V. Conclusion

The results in this note show that the concepts introduced in [6] for solving the minimal communication problem to achieve state disambiguation can be adapted to solve the more general problem of "essential transitions." This adaptation required the introduction of the property of legality, which captures the requirements pertaining to essential transitions. Several issues remain open for future research. The determination of well-posed sets of essential transitions in specific decentralized supervisory control or fault diagnosis problems is of particular interest. This was done in an intuitive manner in the decentralized control example in the note. Systematic procedures for generating these sets when coobservability is violated are currently being investigated. Also, the problem of synthesizing minimal communication maps in multiagent problems (three or more agents) is entirely open and most likely quite challenging.

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PID Stabilization of MIMO Plants

A. N. Gündüz and A. B. Özgüler

Abstract—Closed-loop stabilization using proportional–integral–derivative (PID) controllers is investigated for linear multiple-input–multiple-output (MIMO) plants. General necessary conditions for existence of PID-controllers are derived. Several plant classes that admit PID-controllers are explicitly described. Plants with only one or two unstable zeros at or close to the origin (alternatively, at or close to infinity) as well as plants with only one or two unstable poles which are at or close to origin are among these classes. Systematic PID-Controller synthesis procedures are developed for these classes of plants.

Index Terms—Integral action, proportional–integral–derivative (PID) controllers.

I. INTRODUCTION

Proportional–integral–derivative (PID) controllers are widely used in many control applications and preferred for their simplicity. Due to their integral action, PID-controllers achieve asymptotic tracking of step-input references. The topic of PID-control is treated extensively in every classical control text, e.g., [11]. In spite of the importance and widespread use of these low-order controllers, most PID design approaches lack systematic procedures and rigorous closed-loop stability proofs. Rigorous synthesis methods are explored recently in, e.g., [8]–[10] and [13].

The simplicity of PID-controllers, which is desirable due to easy implementation and from a tuning point-of-view, also presents a major restriction: PID-controllers can control only certain classes of plants. The problem of existence of stabilizing PID-controllers, which is practically very relevant (see [3]), is unfortunately not easy to solve. To gain insight into the problem’s difficulty, note that the existence of a stabilizing PID-controller for a plant $G(s)$ is equivalent to that of a constant stabilizing output feedback for a transformed plant. Alternatively, the problem can be posed as determining existence conditions of a stable and fixed-order controller for $G(s)/(s+1)/s$, which is also a difficult problem [1], [14]. The restriction on the controller order is a further major difficulty. Strong stabilizability of the plant is a necessary condition for existence of PID-controllers, but it is not sufficient; e.g., $G(s) = 1/(s-p)$ cannot be stabilized using a PID-controller for any $p > 0$, although the extended plant $G(s)/(s+1)/s$ is stabilizable using a stable controller (whose inverse is also stable).

The goal of this note is to find sufficient conditions on PID stabilizability, and hence, to identify plant classes that admit PID-controllers. Furthermore, explicit construction of the PID parameters for such plant classes is explored, leading to systematic controller synthesis procedures for linear, time-invariant (LTI), multiple-input–multiple-output (MIMO) plants of arbitrarily high order using the standard unity-feedback system shown in Fig. 1. The results obtained here explore conditions for PID stabilizability of general MIMO unstable plants without

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A. N. Gündüz is with the Electrical and Computer Engineering Department, University of California, Davis, CA 95616 USA (e-mail: angunduz@ucdavis.edu).
A. B. Özgüler is with the Electrical and Electronics Engineering Department, Bilkent University, Ankara 06800, Turkey (e-mail: oguzler@ee.bilkent.edu.tr).
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Fig. 1. Unity-feedback system \(\text{Sys}(G, C)\).

any restrictions on the plant order. Even in the single-input–single-output (SISO) case, explicit descriptions of high-order plant classes that admit PID-controllers are not available. Computational PID-stabilization methods of “efficient search” in the parameter space were recently developed for SISO delay-free systems (e.g., [12]), and some of these techniques were extended to first-order, scalar, single-delay systems [13]. Although some of the conditions on existence of PID-controllers could be derived for SISO plants using root-locus arguments or via a generalization of the Hermite–Riethmüller theorem [2], [6], [7], [15], they would not extend to the MIMO case and would not lead to explicit synthesis procedures. The results here emphasize systematic designs with freedom in the design parameters.

Section II gives preliminary definitions and the basic necessary conditions for stabilizability using PID-controllers. A two-step construction of stabilizing controllers is used as the basis of our synthesis method, which first constructs a proportional–derivative controller and then adds an integral term. Section III has the main results, where a novel use of the small gain theorem leads to identifying plant classes that are stabilizable using PID-controllers. The plants in Section III-A have restrictions on their blocking zeros in the region of instability, which leaves the pole locations completely free. The plants in Section III-B have restrictions on the unstable poles, which allows complete freedom in the zero locations.

The following notation is used: Let \(\mathbb{C}\), \(\mathbb{R}\), and \(\mathbb{R}_+\) denote complex, real, and positive real numbers, respectively. The extended closed right-half complex plane is \(\mathbb{U} = \{s \in \mathbb{C} | \Re(s) \geq 0 \} \cup \{\infty\}\); \(\mathbb{R}_p\) denotes real proper rational functions of \(s\); \(\mathbb{S} \subseteq \mathbb{R}_p\) is the stable subset with no poles in \(\mathbb{U}\); \(\mathcal{M}(\mathbb{S})\) is the set of matrices with entries in \(\mathbb{S}\); \(I_n\) is the \(n \times n\) identity matrix. The \(\mathbb{R}_\infty\)-norm of \(M(s) \in \mathbb{S}\) is \(\|M\| := \sup_{s \in \mathbb{U}} |M(s)|\), where \(\|\cdot\|\) is the maximum singular value and \(\delta M\) is the \(\mathbb{S}\)-boundary of \(M\). We drop \(s\) in transfer matrices such as \(G(s)\) wherever this causes no confusion. We use left coprime factorization over \(\mathbb{S}\), i.e., for \(G \in \mathbb{R}_p^{n \times n}\), \(G = Y^{-1}X\) denotes a left coprime-factorization (LCF), where \(X, Y \in \mathbb{S}\), det \(Y(\infty) \neq 0\).

II. PID STABILIZATION CONDITIONS

Consider the LTI MIMO unity-feedback system \(\text{Sys}(G, C)\) shown in Fig. 1, where \(G \in \mathbb{R}_p^{n \times n}\) is the plant’s transfer function and \(C \in \mathbb{R}_p^{n \times n}\) is the controller’s transfer function. Assume that \(\text{Sys}(G, C)\) is well posed, \(G\) and \(C\) have no unstable hidden modes, and \(G \in \mathbb{R}_p^{n \times n}\) is full (normal) row rank. We consider a realizable form of proper PID-controllers given in (1), where \(K_p, K_i, K_d \in \mathbb{R}_p^{n \times n}\) are the proportional, integral, derivative constants, respectively, and \(\tau \in \mathbb{R}_+\) [3]

\[
C_{\text{pid}} = K_p + K_i \frac{s}{s + K_d \tau} + \frac{K_d \tau}{s + 1}.
\]

For implementation, a (typically fast) pole is added to the derivative term so that \(C_{\text{pid}}\) in (1) is proper. The integral action in \(C_{\text{pid}}\) is present when \(K_i \neq 0\). The subsets of PID-controllers obtained by setting one or two of the three constants equal to zero are denoted as follows: (1) becomes a PI-controller \(C_{\text{pi}}\) when \(K_d = 0\), an ID-controller \(C_{\text{id}}\) when \(K_p = 0\), a PD-controller \(C_{\text{pd}}\) when \(K_i = 0\), a P-controller \(C_p\) when \(K_p = K_d = 0\), and a D-controller \(C_d\) when \(K_p = K_d = 0\).

Definition 1:
1) \(\text{Sys}(G, C)\) is said to be stable iff the transfer function from \((r, v)\) to \((y, w)\) is stable.
2) \(C\) is said to stabilize \(G\) iff \(C\) is proper and \(\text{Sys}(G, C)\) is stable.
3) \(G \in \mathbb{R}_p^{n \times n}\) is said to admit a PID-controller iff there exists \(C = C_{\text{pid}}\) as in (1) such that \(\text{Sys}(G, C_{\text{pid}})\) is stable.

We say that \(G\) is stabilizable by a PID-controller and \(C_{\text{pid}}\) is a stabilizing PID-controller.

Let \(G = Y^{-1} X\) be any LCF and \(C = N_c D_c^{-1}\) be any RCF; for \(G \in \mathbb{R}_p^{n \times n}\), \(X, Y \in \mathcal{M}(\mathbb{S})\), det \(Y(\infty) \neq 0\), and for \(C \in \mathbb{R}_p^{n \times n}\), \(N_c, D_c \in \mathcal{M}(\mathbb{S})\), det \(D_c(\infty) \neq 0\). Then, \(C\) stabilizes \(G\) if and only if \(M := YD_c + XN_c \in \mathcal{M}(\mathbb{S})\) is unimodular [4], [14]. We now examine necessary conditions for PID stabilizability. Note here that, since a general PID-controller contains a pole at the origin and is hence unstable, the result given in part 2) of Lemma 1 is not obvious.

Lemma 1: (Necessary Conditions for Existence of PID): Let \(G \in \mathbb{R}_p^{n \times n}\). Let \(\text{rank}(G(s)) = n_y\).
1) If \(G\) admits a PID-controller such that the integral constant \(K_i \in \mathbb{R}_p^{n \times n}\) is nonzero, then \(G\) has no transmission zeros at \(s = 0\) and \(\text{rank}(K_i) = n_y\).
2) If \(G\) admits a PID-controller, then \(G\) is strongly stabilizable. Although several PID-controller synthesis methods exist for stable plants, which obviously admit PID-controllers, Proposition 1 gives a method applicable to MIMO plants.

Proposition 1: (PID-Controller Synthesis for Stable Plants): Let \(H \in \mathbb{S}^{n \times n}\) and \(\text{rank}(H(s)) = n_y \leq n_u\). If the integral term is to be nonzero, also let \(\text{rank}(H(0)) = n_y\) and let \(H^t(0)\) be a right inverse of \(H(0)\). For any of the PI, D, or PD terms in \(C_{\text{pid}}\) to be nonzero, choose the corresponding \(\Delta_p, \Delta_d,\) and \(\Delta_i = 1\); to make any of these terms zero, choose the corresponding \(\Delta_p, \Delta_d,\) and \(\Delta_i = 0\). Choose any \(K_p, K_d \in \mathbb{R}_p^{n \times n}\), \(\tau \in \mathbb{R}_+\). Choose any \(\gamma \in \mathbb{R}_+\) satisfying

\[
\gamma \leq \left\| H(s) \left( \Delta_p K_p + \Delta_d K_d \Delta \frac{s + 1}{s + 1} + \Delta_i \frac{H(s) H^t(0) - I}{s} \right) \right\|^2.
\]

Let \(K_i = \gamma K_p, K_d = \gamma K_d,\) and \(K_i = \gamma H(0)\); then

\[
C_{\text{pid}} = \Delta_p \gamma K_p + \Delta_d \gamma K_d \frac{s + 1}{s + 1} + \Delta_i \gamma K_i \frac{s + 1}{s + 1},
\]

is a PID-controller that stabilizes \(G\).

Lemma 2 states that if a stabilizing \(C_p, C_i,\) and \(C_d\) exist for the plant \(G\), then it is possible to find suitable choices for the remaining constants and extend to stabilizing PI, PD, and PID-controllers.

Lemma 2: Let \(G \in \mathbb{R}_p^{n \times n}\).
1) Two-step controller synthesis [14, Th. 5.3.10]: Suppose that \(C_p\) stabilizes \(G\) and \(C_i\) stabilizes \(H := G(I + C_p G)^{-1} \in \mathcal{M}(\mathbb{S})\). Then, \(C = C_p + C_i\) also stabilizes \(G\).
2) PID-controllers constructed from subsets: If \(G\) admits a subset of a PID-controller where at least one of the three constants \(K_p, K_d,\) and \(K_i\) is nonzero, then \(G\) admits a PID-controller such that any two or all three of the three constants are nonzero. The integral constant \(K_i\) is nonzero only if \(G\) has no transmission zeros at \(s = 0\).
3) Two-step PID-controller synthesis: Let \(G\) have no transmission-zeroes at \(s = 0\). Suppose that there exists a PD-controller \(C_{\text{pd}}\) stabilizing \(G\). Then, \(C_{\text{pid}} = C_{\text{pd}} + K_i/s\) also stabilizes \(G\), where \(C_{\text{ih}} = K_i/s\) is any I-controller that stabilizes \(H_{\text{pd}} := G(I + C_{\text{pd}} G)^{-1}\). In particular, \(C_{\text{ih}} = K_i/s\) can be chosen as

\[
K_i \frac{K_i}{s} = \frac{\gamma H_{\text{pid}}(0) h}{s} = \frac{\gamma [G(0) + K_p]}{s}
\]
for any $\gamma \in \mathbb{R}_+$ satisfying

$$0 < \gamma < \left\| \frac{H_{pd}(s)H_{pd}(0)^t - I}{s} \right\|^{-1}. \tag{5}$$

Condition (5) on the scaling factor $\gamma$ used in $K_i/s$ proposed in Lemma 2 can also be expressed as

$$\gamma \left\| (I+GC_{pd})^{-1} \frac{G(s)H_{pd}(0)^t - C_{pd}}{s} - \frac{H_{pd}(s)K_{d}}{\tau s+1} \right\|^{-1}. \tag{6}$$

III. PLANT CLASSES THAT ADMIT PID-CONTROLLERS

We investigate specific unstable plant classes that admit PID-controllers and propose synthesis methods. By Lemma 1, plants that admit PID-controllers are necessarily strongly stabilizable. Section III-A deals with various plants satisfying the parity-interlacing property, with restrictions on the unstable region blocking zeros but no restrictions on the location of the poles. By contrast, the PO-poles are restricted in Section III-B in order to allow complete freedom in the zero locations.

The unstable plant classes considered in this section are all square $(n_y = n_u)$ and full-rank, i.e., $G$ satisfies $G \in \mathbb{R}_{p \times q}^{n_y \times n_u}$, (normal) rank $G(s) = n_y$.

1) For unstable plants with no zeros in $\mathcal{U}$ including infinity ($G^{-1} \in \mathbb{S}^{n_y \times n_y}$), there exist $P$, $I$, $D$, (hence, PID, PI, and PID) controllers.

2) For unstable plants with one or two blocking zeros in $\mathcal{U}$ (including $s = 0, \infty$) satisfying certain norm bounds, there exist PID-controllers; when neither one of the zeros is at $s = 0$, there exist PID-controllers.

3) For unstable plants with one or two poles in $\mathcal{U}$ (including $s = 0$) satisfying certain norm bounds, there exist PID-controllers.

A. Unstable Plants With Restrictions on the $\mathcal{U}$-Zeros

Unstable plants without blocking zeros in the unstable region $\mathcal{U}$ (including infinity), which are obviously strongly stabilizable, admit PID-controllers. Plants that have (one or two) real-axis blocking zeros in $\mathcal{U}$ also admit PID-controllers under certain sufficient conditions on these zeros.

1) Unstable Plants With No $\mathcal{U}$-Zeros: Let $G \in \mathbb{R}_{p \times q}^{n_y \times n_u}$ have no transmission-zeros in $\mathcal{U}$ (including infinity); hence, $G$ satisfies the necessary condition in part 1) of Lemma 1 for existence of PID-controllers with nonzero $K_i$. Therefore, $G$ has an LCF $G^{-1} = G^{-1}(s) = (G^{-1} - I)^{-1}$. Let $K_i(s) = K_i(s) + \gamma I$, where $\gamma > 0$. Proposition 2 shows that $G$ admits $P$, $I$, $D$, and PID-controllers.

Proposition 2: Choose $K_p, K_d \in \mathbb{R}^{n_u \times n_y}$, such that $\det(G^{-1} + K_p + \gamma^{-1} K_d) \neq 0$. Let $W_{pd} := G^{-1}(s) + K_p + (K_d s/\tau s + 1)$. Then

$$C_{pid} = K_p + \frac{\gamma}{\tau s + 1} \left[ G(s)^{-1} + K_p + \gamma^{-1} K_d \right]$$

stabilizes $G$ for any $\theta \in \mathbb{R}_+$, satisfying

$$\theta > \left\| \frac{W_{pd} \left( G(s)^{-1} + K_p + \gamma^{-1} K_d \right)^{-1} - I}{s} \right\|. \tag{7}$$

2) Unstable Plants With Positive Real Zeros Including Zero and Infinity: Let $G$ have no transmission zeros in $\mathcal{U}$ other than $\ell \in \{1, 2\}$ (one or two) real-axis blocking zeros (at $s = z_j \in \mathbb{R}$, $z_j \geq 0$, $j \in \{1, 2\}$); $G$ may have any number of transmission zeros in the stable region. The condition in 1) of Lemma 1 for existence of PID-controllers with nonzero $K_i$ is satisfied only when $z_j \neq 0$. The poles of $G$ are completely arbitrary, except that we assume $G$ has no poles at $s = 0$ if there is a zero close to the origin.

We consider two cases where the real-axis zeros at $z_j \geq 0$ are either small, including $z_j = 0$, or "large," including infinity.

Case I: Let $z_j \in \mathbb{R}$, $z_j \geq 0$, $z_j \leq z_{j+1}$, with $a_j \in \mathbb{R}_+, j \in \{1, 2\}$, let

$$y_j := \prod_{j=1}^{\ell} y_j = \prod_{j=1}^{\ell} (a_j s^\ell) + 1 \quad x := \prod_{j=1}^{\ell} x_j = \prod_{j=1}^{\ell} (s - z_j). \tag{9}$$

Let $G$ have an LCF $G = Y^{-1}X = ((x/y)G^{-1})^{-1}((x/y)I)$. Let $G$ have no poles at $s = 0$. Under these assumptions in Proposition 3, $G$ admits PID-controllers if upper bounds are imposed on the zeros; if $\ell = 1$, $G$ also admits P-controllers; $G$ admits PI and PID-controllers only if $z_j \neq 0$. Let $\ell = 1$, some plants (e.g., $G = (s - z)/(s - p)$, $z, p > 0$) do not admit D- and I-controllers. If $\ell = 2$, some plants (e.g., $G = (s - z_1)(s - z_2)/(s^2 - p)$, $z_1, z_2, p > 0$) do not admit PID, D, and I-controllers.

Proposition 3: Let $G$ have no poles at $s = 0$. Let $\ell \in \{1, 2\}$. With $x, y$ as in (9), let $Y = ((x/y)G^{-1})^{-1}((x/y)I)$. Let $\Phi \in \mathcal{M}(S)$, where $z_j \in \mathbb{R}$, $z_j \geq 0$. Define $C_1 := 1 + (k_d s/(\tau s + 1))$ and define

$$\Phi_1 = C_1^{-1} x G^{-1}(s) Y(0) - 1. \tag{10}$$

If $0 \leq z_j \leq \|\Phi_1 s\|^{-1}$, then for any $\alpha \in \mathbb{R}_+$ satisfying (11), the PD-controller $C_2$ in (12) stabilizes $G$.

$$\alpha < \left\| \Phi_1 s \right\|^{-1} - z_j, \tag{11}$$

$$C_1 = K_p + K_d s \frac{1}{\tau s + 1} \frac{1}{1 + \alpha \tilde{C}_1 Y(0)} \tag{12}$$

If $\ell = 2$, choose any $K_2 \in \mathbb{R}_+$. Define

$$\Phi_2 := (k_d s + 1)^{-1} x G^{-1}(s) Y(0) - 1. \tag{13}$$

If $2(z_1 + z_2) \leq \||\Phi_2 s\|^{-1}$, then for any $\alpha, \beta \in \mathbb{R}_+$ satisfying (14), choose $\alpha \in \mathbb{R}_+$, and then $Y = 2(z_1 + z_2) + \alpha + \beta, \rho := \alpha \beta + \alpha z_2 + \beta z_1, \tau := \rho^{-1} s \frac{1}{2(z_1 + z_2) + \alpha + \beta} \frac{1}{\tau s + 1} \frac{1}{1 + \alpha \tilde{C}_2 Y(0)} \tag{15}$$

If $z_j \neq 0$, let $C_{pid} = C_{pd}$ as in (12) or (15) for $\ell = 1$ or $\ell = 2$, respectively. Choose any $\gamma \in \mathbb{R}_+$ satisfying (5), then, a PID-controller that stabilizes $G$ is given by

$$C_{pid} = C_2 + \frac{\gamma}{s} \left( G^{-1}(s) + K_{pd} \right) \tag{16}$$

Equation (16) becomes $C_{pid} = C_1 + ((z_1 + z_2) \gamma/s) G^{-1}(0)$ for $\ell = 1$ and $C_{pid} = C_2 + (\rho^{-1}(s + z_1, z_2) \gamma/s) G^{-1}(0)$ for $\ell = 2$. \triangle
Case 2): With \( z_j \in \mathbb{R}, 0 < z_j \leq \infty \) and \( a_j \in \mathbb{R}, a_j > 0 \), \( j \in \{1, \ell\} \), let
\[
\tilde{y} := \left( \int_{j=1}^{\ell} y_j \right) \left( \frac{s + a_j}{t} \right) \quad \tilde{x} := \left( \int_{j=1}^{\ell} x_j \right) \left( \frac{1 - s}{t} / z_j \right).
\]
(17)

Let \( G \) have an LCF \( G = Y^{-1}X = \left((\tilde{x}/\tilde{y})G^{-1}\right)^{-1}((\tilde{x}/\tilde{y})T) \). Proposition 4 shows that, with lower bounds on the zeros, plants in this class admit PD and PID-controllers; if \( \ell = 1 \), they also admit P- and PI-controllers. If \( \ell = 1 \), some plants \( e.g., G = 1/(s - p), p > 0 \) do not admit D- and I-controllers. If \( \ell = 2 \), some plants \( e.g., G = 1/(s^2 - p), p > 0 \) do not admit P-, D-, and I-controllers.

**Proposition 4:** Let \( \ell \in \{1, 2\} \). With \( \tilde{x}, \tilde{y} \) as in (17), let \( Y = (\tilde{x}/\tilde{y})G^{-1} \in \mathcal{M}(\mathbb{S}) \), where \( z_j \in \mathbb{R}, 0 < z_j \leq \infty \), and \( a_j \in \mathbb{R}, a_j > 0 \), \( j \in \{1, \ell\} \). Let \( Y(\infty)^{-1} = \tilde{x}^{-2}\tilde{y}G(s)_{s=\infty} \). If \( \ell = 1 \), choose any \( k_p, k_i \in \mathbb{R}_+ \). Define \( C_1 := (\tau_i s + k_i)/(\tau_i s + 1) \) and define
\[
\Psi_1 := \tilde{C}_1^{-1}2G^{-1}(s)Y(\infty)^{-1} - s I.
\]
(18)

If \( \|\Psi_1\| < \infty \), then, for any \( \alpha \in \mathbb{R}_+ \) satisfying (19), the PD-controller \( C_1 \) in (20) stabilizes \( G \)
\[
\alpha > \left( \frac{1}{2} \right) \|\Psi_1\|/|z_1|^{-1} \|\Psi_1\|, \quad \alpha > \left( \alpha \right) \]
\[
C_1 = k_{p_1} + \frac{k_i \tau_i}{\tau_i s + 1} \left[ \frac{\alpha}{1 + \alpha/z_1} \right] Y(\infty).
\]
(19)

If \( \ell = 2 \), choose any \( k_p \in \mathbb{R}_+ \). Define \( \Psi_2 := (s + k_2)^{-1}xG^{-1}(s)Y(\infty)^{-1} - s I \).
\[
\alpha > \left( \frac{1}{2} \right) \|\Psi_2\|/|z_1|^{-1} \|\Psi_2\|, \quad \alpha > \left( \alpha \right) \]
\[
C_2 = k_{p_2} + \frac{k_{i_2}}{\tau_i s + 1} \left[ \frac{\alpha}{1 + \alpha/z_1} \right] Y(\infty).
\]
(20)

Remark 1:
1) In Proposition 3, when \( \ell = 1 \), choosing \( k_{p_1} = 0 \) gives a P-controller \( C_p = (z_1 + \alpha)^{-1}Y(0) \) in (12); if \( z_1 \neq 0 \), then (16) becomes a PI-controller \( C_{p_1} = (z_1 + \alpha)^{-1}(-z_1 + (\alpha/s)G^{-1})(0) \). When \( \ell = 2 \), let \( \alpha = 0 \). \( C_{p_2} = k_{i_2}/\tau_i \) in (15) becomes a PI-controller \( C_{p_2} = k_{i_2}/\tau_i \) for \( G^{-1}(s)Y(\infty)^{-1} \).
2) By Proposition 3 (and by the dual Proposition 4), any unstable plant with (up to) two blocking zeros at \( z_1 = z_2 = 0 \) (or \( z_1 = z_2 = \infty \)) and any number of zeros in the stable region can be stabilized using PID-controllers such that the norm bounds \( z_1 < |\Psi_1/s|^{-1} 0 \). \( \Psi_1/s \) is satisfied only if \( z_1 < |\Psi_1/s|^{-1} \). The smallest positive real pole \( p_{\text{min}}Gamma \) becomes \( G^{-1}(p_{\text{min}}Gamma) = 0 \) implies (by the \( H_{\infty} \)-norm definition) that \( |\Psi_1/s|^{-1} \geq 1 \).
3) For some insight on the norm bounds of Proposition 3, we observe that when \( G \) has only one \( U \)-zero \( z_1 \) (\( \ell = 1 \)), the bound \( 0 < z_1 < |\Psi_1/s|^{-1} \) is satisfied only if \( z_1 < \infty \) and \( z_1 < |\Psi_1/s|^{-1} \). The smallest positive real pole \( p_{\text{min}}Gamma \) becomes \( G^{-1}(p_{\text{min}}Gamma) = 0 \) implies (by the \( H_{\infty} \)-norm definition) that \( |\Psi_1/s|^{-1} \geq 1 \).

\[ z_1 < p_{\text{min}}Gamma \] as claimed. Similarly, the norm bound \( z_1 > |\Psi_1/s|^{-1} \) in Proposition 4 is satisfied only if \( z_1 > \infty \) is to the right of the largest positive real pole \( p_{\text{max}}Gamma \) because \( G^{-1}(p_{\text{max}}Gamma) = 0 \) implies \( z_1 > |\Psi_1/s|^{-1} \).
4) Consider, for example, the simple scalar plant \( G = (s - z_1)(s - z_2)/(s^2 - 81) \). This has two \( U \)-zeros \( z_1, z_2 > 0 \). Following Proposition 3, for an arbitrary choice \( k_2 > 0 \), say \( k_2 = 0.09 \), if \( 2(z_1 + z_2) < 3.0043 \), then PD- and PID-controllers exist.

3) Unstable Plants With Complex Zeros:
Let \( G \) have no transmission-zeros in \( \ell \) other than a complex-conjugate pair of blocking zeros at \( z_1 = z_2 \in \mathbb{U} \), but \( G \) may have any number of transmission zeros in the stable region. The poles of \( G \) are completely arbitrary except that we assume \( G \) has no poles at \( s = 0 \) if there is a \( j \)-axis zero close to the origin. We consider two cases where the complex-conjugate zeros are either “small,” including zero, or “large,” including infinity. Let
\[
y := (s + g)^{2} x := \sum_{j=1}^{\ell}(s - z_j) = s^{2} - 2fs + g^{2}
\]
(24)
where \( z_1 = z_2 \in \mathbb{U}, f, g \in \mathbb{R}_+, f \geq 0, \) and \( f < g \). Write \( G \) as \( G = Y^{-1}X = ((x/y)G^{-1})(x/y)T \). The condition in 1) of Lemma 1 for existence of PID-controllers with nonzero \( K_1 \) is satisfied since \( g \neq 0 \), i.e., \( G \) has no transmission zeros at \( s = 0 \in \ell \). Part 1) of Proposition 5 shows that with sufficient conditions that impose upper bounds on the zeros, plants that have no poles at \( s = 0 \) admit PI-, PD-, and PID-controllers. Some plants \( e.g., G = (s^2 + g^2)/(s^2 - p) \), \( p > 0 \) do not admit P- and D-controllers. Part 2) of Proposition 5 shows that with sufficient conditions that impose lower bounds on the zeros, plants in this class admit PID-controllers.

**Proposition 5:**
1) Let \( G \) have no poles at \( s = 0 \). With \( x, y \) as in (24), let \( Y = (x/y)G^{-1} \in \mathcal{M}(\mathbb{S}) \). Choose any \( k_p, k_i \geq 0, \) and \( \tau > 0 \). Define \( C := k_i + (1/s) + (k_2s/\tau s + 1) \) and
\[
\Psi := \frac{x}{x(0)G^{-1}(s)G(0)} - I.
\]
(25)
If \( 2(f + g) < |\Psi| \), then a PID-controller that stabilizes \( G \) is given by \( 1/(2f + g)C x(0)G^{-1}(s)I \), i.e.,
\[
C_{p} = \frac{1}{2(f + g)} \left( k_p + \frac{1}{s} + k_2s/\tau s + 1 \right) x(0)G(0)^{-1}.
\]
(26)
2) Let \( Y(\infty)^{-1} = y^{-1}yG(s)_{s=\infty} \). Choose any \( \chi(\epsilon) := s^2 + k_2s + k_3, k_2, k_3 \in \mathbb{R}_+ \). Define \( \Psi := \chi^{-1}xG^{-1}(s)Y(\infty)^{-1} - I \).
(27)
If \( g^2/(2f + g) > |\Psi| \), then for any \( \alpha \in \mathbb{R}_+ \) satisfying
\[
0 < \alpha < \frac{1}{\alpha} |\Psi|^{-1} - \frac{2(f + g)}{g^2}
\]
(28)
\[
C_{p} = \frac{\chi}{\alpha} \frac{x}{s} G(0) - I.
\]
(29)

B. Unstable Plants With Restrictions on the \( U \)-Poles
The restrictions on the unstable poles are completely due to the restrictions on the \( U \)-zeros in Section III-A. We show that plants that
have (one or two) \( \ell \)-poles admit PID-controllers under certain sufficient conditions.

Let \( G \in \mathfrak{R}_p^{\times \times n_y} \), rank \( G = n_y \), and let \( G \) have no transmission zeros at \( s = 0 \). Let \( G \) have any number of poles in the stable region. Other than \( \ell \in \{1, 2\} \) (one or two) \( \ell \)-poles at \( p_1 \in \mathbb{R} \) and \( p_2 \in \mathbb{R} \), let \( G \) have no poles in the unstable region \( \mathbb{U} \). If \( \ell = 1, p_1 \in \mathbb{R}, p_1 \geq 0 \). If \( \ell = 2, p_1, p_2 \in \mathbb{U} \) may be real or complex. The poles at \( p_1 \) or \( p_2 \) may appear in some or all entries of \( G \). With \( a_j \in \mathbb{R}_+, j \in \{1, \ell\} \), let

\[
y := \prod_{j=1}^{\ell} y_j = \prod_{j=1}^{\ell} (a_j s + 1) \quad n := \prod_{j=1}^{\ell} (s - p_j).
\] (30)

Let \( G \) have an LCF \( G = Y^{-1} X = (n/y) I - (n/y) G \), where rank \( X(p_j) = \text{rank} G(s) \mid s = p_j = n_y, j \in \{1, \ell\} \). Furthermore, since \( G \) has no transmission zeros at \( s = 0 \), rank \( X(0) = \text{rank} G(s) \mid s = 0 = n_y \). We consider the following two cases of real and complex conjugate pairs of poles.

Case 1) The unstable poles are real, i.e., \( p_j \in \mathbb{R}, p_j \geq 0, j \in \{1, \ell\} \). Part 1 of Proposition 6 shows that under certain assumptions, plants in this class admit PID- and PD-controllers; if \( \ell = 1 \), they also admit PI- and PI-controllers. If at least one \( p_j = 0 \), then \( G \) does not admit D-controllers since the plant pole at \( s = 0 \) would then cancel the zero in \( C_d \). If \( \ell = 2 \), some plants (e.g., \( G = \frac{1}{s + \alpha + \beta} \) do not admit I-controllers. For \( p > 0 \), some plants (e.g., \( G = \frac{1}{s \alpha - \beta} \), \( p > 0 \) do not admit D- and I-controllers. If \( \ell = 2 \), some plants (e.g., \( G = \frac{1}{s + p_1 s + p_2} \), \( p_1 \geq 0, p_2 \geq 0 \) do not admit P- or I-controllers.

Case 2) The two poles are a complex conjugate pair, i.e., \( p_1 = \bar{p}_2, n = s^2 - (p_1 + p_2) s + p_1 p_2 = s^2 - 2 f s + g^2, f \geq 0, g > 0, f < g \). In this case, \( X(0) = g^2 G(0) \). Part 2 of Proposition 6 shows that under certain assumptions, plants in this class admit PD-, PI-, and PID-controllers. Some plants (e.g., \( G = \frac{1}{s^2 + g} \), \( g > 0 \) do not admit P- or I-controllers.

Proposition 6: Let \( \ell \in \{1, 2\} \). With \( n, y \) as in (30), let \( X = (n/y) G \in \mathfrak{M}(S) \), where \( p_j \in \mathbb{U} \). Let rank \( X(p_j) = \text{rank} G(s) \mid s = p_j = n_y, j \in \{1, \ell\} \). Let \( X(0) = n_y G(s) \mid s = 0 \). Define \( C_1 = 1 + \frac{F(s)}{\tau s + 1} \), and define

\[
\Gamma_1 := n G(s) \hat{C}_1 X(0)^{-1} - I.
\] (31)

If \( 0 < p_1 < ||\Gamma_1/s||^{-1} \), then for any \( \alpha \in \mathbb{R} \), satisfying (32), the PD-controller \( C_1 \) in (33) stabilizes \( G \)

\[
0 < \alpha < ||\Gamma_1/s||^{-1} - p_1, \text{ (32)}
\]

\[
C_1 = K_{p1} + \frac{K_{d1} s}{\tau s + 1} = (p_1 + \alpha) \left( I + \frac{F(s)}{\tau s + 1} \right) X(0)^{-1} \quad \text{ (33)}
\]

If \( \ell = 2 \), choose any \( \tau_2 \geq 0 \). Define

\[
\Gamma_2 := (\tau_2 s + 1)^{-1} n G(s) X(0)^{-1} - I. \text{ (34)}
\]

If \( 2 (p_1 + p_2) < ||\Gamma_2/s||^{-1} \), then for any \( \alpha, \beta \in \mathbb{R} \), satisfying (35), the PID-controller \( C_2 \) in (36) stabilizes \( G \), where \( \eta := 2 (p_1 + p_2) + \alpha + \beta, p := \alpha + \alpha p_2 + \beta p_1, K_{p2} = p X(0)^{-1} \), and \( K_{d2} = (\gamma - \rho_2 X(0)^{-1} \)

\[
0 < \alpha + \beta < ||\Gamma_2/s||^{-1} - 2 (p_1 + p_2) \text{ (36)}
\]

Choose any \( \gamma \in \mathbb{R} \), satisfying (5). Then, a PID-controller that stabilizes \( G \) is given by (16), where \( C_{p2} = C_1 + \frac{K_{p2}}{\tau_2 s + 1} \), \( K_{p2} \) as in (33) or (36) for \( \ell = 1 \) or \( \ell = 2 \), respectively; (16) becomes \( C_{p2} = C_1 + (\gamma \alpha/s) X(0)^{-1} \) for \( \ell = 1 \), and \( C_{p2} = C_2 + (\gamma (p_1 p_2 + p)/s) X(0)^{-1} \) for \( \ell = 2 \).

Case 2) Let \( p_1 = p_2 \in \mathbb{C} \), then \( s^2 - (p_1 + p_2) s + p_1 p_2 = s^2 - 2 f s + g^2, f \geq 0, g > 0, f < g \). Choose any \( \gamma \in \mathbb{R} \). Define \( I \) as in (34). If \( f + 2 g < ||\Gamma_2/s||^{-1} \), then for any \( \beta \in \mathbb{R} \), satisfying (37), the PD-controller in (38) stabilizes \( G \), where \( K_\beta = (\beta - f) g X(0)^{-1} \), \( K_d = (\beta + f + 2 g - \tau_2 (\beta - f) g) X(0)^{-1} \)

\[
0 < \beta < ||\Gamma_2/s||^{-1} - (f + 2 g) \text{ (37)}
\]

\[
C_{p2} = K_p + \frac{K_{d2}}{\tau_2 s + 1} = \left[ \frac{(\beta + f + 2 g) s + (\beta - f) g X(0)^{-1}}{\tau_2 s + 1} \right] \frac{G(0)^{-1}}{g^2} \text{ (38)}
\]

Choose any \( \gamma \in \mathbb{R} \). With \( H_{s2} = 0 \), a PID-controller that stabilizes \( G \) is given by

\[
C_{p2} = C_1 + \frac{\gamma (\beta + f - g) X(0)^{-1}}{s} \text{ (39)}
\]

Remarks 2:

1) In part 2) of Proposition 6, if \( 2 (f + g) < ||\Gamma_1/s||^{-1} \), then choosing \( \beta = f \) in (37), \( C_{p2} \) in (38) becomes a D-controller \( C_d = (K_{d1}/(\tau_2 s + 1)) = 2 (f + g) G(0)^{-1} s/g^2 (\tau_2 s + 1) \) and (38) becomes an ID-controller \( C_{id} = C_d + (\gamma G(0)^{-1})/s \).

2) By Proposition 6, any plant with (up to) two poles at \( p_1 = p_2 = 0 \) and any number of poles in the stable region, with no restrictions on the location of the zeros, can be stabilized using PID-controllers since the norm bounds \( p_1 < ||\Gamma_1/s||^{-1} \) or \( 2 (p_1 + p_2) < ||\Gamma_2/s||^{-1} \) are obviously satisfied.

3) The norm bounds of Proposition 6 can be interpreted as follows: When \( G \) has only one \( \ell \)-pole \( p_1 (\ell = 1) \), the bound \( 0 < p_1 < ||\Gamma_1/s||^{-1} \) is satisfied only if \( p_1 \) is closer to the origin than the smallest positive real blocking zero \( z_\min \) since \( G(z_\min) = 0 \) implies that \( ||\Gamma_1/s|| \geq 1/z_\min \).

4) The time constant \( \tau \) in the derivative term \( K_{d}/s + (\tau s + 1) \) is completely free in most of the propositions (Propositions 1 and 2; one unstable blocking-zero case of Proposition 6).

IV. CONCLUSION

In this note, we showed the existence of stabilizing PID-controllers for several LTI MIMO plant classes. We proposed systematic PID-controller synthesis procedures that guarantee robust closed-loop stability. We achieved stabilizing PID-controller designs with freedom in the design parameters that can be used towards satisfaction of performance criteria. Some of these results were recently extended to delay differential systems in [5]. Other future goals of this study include identifying other classes of PID stabilizable plants and incorporation of performance issues into design.
APPENDIX

PROOFS

Proof of Lemma 1:

1) Writing \( C_{\text{pid}} = N_c D_c^{-1} = [(K_p + (K_s s)/(\tau_s + 1)))(s^2 + e)[s I/(s + e)] = [(s^2 + e)C_{\text{pid}}]s I/(s + e) \) (for any \( e \in \mathbb{R}_+ \)) and \( G = Y^{-1} X, M_c = Y D_c + X N_c \), unimodular implies \( \text{rank}(M_c) = n_y = \text{rank}(X)(0)K_c \); hence, \( \text{rank}(X) = n_y \) (equivalently, \( G \) has no transmission zeros at \( s = 0 \)), and \( \text{rank}(N_c) = n_y \).

2) For all \( z_i > 0 \), det \( D_z(z_i) = \text{det}(z_i/z_i + e) > 0 \). Now \( M_c \) unimodular implies \( \text{det}(M_c(z_i)) = \text{det}(Y(z_i)) \text{det}(D_z(z_i)) \) has the same sign for all \( z_i \in \mathbb{R}^+ \) such that \( X(z_i) = 0 \); equivalently, \( \text{det}(Y(z_i)) \) has the same sign at all blocking zeros of \( G \); hence, \( G \) is strongly stabilizable [14].

Proof of Proposition 1: Write \( C_{\text{pid}} = N_c D_c^{-1} \), with \( D_c = (1 - (\gamma/(s + \gamma)) \Delta_s I) \) and \( N_c = C_p D_c (D_c = I \) if the integral term is absent \). Then, \( C_{\text{pid}} \) in (3) stabilizes \( H_c \) since \( M_{\text{pid}} := D_c + H C_{\text{pid}} D_c := I + \gamma \Delta_s (K_s + K_s (\gamma s + 1)) D_c + (s^2 + \gamma) \Delta_s (H_c H_c(0) - I/s) \) is unimodular.

Proof of Proposition 2: 1) It follows from [14, Th. 5.3.10].

2) Suppose \( C_p = C_p \) or \( C_g = C_g \) is a P-, or D-, or I-controller stabilizing \( G \); equivalently, \( G = (I + C G)^{-1} \in \mathcal{M}(S) \). The rank of \( G \) is equal to \( \text{rank}(G) = n_y \). When \( C_p = C_p \) or \( C_g = C_g \), if \( G \) has no transmission zeros at \( s = 0 \), i.e., \( \text{rank}(X(0)) = n_y \), then \( \text{rank}(G) = \text{rank}(Y(X_c)^{-1})X(0) = \text{rank}(X(0)) = n_y \).

By Proposition 1, there exists a P-, D-, 1, PI-, ID-, and PID-controller \( C_h \) for \( H_c \) in \( \mathcal{M}(S) \). When \( C_p = C_g \), by Proposition 1, there exists a P-, or D-, or PI-controller \( C_i \) for \( H_c \). By 1), \( C_p + C_i \) stabilizes \( G \) and is a PI-, PD-, or PID-controller.

3) By 1), if \( C_{\text{pid}} \) stabilizes \( G \), then \( C_{\text{pid}} + C_i \), also stabilizes \( G \). By Proposition 1, choosing \( K_p = K_i = 0 \), an I-controller that stabilizes \( H_p \) is given by (3), where \( \gamma > 0 \) satisfies (2), equivalently, (5).

Proof of Proposition 2: By assumption, \( K_p, K_i, \) and \( \tau \) are such that \( W_{\text{pid}}^{-1}(\infty) \) exists. By (8), \( C_{\text{pid}} \) stabilizes \( G \) since \( M_{\text{pid}} := [(s^2 + \theta G)^{-1}] = (s^2 + \theta G) \) \( W_p \) \( W_p^{-1}(\infty) = [I + (1/\theta) \text{s}] W_p \) \( W_p^{-1}(\infty) - I \) \( W_p^{-1}(\infty) \) is unimodular.

Proof of Proposition 3: If \( \delta = 1 \), define \( y_o := (s + \alpha) \) and \( M_1 := XC_1 + Y = \{(x_1/y_1) + YC_1^{-1}C_1 = (x_1/y_1) + (1)/\text{s}/y_1 (Y(0))^{-1}C_1 \} \) \( \| I - (1/\text{s}/y_1) \text{s}(\tau_s + 1/\text{s}) \| W_p \) \( W_p^{-1}(\infty) - I \) \( W_p^{-1}(\infty) \) is unimodular.

Proof of Proposition 4: If \( \delta = 1 \), define \( y_o := (s + \alpha) \) and \( M_1 := XC_1 + Y = \{(x_1/y_1) + YC_1^{-1}C_1 = (x_1/y_1) + (1)/\text{s}/y_1 (Y(0))^{-1}C_1 \} \) \( \| I - (1/\text{s}/y_1) \text{s}(\tau_s + 1/\text{s}) \| W_p \) \( W_p^{-1}(\infty) - I \) \( W_p^{-1}(\infty) \) is unimodular.

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Optimal $\mathcal{H}_2$ Filtering in Networked Control Systems With Multiple Packet Dropout

Mehrdad Sahebsara, Tongwen Chen, and Sirish L. Shah

Abstract—This note studies the problem of optimal $\mathcal{H}_2$ filtering in networked control systems (NCSs) with multiple packet dropout. A new formulation is employed to model the multiple packet dropout case, where the random dropout rate is transformed into a stochastic parameter in the system's representation. By generalization of the $\mathcal{H}_2$-norm definition, new relations for the stochastic $\mathcal{H}_2$-norm of a linear discrete-time stochastic parameter system represented in the space–time form are derived. The stochastic $\mathcal{H}_2$-norm of the estimation error is used as a criterion for filter design in the NCS framework. A set of linear matrix inequalities (LMIs) is given to solve the corresponding filter design problem. A simulation example supports the theory.

Index Terms—$\mathcal{H}_2$-norm, networked control system (NCS), optimal filtering, packet dropout, stochastic systems.

I. INTRODUCTION

Many modern control methods employing the state feedback strategy use state–space formulation. State feedback is applicable under the implicit assumption that all state variables are measurable. However, in practice, some state variables may not be directly accessible or the corresponding sensing devices may be unavailable or very expensive. In such cases, state filters or state estimators are used to give an estimate of the unavailable states.

Networked control systems (NCSs) have gained attention during last few years (e.g., see [7], [8], [12], [14], [19], and references therein). Compared to using the conventional point-to-point system connection, using an NCS has advantages like easy installation and reduced setup, wiring, and maintenance costs. In an NCS, data travel through the communication channels from the sensors to the controller and from the controller to the actuators. Data packet dropout, a kind of uncertainty that may happen due to node failures or network congestion, is a common problem in networked systems. The dropouts happen randomly. Because of random dropout, classical estimation and control methods cannot be used directly. Dropouts can degrade system performance and increase the difficulty of filtering and estimation.

Even though most research conducted on NCSs considers random delay, the closely related random packet dropout has not been well studied and only in last few years has been the focus of some research studies. In fact, systems with packet dropout, uncertain observation, or missing measurements have been studied for a long time (e.g., see [5], [6], [11], [13], [17], [18], and references therein). All of these studies consider the case with uncertainty only in one link and it is not straightforward to extend it to an NCS framework where uncertainty is present both from the sensors to the controller and from the controller to the actuators. Also, in most studies (e.g., see [11], [17], and [18]), the main derivations are given for the case when previous dropout information is given. To the best of our knowledge, no work has been conducted regarding filtering in NCS with multiple packet dropouts, but the problem of stabilization and control has been studied recently in packet dropout systems (e.g., see [9], [10], [20], and references therein). In some of these studies, only sensor data dropouts are studied [9], [20]. While [9] considers adaptive genetic algorithms and simulated annealing algorithms, guaranteed cost control, and the state feedback controller, other references consider switched systems and Markov chains to solve the problem. The main problem in working with Markov chains is the unknown Markov states. Identifying the number of states of the Markov chain and their transient probability by using hidden Markov models are other issues in the research on NCSs.

The problem of optimal $\mathcal{H}_2$ filtering has been tackled in deterministic cases (see, e.g., [4] and [15]). The problem of stochastic packet dropout has also been studied in sensor delay system [16], but, to the best of our knowledge, optimal $\mathcal{H}_2$ filtering has not been studied in NCSs with multiple packet dropout.

In this note, we consider the problem of optimal $\mathcal{H}_2$ filtering in an NCS with multiple packet dropout. A new formulation is proposed to formulate the NCS with multiple random packet dropout. By generalization of the $\mathcal{H}_2$-norm definition, new relations for the stochastic $\mathcal{H}_2$-norm of a linear discrete-time stochastic parameter system represented in the space–time form are derived. The new derivations enable us to consider estimation and filtering of the NCS as a generalization of the classical case. To solve the filtering problem, the filter gains are designed so that the $\mathcal{H}_2$-norm of the estimation error is minimized. As dropout rates are stochastic, the problem formulation leads to a system with stochastic parameters. Thus, the stochastic $\mathcal{H}_2$-norm ($\mathcal{H}_2$-norm) of the estimation error is considered as a measure to minimize. With both deterministic and stochastic inputs present in the NCS framework, a weighted $\mathcal{H}_2$-norm is defined and used. The filtering problem is transformed into a convex optimization problem through a set of linear matrix inequalities (LMIs) that can be solved by using existing numerical methods.