

Incorporating Robustness Requirements Into Antiwindup Design

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Abstract—This paper treats the problem of synthesizing antiwindup compensators that are able to handle plant uncertainty in addition to controller saturation. The uncertainty considered is of the frequency-weighted additive type, often encountered in linear robust control theory, and representative of a wide variety of uncertainty encountered in practice. The main results show how existing linear matrix inequality based antiwindup synthesis algorithms can be modified to produce compensators that accommodate uncertainty better. Embedded within these results is the ever-present performance–robustness tradeoff. A remarkable feature is that the often criticized internal model control antiwindup solution emerges as an “optimally robust” solution. A simple example demonstrates the effectiveness of the modified algorithms.

Index Terms—Antiwindup, constrained control, robust control.

I. INTRODUCTION

MODEL uncertainty and actuator saturation are two troublesome phenomena frequently encountered by control engineers. The problem of ensuring robustness in the presence of model uncertainty in particular has occupied the control community for many years. Other researchers have devoted much time to the study of actuator saturation, yet, commonly—and remarkably—these problems have been considered in isolation and there are few attempts in the literature to unify some of the results (perhaps the most comprehensive account is given in [1]). It may be argued that actuator saturation could be modelled as a particular type of uncertainty, and therefore it could be handled by standard robust control techniques. However, designs based on this mantra would probably be inherently “low gain” in nature and lead to unduly conservative designs, with restricted small-signal performance.

Conversely, the studies of systems subject to actuator saturation have typically chosen to ignore model uncertainty. This has been the case particularly with the antiwindup community, where the implicit assumption has been that the antiwindup compensated constrained system will inherit the robustness of its unconstrained counterpart. While it appears logical to require nominal linear robustness, this appears to be more of a necessary than sufficient condition to ensure robustness of the overall antiwindup compensated nonlinear system.

The goal of this paper is to develop antiwindup compensators that can be implemented in real systems, which will inevitably contain uncertainty, with some confidence. This paper

takes as its inspiration the work of [1], which contains a collection of papers that treat the problem of uncertain, saturated systems in a systematic way. However, most of these papers deal with one-step solutions—rather than the two-step antiwindup approach—and a particular type of uncertainty. Specifically, the uncertainty used is normally parametric or state-space uncertainty, generally of the form

$$\dot{x} = Ax + B\text{sat}(u) + \underbrace{\Delta(x, t)}_{\text{uncertainty}}. \quad (1)$$

Although this type of parametric uncertainty is certainly useful, in practice it is quite limited in scope and is not very useful for capturing *unmodelled dynamics*, which can be more of an obstacle than their modelled, but uncertain, counterparts.

This paper is based on [2] and continues the work of [3] in a logical way, with more focus on constructive algorithms for antiwindup synthesis. It is important to remark that the motivation for this paper, and also for that of [3], is practical: in our experience in the aerospace and hard-disk control fields, it has really been the unmodelled dynamics that have caused the most difficulty in appropriate controller synthesis and, while present, the parametric uncertainty has played a less prominent role (it is often easily countered by large enough low frequency gain).

The aims of this paper are twofold. First, it aims to bring robustness to the fore in antiwindup compensation, where, except for [4], it has had little prominence. Secondly, it aims to promote the use of a type of uncertainty closer to that often used in practice (and close to that used in linear robust control theory, which has shown itself to be successful in recent years).

The notation used in this paper is standard. In particular, we define the induced \mathcal{L}_2 norm, or finite \mathcal{L}_2 gain, of an operator \mathcal{H} as $\|\mathcal{H}\|_{i,2} := \sup_{0 \neq x \in \mathcal{L}_2} (\|\mathcal{H}x\|_2) / (\|x\|_2)$, where $\|x\|_2 = \sqrt{\int_0^\infty \|x\|^2 dt}$ is the \mathcal{L}_2 norm of the vector $x(t)$ and $\|x\|$ is its Euclidean norm. The \mathcal{H}_∞ norm for a (stable) linear operator P is defined as $\|P\|_\infty := \sup_\omega \bar{\sigma}[P(j\omega)]$, where $\bar{\sigma}(\cdot)$ denotes the maximum singular value and $P(j\omega)$ is the frequency response matrix associated with the linear operator P . We do not explicitly distinguish between a linear operator and its transfer function. Equivalently, the \mathcal{H}_∞ norm may be defined as $\|P\|_\infty = \|P\|_{i,2}$. We denote an asymptotically stable, rational transfer function/linear operator P by $P \in \mathcal{RH}_\infty$. A feedback system is said to be “well posed” when there exists a unique solution to the feedback equations. For linear systems, a simple condition for this can be found in, for example, [5].

II. A GENERAL ANTIWINDUP FRAMEWORK

It is convenient to study the robustness of antiwindup (AW) compensators in a framework similar to that proposed in [6] (see

Manuscript received November 1, 2005; revised October 6, 2006. Recommended by Associate Editor S. Tarbouriech.

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Digital Object Identifier 10.1109/TAC.2007.906185

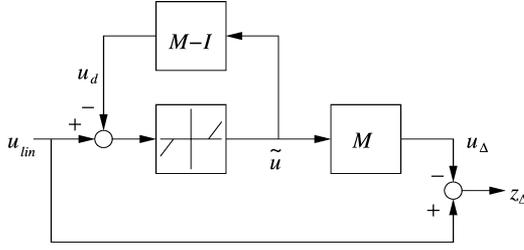


Fig. 6. Robustness optimization for general antiwindup schemes: graphical representation of \mathcal{T}_r .

IV. GENERAL CASE

A. A Stability Robustness Criterion

From Fig. 4, we have that

$$\begin{aligned} y_{\text{lin}} &= G_1 d + G_2 u_{\text{lin}} + W_2 \Delta_G [u_{\text{lin}} - M\mathcal{F}(u_{\text{lin}})] \\ &= G_1 d + G_2 u_{\text{lin}} + W_2 \tilde{\Delta}_G(u_{\text{lin}}) \end{aligned} \quad (14)$$

where, as before, $\mathcal{F}(u_{\text{lin}})$ is the map from u_{lin} to \tilde{u} . Carrying out a small gain analysis, we see that the system is robust against all additive perturbations such that

$$\|\tilde{\Delta}_G\|_{i,2} = \|\Delta_G[I - M\mathcal{F}(u_{\text{lin}})]\|_{i,2} < \frac{1}{\delta}. \quad (15)$$

So nominal robustness is retained if $\|[I - M\mathcal{F}(u_{\text{lin}})]\|_{i,2} \leq 1$. However, as $\mathcal{F}(u_{\text{lin}}) = 0$ around $u_{\text{lin}} = 0$ (as it contains the dead zone), $\|[I - M\mathcal{F}(u_{\text{lin}})]\|_{i,2}$ can never be strictly less than unity. Again this conclusion coincides with our prior discussion, as we could not expect an antiwindup scheme to yield greater robustness margins than the linear system upon which it is constructed. This also serves as justification for the IMC scheme, although it is unlikely to be the *unique* compensator that achieves this optimality.

B. Stability Robustness Optimization

In order to preserve as much robustness as possible, we would like to minimize $\|\mathcal{T}_r\|_{i,2} := \|[I - M\mathcal{F}(u_{\text{lin}})]\|_{i,2}$, preferably making it as close to unity as possible. Diagrammatically, this can be shown as Fig. 6, where we want to minimize the \mathcal{L}_2 gain from u_{lin} to z_Δ , where $z_\Delta := \mathcal{T}_r(u_{\text{lin}})$ is the signal feeding into the uncertainty, as shown in Fig. 4. If we can guarantee that $\|\mathcal{T}_r\|_{i,2} < \gamma$, small gain analysis of Fig. 4 then shows that the system will be asymptotically stable providing that the nonlinear loop $\mathcal{F}(\cdot)$ is asymptotically stable $G_2(s)M(s) \in \mathcal{RH}_\infty$ and the uncertainty $\Delta_G(s) \in \mathbf{\Delta}_{\gamma\delta}$. Note that when $\gamma = 1$, $\mathbf{\Delta}_{\gamma\delta} = \mathbf{\Delta}$ and robust stability equal to that of the linear system is obtained. Thus the goal of our optimization procedure should be to minimize γ , making it as close to unity as possible. As mentioned in [8], this optimization is typically a difficult optimization problem to solve, so instead we shall be content to ensure a certain \mathcal{L}_2 gain bound holds for the map \mathcal{T}_r .

We shall consider robustness optimization using full-order antiwindup compensators. As suggested in [7], we choose $M(s)$ to be part of a right-coprime factorization of $G_2(s) = N(s)M^{-1}(s)$ (this in fact is a dual result to that

of [18], where antiwindup is described as a left coprime factorization of the controller) and attempt to choose a particular factorization such that robustness is optimized. Given the plant realization $G_2(s) \sim (A_p, B_p, C_p, D_p)$, all full-order right coprime factorizations (up to multiplication by an invertible matrix) can be described [7] as

$$\begin{bmatrix} M(s) \\ N(s) \end{bmatrix} \sim \begin{bmatrix} A_p + B_p F & B_p \\ F & I \\ C_p + D_p F & D_p \end{bmatrix} \quad (16)$$

where F is chosen such that $A_p + B_p F$ is a Hurwitz matrix. This choice of coprime factorization prevents implicit equations (algebraic loops) from appearing in the nonlinear loop (as depicted in Figs. 2 and 4, for example) as $M(s) - I$ is strictly proper. The coprime factor representation implies that $G_2(s)M(s) = N(s) \in \mathcal{RH}_\infty$, so, in order to guarantee robust stability for all $\Delta_G \in \mathbf{\Delta}$, it suffices to ensure that $\|\mathcal{T}_r\|_{i,2} < \gamma$ (for $\gamma \leq 1$) and the nonlinear loop $\mathcal{F}(\cdot)$ is well posed. This will indeed be the case if the following inequality holds for sufficiently small γ :

$$J = \frac{d}{dt} x' P x + \|z_\Delta\|^2 - \gamma^2 \|u_{\text{lin}}\|^2 < 0 \quad \forall [x' \quad u_{\text{lin}}'] \neq 0 \quad (17)$$

where x is the state vector associated with the realization of $[M(s) - I, M(s)]'$ and P is some positive definite symmetric matrix. As shown in, for example, [19] and [8], this ensures that the \mathcal{L}_2 gain from u_{lin} to z_Δ is less than γ and that the system in Fig. 6 is asymptotically stable.

As the dead-zone nonlinearity belongs to Sector $[0, I]$ (see [20, ch. 10]), we make use of the inequality

$$2\tilde{u}' X (u_{\text{lin}} - u_d - \tilde{u}) \geq 0 \quad (18)$$

where the ‘‘multiplier’’ matrix $X := \text{diag}(\chi_1, \dots, \chi_m) > 0$ is chosen diagonal as (18) is in fact m scalar inequalities—see, for example, [21, ch. 12]. This can be adjoined to (17) using the S-procedure to obtain the modified function

$$\tilde{J} = \frac{d}{dt} x' P x + \|z_\Delta\|^2 - \gamma^2 \|u_{\text{lin}}\|^2 + 2\tilde{u}' X (u_{\text{lin}} - u_d - \tilde{u}). \quad (19)$$

If $\tilde{J} < 0 \forall [x' \quad \tilde{u}' \quad u_{\text{lin}}'] \neq 0$, this implies that $J < 0$. Evaluating \tilde{J} in a similar manner to [8] (see also [9] and [10] for the discrete and sampled data cases) yields the following linear matrix inequality (LMI):

$$\begin{bmatrix} Q A_p' + A_p Q + L' B_p' + B_p L & B_p U - L' & 0 & L' \\ \star & -2U & I & U \\ \star & \star & -\gamma I & -I \\ \star & \star & \star & -\gamma I \end{bmatrix} < 0 \quad (20)$$

in the variables $Q := P^{-1} > 0, X^{-1} =: U = \text{diag}(\nu_1, \dots, \nu_m) > 0, L, \gamma > 0$.

Satisfaction of this LMI means that (19) is satisfied and hence that the \mathcal{L}_2 gain from u_{lin} to z_Δ is less than γ and a suitable choice of F is given by $F = LQ^{-1}$. A small gain analysis of Fig. 4 reveals that, as $\|\mathcal{T}_r\|_{i,2} < \gamma$ and

$\|(I - K_2G_2)^{-1}K_2W_2\|_\infty = \delta$, then our system will be asymptotically stable for all $\Delta_G \in \Delta_{\gamma\delta}$.

Note from the lower right 2×2 block of the LMI (20)

$$\begin{bmatrix} -\gamma I & -I \\ \star & -\gamma I \end{bmatrix}$$

that, as anticipated earlier, the \mathcal{L}_2 gain can be no less than unity, which is achieved for the IMC scheme.

C. Optimization for Robustness and Performance

The primary goal of antiwindup compensation is to provide performance improvement during saturation, but optimizing the LMI (20) alone does not guarantee this. Indeed, there is little point in optimizing (20) when an optimal solution can be found by inspection as the IMC antiwindup solution. The real use of (20) and the arguments of the previous subsection is to use them in conjunction with performance optimization, the goal being to optimize performance and robustness together, although there will often be a tradeoff.

In [8], it was argued that \mathcal{T}_p , the map from u_{lin} to y_d was central to the “true goal” of antiwindup compensation: if the induced norm of this operator was minimized, the deviation of the system’s nonlinear behavior during and after saturation would be minimized. This paper [8] solved this problem with $\Delta_G = 0$, in the \mathcal{L}_2 sense, for static and low-order compensators (similar treatments for discrete-time AW compensators were given in [9] and [10]).

More realistically, we would really like to optimize some weighted combination of \mathcal{T}_p and \mathcal{T}_r . This can be accomplished by ensuring that

$$\left\| \begin{bmatrix} W_p^{\frac{1}{2}} y_d \\ W_r^{\frac{1}{2}} z_\Delta \end{bmatrix} \right\|_2 < \gamma \|u_{\text{lin}}\|_2 \quad (21)$$

where W_p and W_r are weighting matrices that reflect the relative importance of performance and robustness, respectively, and are chosen by the designer. Taking advantage of the sector bound (18), we are sure that (21) holds if

$$J = \frac{d}{dt} x' P x + \left\| \begin{bmatrix} W_p^{\frac{1}{2}} y_d \\ W_r^{\frac{1}{2}} z_\Delta \end{bmatrix} \right\|^2 - \gamma^2 \|u_{\text{lin}}\|^2 + 2\tilde{u}' X(u - \tilde{u}) < 0 \quad \forall [x' \ \tilde{u}' \ u'_{\text{lin}}]' \neq 0. \quad (22)$$

As before, this inequality holds if the LMI as shown in (23) at the bottom of the page (in the variables $Q > 0, U = \text{diag}(\nu_1, \dots, \nu_m) > 0, L, \gamma > 0$) is satisfied. The derivation of this LMI is carried out in a similar way to that of the previous

section, in the spirit of that done in [8]. Again the matrix F is recovered through $F = LQ^{-1}$. The presence of the weighting matrix W_r ensures that $\|\mathcal{T}_r\|_{i,2} < \gamma/\sqrt{\underline{\sigma}(W_r)}$ and hence that asymptotic stability of Fig. 4 (and hence Fig. 3) is guaranteed for all perturbations $\Delta_G(s) \in \Delta_{\tilde{\gamma}\delta}$, where

$$\Delta_{\tilde{\gamma}\delta} := \left\{ \Delta_G \in \mathcal{RH}_\infty : \|\Delta_G\|_\infty < \frac{1}{\tilde{\gamma}\delta} \right\} \quad (24)$$

and $\tilde{\gamma} := \gamma/\sqrt{\underline{\sigma}(W_r)}$. Obviously, if $W_r = I$, $\Delta_{\tilde{\gamma}\delta} = \Delta_{\gamma\delta}$.

Remark 1: Note that satisfaction of LMI (23) ensures that, with no uncertainty $\Delta_G = 0$, the nonlinear loop in Fig. 4 is globally asymptotically stable and \mathcal{T}_p is finite \mathcal{L}_2 gain stable. Thus as Figs. 4 and 3 are mathematically equivalent, this ensures global asymptotic and finite \mathcal{L}_2 gain of the overall saturated system. Similarly, when $\Delta_G \neq 0$, satisfaction of the LMI (23) ensures that the system in Fig. 3 is globally asymptotically stable with finite \mathcal{L}_2 gain providing that $\|\Delta_G\|_\infty \leq (\sqrt{\underline{\sigma}(W_r)})/(\gamma\delta)$. Thus providing $\tilde{\gamma} \approx 1$ (for example, if $W_r = I$ and $\gamma \approx 1$), the saturated system with AW is robust against a similar class of uncertainties (Δ) to the nominal linear system. \square

Remark 2: Another advantage of using LMI (23) to synthesize full-order compensators is that it tends to prevent fast poles appearing in the compensator dynamics. If a robustness weight (W_r) were not included in the optimization—or if W_r were only chosen small—the poles of the antiwindup compensator would tend to be rather fast, lying far to the left of the imaginary axis. Obviously this would require a very high sampling frequency for implementation, which is not always possible in practice. However, when simultaneously optimizing performance and robustness using (23), the poles are placed in regions more comparable to that of the controller. To see this, note that LMI (23) equivalent to the matrix inequality (25), as shown at the bottom of the next page, is satisfied.

A necessary condition for this to hold is obtained from the (1, 1) term, which with $L = FQ$ is

$$Q(A_p + B_p F)' + (A_p + B_p F)Q + \gamma^{-1} Q F' B_p' W_r B_p F Q < 0. \quad (26)$$

Note that when $W_r = 0$ (i.e., robustness is not included), the matrix F is not constrained in magnitude, but when $W_r \neq 0$, it must be sufficiently “small” to ensure that (26) holds. This feature is reminiscent of solving “singular” \mathcal{H}_∞ problems with LMIs, where poles tend to get placed far from the imaginary axis. Of course in the Riccati-based methods, these singular problems are not solvable, which normally prevents the appearance of these large poles. \square

$$\begin{bmatrix} QA'_p + A_p Q + L'B'_p + B_p L & B_p U - L' & 0 & QC'_p + L'D'_p & L' \\ \star & -2U & I & UD'_p & U \\ \star & \star & -\gamma I & 0 & -I \\ \star & \star & \star & -\gamma W_p^{-1} & 0 \\ \star & \star & \star & \star & -\gamma W_r^{-1} \end{bmatrix} < 0. \quad (23)$$

D. Stability Robustness of the Work in [9]

The work in [9] and [8] advocates only the optimization of antiwindup performance, that is, the minimization of the \mathcal{L}_2 gain of the operator \mathcal{T}_p . By setting $W_r = 0$ and solving the LMI (23), we obtain a full-order compensator that only optimizes this performance. This approach is the continuous-time counterpart to the discrete-time approach of [9]. As argued in [8] and [9], this operator is central to obtaining desirable antiwindup behavior. It is interesting to examine whether this approach has any intrinsic robustness properties.

Suppose now that we consider output multiplicative uncertainty instead of additive uncertainty, that is, $\tilde{G}(s) = (I + \Delta_o(s))G_2(s)$, or, equivalently, that $W_2(s)\Delta_G(s) = \Delta_o(s)G_2(s) = W_3(s)\tilde{\Delta}_o(s)G_2(s)$. Here, $W_3(s) \in \mathcal{RH}_\infty$ represents the “known” part of the output multiplicative uncertainty and $\tilde{\Delta}_o \in \mathbf{\Delta}$ represents the unknown but bounded part.² In this case, our expression for y_{lin} becomes

$$y_{\text{lin}} = G_1 d + G_2 u_{\text{lin}} + W_3 \tilde{\Delta}_o G_2 (u_{\text{lin}} - M\mathcal{F}(u_{\text{lin}})). \quad (27)$$

We are sure that the system is robustly stable when $\mathcal{F}(u_{\text{lin}}) = 0$, as this is a property of the nominal linear system. Therefore, the smaller we can make the extra term $-W_3 \tilde{\Delta}_o G_2 M\mathcal{F}(u_{\text{lin}})$, the closer to nominal robustness we shall be. We can do nothing about $\tilde{\Delta}_o$, so the logical approach is to make

$$\|G_2 M\mathcal{F}(u_{\text{lin}})\|_{i,2} = \|N\mathcal{F}(u_{\text{lin}})\|_{i,2} \geq \|G_2\|_\infty \quad (28)$$

as small as possible, where the last inequality is a consequence of the \mathcal{L}_2 gain of a saturated system being bounded from below by that of the open-loop plant—see [19] and [8] for details. Note that as in [9], because $G_2(s) = N(s)M^{-1}(s)$ is a right coprime factorization of the plant $G_2(s)$, the quantity in (28) is exactly our performance operator norm $\|\mathcal{T}_p\|_{i,2}$. Therefore the minimization of $\|\mathcal{T}_p\|_{i,2}$ not only leads to desirable antiwindup performance but also endows the saturated system with some indirect robustness when the uncertainty is of the output multiplicative type. Note that the robustness is not *guaranteed* to approach that of the linear system and that multiplicative uncertainty is a narrower class of uncertainty than the additive type we have been considering heretofore.

Nevertheless, it does appear to explain some of the results of [22], where a discrete-time version of the results of [8] was implemented on a hard-disk system. In that work, few robustness problems were encountered, and the above analysis goes some way to explaining this.

²In terms of our *additive* uncertainty, observe that we now have two “known” parts either side of the unknown part, i.e., $W_3(s)\tilde{\Delta}_o(s)G_2(s)$, unless some of these transfer function matrices commute.

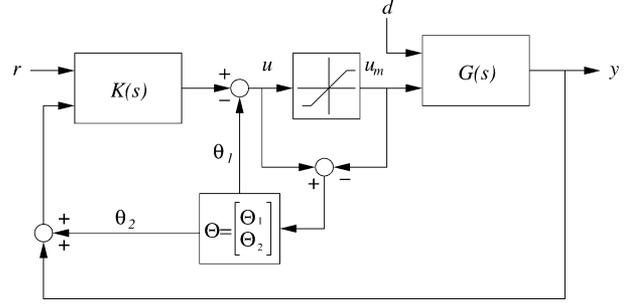


Fig. 7. Structure of static antiwindup scheme.

V. ROBUST STATIC ANTIWINDUP SYNTHESIS

Static antiwindup compensators are probably the most commonly used form of antiwindup compensator; they are simple and computationally efficient to implement. However, it is well known that static antiwindup compensators are not able to achieve global quadratic stability for all plant-controller combinations (see [19], for example). Thus, it is obvious that even if a static compensator existed for the nominal plant, *without uncertainty present*, there is no guarantee that one would exist for the *uncertain plant*. Therefore, it is more difficult to study the *robust* stability of static antiwindup compensators from the outset.

A. A Decoupled Representation

In line with the results presented earlier, to study the robustness of static antiwindup compensators, we first interpret them in the decoupled structure of [7]. Consider Fig. 7, where $\Theta = [\Theta_1 \ \Theta_2]'$ is the static antiwindup compensator. Comparing this to Fig. 1, it follows that in order to realize this compensator in decoupled form, this is equivalent to choosing $M(s)$ [8] as

$$M(s) = (I - K_2(s)G_2(s))^{-1}(-K_2(s)G_2(s)\Theta_2 + \Theta_1 + I). \quad (29)$$

This is exactly the same approach taken in [8], where only performance was considered. Indeed, to combine performance objectives with robustness objectives, we need to consider Fig. 4 and attempt to find the $M(s)$ that minimizes y_d and z_Δ , in the same manner as before except with $M(s)$ restricted to the structure given in (29).

As in [8], one can find a realization of the key transfer functions to be considered as

$$\begin{bmatrix} M(s) - I \\ N(s) \\ M(s) \end{bmatrix} \sim \begin{cases} \dot{\tilde{x}} = \bar{A}\tilde{x} + (B_0 + \bar{B}\Theta)\tilde{u} \\ u_d = \bar{C}_1\tilde{x} + (D_{01} + \bar{D}_1\Theta)\tilde{u} \\ y_d = \bar{C}_2\tilde{x} + (D_{02} + \bar{D}_2\Theta)\tilde{u} \\ u_\Delta = \bar{C}_1\tilde{x} + (\bar{\Gamma} + \bar{D}_1\Theta)\tilde{u} \end{cases} \quad (30)$$

where the state-space matrices can be found in Appendix II.

$$\begin{bmatrix} QA'_p + A_p Q + L'B'_p + B_p L + \gamma^{-1}L'W_r L & B_p U - L' + \gamma^{-1}L'W_r U & 0 - \gamma^{-1}L'W_r & QC'_p + L'D'_p \\ * & -2U + \gamma^{-1}UW_r U & I - \gamma^{-1}UW_r & U D'_p \\ * & * & -\gamma I + \gamma^{-1}W_r & 0 \\ * & * & * & -\gamma W_p^{-1} \end{bmatrix} < 0 \quad (25)$$

In order to ensure that the antiwindup system in Fig. 4 is robust to uncertainty within the ball $\Delta_{\gamma\delta}$, it is sufficient to ensure that (21) holds, where as before W_p and W_r are some positive definite weighting matrices and $z_\Delta = u_{\text{lin}} - u_\Delta$. Similarly to the previous section, this holds if the following inequality is satisfied:

$$J = \frac{d}{dt} \bar{x}' P \bar{x} + \left\| \begin{array}{c} W_p^{\frac{1}{2}} y_d \\ W_r^{\frac{1}{2}} z_\Delta \end{array} \right\|^2 - \gamma^2 \|u_{\text{lin}}\|^2 + 2\tilde{u}' W(u - \tilde{u}) < 0 \quad \forall [\bar{x}' \quad \tilde{u}' \quad u'_{\text{lin}}]' \neq 0. \quad (31)$$

After simplification, this inequality can be reposed as (32) as shown at the bottom of the page, which is an LMI in variables $Q > 0$, diagonal $U > 0, \gamma > 0$, and matrix L . If this LMI is satisfied, a globally stabilizing compensator, ensuring the \mathcal{L}_2 gain bound (21) is satisfied, is given by $\Theta = LU^{-1}$. As before, this implies that the system in Fig. 4 is globally asymptotically stable for all $\Delta_G \in \Delta_{\gamma\delta}$. Satisfaction of this inequality also guarantees well-posedness, although the details are fairly tedious [8].

B. Local Static Results

For arbitrary plant-controller combinations, the LMI (32) may not be feasible and hence a robust, *globally* quadratically stabilizing, static AW compensator will not be computable. However, there is a small relaxation that can be made to allow the synthesis of a *locally* stabilizing robust static AW compensator. We therefore modify our objectives and instead seek to ensure that our AW compensator bestows robust stability for all $\Delta_G \in \Delta_{\gamma\delta}$, where

$$\Delta_{\gamma\delta} := \left\{ \Delta \in \mathcal{RH}_\infty : \|\Delta\|_\infty < \frac{1}{\gamma\delta} \right\} \quad (33)$$

providing that the magnitude of the control signal satisfies the bound

$$|u_i(t)| \leq \beta_i \bar{u}_i \quad \forall i \quad \forall t \geq 0, \quad \beta_i > 1. \quad (34)$$

This implies we are locally robust only providing our control signal remains sufficiently small, which can also be interpreted

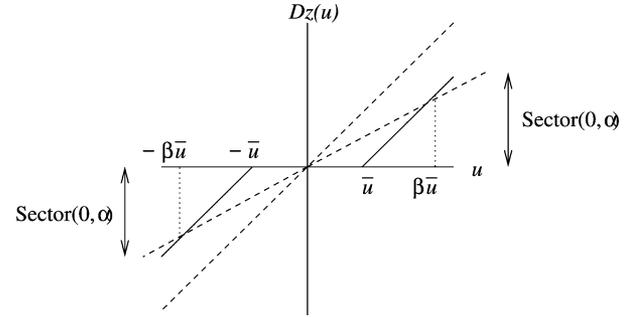


Fig. 8. Reduced sector bounds.

as the states of the system's remaining within some sufficiently small region of the state-space.

The approach we take is essentially the same as advocated in [23] (see also [24]), where we assume the dead-zone nonlinearity inhabits a smaller sector than the $\text{Sector}[0, I]$, namely, $\text{Sector}[0, \alpha I], \alpha \in (0, 1)$. This can be interpreted as shown in Fig. 8. First consider the scalar case: notice that the standard dead zone inhabits $\text{Sector}[0, 1]$, as its graph lies between the gradient zero and gradient one line. However, if we assume that the control signal u never exceeds a certain bound, as described by (34), it is evident that the graph of the dead zone lies within a narrower sector, namely, $\text{Sector}[0, \alpha]$, where $\alpha = \beta/(\beta - 1)$. For the multivariable case, we can follow a similar argument for all $i = \{1, \dots, m\}$ elements of the dead zone and then define $\mathcal{A} := \text{diag}(\alpha_1, \dots, \alpha_m)$. Thus, providing the components of u are sufficiently small, i.e., $u_i \leq \beta_i \bar{u}_i = 1/(1 - \alpha_i) \bar{u}_i \quad \forall i$, then $Dz \in \text{Sector}[0, \mathcal{A}]$.

If we assume that $Dz \in \text{Sector}[0, \mathcal{A}]$ —which is equivalent to the bound (34) holding—then it follows [20] that the following inequality holds:

$$\tilde{u}' X [\mathcal{A}(u_{\text{lin}} - u_d) - \tilde{u}] \geq 0 \quad (35)$$

where again X is some positive definite diagonal matrix.

Using this assumption, the synthesis LMI for a locally robust static AW compensator is given by (36) as shown at the bottom of the page, where $Q > 0, U = \text{diag}(\nu_1, \dots, \nu_m) > 0, L, \gamma >$

$$\begin{bmatrix} Q\bar{A}' + \bar{A}Q & B_0U + \bar{B}L - Q\bar{C}'_1 & 0 & Q\bar{C}'_2 & -Q\bar{C}'_1 \\ * & -2U - D_{01}U - \bar{D}_1L - U\bar{D}'_{01} - L'\bar{D}'_1 & I & U\bar{D}'_{02} + L'\bar{D}'_2 & -U\bar{\Gamma}' - L'\bar{D}'_1 \\ * & * & -\gamma I & 0 & I \\ * & * & * & -\gamma W_p^{-1} & 0 \\ * & * & * & * & -\gamma W_r^{-1} \end{bmatrix} < 0 \quad (32)$$

$$\begin{bmatrix} Q\bar{A}' + \bar{A}Q & B_0U + \bar{B}L - Q\bar{C}'_1\mathcal{A} & 0 & Q\bar{C}'_2 & -Q\bar{C}'_1 \\ * & -2U - \mathcal{A}(D_{01}U + \bar{D}_1L) + (U\bar{D}'_{01} + L'\bar{D}'_1)\mathcal{A} & \mathcal{A} & U\bar{D}'_{02} + L'\bar{D}'_2 & -U\bar{\Gamma}' - L'\bar{D}'_1 \\ * & * & -\gamma I & 0 & I \\ * & * & * & -\gamma W_p^{-1} & 0 \\ * & * & * & * & -\gamma W_r^{-1} \end{bmatrix} < 0 \quad (36)$$

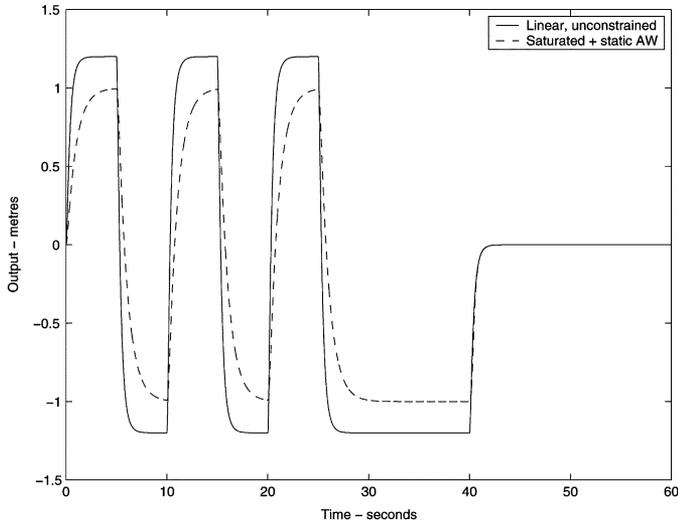


Fig. 10. Response of unperturbed system with optimal static antiwindup.

there appears to be a phase lag between the reference and the output.

To improve the behavior of the system, static antiwindup compensation as suggested in [8] is introduced. This antiwindup compensation minimizes the difference between the nominal linear system's response and the saturated system's response, i.e., $\|\mathcal{T}_p\|_{i,2}$ only is minimized. The \mathcal{L}_2 gain of \mathcal{T}_p , γ is calculated as $\gamma \approx 24$ and a static antiwindup compensator achieving this bound is computed as

$$\Theta = \begin{bmatrix} -0.1909 \\ 0.1402 \end{bmatrix}. \quad (41)$$

This compensator was implemented as shown in Fig. 7. Note that the unperturbed system is quadratically stabilizable by static antiwindup compensation, but we cannot be sure that the perturbed system also has this desirable property (in fact it does not). Fig. 10 shows the response of the system with static antiwindup [i.e., the feedback interconnection of the plant (39) and the controller (40) with the AW compensator (41) and saturation] and it can be seen that the performance of the system has improved: the system output is now in phase with the reference demand, although the infeasibility of the reference means it is not possible for the output to track the input with the correct magnitude. The robust and full order antiwindup compensators introduced in the following sections both yield a similar response to that in Fig. 10 when used on the nominal system $G(s)$.

B. The Perturbed System

The true, or perturbed, plant $\tilde{G}(s) = G(s) + \Delta_G(s)$ is that given in [27]. This has a large resonant peak and is described by the following state-space matrices:

$$\left[\begin{array}{c|c} \tilde{A}_p & \tilde{B}_p \\ \hline \tilde{C}_p & \tilde{D}_p \end{array} \right] = \left[\begin{array}{cc|c} 0 & 1 & 0 \\ -10 & -0.01 & 10 \\ \hline 1 & 0 & 0 \end{array} \right]. \quad (42)$$

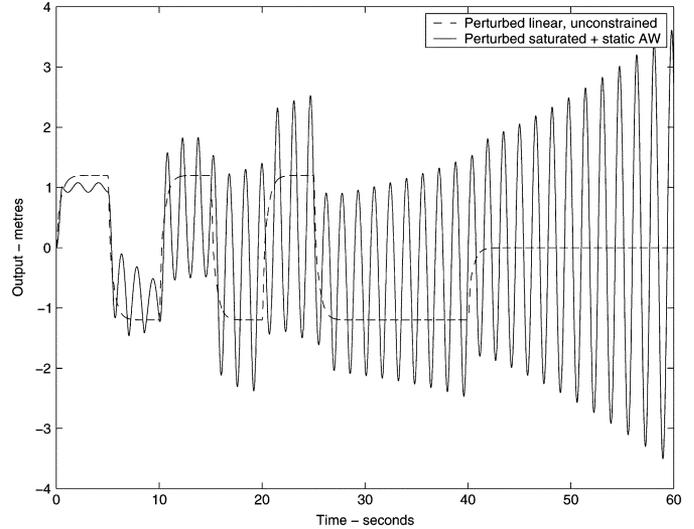


Fig. 11. Response of perturbed system with optimal static antiwindup.

The dashed line in Fig. 11 shows the response of this perturbed plant using the same controller as before and no input saturation [i.e., the feedback interconnection of the plant (42) and the controller (40) without saturation]; the controller yields a similar type of performance as before and hence can be considered satisfactory. However, when input saturation is introduced, the static antiwindup compensator (41), which was designed according to the method in [8] *not considering uncertainty*, actually drives the system unstable, as depicted by the solid line in Fig. 11. In fact, this static antiwindup is worse than *no antiwindup at all* (see Fig. 13), which at least remains stable, although significant phase lag and oscillations are observed. Note that for this perturbed plant and controller, most static antiwindup [28], [8], [19] methods³ are not feasible *in a global sense*, so we cannot expect it to stabilize the system in question for any reference input.

To overcome this problem, we choose $W_p = 0.001$ and $W_r = 1$ and synthesize a full order robust dynamic compensator according to the LMI (23). This yields the matrix F as

$$F = [0.2242 \quad 0.0446] \times 10^{-4}, \quad \gamma \approx 1. \quad (43)$$

In this case, we have essentially the IMC solution, as F is almost zero. As $\gamma \approx 1$, we can expect to recover the robustness results of the linear system. Indeed Fig. 12 shows the system's response with this AW compensator [i.e., the plant (42), the controller (40), and the compensator gain (43)] and we can see that the system is stable. Although the response is not as good as for the unperturbed system, it is substantially better than that of the static AW compensation. It should also be mentioned that this antiwindup compensator yields similar performance to the optimal static AW compensator when applied to the unperturbed plant; that is, the AW dynamic compensator designed using (43) and implemented on the unperturbed plant (39) gives similar performance to the compensator (41) implemented for the same unperturbed plant. In other words, the antiwindup compensator

³These methods all use quadratic Lyapunov functions and sector bounds. Static methods based on alternative Lyapunov functions or "sector bounds" may, however, be feasible.

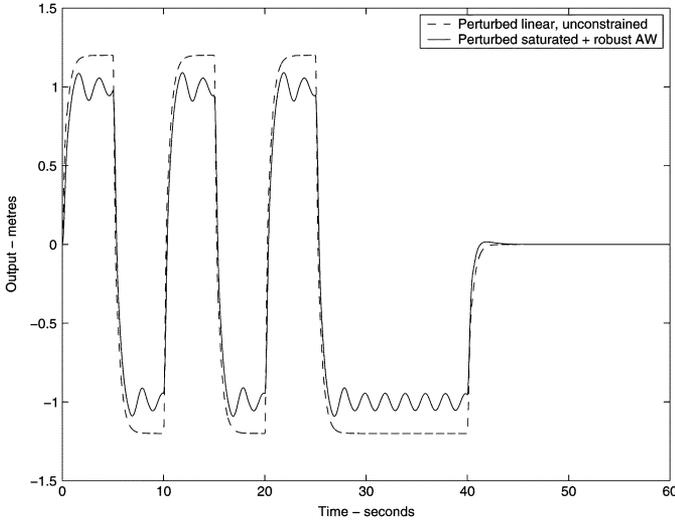


Fig. 12. Response of perturbed system with full-order robust antiwindup.

designed using (43) works well for both perturbed and unperturbed plants.

C. A Robust Static Design

The LMI (36) can be used to provide a static compensator that is guaranteed to be locally robust to perturbations to $G(s)$ providing the saturation level is sufficiently low, i.e., $u_i < \beta_i \bar{u}_i \forall i$, and hence, locally $Dz(\cdot) \in \text{Sector}[0, \mathcal{A}]$. Recall that because the perturbed plant $\tilde{G}(s)$ considered here is not globally quadratically stabilizable by static AW, this motivates the use of a local result instead. In this case, $\mathcal{A} = \alpha$ (single-input single-output system) and we found that many choices of the free parameter $\alpha \in (0, 1)$ could enable us to synthesize a locally robust static antiwindup compensator. However, a value close to unity seemed to provide few robustness improvements, as essentially, we were close to global AW; thus although we could compute a static compensator, its large γ value was indicative of poor robustness. Conversely, a value close to zero led to a nonlinear loop close to being ill-posed. A good compromise seemed to be $\alpha = 0.6$, which guaranteed robustness provided that the control law did not exceed a magnitude of $1/(1 - \alpha) = 2.5$. Using $\alpha = 0.6$ with performance and robustness weights $W_r = 1, W_p = 0.5$, yielded the following static antiwindup compensator:

$$\Theta = \begin{bmatrix} -0.9689 \\ 0.0001 \end{bmatrix}. \quad (44)$$

Fig. 13 shows a comparison between nominal responses, responses with uncertainty and no antiwindup, and the response with uncertainty and robust static antiwindup (all constrained). In contrast to the nonrobust static AW response (Fig. 11) for this input, reasonably good performance is obtained: although some oscillatory behavior is observed, the phase lag is minor (compared to no AW), and stability is maintained. However, the local nature of the results is revealed in Fig. 14, where a larger reference demand of magnitude 2.4 causes the system with the *locally robust* static AW to become unstable. Incidentally, for this reference demand, equipping the system with no

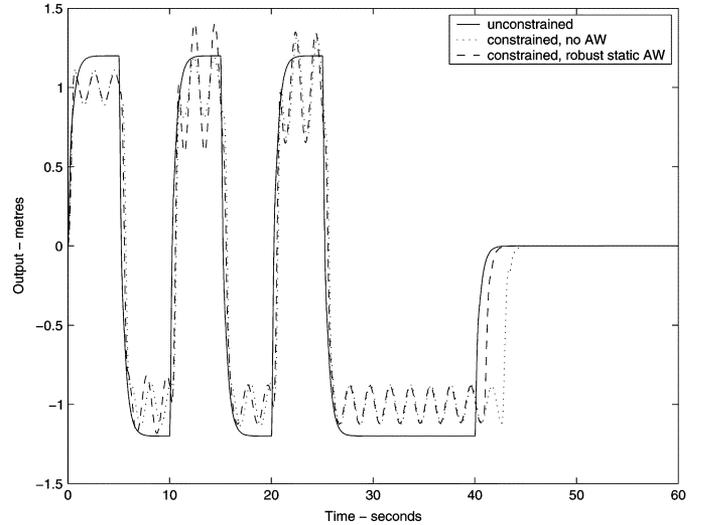


Fig. 13. Response of perturbed system with robust static AW.

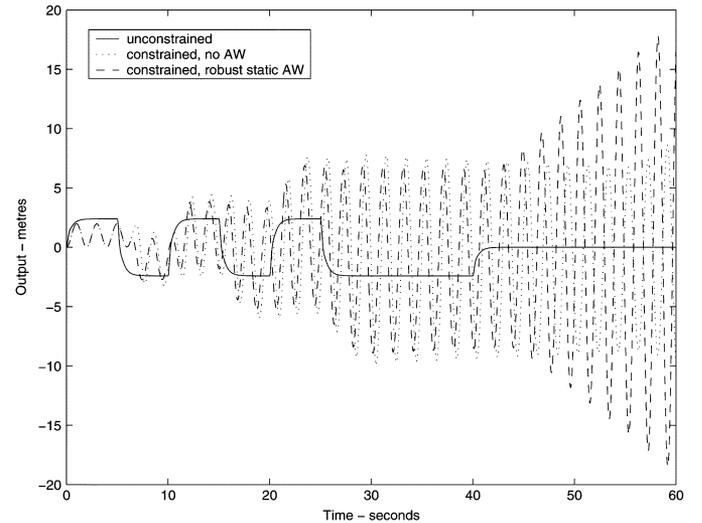


Fig. 14. Response of perturbed system with robust static AW: large reference demand.

antiwindup also causes instability. The conservatism of the local sector bound should also be mentioned: although in theory the local results only hold for $|u| < 2.5$, in fact they appeared to hold for $|u| < 4000$; only when this was exceeded (as in the case for a reference demand of 2.4) were stability problems encountered.

D. Other Antiwindup Compensators

As discussed in Section IV-D, the full-order antiwindup compensation method obtained by setting $W_r = 0$ and solving the LMI (23) can, in a certain sense, provide a robust antiwindup solution. This type of antiwindup solution is the continuous time counterpart of the discrete-time full-order compensator described in [9]. We designed a full-order antiwindup compensator according to this approach, choosing $W_p = 1$ and $W_r = 0.0001 \approx 0$ and then solving the LMI (23). The optimal gain matrix F was given by

$$F = [-1.3138 \quad -0.1424] \times 10^4, \quad \gamma \approx 1. \quad (45)$$

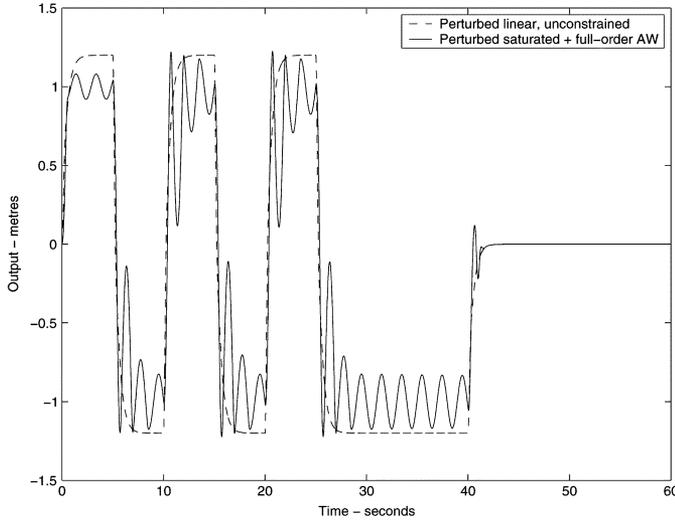


Fig. 15. Response of perturbed system with full-order antiwindup.

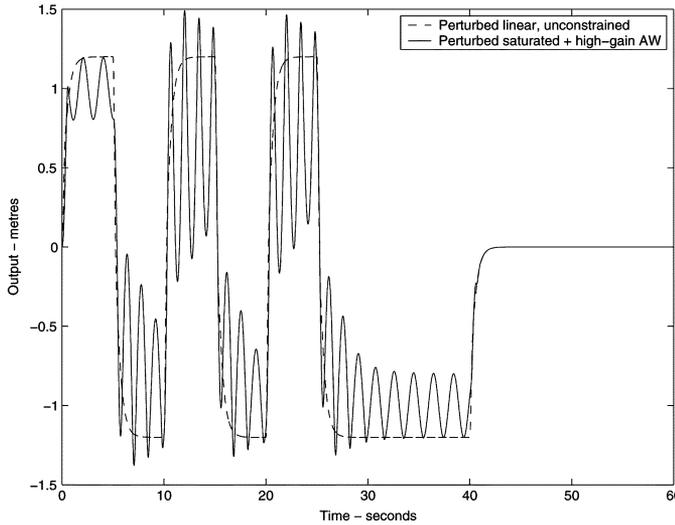


Fig. 16. Response of perturbed system with high-gain antiwindup.

It is important to realize that, in this case, the *guaranteed* robustness margin of the system is now given by $\gamma/\sqrt{W_r} \approx 100$, although this appears to be a conservative estimate. Again note that this places the poles of the AW compensator far into the left half-plane and, therefore, this would require a fast sampling frequency for correct implementation. Fig. 15 shows the response of the saturated system using this compensator. The stable response is indicative of the scheme's intrinsic robustness properties, although it does appear to be more oscillatory than that of the robust antiwindup compensator.

For purposes of comparison and due to its industrial popularity, we also tested a high-gain antiwindup compensator [13], [14]. In terms of Fig. 7, we chose $\Theta = [0 \ 14]'$. No formal performance or stability guarantees accompany this scheme. Nevertheless the response of the unperturbed system using this antiwindup scheme is very similar to the response with the optimal static scheme (Fig. 10). When this compensator was tested on the perturbed plant (Fig. 16) stability was also

obtained, although we note the very oscillatory response, which may be indicative of poor robustness. Its response was certainly not as good as that of the robust antiwindup compensator shown in Fig. 12, although it was obviously superior to that of the optimal static scheme shown in Fig. 11.

VII. CONCLUSION

This paper has introduced a framework for synthesizing robust antiwindup compensators for open-loop plants subject to additive uncertainties. The problem was posed in a similar way to that of linear \mathcal{H}_∞ control theory, and its solution appears as a set of LMIs of a similar type to those proposed in [8]. The attractive feature of the proposed solution is that the class of uncertainties considered are those that are routinely used by control practitioners, which underlines the practical relevance of the results. As an important aside, we have also demonstrated the optimal robustness of the much-denigrated IMC antiwindup strategy.

We note that many of the simple antiwindup schemes also seem to be quite robust in practice. For example, the Hanus scheme [12], which is a particular type of *static* AW compensator, has been the practitioners' method of choice for some time ([14] and [29] contain examples). Using the static results contained in this paper, it is possible to analyze the robustness of such a scheme, albeit typically locally. As the results here use the framework of [7], in which most linear AW schemes can be interpreted, the robustness results here should find wide applicability.

Finally, it should be mentioned that recently the so-called weakened antiwindup scheme has been introduced by Galeani *et al.* (see [30]), which promises to offer improved robustness to saturated systems but at the expense of slightly adjusting the linear loop. This should reveal further avenues of research.

APPENDIX I

PROOF THAT LINEAR ROBUSTNESS IS OPTIMAL

In order to prove that linear robustness is optimal, we need to prove that the induced norm of the nonlinear operator $\mathcal{T}_r : u_{\text{lin}} \mapsto z_\Delta$ can never be less than unity. i.e.,

$$\|\mathcal{T}_r\|_{i,2} = \sup_{0 \neq u_{\text{lin}} \in \mathcal{L}_2} \frac{\|u_{\text{lin}} - M\mathcal{F}(u_{\text{lin}})\|_2}{\|u_{\text{lin}}\|_2} \geq 1. \quad (46)$$

To see that this is the case, assume the opposite: $\|\mathcal{T}_r\|_{i,2} < 1$. Now choose a signal $u_{\text{lin}}(t) \in \mathcal{L}_2$ such that $u_{\text{lin}}(t) \leq \bar{u} \forall t$; such a signal always exists for $\bar{u}_i > 0$. Then it follows that $\mathcal{F}(u_{\text{lin}}) = 0 \forall t$ and thus that

$$\|\mathcal{T}_r\|_{i,2} = \sup_{0 \neq u_{\text{lin}} \in \mathcal{L}_2} \frac{\|u_{\text{lin}}\|_2}{\|u_{\text{lin}}\|_2} = 1. \quad (47)$$

Thus we have a contradiction and hence $\|\mathcal{T}_r\|_{i,2} \geq 1$.

APPENDIX II

STATE-SPACE MATRICES FOR ROBUST STATIC ANTIWINDUP

Given a controller $K(s) = [K_1(s) \ K_2(s)] \sim (A_c, [B_{cr} \ Bc], C_c, [D_{cr} \ Dc])$ and plant $G_2(s) \sim$

$$\begin{bmatrix} \bar{x} \\ \tilde{u} \end{bmatrix}' \begin{bmatrix} \bar{C}'_{1,i} \bar{C}_{1,i} & \bar{C}'_{1,i} (D_{01} + \bar{D}_1 \Theta)_i \\ (D_{01} + \bar{D}_1 \Theta)'_i \bar{C}_{1,i} & (D_{01} + \bar{D}_1 \Theta)'_i (D_{01} + \bar{D}_1 \Theta)_i \end{bmatrix} \begin{bmatrix} \bar{x} \\ \tilde{u} \end{bmatrix} \leq \beta_i^2 \bar{u}_i^2 \quad \forall i \quad (51)$$

$$\begin{bmatrix} \bar{x} \\ \tilde{u} \end{bmatrix}' \begin{bmatrix} \bar{C}'_{1,i} \bar{C}_{1,i} & \bar{C}'_{1,i} (D_{01} + \bar{D}_1 \Theta)_i \\ (D_{01} + \bar{D}_1 \Theta)'_i \bar{C}_{1,i} & (D_{01} + \bar{D}_1 \Theta)'_i (D_{01} + \bar{D}_1 \Theta)_i \end{bmatrix} \begin{bmatrix} \bar{x} \\ \tilde{u} \end{bmatrix} \leq \beta_i^2 \bar{u}_i^2 \bar{x}' P \bar{x} / c_{\max} \quad \forall i \quad (52)$$

$$\bar{x}' P \bar{x} - \frac{c_{\max}}{\beta_i^2 \bar{u}_i^2} \begin{bmatrix} \bar{x} \\ \tilde{u} \end{bmatrix}' \begin{bmatrix} \bar{C}'_{1,i} \bar{C}_{1,i} & \bar{C}'_{1,i} (D_{01} + \bar{D}_1 \Theta)_i \\ (D_{01} + \bar{D}_1 \Theta)'_i \bar{C}_{1,i} & (D_{01} + \bar{D}_1 \Theta)'_i (D_{01} + \bar{D}_1 \Theta)_i \end{bmatrix} \begin{bmatrix} \bar{x} \\ \tilde{u} \end{bmatrix} \geq 0 \quad \forall i \quad (53)$$

$$\begin{bmatrix} P & & & \bar{C}'_{1,i} \\ * & 2X + X\mathcal{A}(D_{01} + \bar{D}_1 \Theta) + (D_{01} + \bar{D}_1 \Theta)' \mathcal{A}X & & \bar{C}'_{1,i} \\ * & & * & \bar{C}'_{1,i} \\ * & & & \beta_i^2 \bar{u}_i^2 / c_{\max} \end{bmatrix} \geq 0 \quad \forall i \quad (54)$$

(A_p, B_p, C_p, D_p) , the state-space matrices for the representation given in (30) are given by

$$\begin{aligned} \bar{A} &:= \begin{bmatrix} A_p + B_p \tilde{\Gamma} D_c C_p & B_p \tilde{\Gamma} C_c \\ B_c \tilde{\Gamma} C_p & A_c + B_c \tilde{\Gamma} D_p C_c \end{bmatrix} \\ B_0 &:= \begin{bmatrix} B_p \tilde{\Gamma} \\ B_c \tilde{\Gamma} D_p \end{bmatrix}, \quad \bar{B} := \begin{bmatrix} B_p \tilde{\Gamma} & -B_p \tilde{\Gamma} D_c \\ B_c \tilde{\Gamma} D_p & -B_c \tilde{\Gamma} \end{bmatrix} \\ \Theta &:= \begin{bmatrix} \Theta_1 \\ \Theta_2 \end{bmatrix} \\ \bar{C}_1 &:= [\tilde{\Gamma} D_c C_p \quad \tilde{\Gamma} C_c], \quad D_{01} := \tilde{\Gamma} D_c D_p \\ \bar{D}_1 &:= [I + \tilde{\Gamma} D_c D_p \quad -\tilde{\Gamma} D_c] \\ \bar{C}_2 &:= [\Gamma C_p \quad \Gamma D_p C_c], \quad D_{02} := \Gamma D_p \\ \bar{D}_2 &:= [\Gamma D_p \quad -\Gamma D_p D_c] \end{aligned}$$

where $\Gamma := (I - D_p D_c)^{-1}$ and $\tilde{\Gamma} := (I - D_c D_p)^{-1}$.

APPENDIX III

GUARANTEEING A REGION OF ASYMPTOTIC STABILITY

Essentially the ideas follow from [23] and [25] or [24], the main difference being that we consider loops with direct feedthrough terms. First assume that $\bar{x} \in \mathcal{E}$ is small enough such that $Dz(\cdot) \in \text{Sector}[0, \mathcal{A}]$. This implies

$$|u_i| = |\bar{C}_1 \bar{x} + (D_{01} + \bar{D}_1 \Theta) \tilde{u}|_i \leq \beta_i \bar{u}_i \quad \forall i \quad (48)$$

which is equivalent to (51). Again, because $\bar{x} \in \mathcal{E}$, it follows that

$$\bar{x}' P \bar{x} / c_{\max} \leq 1 \quad \forall \bar{x} \in \mathcal{E}. \quad (49)$$

Thus it follows that (51) holds for all $\bar{x} \in \mathcal{E}$ if (52) holds, which after rearrangement is equivalent to (53). Rewriting the sector condition as

$$2\tilde{u}' X [\tilde{u} + \mathcal{A} \bar{C}_1 \bar{x} + \mathcal{A}(D_{01} + \bar{D}_1 \Theta) \tilde{u}] \leq 0 \quad (50)$$

and using the S-procedure, a sufficient condition for (53) to hold (and therefore for (51) to hold) is, after applying the Schur complement, obtained as shown in (51)–(54) at

the top of the page. Using the congruence transformation $\text{diag}(P^{-1}, X^{-1}, I) = \text{diag}(Q, U, I)$, we arrive at (37).

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