TRACKING WITH BOUNDED ACTUATORS: SCHEDULED CONTROLLERS

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Abstract: We extend a class of scheduled controllers, that were originally developed to address actuator saturation in the disturbance attenuation problem, to the problem of tracking of generally unknown signals. The main assumption is that some (possibly conservative) bounds be known for peak magnitude and rate of the tracking signal. Both state feedback for and static output feedback controllers are presented, though the choice of the approach to obtain the latter is left to the user. The solvability conditions and effectiveness are shown through an example.

Keywords: Saturation, tracking, scheduled controllers

1. INTRODUCTION

We extent the results of Kose and Jabbari (2003), which concerned design of scheduled controllers for disturbance attenuation with bounded actuators, to tracking problems. There is a large body of work in set-point tracking or tracking signals that are generated by known (or partially known) dynamics, often leading conditions on optimality, perfect tracking etc. Given space limitations, a proper review of literature is not feasible, but Lin, et al (1998) and Stoorvogel, et al (1999), Liu, et al. (2001), and their references can be consulted for much of the relevant literature.

Here, we focus on tracking generally unknown time-varying signals, as long as there are some possibly conservative bounds for their peak magnitude and peak rate. The objective is to design controllers that reduce the norm (e.g., peak) of the error between the output and the reference signal, since perfect tracking is not typically possible due to the unknown generating dynamics of the reference signal, such as the operator's stick commands. To reduce the conservatism (associated with the worst case nature of results, uncertainty on the bounds, saturation limit constraint, etc) we employ a scheduling scheme in which a family of controllers are obtained beforehand and, during the operation, the most aggressive controller feasible is used.

The basic idea behind scheduling is same as Gutman and Hagander (1985) and Kose and Jabbari (2003); when the relevant states (used in the control law) are large, small gains are unavoidable, but as these states get smallers, larger and more aggressive gains are used. To make scheduling more effective in tracking problems, the controller can be based on the error signal, since some of the states do not go to zero even when the reference signal is tracked perfectly. Such a controller is of particular interest in the case of bounded

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actuators. Also, the reference signal often enters the system as a external and unknown signal. The performance guarantees thus depend on a worst case description of the reference signal; i,e., modest reference signals (in magnitude and rate) might lead to lower controller gains and lower performance guarantees. The peaks (or worst) bounds are used to establish a stable closed loops with some performance guarantees even for the worst case, though such a reference signal (and thus the controllers) might never be encountered.

By incorporating the error signal into the state vector, the derivative of the reference signal enters the model explicitly, allowing a more transparent relationship between performance guarantees and potential bandwidth of the signal. Also, when only the error signals are used in the controller, the resulting controller will appear as a static output feedback (SOF) structure. While a general dynamic compensator is possible, here we focus on scheduling a family of state feedback problems and its extension to developing a family of static output feedback controllers. Finding optimal SOF controllers remains an unsolved problem, and a host of ah-hoc techniques have been proposed. For our purposes, we need a family of increasingly aggressive controllers, keyed to a Lyapunov function, and the specific technique used to obtain these controllers is not critical. The proposed approach is applied to an example. The results are compared with state feedback case, to evaluate the particular SOF technique used, a few extensions are discussed as well.

2. PRELIMINARIES

We first re-write the model so that the tracking error is part of the state vector, as in Krikelis and Barkas (1984) or Tarbouriech et al (2000), where it was used for anti-windup designs for the set-point tracking problem. Consider the following transformed state space model

$$\begin{cases} \dot{\hat{x}} &= A\hat{x} + B_2 u \\ y &= C\hat{x} = [I \;\; 0]\hat{x} = \hat{x}_1 \\ e &= y - y_r = \hat{x}_1 - y_r \end{cases} \tag{1}$$

Next, we change the state vector as the following

$$x = \begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \end{bmatrix} - \begin{bmatrix} y_r \\ 0 \end{bmatrix} = \begin{bmatrix} \hat{x}_1 - y_r \\ \hat{x}_2 \end{bmatrix} = \begin{bmatrix} e \\ \hat{x}_2 \end{bmatrix} \tag{2}$$

which leads to the state space model:

$$\begin{aligned} \dot{x} &= Ax + B_2 u - \begin{bmatrix} \dot{y}_r \\ 0 \end{bmatrix} + \begin{bmatrix} A_{11} \\ A_{21} \end{bmatrix} y_r \\ &= Ax + B_2 u + \begin{bmatrix} A_{11} & -I \\ A_{21} & 0 \end{bmatrix} \begin{bmatrix} y_r \\ \dot{y}_r \end{bmatrix} \\ &= Ax + B_2 u + B_1 w \end{aligned} \tag{3}$$

where $w^T(t) = [y_r^T(t) \; \dot{y}_r^T(t)]$. For simplicity, we discuss the single input case only. The generalization to the multi-input case is immediate. Note that this model has the tracking error as state, which facilitates a more desirable scheduling. Also, the rate of change in the command explicitly enters the model, which will allow us to study the effect of this derivative (related to the bandwidth of the command signal) on the overall performance of the controller. Naturally, we assume that saturation bound, u_{lim} is known.

Our main interest is to obtain desirable performance, in terms of L_2 gain or peak to peak gain from this w to the following controlled output:

$$z = Cx = e = y - y_r \tag{4}$$

for the class of disturbances with known peak or energy bounds. For brevity, we will focus on generally unknown signals with known bounds as in

$$w^T(t)\, w(t) \leq w_{max}^2. \tag{5}$$

The bound concerns the $\mathbb{R}^n$ norm of the combination of y_r and $\dot{y}_r$ (i.e., $w^T(t) = [y^T(t) \; \dot{y}_r^T(t)]$). As such, it allows a more realistic trade-off; i.e., tracking signals with larger rates (e.g., bandwidth) effectively increases the size of w_{max}, with a commensurate impact on the overall performance. As shown below, when $\dot{y}_r$ is available in real time, compensators can establish different performance guarantees depending on how much of w_{max} is from y_r or from $\dot{y}_r$.

3. STATE FEEDBACK

We start with state feedback controllers in part to discuss different controller structures with most clarity, and in part to establish the upper limits of performance that can be obtained. For this, consider the following controller structure

$$u = Kx + Hw = Kx + Hm\, y_r + Hd\, \dot{y}_r \tag{6}$$

where the most general case; i.e., when $Hm \neq 0$ and $Hd \neq 0$ requires availability of both y_r and $\dot{y}_r$, on-line. While y_r is often available, it is unlikely that $\dot{y}_r$ can be used in most applications. As a result, in most cases we discuss different controller structures separately. In certain unconstrained and single objective problems (e.g., the basic L_2 gain problem) the guaranteed bounds are not improved with the use of the Hw term. As shown below, the constrained nature of the problem (and the fact that many of the problems become multi-objective) results in better performance when the Hw term is used in the controller.

With this we can state the preliminary and background material needed for the remainder of this paper. Given the bound given in (5), an easy way to obtain estimate for the invariant set is through a Lyapunov function of the form

$$V(x) = x^T P x \qquad (7)$$

and making sure that its derivative, taking along the path of the motion, satisfies

$$\dot{V} + \alpha(V - w^T w) < 0 \qquad (8)$$

for some scalar $\alpha > 0$, which would imply that $V(x) \leq w_{max}^2$ is an invariant set for the (closed loop) of the system. It is well known that (8) is established by

$$\begin{pmatrix} PA_{cl} + A_{cl}^T P + \alpha P & PB_{cl} \\ B_{cl}^T P & -\alpha I \end{pmatrix} < 0 \qquad (9)$$

where A_{cl} and B_{cl} are the closed loop matrices resulting from (3) and (6), and α is scalar whose choice is optimized thought a simple linear search. Along with the invariant set inequality, the following matrix inequalities are used to ensure that saturation bounds are not violated in the invariant set. For the simple case of $u = Kx$, we use

$$\begin{bmatrix} P & K^T \\ K & \frac{u_{lim}^2}{w_{max}^2} \end{bmatrix} > 0 \qquad (10)$$

while for $u = Kx + Hw$, we use

$$\begin{pmatrix} \sigma P & 0 & K^T \\ 0 & (1-\sigma)I & H^T \\ K & H & \frac{u_{lim}^2}{w_{max}^2} I \end{pmatrix} > 0 \qquad (11)$$

for some positive scalar σ. For brevity, we establish the claim for (11), since (10) can be considered a special case. First, we apply the Schur formula to (11) to get

$$\begin{pmatrix} \sigma P & 0 \\ 0 & (1-\sigma)I \end{pmatrix} - \begin{pmatrix} K^T \\ H^T \end{pmatrix} \frac{w_{max}^2}{u_{lim}^2} \; [K \;\; H] > 0 \qquad (12)$$

and pre and post multiplying (12) with $[x \;\; w]$ we get

$$\sigma x^T P x + (1-\sigma) w^T w > \frac{w_{max}^2}{u_{lim}^2} \; |u(t)|^2 \qquad (13)$$

which in light of having $V = x^T P x \leq w_{max}^2$ and (5) becomes

$$u_{lim} > |u(t)|.$$

If we had information regarding the portion of w_{max}^2 that is from the rate terms (i.e., η in $\dot{y}_r^T \dot{y}_r \leq (1-\eta) w_{max}^2$), then we could replace the middle diagonal block of (11) by

$$\begin{pmatrix} \frac{1-\sigma-\sigma_1}{\eta} & 0 \\ 0 & \frac{\sigma_1}{1-\eta} \end{pmatrix}$$

and use σ_1 to improve the results (with minimal additional burden). Naturally, if $\dot{y}_r$ is not available in either cases and we use $u = Kx + Hm\, y_r$, we simply replace H in (11) - or its modified form if we have η - with $[Hm \; 0]$ and reduce its dimension.

The state feedback problem is then attempted through the standard approach; i.e., congruent transformations that result in a set of matrix inequalities in terms of $Q = P^{-1}$ and $Y = KP^{-1}$. Depending on the performance index and the assumptions on the tracking signal, the search is either convex or close to it (e.g., a convex search modulo one or two scalar variables such as α or σ above). Performance can be ensured by adding the appropriate matrix inequality. For example, L_2 can be minimized by adding the corresponding bonded real inequality, which results in a multi-objective problem (see below).

4. STATIC OUTPUT FEEDBACK

Since the error $y - y_r$ is a part of the state vector, and it often the main objective, it would be desirable to design a compensator that only uses this part of the state vector. As discussed below, this is even more desirable for scheduled controllers. This leads to a typical static output feedback (SOF) structure. As mentioned earlier, this is often a hard problem for which a variety of potentially conservative approaches are available. Here, we suggest an approach that can be used in conjunction with many of these techniques. We consider control laws of the form

$$u = Ke + Hm y_r + Hd\, \dot{y}_r \qquad (14)$$

where Hm and/or Hd can be zero in different cases. Without any loss of generality, we partition the main Lyapunov matrix according to the dimension of e ; i.e., $x^T = [e^T \;\; \hat{x}_2^T]$

$$Q = \begin{bmatrix} S & SN \\ N^T S & R \end{bmatrix} \Rightarrow P = Q^{-1} = \qquad (15)$$

$$\begin{bmatrix} S^{-1} + N(R - N^T S N)^{-1} N^T & -N(R - N^T S N)^{-1} \\ -(R - N^T S N)^{-1} N^T & (R - N^T S N)^{-1} \end{bmatrix}$$

Following standard projection, we have:

$$x_{cl} \in \{x_{cl} : x_{cl}^T P x_{cl} \leq w_{max}^2\} \Rightarrow \qquad (16)$$

$$e \in \mathcal{E}(S^{-1}, w_{max}^2) = \{e : e^T S^{-1} e \leq w_{max}^2\} \qquad (17)$$

where $e = x_1 = y - y_r$. As before, we will use a reachable/invaraint matrix inequality that establishes $\{x : x^T P x \le w_{max}^2\}$ which means

$$e^T S^{-1} e \le w_{max}^2, \quad \text{for} \quad w^T(t)w(t) \le w_{max}^2 \qquad (18)$$

which will be combined with a performance inequality and a constraint inequality to enforce the saturation bound.

Case i: $u = Ke$: We use

$$u = Ke = YS^{-1}\, e \qquad (19)$$

for which we can avoid saturation inside of $\mathcal{E}(S^{-1}, w_{max}^2)$ if

$$\begin{bmatrix} S & Y^T \\ Y & \dfrac{u_{lim}^2}{w_{max}^2} \end{bmatrix} > 0 \qquad (20)$$

Case ii: $u = Ke + Hw$: We use

$$u = YS^{-1}e + [Hm \;\; Hd] \begin{pmatrix} y_r \\ \dot{y}_r \end{pmatrix} = YS^{-1}e + Hw$$

which gives the closed loop system

$$\dot{x} = [A + B_2K(I\ 0)]x + (B_1 + B_2H)w \qquad (21)$$

Using the same logic as before, it is straight forward to establish this sufficient condition for avoiding actuator saturation in the invariant set:

$$\begin{pmatrix} \sigma S & 0 & Y^T \\ 0 & (1-\sigma)I & H^T \\ Y & H & \dfrac{u_{lim}^2}{w_{max}^2} I \end{pmatrix} > 0 \qquad (22)$$

All other variations; e.g. having $Hd = 0$ or using η to bound the portion of y_r (e.g., $y_r^T y_r \le \eta w_{max}^2$), follow similarly and are not repeated due to space limitations. Generally, the search for the variables are not convex and difficult. Often some form of trial and error or sufficient condition can be used to obtain solution (see Fujimori, 2004). For example, in the structure used above, one can use a typical relaxation algorithms and iterate over variable N and the rest of the variables (though other techniques can also be used with ease).

5. SCHEDULING

Assuming a conservative bound for the disturbance, assuring results for the worse case or use of static output feedback often entails significant conservatism. Along the lines of Kose, Jabbari, (2003), we use scheduling to use a more aggressive controller, when the relevant states (that are used in the controller) are small and larger gains are possible. We start with state feedback. The steps are the following: (i) Obtain a family of nested ellipsoids: $\mathcal{E}(P, 1/r_i) = \{x : x^T P x \le \frac{1}{r_i} w_{max}^2\}$, with $1 = r_1 < \cdots < r_l$, and establish the largest set $\mathcal{E}(P, 1/r_1)$ as the invariant set, (ii) In each ellipsoid, obtain controller gains (e.g, K_i, H_i) to minimize the performance in the ellipsoid i, while guaranteeing that $|u(t)| \le u_{lim}$. Note that the performance will be better for larger values of r_i.

Since the technical details are beyond the space limitations, we only discuss the salient features of the proposed scheduling. The inequality in (9), to establish the invariant set, along with one of the constraint inequalities (e.g., (11)) and a performance inequality are used for the largest ellipsoid with lowest gain controller K_1 (with lowest guaranteed performance) corresponding to the worst case. Then for each smaller ellipsoid, two (additional) matrix inequalities are used, one for performance and one to enforce the saturation limit. For the state feedback problem this is sufficient since by measuring the state, we can ensure that K_1 – which establishes the invariant set – is used when the state enters the largest ellipsoid.

For the static output feedback, many of the same steps follow: we establish a large invariant set associated with the worst case performance (as in previous section) and divide it into a family of nested ellipsoids. For each smaller ellipsoid, we solve a performance and a constraint inequality associated with the SOF structure. For example, the L_2 gain inequality is

$$\begin{pmatrix} AQ + B_2K_iCQ + * & B_1 + B_2H_i & QC^T \\ B_1^T + H_i^T B_2^T & -\gamma_i I & 0 \\ CQ & 0 & -\gamma_i I \end{pmatrix} \qquad (23)$$

where " $*$ " denotes symmetric half and given the structure of C and Q, the term BK_iCQ can be written as $C = B_2Y_i[I\ N]$. The scheduling, however, is now based on e and not x; i.e., we find controllers that do not saturate in $\mathcal{E}(S^{-1}, 1/r_i) = \{e : e^T S^{-1} e \le \frac{1}{r_i} w_{max}^2\}$. The constraint inequality corresponding to (22) is thus

$$\begin{pmatrix} r_i\ \sigma S & 0 & Y_i^T \\ 0 & (1-\sigma)I & H_i^T \\ Y_i & H_i & \dfrac{u_{lim}^2}{w_{max}^2} I \end{pmatrix} > 0 \qquad (24)$$

where larger r_i relaxes the constraint allowing for lager gain which in turn can improve the performance. Unlike state feedback, however, we need to ensure that none of these controllers violate the invariant set. This can be accomplished either by using the invariant LMI for each controller (as was done in Kose and Jabbari (2003)), or by a standard application of S-procedure, which is often less conservative. Much of the details is omitted due to space limitations.

The search for variables can be accomplished through a single stage, combining all inequalities and unknown variables, or through a sequential approach where the low gain (outer ellipsoid) problem is solved for P and K_1, for example, and the variables associated with inner ellipsoids are obtained one ellipsoid at a time. In general the sequential approach is somewhat more conservative, but has considerably lower computational burden (though the specific tradeoffs often depends on which controller structure is used). This is the approach used in the example below.

Remark 1. For simplicity, we have focused on the case of constant $P = Q^{-1}$, though general parameter varying $Q(r)$, as in Kose and Jabbari (2003) is relatively straight forward (albeit with a large increase in notational complexity). Similarly, we present results for the simplest case in which the controller is switched from one gain to another. A smooth version can be used with ease following the technique used in Kose and Jabbari (2003), in which the controller is a continuous linear spline function using K_i as basis or 'knots'. The notation, however, is rather involved and is thus omitted here.

Remark 2. A variety of techniques can be used to obtain the solution to the SOF problem (e.g., those used in Prempain and Postlethwaite (2001), Crusius and Trofino (1999), etc). For the results here, we only solve one SOF problem for the low gain matrix. Once this is accomplished, we essentially keep the main Lyapunov matrix constant, and solve for the gains in smaller ellipsoid sequentially. Beyond the first step, the problem is convex and easy.

6. EXAMPLE

Du to space considerations, we consider a simple second order system as in

$$\dot{x} = \begin{bmatrix} 2 & 1 \\ 0 & 1 \end{bmatrix} x + \begin{bmatrix} 0.5 \\ 1 \end{bmatrix} u \qquad (25)$$

with $u_{lim} = 15$, and $w^T(t)w(t) \leq w_{max}^2 = 1$.For brevity, we denote $u = Kx$ as case (i), $u = Kx + Hw$ as case (ii) and $u = Kx + Hmy_r$ as case (iii). When attempting static output feedback, similar notation will be used (i.e., $u = Ke$ as case (SOF-i), etc). The results for the state feedback and static output feedback, for the LTI cases, are presented in Table 1.

Table 1 L_2 gain from w to e (γ) : LTI cases

Case	(i)	(ii)	(iii)
State Feedback	0.1488	0.0023	0.0341
SOF	0.1861	0.0192	0.1024

The best results are where $\dot{y}_r$ is also available, though in almost all cases this signal will be unavailable and the other two cases are more realistic. Also note that case (iii) does not require any new measurements (x or e already assumes that y_r is available) and the significant improvement is due to extra flexibility in choice of gain matrices, though it involves another scalar variable that enters the matrix inequalities often nonlinearly (i.e., σ). As a result, we focus on the state feedback and output feedback forms of cases (i) and (iii).

We next attempt the scheduling approach discussed earlier. For brevity, we show results for 2 cases. The first row in Table 2 shows the results of the the simplest form; i.e., state feedback of the form $u = Kx$, while the second row corresponds to the controller of the form $u = Ke + Hmy_r$. Here, by L_2 gain of a controller we mean the energy gain of the closed loop system if the same controller was used throughout (- otherwise a time averaged from. i.e., $\int z^T z\, dt \leq \int \gamma(t) w^T w\, dt$ applies). Notice the significant improvement in the value of the L_2 gain for larger r_i (i.e., smaller ellipsoids), while the gains are. Often, the gains associated to x_2 are considerably smaller than those for e. This motivates finding controllers that use e only, even if the whole state vector is available. This leads to the results on the second row of the table.

Table 2 L_2 gain from w to e (γ_i) : Scheduled controllers

r_i	1	5	25	125
SF-Case (i)	0.149	0.096	0.061	0.061
SOF-Case (iii)	0.102	0.071	0.048	0.031

Finally, we show a representative sample for the controller performance. Consider the the reference signal $y_r = 0.56[sin(t) + sin(0.5t)]$ (which meets $w^T(t)w(t) \leq 1$). Figure 1 shows the tracking error for the two controllers in Table 2 and the non-scheduled form of Case (i) of Table 1. We show the tracking error since $y(t)$ is hard to distinguish from y_r. Figure 2 shows the index of the controller. Here index i refers to the gain k_i, which is used in the ellipsoid which is $\frac{1}{r_i}$ times smaller than the invariant ellipsoid. For these simulations, we have used $r_i = 5^{i-1}$. Note that once the peak error is reduced (due to slowing of $\dot{y}_r$ near the peaks of of y_r), the controller switches to higher gains to improve the performance. Also, since the reference signal is a smooth signal, the state vector is never in the largest set which has index of 1.

7. INTEGRATOR

The case of $y_r = const$ often receives special attention. Almost universally, an integrator is included in the controller to establish zero steady sate error. In the following, we incorporate an integrator to obtain a controller that ensures zero

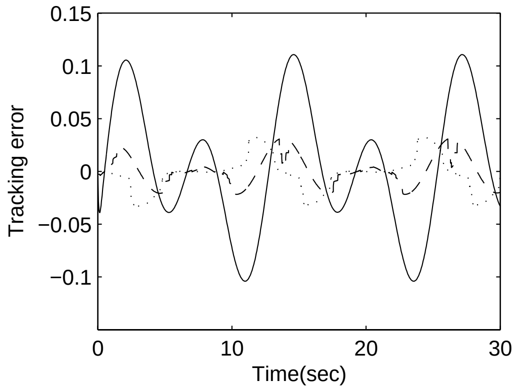

Fig. 1. Tracking errors: FS-nonscheduled (solid), FS-scheduled (dashed), SOF-scheduled (dotted)

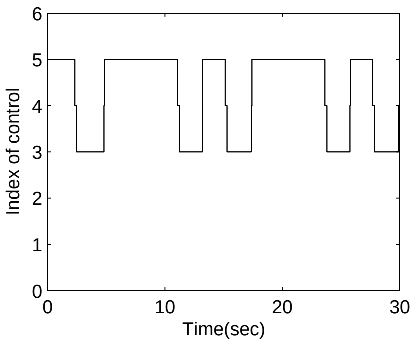

Fig. 2. Control Index for Case SOF-(iii)

steady state error when y_r is constant, while it also provides desirable performance (e.g., low L_2 gain) for time varying reference signal. We augment the model with an integrator

$$\begin{aligned} \dot{q} &= e \\ \dot{x} &= Ax + B_2 u + B_1 w \end{aligned} \tag{26}$$

where, as before, $x^T = [e^T \quad \hat{x}_2^T]$. This gives

$$\begin{pmatrix} \dot{q} \\ \dot{e} \\ \dot{x}_2 \end{pmatrix} = \begin{pmatrix} 0 & I & 0 \\ 0 & A_{11} & A_{12} \\ 0 & A_{21} & A_{22} \end{pmatrix} \begin{pmatrix} q \\ e \\ x_2 \end{pmatrix} + \begin{pmatrix} 0 \\ B_{21} \\ B_{22} \end{pmatrix} u + \begin{pmatrix} 0 & 0 \\ A_{11} & -I \\ A_{21} & 0 \end{pmatrix} w$$

with both e and q are assumed to be measured (though assuming q available is not critical); i.e.,

$$C = \begin{bmatrix} I & 0 & 0 \\ 0 & I & 0 \end{bmatrix} \tag{27}$$

where we try to have a controller of the kind

$$u = K_I q + K_P e + K_x x_2 + Hm y_r + Hd \dot{y}_r \tag{28}$$

for the full state feedback. For static output feedback we set $K_x = 0$ Note that the model above include the whole vector w (with $w^T = [y_r^T \ \dot{y}_r^T]$). This allows us to evaluate this approach for tracking signals that may not be constant all the times (and for constant y_r we can simply remove the terms associated with $\dot{y}_r$).

REFERENCES

Acikmese, A.B., Corless, M., (2002), "Robust output tracking for uncertain/nonlinear systems subject to almost constant disturbances," *Automatica*, vol 38, (2002), pp 1919-26.

Cao, Y.Y., Lin, Z., Ward, D.G., (2004) "Anti-windup desgin of output tracking systems subject to actuator saturation and constant disturbances," *Automatica*, vol 40, pp 1221-28.

Crusius, A.R., Trofino, A., (1999), "Sufficient LMI conditions for output feedback contorl problems," *IEEE Trans. on AC*, 1999, vol 44, no 5, pp. 1053-56.

Fujimori, A. (2004), "Static output feedback using substitutive LMI," *IEEE Trans. on AC*, vol 49, no 6, pp. 995-1000.

Gutman, P.O., Hagander, P., (1985) "A New Design of Constrained Controllers for Linear Systems," *IEEE Transactions on Automatic Control*, Vol.30, No.1, 1pp.22-33.

Henrion, D., Garcia, G., and Tarbouriech, S. (1998), " Piecewise-linear Robust Control of Systems with Input Constraints," *Proceedings of the ACC*, pp.3545-49.

Kappor, N., Teel, A.R., Daoutidis, P., (1998), "An Anti-windup Design for Linear Systems with Input Saturation," *Automatica*, Vol.34, No.5, pp.559-574.

Kose, I.E., Jabbari, F., (2003), " Scheduled controllers for linear systems with saturating actuator," *Automatica*, Vol. 39, No 8, pp. 1377-87.

Kothare, M., Campo, P.J., Morari, M., Nett, C.N., (1994), "A Unified Framework for the Study of Anti-windup Designs," *Automatica* Vol. 30, No. 12, 199, pp. 1869-1883.

Krikelis, N.J., Barkass, S.K., 1984, "Design if tracking systems subject to actuator saturation and integration wind-up," *Int. J. of Control*, 1984, vol. 39, no. 4, pp. 667-682.

Lin, Z., (1998),"Global Control of Linear Systems with Saturating Actuators," *Automatica*, Vol.34, No.7, 1998, pp.897-905.

Lin, Z., Pather, M., Banda, S. (1998), "Toward improvement of tracking performance - nonlinear feedback for linear systems," *Int. J. of Control*, 1998, vol. 70, no. 1, pp. 1-11.

Liu, K., Chen, B.M., Lin, Z., (2001), "On the problem of robust and prefect tracking for linear systems with extenal disturbances", *Int. J. of Control*, 2001, vol 74, No 2, pp 158-174.

Prempain, E., Postlethwaite, I., (2001), "Static output feedback stabilization with H_∞ performane for a class of plants, " *Systems and Control Letters*, Vol 43, pp 159-166.

Stoorvogel, A.A., Saberi, A., (1999), "Output Regulation of Linear Plants with Actuators Subject to Amplitude and Rate Constraints", *Int. Jour. of Robust and Nonlinear Control*, Vol.9, No.10, pp.631-657.

Tarbouriech, S. Pittet, C. Burgat, C., (2000), "Output tracking problem for systems with input saturation via nonlinear integrating actions," *Int. J. Robust and Nonlinear Contorl*, vol. 10, pp. 489-512.