stabilization of the system (VI.1) around the operational point (VI.2) for the two gains  $K$  and  $K_{\text{MB}}$  in (VI.3) and (VI.4), respectively. We used a network induced delay that randomly changed in the proposed range, i.e., between zero and five. In Figure 2, the error norm between the system state and  $x_{\text{op}}$  is shown. We can observe that both controllers achieve the task in a similar time, under the same circumstances. Finally, in Figure 3, the response of the system (VI.1) to the control process using  $K$  in (VI.3) is illustrated for reference.

## B. Scenario 2: Noisy Data and Unknown Delay

In order to evaluate the robustness of the proposed approach under corrupted measurements, we add an uniform distributed random signal  $\delta(k)$  to each measurement  $x(k)$  and  $u(k)$  of the system (VI.1). The range of  $\delta(k)$  corresponds to  $[1 \times 10^{-4}, 1 \times 10^{-4}]$ . As before, for the control design we assume a constant and unknown input delay  $h_2$  as well as  $h_1 = 0$ , with the same delay upper bound  $\bar{h} = 8$ . For this section, and because we are dealing with data corrupted by

<sup>1</sup>CVX can parse LMIs with equality constraints and process them as a semidefinite program. Therefore, including (IV.3) is straightforward in this case. For noisy data, a numerically more robust approach consists in jointly minimizing the norms  $\|U_{h_2,\{0\}}Q_2\|_2$ ,  $\|U_{h_2,\{0\}}Q_3\|_2$ ,  $\|X_{h_1,\{0\}}Q_1\|_2$ ,  $\|X_{h_1,\{0\}}Q_3\|_2$ ,  $\|X_{\{0\}}Q_1\|_2$ ,  $\|X_{\{0\}}Q_2\|_2$ ,  $\|X_{h_1,\{0\}}Q_2 - X_{\{0\}}Q_3\|_2$  subject to the LMIs (IV.1) and (IV.2), which is a convex problem. If needed, the norm minimization can be transformed into a semidefinite program following [32].

noise, we set  $T = 50$ . Now, in order to determine the input delay, and following Proposition V.2, we need to estimate the upper bounds

$$r_{X_{\{1\}}^\delta} \geq \|X_{\{1\}}^\delta\|_2, \quad r_{W_0^{\delta,11}} \geq \|W_0^{\delta,11}\|_2, \quad r_{W_0^{\delta,12}} \geq \|W_0^{\delta,12}\|_2.$$

Since the noise follows a uniform distribution, it is bounded in magnitude. We can find the required upper bounds by using the Frobenius norm with the maximum value for each component:

$$\begin{aligned} \|X_{\{1\}}^\delta\|_2 &\leq 10^{-4} \cdot \sqrt{n \cdot T} = 1.41 \times 10^{-3} =: r_{X_{\{1\}}^\delta}, \\ \|W_0^{\delta,11}\|_2 &\leq 10^{-4}(n+m) = 6 \times 10^{-4} =: r_{W_0^{\delta,11}}, \\ \|W_0^{\delta,12}\|_2 &\leq 10^{-4} \sqrt{(n+m)(T-m-n)} \\ &= 1.62 \times 10^{-3} =: r_{W_0^{\delta,12}}. \end{aligned}$$

We proceed to compute the distance  $d_{(i,j)}(X_{\{1\}})$  given in (V.7) and the criterion given in (V.10), but with noisy data. The results are shown in Table II. In contrast to the noise-free scenario in Section VI-A, the distance value for  $i = 3$ , i.e., the correct delay length, is not as close to zero as before. Still, using the criterion derived in Proposition V.2, we can correctly identify the input delay as  $h_2 = 3$  since it is the only case in which the criterion (V.10) is satisfied.

To guarantee a robust closed-loop performance despite the presence of noise, we seek to employ Theorem V.3 for the controller synthesis. Thus in order to proceed, we need to estimate a bound for  $\|\Delta_{[\cdot]}\|_2$  in (V.12). From (V.12), we have

$$\|\Delta_{[\cdot]}\|_2 \leq \| [B \quad A_0] \|_2 \|W_0^\delta\|_2 + \|X_{\{1\}}^\delta\|_2.$$

Again, using a bound over the Frobenius norm, we obtain

$$\|W_0^\delta\|_2 \leq 10^{-4} \cdot \sqrt{(m+n) \cdot T} = 1.73 \times 10^{-3}. \quad (\text{VI.5})$$

To estimate the norm of the system matrices we use the relation

$$\begin{aligned} \| [B \quad A_0] \|_2 &= \| X_{\{1\}}^{\text{nom}} (W_0^{\text{nom}})^\dagger \|_2 \\ &\leq \left( \|X_{\{1\}}\|_2 + r_{X_{\{1\}}^\delta} \right) \frac{\sqrt{2} \|W_0^\dagger\|_2}{1 - r_{W_0^\delta} \|W_0^\dagger\|_2} \leq 102, \end{aligned} \quad (\text{VI.6})$$

where the last step follows from the upper bound for  $\|(W_0^{\text{nom}})^\dagger\|_2$  given in [30, Lem. 3.1]. Using (VI.5) and (VI.6) we obtain

$$\|\Delta_{[\cdot]}\|_2 \leq 0.191.$$

Thus we have  $\|\Delta_{[\cdot]}\|_2 \leq \alpha$  with  $\alpha = 0.191$ . Now, we can compute the feedback gain  $K$  using Theorem V.3 and CVX [33], which for  $\alpha = 0.251$ ,  $\varepsilon = 70$  and  $\lambda = 1.42 \times 10^3$  yields

$$K = \begin{bmatrix} 0.736 & -0.623 & 9.17 \times 10^{-2} & -1.07 \\ 2.35 & -7.99 \times 10^{-3} & 1.70 & -1.17 \end{bmatrix}. \quad (\text{VI.7})$$

To test this new feedback gain, we use the same setting as in Section VI-A, i.e., the stabilization around the operational point  $x_{\text{op}}$  in (VI.2) with uncertain network induced delay. The results of this simulation are presented in Figure 4. As can be seen, despite  $K$  being computed using data corrupted by noise, the stabilization is achieved. This demonstrates the robustness of the proposed approach with respect to noisy data.

TABLE II

COMPUTED VALUES OF THE DISTANCE  $d_{(i,j)}(X_{\{1\}})$  GIVEN IN (V.7) AND THE UPPER BOUND GIVEN IN (V.10) FOR THE DIFFERENT VALUES OF  $i$ , WITH  $j = 0$  AND  $T = 50$ , WITH THE DATA RECORDED FROM THE SYSTEM (VI.1) CORRUPTED BY NOISE.

$i$	0	1	2
$d_{(i,j)}(X_{\{1\}})$ (V.10)	1.81	1.35	$6.71 \times 10^{-1}$
	$1.40 \times 10^{-1}$	$1.35 \times 10^{-1}$	$1.30 \times 10^{-1}$
$i$	3	4	5
$d_{(i,j)}(X_{\{1\}})$ (V.10)	$6.88 \times 10^{-4}$	$5.68 \times 10^{-1}$	1.03
	$1.27 \times 10^{-1}$	$1.29 \times 10^{-1}$	$1.36 \times 10^{-1}$
$i$	6	7	8
$d_{(i,j)}(X_{\{1\}})$ (V.10)	1.40	1.69	2.14
	$1.56 \times 10^{-1}$	$1.81 \times 10^{-1}$	$1.50 \times 10^{-1}$

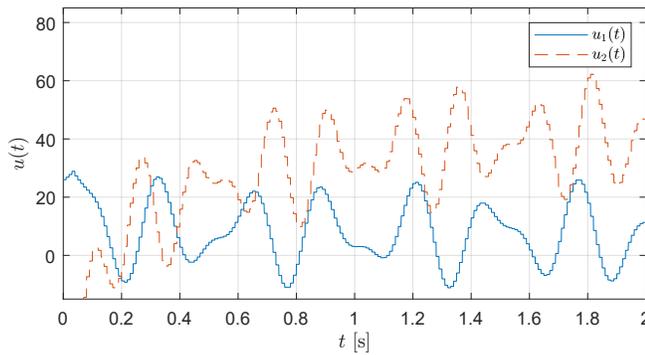


Fig. 1. Input signal  $u(t)$  used to excite the batch reactor in (VI.1).

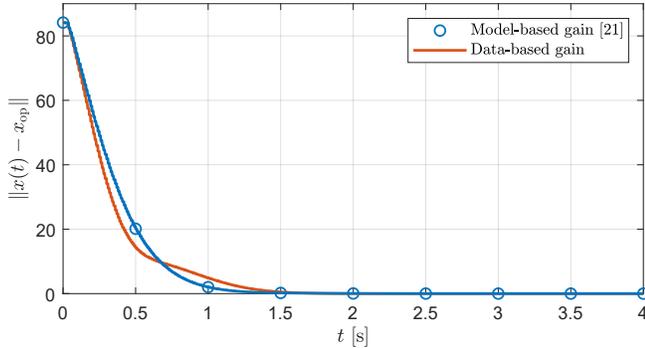


Fig. 2. Norm of the error between the state of the batch reactor (VI.1) and the operational point  $x_{op}$  in (VI.2) for the control gain  $K$  in (VI.3) computed using data and the gain  $K_{MB}$  in (VI.4) computed following [21, Chap. 6].

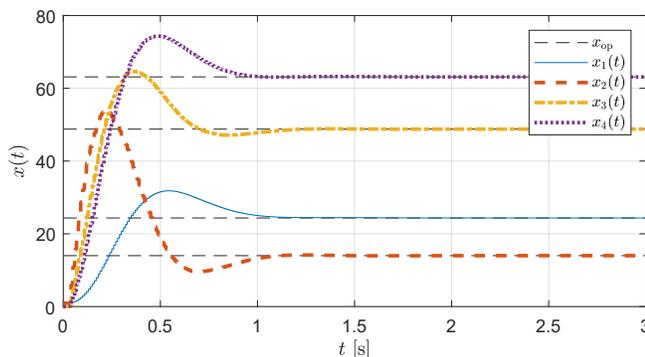


Fig. 3. Response of the batch reactor in (VI.1) to the stabilization process around the operational point  $x_{op}$  in (VI.2) using  $K$  in (VI.3).

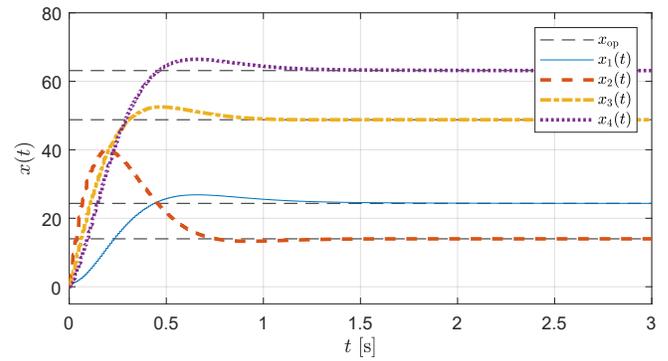


Fig. 4. Response of the batch reactor in (VI.1) to the stabilization process around the operational point  $x_{op}$  in (VI.2) using  $K$  in (VI.7) computed with noisy data.

## VII. CONCLUSIONS

In this work we have presented a method for designing robust controllers for LTD-TDSs relying exclusively on input-state data recorded from the system, i.e., avoiding the system modeling. We have provided explicit data-dependent formulas to compute state feedback gains for stabilization, guaranteed cost control and  $H_\infty$  control. By accounting on possible noise and unknown constant delays in the recorded data, the method ensures closed-loop stability of the system with the computed gain even under such circumstances.

Differently from other methods based on data [13], [16], we have investigated robustness against uncertain delays. The proposed design approach provides stability guarantees on the closed-loop system through a robust control design. Furthermore, the amount of data required for the control design is relatively small as is shown in the numerical example.

Future work will be geared towards the implementation and experimental validation of the reported results in real-world applications, such as traffic control or power systems operation. Likewise, we plan to investigate extensions to nonlinear systems, possibly by incorporating prior system knowledge as recently proposed in [17].

## APPENDIX

*Proof of Proposition III.1.* The matrices  $A_0$ ,  $A_1$  and  $B$  of the system (II.1) are related through data by

$$X_{\{1\}} = [B \ A_1 \ A_0] W_0. \quad (\text{A.6})$$

*Sufficiency:* Since by assumption  $W_0$  has full-row rank, we obtain from (A.6) that

$$[B \ A_1 \ A_0] = X_{\{1\}} W_0^\dagger. \quad (\text{A.7})$$

Using (A.7) in (III.1) yields (III.3).

*Necessity:* If  $\text{rank}(W_0) < m + 2n$ , then (A.6) together with the Rouché-Capelli theorem [34] implies that the matrices  $A_0$ ,  $A_1$  and  $B$  of the system (II.1) cannot be determined univocally.

Therefore, the condition  $\text{rank}(W_0) = m + 2n$  is necessary and sufficient to represent the open-loop system through data.

□□□

$$\Phi_1 = \begin{bmatrix} R_1 + R_2 & -R_1 + S_{12,1} & -R_2 + S_{12,2} & -S_{12,1} - S_{12,2} & 0 \\ \star & 2R_1 - S_{12,1} - S_{12,1}^\top & 0 & -R_1 + S_{12,1} & 0 \\ \star & \star & 2R_2 - S_{12,2} - S_{12,2}^\top & -R_2 + S_{12,2} & 0 \\ \star & \star & \star & R_1 + R_2 & 0 \\ \star & \star & \star & \star & 0 \end{bmatrix}. \quad (\text{A.1})$$

$$\Phi_2 = \begin{bmatrix} \Phi_{2,11} & P_2^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} & P_2^\top X_{\{1\}} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} & 0 & -P_2^\top + \left( P_3^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \right)^\top \\ \star & 0 & 0 & 0 & \left( P_3^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} \right)^\top \\ \star & \star & 0 & 0 & \left( P_3^\top X_{\{1\}} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} \right)^\top \\ \star & \star & \star & 0 & 0 \\ \star & \star & \star & \star & -(P_3 + P_3^\top) \end{bmatrix}, \quad (\text{A.2})$$

$$\Phi_{2,11} = P_2^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} + \left( P_2^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \right)^\top.$$

*Proof of Lemma III.2.* The closed-loop system (III.4) can be rewritten as

$$x(k+1) = [B \quad A_1 \quad A_0] \begin{bmatrix} K & 0 & 0 \\ 0 & I_n & 0 \\ 0 & 0 & I_n \end{bmatrix} \begin{bmatrix} x_{h_2}(k) \\ x_{h_1}(k) \\ x(k) \end{bmatrix}.$$

*Sufficiency:* Since by assumption  $\text{rank}(W_0) = m + 2n$ , one has that

$$\text{rank} \left( \left[ \begin{array}{c|c} \begin{bmatrix} K & 0 & 0 \\ 0 & I_n & 0 \\ 0 & 0 & I_n \end{bmatrix} & W_0 \end{array} \right] \right) = \text{rank}(W_0).$$

Thus, by the Rouché-Capelli theorem [34], there exists a  $T \times 3n$  matrix  $G_K$ , such that (III.6) holds. Therefore,

$$[B \quad A_1 \quad A_0] \begin{bmatrix} K & 0 & 0 \\ 0 & I_n & 0 \\ 0 & 0 & I_n \end{bmatrix} = [B \quad A_1 \quad A_0] W_0 G_K \\ = X_{\{1\}} G_K,$$

where the relation (A.6) has been used.

*Necessity:* If  $\text{rank}(W_0) < m + 2n$ , then by the Rouché-Capelli theorem [34], not for any matrix  $\text{diag}(K, I_n, I_n)$  there is a matrix  $G_K$  that satisfies (III.6).

This proves the main claim of Lemma III.2. The explicit formula for  $K$  in (III.7) is obtained from (III.6) by considering the definition of  $W_0$  in (III.2).  $\square\square\square$

*Proof of Theorem IV.2.* Inspired by the procedure of [21, Sec. 6.1.3.2], we propose the following Lyapunov-Krasovskii functional

$$V(k) = V_P(k) + V_S(k) + \sum_{i=1}^2 V_{R,i}(k), \quad (\text{A.8})$$

$$V_P(k) = x^\top(k) P x(k),$$

$$V_S(k) = \sum_{j=k-\bar{h}}^{k-1} x^\top(j) S x(j),$$

$$V_{R,i}(k) = \bar{h} \sum_{m=-\bar{h}}^{-1} \sum_{j=k+m}^{k-1} \bar{y}^\top(j) R_i \bar{y}(j),$$

$$\bar{y}(j) := x(j+1) - x(j),$$

where  $P > 0, S > 0, R_i > 0$ , with  $i = \{1, 2\}$ , are  $n \times n$  matrix variables. We are interested in computing  $V(k+1) - V(k)$ , which can be done by taking the difference of each of its components. By direct calculation, the following difference equations are obtained:

$$V_P(k+1) - V_P(k) = x^\top(k+1) P x(k+1) - x^\top(k) P x(k), \\ V_S(k+1) - V_S(k) = x^\top(k) S x(k) - x_{\bar{h}}^\top(k) S x_{\bar{h}}(k), \\ V_{R,i}(k+1) - V_{R,i}(k) = \bar{h}^2 \bar{y}^\top(k) R_i \bar{y}(k) \\ - \bar{h} \sum_{j=k-\bar{h}}^{k-1} \bar{y}^\top(j) R_i \bar{y}(j).$$

The summation term in  $V_{R,i}(k+1) - V_{R,i}(k)$  is rewritten as

$$\bar{h} \sum_{j=k-\bar{h}}^{k-1} \bar{y}^\top(j) R_i \bar{y}(j) = \\ \bar{h} \sum_{j=k-\bar{h}}^{k-1-h_i(k)} \bar{y}^\top(j) R_i \bar{y}(j) + \bar{h} \sum_{j=k-h_i(k)}^{k-1} \bar{y}^\top(j) R_i \bar{y}(j).$$

$$\Phi = \begin{bmatrix} \Phi_{11} & \Phi_{12} & \Phi_{13} & -S_{12,1} - S_{12,2} & \Phi_{15} \\ \star & 2R_1 - S_{12,1} - S_{12,1}^\top & 0 & -R_1 + S_{12,1} & - \left( P_3^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} \right)^\top \\ \star & \star & 2R_2 - S_{12,2} - S_{12,2}^\top & -R_2 + S_{12,2} & - \left( P_3^\top X_{\{1\}} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} \right)^\top \\ \star & \star & \star & R_1 + R_2 + S & 0 \\ \star & \star & \star & \star & \Phi_{55} \end{bmatrix}, \quad (\text{A.3})$$

$$\Phi_{11} = P - S + (1 - \bar{h}^2)(R_1 + R_2) - P_2^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} - \left( P_2^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \right)^\top,$$

$$\Phi_{12} = -R_1 + S_{12,1} - P_2^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix}, \quad \Phi_{13} = -R_2 + S_{12,2} - P_2^\top X_{\{1\}} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix},$$

$$\Phi_{15} = \bar{h}^2(R_1 + R_2) + P_2^\top - \left( P_3^\top X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \right)^\top, \quad \Phi_{55} = -P - \bar{h}^2(R_1 + R_2) + (P_3 + P_3^\top).$$

$$\bar{\Phi} = \begin{bmatrix} \bar{\Phi}_{11} & \bar{\Phi}_{12} & \bar{\Phi}_{13} & -\bar{S}_{12,1} - \bar{S}_{12,2} & \bar{\Phi}_{15} \\ \star & 2\bar{R}_1 - \bar{S}_{12,1} - \bar{S}_{12,1}^\top & 0 & -\bar{R}_1 + \bar{S}_{12,1} & - \left( \varepsilon X_{\{1\}} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} \bar{P}_2 \right)^\top \\ \star & \star & 2\bar{R}_2 - \bar{S}_{12,2} - \bar{S}_{12,2}^\top & -\bar{R}_2 + \bar{S}_{12,2} & - \left( \varepsilon X_{\{1\}} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} \bar{P}_2 \right)^\top \\ \star & \star & \star & \bar{R}_1 + \bar{R}_2 + \bar{S} & 0 \\ \star & \star & \star & \star & -\bar{P} - \bar{h}^2(\bar{R}_1 + \bar{R}_2) + \varepsilon(\bar{P}_2 + \bar{P}_2^\top) \end{bmatrix}, \quad (\text{A.4})$$

$$\bar{\Phi}_{11} = \bar{P} - \bar{S} + (1 - \bar{h}^2)(\bar{R}_1 + \bar{R}_2) - X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \bar{P}_2 - \left( X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \bar{P}_2 \right)^\top,$$

$$\bar{\Phi}_{12} = -\bar{R}_1 + \bar{S}_{12,1} - X_{\{1\}} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} \bar{P}_2, \quad \bar{\Phi}_{13} = -\bar{R}_2 + \bar{S}_{12,2} - X_{\{1\}} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} \bar{P}_2,$$

$$\bar{\Phi}_{15} = \bar{h}^2(\bar{R}_1 + \bar{R}_2) + \bar{P}_2 - \varepsilon \left( X_{\{1\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \bar{P}_2 \right)^\top.$$

$$\tilde{\Phi}_2 = \begin{bmatrix} P_2^\top \Delta_{[\cdot]} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} + \left( P_2^\top \Delta_{[\cdot]} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \right)^\top & P_2^\top \Delta_{[\cdot]} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} & P_2^\top \Delta_{[\cdot]} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} & 0 & \left( \varepsilon P_2^\top \Delta_{[\cdot]} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \right)^\top \\ \star & 0 & 0 & 0 & \left( \varepsilon P_2^\top \Delta_{[\cdot]} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} \right)^\top \\ \star & \star & 0 & 0 & \left( \varepsilon P_2^\top \Delta_{[\cdot]} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} \right)^\top \\ \star & \star & \star & 0 & 0 \\ \star & \star & \star & \star & 0 \end{bmatrix}. \quad (\text{A.5})$$

By applying Jensen's inequality twice [21, Sec. 6.1.3.2], one obtains

$$\begin{aligned} \bar{h} \sum_{j=k-\bar{h}}^{k-1-h_i(k)} \bar{y}^\top(j) R_i \bar{y}(j) &\geq \\ \frac{\bar{h}}{\bar{h}-h_i(k)} [x_{h_i(k)}(k) - x_{\bar{h}}(k)]^\top R_i [x_{h_i(k)}(k) - x_{\bar{h}}(k)], \\ \bar{h} \sum_{j=k-h_i(k)}^{k-1} \bar{y}^\top(j) R_i \bar{y}(j) &\geq \\ \frac{\bar{h}}{h_i(k)} [x(k) - x_{h_i(k)}(k)]^\top R_i [x(k) - x_{h_i(k)}(k)]. \end{aligned}$$

Furthermore, by invoking the reciprocally convex approach [21, Lem. 3.4] one has

$$\begin{aligned} \bar{h} \sum_{j=k-\bar{h}}^{k-1} \bar{y}^\top(j) R_i \bar{y}(j) &\geq \\ \frac{\bar{h}}{\bar{h}-h_i(k)} [x_{h_i(k)}(k) - x_{\bar{h}}(k)]^\top R_i [x_{h_i(k)}(k) - x_{\bar{h}}(k)] \\ + \frac{\bar{h}}{h_i(k)} [x(k) - x_{h_i(k)}(k)]^\top R_i [x(k) - x_{h_i(k)}(k)] &\geq \\ \begin{bmatrix} x(k) - x_{h_i(k)}(k) \\ x_{h_i(k)}(k) - x_{\bar{h}}(k) \end{bmatrix}^\top \begin{bmatrix} R_i & S_{12,i} \\ \star & R_i \end{bmatrix} \begin{bmatrix} x(k) - x_{h_i(k)}(k) \\ x_{h_i(k)}(k) - x_{\bar{h}}(k) \end{bmatrix}, \end{aligned} \quad (\text{A.9})$$

for any  $S_{12,i} \in \mathbb{R}^{n \times n}$  satisfying

$$\begin{bmatrix} R_i & S_{12,i} \\ \star & R_i \end{bmatrix} \geq 0.$$

Consider the short-hand

$$\chi^\top(k) = \left[ x^\top(k), x_{h_1(k)}^\top(k), x_{h_2(k)}^\top(k), x_{\bar{h}}^\top(k), x^\top(k+1) \right]. \quad (\text{A.10})$$

Then, by using (A.9) and (A.10) we obtain

$$\begin{aligned} \sum_{i=1}^2 V_{R_i}(k+1) - V_{R_i}(k) &\leq \bar{h}^2 \bar{y}^\top(k) (R_1 + R_2) \bar{y}(k) \\ &\quad - \chi^\top(k) \Phi_1 \chi(k), \end{aligned} \quad (\text{A.11})$$

where  $\Phi_1$  is given in (A.1).

Now, consider (III.5) and (III.6) in Lemma III.2. From them it follows that

$$\begin{aligned} BK &= X_{\{1\}} G_K \begin{bmatrix} I_n & 0 & 0 \end{bmatrix}^\top, \\ A_1 &= X_{\{1\}} G_K \begin{bmatrix} 0 & I_n & 0 \end{bmatrix}^\top, \\ A_0 &= X_{\{1\}} G_K \begin{bmatrix} 0 & 0 & I_n \end{bmatrix}^\top, \end{aligned} \quad (\text{A.12})$$

subject to

$$\begin{aligned} 0 &= U_{h_2, \{0\}} G_K \begin{bmatrix} 0 & I_n & 0 \end{bmatrix}^\top = U_{h_2, \{0\}} G_K \begin{bmatrix} 0 & 0 & I_n \end{bmatrix}^\top, \\ 0 &= X_{h_1, \{0\}} G_K \begin{bmatrix} I_n & 0 & 0 \end{bmatrix}^\top = X_{h_1, \{0\}} G_K \begin{bmatrix} 0 & 0 & I_n \end{bmatrix}^\top, \\ 0 &= X_{\{0\}} G_K \begin{bmatrix} I_n & 0 & 0 \end{bmatrix}^\top = X_{\{0\}} G_K \begin{bmatrix} 0 & I_n & 0 \end{bmatrix}^\top. \end{aligned} \quad (\text{A.13})$$

Now, combining the descriptor method [21, Sec. 3.5.2] with the data-based representation of the system matrices in (A.12), we obtain

$$\begin{aligned} 0 &= 2 \left[ x^\top(k) P_2^\top + x^\top(k+1) P_3^\top \right] \times \\ &\quad \times \left[ X_{\{1\}} G_K \left[ x_{h_2(k)}^\top(k), x_{h_1(k)}^\top(k), x^\top(k) \right]^\top - x(k+1) \right] \\ &= -2x^\top(k) P_2^\top x(k+1) - 2x^\top(k+1) P_3^\top x(k+1) \\ &\quad + 2 \left( x^\top(k) P_2^\top + x^\top(k+1) P_3^\top \right) X_{\{1\}} G_K \times \\ &\quad \times \left( \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} x_{h_2(k)}(k) + \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} x_{h_1(k)}(k) + \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} x(k) \right), \end{aligned} \quad (\text{A.14})$$

where  $P_2 \in \mathbb{R}^{n \times n}$  and  $P_3 \in \mathbb{R}^{n \times n}$  are matrix variables. By using the short-hand (A.10), we can rewrite (A.14) as the quadratic form

$$0 = \chi^\top(k) \Phi_2 \chi(k), \quad (\text{A.15})$$

with  $\Phi_2$  given in (A.2).

Now, retaking the calculation  $V(k+1) - V(k)$ , by considering (A.11), (A.15), and adding  $\Phi_1$  and  $\Phi_2$  we obtain

$$V(k+1) - V(k) \leq -\chi^\top(k) \Phi \chi(k), \quad (\text{A.16})$$

where  $\Phi$  is given in (A.3).

By recalling that both  $G_K$  and  $K$  are design parameters, an inspection of  $\Phi$  in (A.3) reveals that it contains nonlinear terms in the decision variables. In order to reformulate  $\Phi$  as an LMI, we follow the standard approach in control design of TDSs via the descriptor method, see [21, Chapter 6]. Since  $\Phi > 0$  in (A.3) implies that  $\Phi_{55} > 0$ , we have  $P_3 + P_3^\top > 0$ , meaning that  $P_3$  is invertible. We choose  $P_3 = \varepsilon P_2$ , where  $\varepsilon > 0$  is a scalar tuning parameter. Then, we define  $\bar{P}_2 = P_2^{-1}$  and the matrices

$$\begin{aligned} \bar{P} &= \bar{P}_2^\top P \bar{P}_2, & \bar{R}_i &= \bar{P}_2^\top R_i \bar{P}_2, \\ \bar{S} &= \bar{P}_2^\top S \bar{P}_2, & \bar{S}_{12,i} &= \bar{P}_2^\top S_{12,i} \bar{P}_2. \end{aligned} \quad (\text{A.17})$$

Consider the congruent transformation  $\bar{\Phi} = \mathbf{P}^\top \Phi \mathbf{P}$ , with

$$\mathbf{P} = \text{diag}(\bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2).$$

The resulting matrix  $\bar{\Phi}$  is given in (A.4), which is linear in all decision variables. Now, inspired by [13], we introduce the auxiliary matrix variables

$$Q_1 = G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} \bar{P}_2, \quad Q_2 = G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} \bar{P}_2, \quad Q_3 = G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \bar{P}_2. \quad (\text{A.18})$$

By considering (III.2) and the restriction (III.6), it holds that

$$\begin{aligned} K \bar{P}_2 &= U_{h_2, \{0\}} G_K \begin{bmatrix} I_n \\ 0 \\ 0 \end{bmatrix} \bar{P}_2 = U_{h_2, \{0\}} Q_1, \\ \bar{P}_2 &= X_{h_1, \{0\}} G_K \begin{bmatrix} 0 \\ I_n \\ 0 \end{bmatrix} \bar{P}_2 = X_{h_1, \{0\}} Q_2, \\ \bar{P}_2 &= X_{\{0\}} G_K \begin{bmatrix} 0 \\ 0 \\ I_n \end{bmatrix} \bar{P}_2 = X_{\{0\}} Q_3. \end{aligned} \quad (\text{A.19})$$

Replacing the previous definitions in  $\bar{\Phi}$  in (A.4) yields the matrix  $\bar{\Phi}$  in (IV.5) and the data-based set of inequalities (IV.1) and (IV.2). The equality restrictions (IV.3) follow from (A.13) and the definitions (A.18) and (A.19). The control gain  $K$  is obtained from (A.19) and is given in (IV.4). Then, if the above-mentioned inequalities hold, we have  $V(k+1) - V(k) < 0$ . By invoking [21, Thm. 6.1], the assertions of the theorem follow.  $\square\square\square$

*Proof of Corollary IV.4.* Consider the Lyapunov-Krasovskii functional given in (A.8) and the short-hand introduced in (A.10). From (A.16) together with the performance output  $z(k)$  given in (IV.7), we have

$$V(k+1) - V(k) + z^\top(k)z(k) \leq -\hat{\chi}^\top(k)\Psi\hat{\chi}(k),$$

where  $\hat{\chi}^\top(k) = [\chi^\top(k), z^\top(k)]$  and

$$\Psi = \begin{bmatrix} \bar{\Phi} & -\bar{\kappa}^\top \\ \star & I_n \end{bmatrix}, \quad (\text{A.20})$$

$$\bar{\kappa} = [L_1 \quad L_2 \quad DK \quad 0 \quad 0].$$

The matrix  $\bar{\Phi}$  is given in (A.3). Consider once more the matrices introduced in (A.17). As before, in order to obtain an LMI, we perform a congruent transformation. Define the block-diagonal matrix  $\mathbf{P}$

$$\mathbf{P} = \text{diag}(\bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2, I_n).$$

Using  $\mathbf{P}$ , we define the congruent matrix  $\bar{\Psi} = \mathbf{P}^\top \Psi \mathbf{P}$ , and, by considering the substitutions (A.18), we obtain the matrix

$$\bar{\Psi} = \begin{bmatrix} \bar{\Phi} & -\bar{\kappa}^\top \\ \star & I_n \end{bmatrix},$$

$$\bar{\kappa} = [L_1 \bar{P}_2 \quad L_2 \bar{P}_2 \quad DK \bar{P}_2 \quad 0 \quad 0],$$

with  $\bar{\Phi}$  given in (A.4). By replacing  $\bar{P}_2$  following (A.19), it results  $\bar{\Psi}$  in (IV.8). By invoking [21, Prop. 6.5], we have that  $J \leq V(x_0) = x^\top(0)Px(0)$  if the data-based inequalities (IV.8) and (IV.9) are satisfied. In addition, we have  $V(0) \leq \delta$ , and by transitivity  $J \leq \delta$ , if (IV.10) is simultaneously satisfied. This is achieved by selecting the feedback gain  $K$  according to (IV.11).  $\square\square\square$

*Proof of Corollary IV.6.* Consider the performance index

$$J_\infty = \sum_{k=0}^{\infty} (z^\top(k)z(k) - \gamma\omega^\top(k)\omega(k)).$$

If one can find a feedback gain  $K$  such that  $J_\infty < 0$ , then the system (IV.12) has an  $L_2$ -gain less than  $\gamma$  [21]. Such gain can be found as follows. Consider the short hand

$$\hat{\chi}_2^\top(k) = [\hat{\chi}^\top(k), \omega^\top(k)].$$

From (A.16) together with the system dynamics (IV.12) and (IV.13), we obtain

$$-\hat{\chi}_2^\top(k)\Gamma\hat{\chi}_2(k) \geq$$

$$V(k+1) - V(k) + (z^\top(k)z(k) - \gamma\omega^\top(k)\omega(k)), \quad (\text{A.21})$$

with

$$\Gamma = \begin{bmatrix} \Psi & \kappa_2^\top \\ \star & \gamma I_n \end{bmatrix}, \quad (\text{A.22})$$

$$\kappa_2 = [-D_0^\top P_2 \quad 0 \quad 0 \quad 0 \quad -\varepsilon D_0^\top P_2 \quad 0],$$

where  $\Psi$  is given in (A.20). As in the proof of Theorem IV.2 and Corollary IV.4, we consider the congruent matrix  $\bar{\Gamma} = \mathbf{P}^\top \Gamma \mathbf{P}$  with

$$\mathbf{P} = \text{diag}(\bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2, I_n, I_n),$$

where  $\bar{P}_2 = P_2^{-1}$ . This done to obtain an LMI from the original BMI. By taking into account (A.17), (A.18) and (A.19), we obtain the matrix  $\bar{\Gamma}$  in (IV.14). Hence, if (IV.14) and (IV.15) are satisfied, the gain given in (IV.16) guarantees a  $L_2$  gain of  $\gamma$  for  $\omega(k) \neq 0$ .  $\square\square\square$

*Proof of Proposition V.2.* Let  $h_1 = j^*$  and  $h_2 = i^*$  hold, and consider the short-hands  $W_{0\star} = W_{0,(i^*,j^*)}$ ,  $W_{0\star}^{\text{nom}} = W_{0,(i^*,j^*)}^{\text{nom}}$  and  $W_{0\star}^\delta = W_{0,(i^*,j^*)}^\delta$ . In the unperturbed case (i.e.,  $W_{0\star}^\delta = 0$  and  $X_{\{1\}}^\delta = 0$ ), the orthogonal distance  $d_{(i^*,j^*)}(X_{\{1\}})$  defined in (V.7) is zero, in other words (V.5) holds. In contrast, in the perturbed case ( $W_{0\star}^\delta \neq 0$ ,  $X_{\{1\}}^\delta \neq 0$ ) we have

$$d_{(i^*,j^*)}(X_{\{1\}}) = \left\| X_{\{1\}} \left( I_T - W_{0\star}^\dagger W_{0\star} \right) \right\|_2$$

$$= \left\| \left( X_{\{1\}}^{\text{nom}} + X_{\{1\}}^\delta \right) \left( I_T - W_{0\star}^\dagger W_{0\star} \right) \right\|_2. \quad (\text{A.23})$$

Therefore, it follows that

$$d_{(i^*,j^*)}(X_{\{1\}}) \leq d_{(i^*,j^*)}(X_{\{1\}}^{\text{nom}}) + d_{(i^*,j^*)}(X_{\{1\}}^\delta). \quad (\text{A.24})$$

From the properties of orthogonal projectors, we immediately have

$$d_{(i^*,j^*)}(X_{\{1\}}^\delta) \leq \|X_{\{1\}}^\delta\|_2 \leq r_{X_{\{1\}}^\delta}. \quad (\text{A.25})$$

Now, we proceed to upper bound  $d_{(i^*,j^*)}(X_{\{1\}}^{\text{nom}})$ , for which we recall (V.2) and consider

$$I_T - W_{0\star}^\dagger W_{0\star} = I_T - (W_{0\star}^{\text{nom}})^\dagger W_{0\star}^{\text{nom}}$$

$$+ (W_{0\star}^{\text{nom}})^\dagger W_{0\star}^{\text{nom}} - W_{0\star}^\dagger W_{0\star}. \quad (\text{A.26})$$

In account of (A.26) and the fact that

$$X_{\{1\}}^{\text{nom}} \left( I_T - (W_{0\star}^{\text{nom}})^\dagger W_{0\star}^{\text{nom}} \right) = 0, \quad (\text{A.27})$$

we have from (A.23) that

$$d_{(i^*,j^*)}(X_{\{1\}}^{\text{nom}}) = \left\| X_{\{1\}}^{\text{nom}} \left( (W_{0\star}^{\text{nom}})^\dagger W_{0\star}^{\text{nom}} - W_{0\star}^\dagger W_{0\star} \right) \right\|_2. \quad (\text{A.28})$$

Therefore, to obtain an upper bound for  $d_{(i^*,j^*)}(X_{\{1\}}^{\text{nom}})$  we need to bound the difference

$$\left\| (W_{0\star}^{\text{nom}})^\dagger W_{0\star}^{\text{nom}} - W_{0\star}^\dagger W_{0\star} \right\|_2.$$

Following [29, Thm. 4.1], and under Assumption V.1, it holds that

$$\left\| (W_{0\star}^{\text{nom}})^\dagger W_{0\star}^{\text{nom}} - W_{0\star}^\dagger W_{0\star} \right\|_2 \leq \psi \left( \frac{\|W_{0\star}^{\delta,12}\|_2 \|W_{0\star}^\dagger\|_2}{1 - \|W_{0\star}^{\delta,11}\|_2 \|W_{0\star}^\dagger\|_2} \right),$$

with  $\psi(\cdot)$  defined in (V.8), and  $\|W_{0\star}^{\delta,11}\|_2$  and  $\|W_{0\star}^{\delta,12}\|_2$  as in (V.3). Given that  $\psi$  is a monotonically increasing function of its argument and that with Assumption V.1 the relation (V.9) holds, we have that

$$1 - \|W_{0\star}^{\delta,11}\|_2 \|W_{0\star}^\dagger\|_2 \geq 1 - r_{W_{0\star}^{\delta,11}} \|W_{0\star}^\dagger\|_2 > 0,$$

it follows that

$$\begin{aligned} \left\| (W_{0\star}^{\text{nom}})^\dagger W_{0\star}^{\text{nom}} - W_{0\star}^\dagger W_{0\star} \right\|_2 &\leq \psi \left( \frac{r_{W_{0\star}^{\delta,12}} \|W_{0\star}^\dagger\|_2}{1 - r_{W_{0\star}^{\delta,11}} \|W_{0\star}^\dagger\|_2} \right) \\ &= \psi(\sigma^*), \end{aligned} \quad (\text{A.29})$$

with  $r_{W_{0\star}^{\delta,11}}$  and  $r_{W_{0\star}^{\delta,12}}$  as in (V.9) and  $\sigma^*$  given in (V.11). Hence, from (A.28) and (A.29), we obtain

$$\begin{aligned} d_{i^*,j^*}(X_{\{1\}}^{\text{nom}}) &\leq \|X_{\{1\}}^{\text{nom}}\|_2 \cdot \psi(\sigma^*) \\ &\leq (\|X_{\{1\}}\|_2 + r_{X_{\{1\}}^\delta}) \cdot \psi(\sigma^*). \end{aligned} \quad (\text{A.30})$$

In account of (A.24) and the bounds (A.25) and (A.30), the upper bound for  $d_{(i^*,j^*)}(X_{\{1\}})$  given in (V.10) follows.

For  $i \neq i^*$  and  $j \neq j^*$ , (A.27) does not hold, and thus, the upper bound for  $d_{(i,j)}(X_{\{1\}})$  given in (V.10) increases.

Once the correct delays are determined, the corresponding open- and closed-loop data-based system representations are obtained via Proposition III.1 and Lemma III.2, respectively.

Finally, if for two distinct pairs  $(i, j)$  the condition (V.10) holds simultaneously, there are two candidates for the delay values and it is not possible to distinguish between them with the derived bound (V.10) and with the available data.  $\square\square\square$

*Proof of Theorem V.3.* As in Theorem V.3, let there exist the matrices  $Q_i$  for  $i = \{1, 2, 3\}$ ,  $\bar{P}$ ,  $S$ ,  $\bar{R}_i$  and  $S_{12,i}$  for  $i = \{1, 2\}$ , and compute  $K$  following (V.16). Following (A.19), define  $\bar{P}_2 = X_{\{0\}} Q_3$ . By using (A.17), the matrices  $P$ ,  $S$ ,  $R_i$  and  $S_{12,i}$  for  $i = \{1, 2\}$  can be computed. With these matrices, the Lyapunov-Krasovskii functional (A.8) can be built and used to analyze the stability of (II.1) with feedback gain  $K$ .

Consider the proof of Theorem IV.2. The effect of the noise impacts the terms introduced by the descriptor method, i.e. (A.14), since

$$\begin{aligned} X_{\{1\}} G_K \begin{bmatrix} x_{h_2(k)}(k) \\ x_{h_1(k)}(k) \\ x(k) \end{bmatrix} - x(k+1) &= \\ (X_{\{1\}} G_K - [BK \ A_1 \ A_0]) \begin{bmatrix} x_{h_2(k)}(k) \\ x_{h_1(k)}(k) \\ x(k) \end{bmatrix} &\neq 0. \end{aligned} \quad (\text{A.31})$$

To account for this mismatch, we compute the error induced by the corrupted data. Consider the nominal part of the data  $X_{\{1\}}^{\text{nom}}$

and  $W_0^{\text{nom}}$ . From Assumption V.1.1 we have  $\text{rank}(W_0^{\text{nom}}) = m + 2n$ . It follows that

$$[B \ A_1 \ A_0] = X_{\{1\}}^{\text{nom}} (W_0^{\text{nom}})^\dagger.$$

From (III.6) in combination with the expression above, we get

$$[BK \ A_1 \ A_0] = X_{\{1\}}^{\text{nom}} (W_0^{\text{nom}})^\dagger W_0 G_K.$$

By using this relation, we obtain

$$\begin{aligned} X_{\{1\}} G_K - [BK \ A_1 \ A_0] &= \\ (X_{\{1\}} - X_{\{1\}}^{\text{nom}} (W_0^{\text{nom}})^\dagger W_0) G_K. \end{aligned} \quad (\text{A.32})$$

Note that  $X_{\{1\}} = X_{\{1\}}^{\text{nom}} + X_{\{1\}}^\delta$  and  $W_0 = W_0^{\text{nom}} + W_0^\delta$ . We further continue from (A.32) as

$$\begin{aligned} X_{\{1\}} - X_{\{1\}}^{\text{nom}} (W_0^{\text{nom}})^\dagger W_0 &= \\ X_{\{1\}}^{\text{nom}} (I_T - (W_0^{\text{nom}})^\dagger W_0) + X_{\{1\}}^\delta &= \\ X_{\{1\}}^{\text{nom}} (I_T - (W_0^{\text{nom}})^\dagger W_0^{\text{nom}} - (W_0^{\text{nom}})^\dagger W_0^\delta) + X_{\{1\}}^\delta &= \\ = -X_{\{1\}}^{\text{nom}} (W_0^{\text{nom}})^\dagger W_0^\delta + X_{\{1\}}^\delta &= \\ = -[B \ A_1 \ A_0] W_0^\delta + X_{\{1\}}^\delta = -\Delta_{[\cdot]} \end{aligned}$$

Then, in order to account for the mismatch (A.31) and to keep the equation (A.14) equal to zero, we add the compensating term

$$2 [x^\top(k) + \varepsilon x^\top(k+1)] P_2^\top \times \Delta_{[\cdot]} G_K \begin{bmatrix} x_{h_2(k)}(k) \\ x_{h_1(k)}(k) \\ x(k) \end{bmatrix}$$

to (A.14). This term can be written as the quadratic form  $\chi^\top(k) \tilde{\Phi}_2 \chi(k)$  with  $\tilde{\Phi}_2$  given in (A.5). Carrying this term along and continuing as in the proof of Theorem IV.2, we obtain

$$V(k+1) - V(k) \leq -\chi^\top(k) (\bar{\Phi} - \tilde{\Phi}_2) \chi(k). \quad (\text{A.33})$$

As before, to obtain an LMI from (A.33), we make use of a congruent transformation. Let  $\mathbf{P} = \text{diag}(\bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2, \bar{P}_2)$ , and consider the congruent transformation

$$\mathbf{P}^\top (\bar{\Phi} - \tilde{\Phi}_2) \mathbf{P} = \bar{\Phi} - \tilde{\Phi}_2,$$

where  $\bar{\Phi}$  is given in (IV.5) and

$$\tilde{\Phi}_2 = \Delta \mathbf{Q} + \mathbf{Q}^\top \Delta^\top, \quad (\text{A.34})$$

with  $\Delta = \text{diag}(\Delta_{[\cdot]}, \Delta_{[\cdot]}, \Delta_{[\cdot]}, \Delta_{[\cdot]}, \Delta_{[\cdot]})$  and  $\mathbf{Q}$  is given in (V.15). Using (A.34), it follows that

$$\tilde{\Phi}_2 \leq \lambda \Delta \Delta^\top + \frac{1}{\lambda} \mathbf{Q}^\top \mathbf{Q} \leq \alpha^2 \lambda I_{5n} + \frac{1}{\lambda} \mathbf{Q}^\top \mathbf{Q},$$

with  $\lambda > 0$  and  $\alpha > 0$  as in Theorem (V.3). Therefore, the negativeness of (A.33) can be ensured if

$$\bar{\Phi} - \alpha^2 \lambda I_{5n} - \frac{1}{\lambda} \mathbf{Q}^\top \mathbf{Q} > 0.$$

By using the Schur complement, the inequality above is transformed into (V.13). Hence, if (V.13) and (V.14) are satisfied, it is ensured that  $V(k+1) - V(k) < 0$  and the origin of (II.1) with feedback  $u(k) = Kx(k)$ , where  $K$  is given in (V.16) is exponentially stable.  $\square\square\square$

## REFERENCES

- [1] A. Bazanella, L. Campestri, and E. D., *Data-Driven Controller Design - The  $H_2$  Approach*. Dordrecht, The Netherlands: Springer Netherlands, 2012. [Online]. Available: doi.org/10.1007/978-94-007-2300-9
- [2] Z.-S. Hou and Z. Wang, "From model-based control to data-driven control: Survey, classification and perspective," *Information Sciences*, vol. 235, pp. 3 – 35, 2013. [Online]. Available: doi.org/10.1016/j.ins.2012.07.014
- [3] I. Markovsky and P. Rapisarda, "On the linear quadratic data-driven control," in *2007 European Control Conference (ECC)*, 2007, pp. 5313–5318.
- [4] I. Markovsky and P. Rapisarda, "Data-driven simulation and control," *International Journal of Control*, vol. 81, no. 12, pp. 1946–1959, 2008.
- [5] U. S. Park and M. Ikeda, "Stability analysis and control design of LTI discrete-time systems by the direct use of time series data," *Automatica*, vol. 45, no. 5, pp. 1265 – 1271, 2009. [Online]. Available: doi.org/10.1016/j.automatica.2008.12.012
- [6] J. Coulson, J. Lygeros, and F. Dörfler, "Data-enabled predictive control: In the shallows of the deepc," in *2019 18th European Control Conference (ECC)*. IEEE, 2019, pp. 307–312.
- [7] —, "Regularized and distributionally robust data-enabled predictive control," in *2019 IEEE 58th Conference on Decision and Control (CDC)*. IEEE, 2019, pp. 2696–2701.
- [8] J. Berberich, J. Köhler, M. A. Müller, and F. Allgöwer, "Data-driven model predictive control with stability and robustness guarantees," *IEEE Transactions on Automatic Control*, 2020.
- [9] B. Pang, T. Bian, and Z. Jiang, "Data-driven finite-horizon optimal control for linear time-varying discrete-time systems," in *2018 IEEE Conference on Decision and Control (CDC)*, 2018, pp. 861–866.
- [10] S. Tu and B. Recht, "The gap between model-based and model-free methods on the linear quadratic regulator: An asymptotic viewpoint," *arXiv preprint arXiv:1812.03565*, 2018. [Online]. Available: arxiv.org/abs/1812.03565
- [11] M. Rotulo, C. De Persis, and P. Tesi, "Data-driven linear quadratic regulation via semidefinite programming," *arXiv preprint arXiv:1911.07767*, 2019. [Online]. Available: arxiv.org/abs/1911.07767
- [12] S. Dean, S. Tu, N. Matni, and B. Recht, "Safely learning to control the constrained linear quadratic regulator," in *2019 American Control Conference (ACC)*. IEEE, 2019, pp. 5582–5588.
- [13] C. De Persis and P. Tesi, "Formulas for data-driven control: Stabilization, optimality and robustness," *IEEE Transactions on Automatic Control*, 2019. [Online]. Available: doi.org/10.1109/TAC.2019.2959924
- [14] C. De Persis and P. Tesi, "Low-complexity learning of linear quadratic regulators from noisy data," *arXiv preprint arXiv:2005.01082*, 2020.
- [15] H. J. van Waarde and M. Mesbahi, "Data-driven parameterizations of suboptimal LQR and  $\mathcal{H}_2$  controllers," *arXiv preprint arXiv:1912.07671*, 2019. [Online]. Available: arxiv.org/abs/1912.07671
- [16] H. J. Van Waarde, J. Eising, H. L. Trentelman, and M. K. Camlibel, "Data informativity: a new perspective on data-driven analysis and control," *IEEE Transactions on Automatic Control*, pp. 1–1, 2020.
- [17] J. Berberich, C. W. Scherer, and F. Allgöwer, "Combining prior knowledge and data for robust controller design," *arXiv preprint arXiv:2009.05253*, 2020.
- [18] J. Berberich and F. Allgöwer, "A trajectory-based framework for data-driven system analysis and control," *arXiv preprint arXiv:1903.10723*, 2019. [Online]. Available: arxiv.org/abs/1903.10723
- [19] J. G. Rueda-Escobedo and J. Schiffer, "Data-driven internal model control of second-order discrete Volterra systems," *arXiv preprint arXiv:2003.14158*, 2020.
- [20] J.-P. Richard, "Time-delay systems: an overview of some recent advances and open problems," *Automatica*, vol. 39, no. 10, pp. 1667–1694, 2003.
- [21] E. Fridman, *Introduction to time-delay systems: Analysis and control*. Springer, 2014.
- [22] K. Liu, A. Selivanov, and E. Fridman, "Survey on time-delay approach to networked control," *Annual Reviews in Control*, vol. 48, pp. 57 – 79, 2019. [Online]. Available: doi.org/10.1016/j.arcontrol.2019.06.005
- [23] S. Formentin, M. Corno, S. M. Savaresi, and L. Del Re, "Direct data-driven control of linear time-delay systems," *Asian Journal of Control*, vol. 14, no. 3, pp. 652–663, 2012. [Online]. Available: doi.org/10.1002/asjc.387
- [24] O. Kaneko, S. Yamamoto, and Y. Wadagaki, "Simultaneous attainment of model and controller for linear time delay systems with the data driven smith compensator\*," *IFAC Proceedings Volumes*, vol. 44, no. 1, pp. 7684 – 7689, 2011. [Online]. Available: doi.org/10.3182/20110828-6-IT-1002.02792
- [25] H. Jiang and B. Zhou, "Data-driven based quadratic guaranteed cost control of unknown linear time-delay systems," in *2019 Chinese Control Conference (CCC)*, 2019, pp. 256–259.
- [26] Y. Liu, H. Zhang, R. Yu, and Q. Qu, "Data-driven optimal tracking control for discrete-time systems with delays using adaptive dynamic programming," *Journal of the Franklin Institute*, vol. 355, no. 13, pp. 5649 – 5666, 2018. [Online]. Available: https://doi.org/10.1016/j.jfranklin.2018.06.013
- [27] Y. Liu, H. Zhang, R. Yu, and Z. Xing, " $H_\infty$  tracking control of discrete-time system with delays via data-based adaptive dynamic programming," *IEEE Transactions on Systems, Man, and Cybernetics: Systems*, pp. 1–8, 2019.
- [28] J. Willems, P. Rapisarda, I. Markovsky, and B. D. Moor, "A note on persistency of excitation," *Systems & Control Letters*, vol. 54, no. 4, pp. 325 – 329, 2005. [Online]. Available: doi.org/10.1016/j.sysconle.2004.09.003
- [29] G. W. Stewart, "On the perturbation of pseudo-inverses, projections and linear least squares problems," *SIAM Review*, vol. 19, no. 4, pp. 634–662, 1977.
- [30] P.-A. Wedin, "Perturbation theory for pseudo-inverses," *BIT Numerical Mathematics*, vol. 13, pp. 217–232, 1973. [Online]. Available: doi.org/10.1007/BF01933494
- [31] G. C. Walsh and Hong Ye, "Scheduling of networked control systems," *IEEE Control Systems Magazine*, vol. 21, no. 1, pp. 57–65, 2001. [Online]. Available: doi.org/10.1109/37.898792
- [32] L. Vandenberghe and S. Boyd, "Semidefinite programming," *SIAM Review*, vol. 38, no. 1, pp. 49–95, 1996. [Online]. Available: doi.org/10.1137/1038003
- [33] M. Grant and S. Boyd, "CVX: Matlab software for disciplined convex programming, version 2.1," mar 2014. [Online]. Available: cvxr.com/cvx
- [34] I. R. Shafarevich and A. O. Remizov, *Linear algebra and geometry*. Springer Science & Business Media, 2012.



**Juan G. Rueda-Escobedo** was born in Mexico. He received the Licenciante-Degree in mechanical engineering, the Master-Degree and the Doctoral-Degree (with honors) in electrical engineering (Automatic Control) from the National Autonomous University of Mexico (UNAM), Cd. Mx., Mexico, in 2013, 2014 and 2018, respectively.

He is a Research Associate with the Chair of Control Systems and Network Control Technology, Brandenburg University of Technology, Cottbus, Germany. His main research interests are observer design for nonlinear and time-varying systems, new parameter identification strategies, adaptive observation and control with enhanced convergence speed.



**Emilia Fridman** (Fellow, IEEE) received the M.Sc. degree in mathematics from Kuibyshev State University, Samara, Russia, in 1981, and the Ph.D. degree in mathematics from Voronezh State University, Voronezh, Russia, in 1986.

From 1986 to 1992, she was an Assistant Professor and an Associate Professor with the Department of Mathematics, Kuibyshev Institute of Railway Engineers, Russia. Since 1993, she has been with Tel Aviv University, Tel Aviv, Israel, where she is currently a Professor of electrical engineering systems. She has held visiting positions at the Weierstrass Institute for Applied Analysis and Stochastics, Berlin, Germany; INRIA, Rocquencourt, France; Ecole Centrale de Lille, Villeneuve-d'Ascq, France; Valenciennes University, Valenciennes, France; Leicester University, Leicester, U.K.; Kent University, Canterbury, U.K.; CINVESTAV, Mexico; Zhejiang University, Hangzhou, China; St. Petersburg IPM, Russia; Melbourne University, Melbourne, VIC, Australia; Supélec, France; and the KTH Royal Institute of Technology, Stockholm, Sweden. She is the author of two monographs including *Introduction to Time-Delay Systems: Analysis and Control* (Birkhauser, 2014). She has published more than 190 articles in international scientific journals. Her research interests include time-delay systems, networked control systems, distributed parameter systems, robust control, singular perturbations, and nonlinear control.

Dr. Fridman is also a member of the Council of the IFAC and IEEE fellow. She was nominated as a Highly Cited Researcher by Thomson ISI in 2014. She serves/served as an Associate Editor for *Automatica*, *SIAM Journal on Control and Optimization*, and *IMA Journal of Mathematical Control and Information*.



**Johannes Schiffer** received the Diploma degree in engineering cybernetics from the University of Stuttgart, Stuttgart, Germany, in 2009, and the Ph.D. degree (Dr.Ing.) in electrical engineering from the Technische Universität (TU) Berlin, Berlin, Germany, in 2015.

He has held appointments as a Lecturer (Assistant Professor) at the School of Electronic and Electrical Engineering, University of Leeds, Leeds, U.K., and as a Research Associate at the Control Systems Group and the Chair of Sustainable Electric Networks and Sources of Energy, TU Berlin. He currently holds the Chair of Control Systems and Network Control Technology, Brandenburgische Technische Universität Cottbus–Senftenberg, Cottbus, Germany. His current research interests include distributed control and analysis of complex networks with application to microgrids and power systems.

Dr. Schiffer and his coworkers received the *Automatica* Paper Prize over the years 2014–2016.