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Asymptotic Stabilization of Delayed Systems with Input and Output Saturations

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Article Info	ABSTRACT
	We consider in this paper the problem of controlling an arbitrary linear delayed system with saturating input and output. We study the stability of such a system in closed-loop with a given saturating regulator. Using input-output stability tools, we formulated sufficient conditions ensuring global asymptotic stability.

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1. INTRODUCTION

The presence of constraints on the input, output or states in industrial systems is illustrated by technological limitations such as limit devices or adjustment elements. These constraints make the control problem more complicated. Most of research activities have been interesting in stabilizing delayed input linear systems during the last two decades [1-6]. The control problem of constrained output systems represents one of actuality subjects. It may have two different aspects. In the first one, the output is really technologically limited, the system model includes an essential nonlinear static element and the system dynamics must be nonlinear because of this limitation. In the second case, the output is just analytically limited but technologically the constraint can be violated. The constraint does not affect the system model and the dynamics can or can not be linear.

In this paper we are focusing on controlling linear delayed systems with both input and output saturation. In this situation, two main questions can be raised: how to develop a saturating regulator in order to stabilize a saturating delayed system? This issue is not yet solved. The second question is most important which is: is the closed-loop system of a given saturating delayed system and a saturating stabilizing regulator asymptotically globally stable?

A similar research activities on the stabilization of a specific class of state saturating systems were formulated by [7] and especially by [8] using linear constrained regulators but in the non-delayed systems case. The class of systems considered in the present paper and the one in [7] and [8] differ in the fact that we are interested in systems with delayed inputs and also the constraint enters the system model. In [8] and [7], the authors supposed that the states are all available which allows using state-feedback regulator, but in our paper we consider an output-feedback regulator because only the system output is available. Finally, we have to notice that the problems in the present paper and in [7] are quite different from the classical control

problem of unsaturating output systems in presence of input saturation [9]. Indeed, in the problem of controlling output saturating delayed systems one has three main features describing the closed-loop system with saturating regulator: (1) the system is nonlinear; (2) the closed-loop system may be asymptotically globally stable even if the system is open-loop strictly unstable at the origin, and (3) all the signals of the closed-loop are bounded. Notice that in the saturating input case the system output is not a priori bounded, for this, the system should not be open-loop strictly unstable. In this paper we are focusing this control problem of systems with saturating input and output. Sufficient conditions for global stability of the resulting closed-loop system are formulated using input-output stability tools [9] - [10].

These conditions will show that the closed-loop system is asymptotically globally stable although the system is strictly unstable. This paper is organized as follows: Section 2 is devoted to formulate the control problem; the controller design is described in Section3; the resulting closed-loop system is analyzed in Section 4; the corresponding stabilization performances are illustrated by simulation in Section 5.

2. CLASS OF CONTROL SYSTEM

The input-output representation of a saturating input-delayed system can be modeled as follows:

$$\hat{x}(s) = \left[1 - A(s)\right]\hat{y}(s) + B(s)e^{-s\tau}\hat{u}(s) \tag{1.1}$$

with $y(t) = sat(y_M, x(t))$ (1.2)

$$A(s) = s^n + a_{n-1}s^{n-1} + \cdots + a_1 s + a_0 \tag{2.1}$$

$$B(s) = b_{n-1}s^{n-1} + \cdots + b_1 s + b_0 \tag{2.2}$$

in presence of input constraint:

$$\left|u(t)\right| \le u_M \tag{3}$$

where $(u(t),y(t))$ are the system input and output and $(\hat{u}(s),\hat{y}(s))$ their Laplace transforms; u_M and y_M are two real positive constants.

It is further assumed that:
A1. $A(s)$ is Hurwitz polynomial,
A2. $(sA(s), B(s))$ are coprime.

Note that in the case of unconstrained output ($y_M = \infty$), the system is controllable with a linear state feedback. Also, $A(s)$ is not necessarily Hurwitz, i.e. the origin can be an unstable equilibrium.

From (1)-(2) it's easily seen that the system is can be represented around the origin by the linearized model:

$$A(s)\hat{y}(s) = B(s)e^{-s\tau}\hat{u}(s) \tag{4}$$

3. CLASS OF STABILIZING REGULATOR

The control design method is the finite spectrum assignment (FSA) which is an extension to time-delay systems case of the standard pole placement design technique. The starting step is an arbitrary choice, by the designer, of a pair of Hurwitz polynomials of the form:

$$\begin{aligned} C(s) &= s^n + c_{n-1}s^{n-1} + \ldots + c_1 s + c_0, \\ \Lambda(s) &= s^n + \lambda_{n-1}s^{n-1} + \ldots + \lambda_1 s + \lambda_0 \end{aligned} \tag{5}$$

There exists a pair of pseudo-polynomials $R(s)$ and $S(s)$ satisfying the Bezout equation:

$$sR(s)A(s) + S(s)B(s)e^{-s\tau} = C(s)\Lambda(s) \tag{6}$$

Following the pole placement technique $R(s)$ and $S(s)$ are the unique solution of the Bezout equation of the form:

$$R(s) = s^{n-1} + \sum_{i=0}^{n-2} R_i(e^{-s\tau})s^i + R_{-1}(s),$$

$$S(s) = \sum_{i=0}^{n} S_i(e^{-s\tau})s^i + S_{-1}(s)$$

where $R_{-1}(s)$ and $S_{-1}(s)$ belong to G, the set of transfer functions of distributed and punctual delay operators (Appendix A in [1]). For $i \geq 0$, $R_i(e^{-s\tau})$ and $S_i(e^{-s\tau})$ belong to $\mathbf{R}[e^{-s\tau}]$, the set of polynomials in $e^{-s\tau}$. Unlike the case of non-delayed systems, the (finite-degree) operators $R(s)$ and $S(s)$ are presently pseudo-polynomials and, consequently, are analytical functions of s.

As $\deg\left(S(s)B(s)e^{-s\tau}\right) \leq 2n-1$, it follows that deg $(sA(s)R(s))=\deg(C(s)\Lambda(s))=2n$ which implies that deg($R(s)$)=n-1, because $\deg(sA(s)) = n+1$ and furthermore as $sA(s)$ and $C(s)\Lambda(s)$ are monic, (i.e. their higher degree term coefficient equals 1).

With all the above notations, the saturated linear regulator is given the alternative form:

$$\hat{v}(s) = \frac{\Lambda(s) - sR(s)}{\Lambda(s)}\hat{u}(s) - \frac{S(s)}{\Lambda(s)}\hat{y}(s) \tag{7.1}$$

$$\hat{u}(s) = sat(\hat{v}(s)) \tag{7.2}$$

This defined regulator is determined by the choice of the polynomials C and Λ.

We are focusing on the following problem: given a delayed system (1-3) and a regulator (7), based on the choice of polynomials C and Λ, is the resulting closed-loop system globally asymptotically stable?

This problem is related to two issues:
a)   Does the stabilizing regulator, for a given system (1-3), exist?
b)   If it does, how can we design it?

To our knowledge these issues are not yet soved.

Remarks: (i) if we temporarily consider that the system (1-2) is not subject to the constraint (3), i.e. ($u_{M}=\infty$).

Then, the above defined regulator reduces to the standard regulator $\hat{u}(s) = -\dfrac{S(s)}{sR(s)}\hat{y}(s)$. If we have also

$y_M=\infty$, then the closed-loop system is transformed to a linear system whose poles are those of C(s).

(ii) From the above system and regulator, it follows that the signals v(t) and x(t) are bounded whatever the C(s) polynomial's choice. So from (1-2) and (7), it follows, for all t, that:

$$v_M = \max\left\{u_M, \left(\sum_{j=0}^{n-1}\left|\frac{r_j}{\lambda_j}\right|\right)u_M + \left(\sum_{j=0}^{n}\left|\frac{s_j}{\lambda_j}\right|\right)y_M\right\}$$

$$x_M = \max\left\{y_M, \left(\sum_{j=0}^{n-1}\left|b_j\right|\right)u_M + \left(\sum_{j=0}^{n}\left|a_j\right|\right)y_M\right\} \tag{8}$$

iii) Due to this structural boundedness, some unstable systems can be globally asymptotically stabilized. But in the case of unconstrained output, the signals are not a priori bounded and the system is globally stabilized only if its poles are all in the right half plane.

4.   CLOSED-LOOP SYSTEM ANALYSIS
First, let us point out a sector property for the saturation function ([9] page 417).
LEMMA 1 Consider an arbitrary positive real β and a real function Φ(β,.) defined as follows:

$$\Phi(\beta,z)=z - sat(\beta,z) \text{ for any real z} \tag{9.1}$$

then, for any $z \in [-z_M, z_M]$ and any $z_M > \beta$ one has:

$$0 \le z.\Phi(\beta, z) \le \alpha z^2 \text{ where } \beta = \frac{z_M - \alpha}{z_M} \tag{9.2}$$

which means that $\Phi(\beta,.)$ belongs to sector $[0\ \beta]$, when restricted to the interval $[-z_M, z_M]$.

The main result is described by the following theorem.

THEOREM 1 Consider the closed-loop control system consisting of system (1-2) submitted to assumptions A1 and A2 and the saturated regulator (7). Then, if one has:

$$\operatorname*{Re}_{0 \le \omega \le \infty} \left(\frac{A(j\omega)}{C(j\omega)} \right) < 1 - \frac{1}{\alpha_u} \text{ and} \tag{10.1}$$

$$\inf_{0 \le \omega \le \infty} \operatorname{Re} \left[\frac{\omega R_{im}(j\omega)}{C(j\omega)\Lambda(j\omega)} \right] < 1 - \frac{1}{\alpha_y} \tag{10.2}$$

$$\gamma_2 \left(\frac{S(s)B(s)}{\Lambda(s)C(s)^2} e^{-s\tau} \right) < \frac{1}{\gamma^0 \gamma^1} \tag{10.3}$$

then, all signals $v(t)$, $u(t)$, $x(t)$ and $y(t)$ belong to L_2.
where γ_p is the L_p-gain of an L_p-stable operator.

$$\alpha_u = \frac{v_M - u_M}{v_M}, \quad \alpha_y = \frac{x_M - y_M}{x_M}, \quad R(j\omega) = R_{re}(\omega) + jR_{im}(\omega) \tag{11.1}$$

$$\gamma_i = \gamma_2 \left(\frac{A(s) - C(s)}{C(s) + (A(s) - C(s)).\alpha_u / 2} \right) \tag{11.2}$$

$$\gamma_{ii} = \gamma_2 \left(\frac{R(s) - C(s)}{C(s) + (R(s) - C(s)).\alpha_y / 2} \right) \tag{11.3}$$

$$\gamma^0 = \alpha_u \cdot \frac{1 + \alpha_u \gamma_i / 2}{1 - \alpha_u \gamma_i / 2}, \quad \gamma^1 = \alpha_y \cdot \frac{1 + \alpha_y \gamma_{ii} / 2}{1 - \alpha_y \gamma_{ii} / 2} \tag{11.4}$$

In the sequel, the notations will be simplified by not writing explicitly the dependence on s of all polynomials and pseudo-polynomials. Also we'll avoid the symbol "^" for the Laplace transforms unless necessary. Thus, depending on the context, the letter x will be either the signal $x(t)$ or its Laplace transform.

PROOF:
Let us define these new errors:
$$\bar{v} = v - u, \bar{x} = x - y \tag{12}$$

By considering all the above notations, equations (1) and (7.1) are written as follows:

$$\bar{v} = -\frac{sR}{\Lambda}u - \frac{S}{\Lambda}y \tag{13.1}$$

$$\bar{x} = -Ay + Be^{-s\tau}u \tag{13.2}$$

Multiplying by-A both sides of (13.1), one has:

$$-A\bar{v} = \frac{sAR}{\Lambda}u + \frac{AS}{\Lambda}y$$

Now, operating S/Λ on both sides of (13.2) yields

$$\frac{S}{\Lambda}\bar{x} = -\frac{SA}{\Lambda}y + \frac{SB}{\Lambda}e^{-s\tau}u$$

Using (6), adding these two last equations gives

$$\frac{S}{\Lambda}\bar{x} - A\bar{v} = Cu \qquad (14)$$

Using the fact that $u = v - \bar{v}$ and rearranging terms, one has:

$$v = \frac{C-A}{C}\bar{v} + \frac{S}{C\Lambda}\bar{x}$$

(14) can be equivalently written as follows:

$$v = \frac{C-A}{C}\bar{v} + \frac{S}{C\Lambda}\bar{x} + \delta_1 \qquad (15)$$

where δ_1 is a transfer function of a signal arising from initial conditions.

As C and Λ are Hurwitz, δ_1 vanishes exponentially, which implies that $\delta_1 \in L_2$.

Operating $Be^{-s\tau}$ on both sides of (13.1) and yields

$$Be^{-s\tau}\bar{v} = -\frac{sBR}{\Lambda}e^{-s\tau}u - \frac{BS}{\Lambda}e^{-s\tau}y \qquad (16.1)$$

Operating sR/Λ on both sides of (13.2) and yields

$$\frac{sR}{\Lambda}\bar{x} = -s\frac{RA}{\Lambda}y + s\frac{RB}{\Lambda}e^{-s\tau}u \qquad (16.2)$$

Using (12), adding (16.1) and (16.2) gives:

$$x = \frac{C\Lambda - sR}{C\Lambda}\bar{x} - \frac{B}{C}e^{-s\tau}\bar{v} + \delta_2 \qquad (17)$$

where $\delta_2 \in L_2$.

Equations (15) and (17) are represented by Figure 1 as the system with feedbacks below where:

$$U_1 = \frac{S}{C\Lambda}\bar{x} + \delta_1; \ U_2 = -\frac{B}{C}e^{-s\tau}\bar{v} + \delta_2$$

This system consists of a main feedback and two internal feedbacks, referred to as feedbacks F1 and F2. The whole system stability analysis will be done in three steps.

Step 1: stability of feedback F1:

The forward pathway of this feedback is a linear time-invariant system with transfer function *(A-C)/C*. The return pathway is the nonlinear operator $\Phi(u_M,.)$ which, using lemma 1, belongs to $[0,\alpha_u]$. Using the circle criterion [9] – [15], one can get that *F1* is L_2-stable if

$$\inf_{0\le\omega\le2\pi}\text{Re}\left(\frac{A(j\omega)-C(j\omega)}{C(j\omega)}\right)>-\frac{1}{\alpha_u} \tag{18.1}$$

Now, let consider the operator G_1 such that $\bar{v}(t)=G_1\left(U_1(t)\right)$. Then, if we apply the loop transformation theorem ([2] pages 341-343), we can easily get the L_2-gain of G_1 as follows:

$$\gamma_2(G_1)=\frac{1+\gamma_i\alpha_u/2}{1-\gamma_i\alpha_u/2}\cdot\alpha_u \tag{18.2}$$

which is nothing but γ^0.

Step 2: stability of feedback F2:
In a similar manner, one can show that F2 is L_2-stable if

$$\inf_{0\le\omega\le2\pi}\text{Re}\left(\frac{C(j\omega)\Lambda(j\omega)-j\omega R(j\omega)}{C(j\omega)\Lambda(j\omega)}\right)>-\frac{1}{\alpha_y} \tag{18.3}$$

Furthermore, let G2 denote the operator $\bar{x}(t)=G_2\left(U_2(t)\right)$ such that

$$\gamma_2(G_2)=\frac{1+\gamma_{ii}\alpha_y/2}{1-\gamma_{ii}\alpha_y/2}\cdot\alpha_y=\gamma^1 \tag{18.4}$$

Step 3: Main feedback stability: applying the small gain theorem on Figure 1, it follows that this feedback is L_2-stable provided that

$$\gamma_2(G_1)\gamma_2(G_2)\gamma_2\left(\frac{SB}{\Lambda C^2}e^{-s\tau}\right)<1$$

which is nothing but the condition (10.3). Then it follows that $U_1, U_2, \bar{x}$ and $\bar{v}$ belong to L_2 as $\delta_1, \delta_2 \in L_2$.

Finally, since feedbacks F1 and F2 are L_2-stable, we deduce from (15) and (17) that $x \in L_2$ and $v \in L_2$.

REMARKS
a) In case where conditions (10.1) and (10.2) hold, the global asymptotic stability at the origin is guaranteed.
   Recall that all the signals are bounded i.e. $|u|\le u_M, |y|\le y_M, |v|\le v_M$ and $|x|\le x_M$. Then global stability means that all signals converge to zero for all initial conditions.
b) The design procedure of the stabilizing regulator could be composed of three steps which are: choosing polynomials C and Λ, solving Bezout equation (6) and computing pseudo-polynomials R and S and finally checking conditions (10.1) and (10.2). If these hold keep the obtained regulator. Else, make a different choice of C and go back to second step.
c) Although conditions (10.1) and (10.2) do not allow characterization of stabilizable systems (in terms of zeros, poles,...) outside the left half plane. This is illustrated by the example in section simulation.
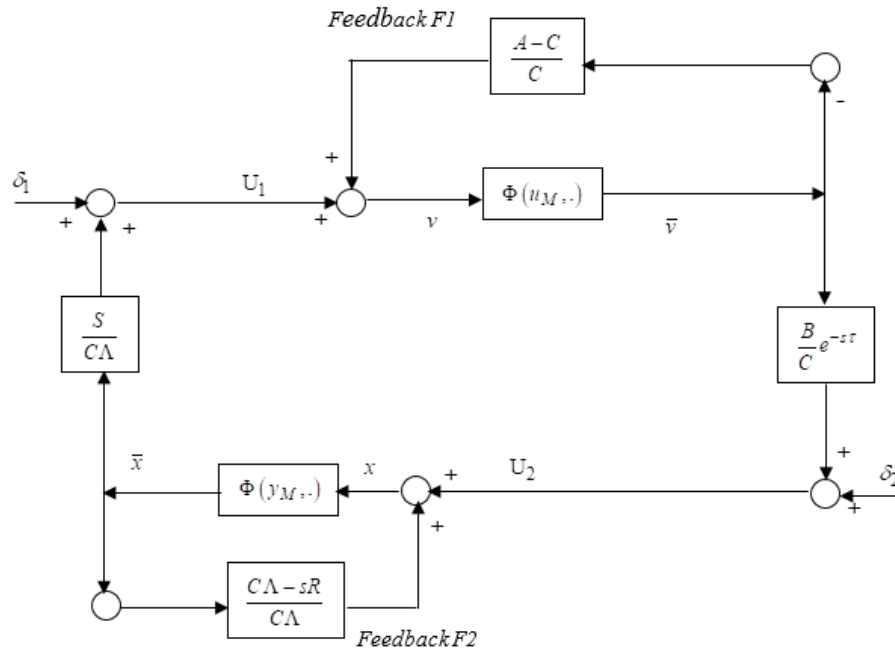
Figure 1. Block diagram of the feedback system described by equations (15) and (17)

5. SIMULATION

We consider the simple saturating linear delayed system described as follows:

$$x(t) = Ay(t) + Bu(t - \tau); \quad y(t) = sat\left(y_M, x(t)\right)$$

with

$$A = s - 1; \quad B = s + 1; \quad \tau = 1s; \quad y_M = u_M = 1 \tag{19}$$

Notice that the system is strictly unstable. By solving (6) to obtain the regulator parameters R and S with polynomials $\Lambda(s) = s + 0.2$ and

$$C(s) = s + c_0 \text{ where } 0 < c_0 < 1, \tag{20}$$

one gets

$$S(s) = s + 0.32e^{-s\tau} \text{ and } R(s) = 1 \tag{21}$$

In the rest of the section, we will show that there exists a set of values of the parameter c_0, so that conditions (10) hold. From (19) and (20) one gets

$$\inf_{0 \le \omega \le 2\pi} \text{Re}\left(\frac{A(j\omega)}{C(j\omega)}\right) = -\frac{1}{c_0} \tag{22}$$

From (8), it can be checked that

$$v_M = \max\left\{u_M, \left(\left|\frac{1}{0.2}\right|\right)u_M + \left(|1|\right)y_M\right\} = 5u_M + y_M = 6$$

$$x_M = \max\left\{y_M, \left(|1|\right)u_M + \left(|1+1|\right)y_M\right\} = u_M + 2y_M = 3$$

Necessarily we have to specify the value of c_0 to get v_M. So according to (11.1), one gets

$$\alpha_u = \frac{v_M - u_M}{v_M} = \frac{5}{6} \text{ and } 1 - \frac{1}{\alpha_u} = -1/5 \tag{23.1}$$

and

$$\alpha_y = \frac{x_M - y_M}{x_M} = \frac{2}{3} \text{ and } 1 - \frac{1}{\alpha_y} = -1/2 \tag{23.2}$$

It is easily checked using (22) and (23.1) that the first part of condition (10.1) is satisfied. Similarly we get using (20) and (21):

$$\inf_{0 \le \omega \le 2\pi} \text{Re}\left(\frac{C\Lambda - j\omega R}{C\Lambda}\right) = \inf_{0 \le \omega \le 2\pi}\left(\frac{\left(\omega^2 + 0.2\right)^2 + 0.72\omega^2}{\left(\omega^2 + 0.2\right)^2 - 0.64\omega^2}\right) > 0 \tag{24}$$

Equations (23.2) and (24) show that condition (10.2) holds for any $0 < c_0 < 1$. Now, one has to compute the involved gains to analyze the condition (10.3).

$$\gamma_i = \gamma_2\left(\frac{A(s) - C(s)}{C(s) + (A(s) - C(s)).\alpha_u / 2}\right)$$

$$= \max_{0 \le \omega \le 2\pi}\left|\frac{(a_0 - c_0)e^{-j\omega}}{1 - (c_0 + (a_0 - c_0).\alpha_u / 2)e^{-j\omega}}\right|$$

$$= 1.5$$

which, using the first part of (11.4) follows that

$$\gamma^0 = \alpha_u \cdot \frac{1 + \alpha_u \gamma_i / 2}{1 - \alpha_u \gamma_i / 2} = \alpha_u \cdot \frac{1 + 0.75\alpha_u}{1 - 0.75\alpha_u} = \frac{0.17}{1.15} \tag{25}$$

$$\gamma_{ii} = \frac{c_0}{1 - c_0(1 - \alpha_y / 2)}$$

$$= \frac{c_0}{1 - 1.25c_0}$$

and we get

$$\gamma^1 = \alpha_y \cdot \frac{1 + \gamma_{ii}\alpha_y / 2}{1 - \gamma_{ii}\alpha_y / 2} = -0.5\frac{1 - 1.75c_0}{1 + 1.75c_0} = \frac{0.875c_0 - 0.5}{1 + 1.75c_0}$$

Furthermore, equations (19)-(21) yield

$$\gamma_2\left(\frac{S(s)B(s)}{\Lambda(s)C(s)^2}e^{-s\tau}\right) = \frac{0.32}{0.2c_0^2} \tag{26}$$

Applying the above theorem, the L_2-stability of the closed-loop system is achieved provided

$$\frac{c_0^2(0.07 + 0.13c_0)}{0.148c_0 - 0.85} < 1$$

This condition is satisfied by the value $0 < c_0 = 0.6 < 1$. The performances of the resulting regulator are illustrated in the following Figure 2.
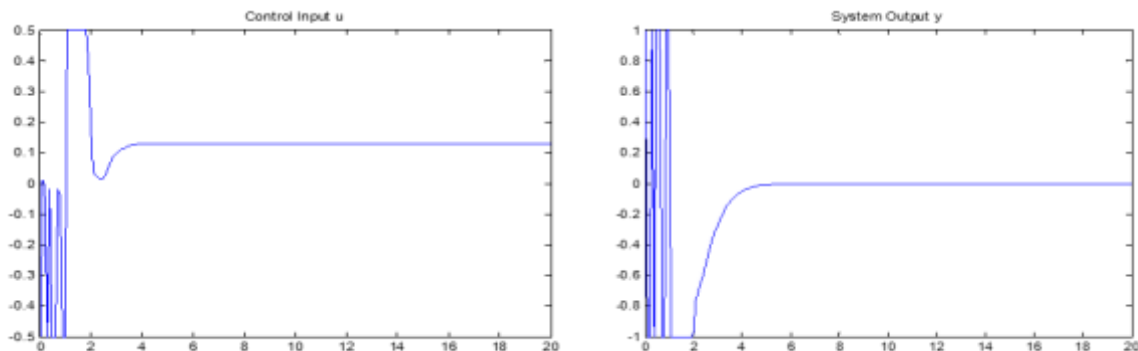

Figure 2. Closed-loop signals

6. CONCLUSION

We have interested in control system including an output saturating delayed linear system and a saturating regulator. We have shown that this association can be represented by a nonlinear feedback schema. Analyzing stability of this feedback leads to examine the closed-loop asymptotic global stability. Using tools of input-output stability approach, sufficient conditions for L_2-stability are then obtained. These conditions that concern both the regulator and the system parameters did not give an easy characterization of the class of systems that can be globally asymptotically stabilized. However, it has been verified that a saturating system that's strictly unstable can be globally asymptotically stabilized.

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