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RESETTING SMITH PREDICTOR FOR THE CONTROL OF UNSTABLE SYSTEMS WITH DELAY [1]

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Abstract: A modification of the Smith Predictor Control schemes is introduced. This modification consists in a periodical resetting of the initial condition of the predictor. It allows to extend the use of these control laws to unstable linear systems with delay. The continuous and discrete implantation of this scheme is considered. The stability of the scheme is proved and sufficient conditions for robustness with respect to parameter uncertainty and delay uncertainty are obtained in the discrete case. An illustrative example is discussed. *Copyright© 2002 IFAC*

Keywords: time-delay systems, discrete systems predictor, robust control.

1. INTRODUCTION

The *Smith Predictor* (SP) (Smith, 1959), (Palmor, 1996), depicted in Figure 1, can be considered as the first option control method for linear processes that have a delay in their input or output. Such delays occur frequently due to transport phenomenons, time consuming information processing or sensors design, among others. The wide acceptance of this control strategy is due to its two step design. The first step consist in designing a control law for the system without delay, with the help of standard control tools, familiar to engineers. The second step is the straightforward obtention of an equivalent controller for the delayed system. Clearly, this procedure does not require any expertise in the control and analysis of delayed systems. Moreover, the discrete time version of the SP (Palmor and Halevi, 1990) answers to the central concern of control engineers that compensators are commonly implemented on digital equipment. However, in the continuous case (Smith, 1959), (Palmor and Halevi, 1983), (Palmor, 1996) as well as in the discrete case (Palmor and Halevi, 1990), the use of the SP is restricted to stable plants.

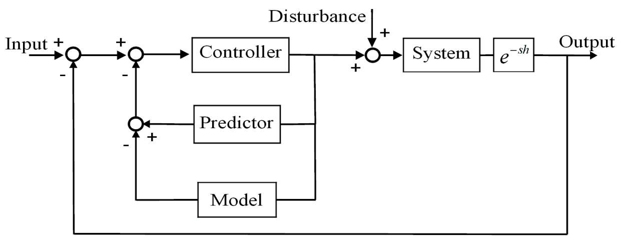

Fig. 1: Smith Predictor control scheme

In recent works, it was shown in the framework of the closely related control scheme of model process control in the spirit of Manitius and Olbrot (1979) that it is possible to obtain a robust closed loop with respect to parameter and delay mismatch in the case of unstable processes by introducing a periodic resetting of the prediction (Mondié *et al.*, 2001b).

These ideas lead naturally to the *Continuous Resetting Smith Predictor (*CRSP) and to the *Discrete Resetting Smith Predictor* (DRSP) which are the object of this paper.

The paper is organized as follows: the CRSP is presented in Section 2. The resetting leads to the design in Section 3 of a DRSP whose stability and robustness properties with respect to parameter mismatch and delay mismatch are established. An

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illustrative example is presented in Section 4 and the paper ends with some concluding remarks.

2. RESETTING SMITH PREDICTOR

In this section we consider observable linear multivariable system with nominal delay h^* in the input described by

$$\dot{x}(t) = A^* x(t) + B^* u(t - h^*) + d(t - h^*) \quad (1)$$
$$y(t) = C^* x(t) \quad (2)$$

where the nominal parameters are $A^* \in \mathbb{R}^{n\times n}, B^* \in \mathbb{R}^{n\times m}, C^* \in \mathbb{R}^{p\times n}$, $d(t)$ is an input bounded disturbance and the initial condition is $x(t) = f(t), t \in [-h^*, 0]$. In a robust analysis framework, the delay and the parameters used in the design, h, A, B and C can be different from h^*, A^* B^* and C^*, respectively.

A prediction $x_p(t)$ for the variable $x(t+h)$, is given by

$$x_p(t) = e^{Ah} x(t) + \int_t^{t+h} e^{A(t+h-\sigma)} B u(\sigma - h) d\sigma.$$

Defining $\tau := \sigma - h$, $x_p(t)$ can be rewritten as

$$x_p(t) = e^{Ah} x(t) + \int_{t-h}^{t} e^{A(t-\tau)} B u(\tau) d\tau. \quad (3)$$

The above depends only on past and present values of $x(t)$ and $u(t)$. Thus, $x_p(t)$ is available at time t.

However, as explained in Palmor (1996) the SP is not suitable for stabilization due to internal instability linked to the unstable prediction. A realization of the control law in the so-called integral form is suggested in Palmor (1996). This approach was also proposed in the framework of spectrum assignment with distributed delay (Manitius and Olbrot, 1979). However, as shown in Mondié *et al.* (2001a), when the integral is calculated with a constant step methods, such implementation produces an unstable behavior if the control law is itself unstable, whatever the precision of the integration method.

The purpose of this work is to allow the Smith predictor to control unstable systems, without loosing its main advantage, namely, the simple design procedure.

3. RESETTING OF THE SMITH PREDICTOR

In this paper, a periodic refreshing of the calculated initial condition is introduced in the Continuous Smith predictor.

A timing device provided by a square signal such that $l := t_{k+1} - t_k > h$, $k = 1, 2, ...$ determines the predictor and model resetting time as shown in the overall scheme depicted in Figure 2.

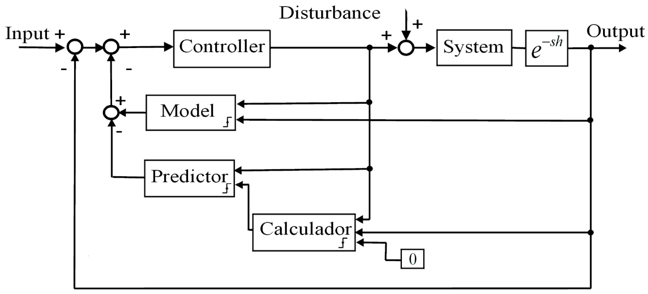

Fig. 2: Resetting Smith predictor.

The calculator block computes an estimate of the state $x(t_i + h)$ for $i = 1, ..., k, k+1, ...$The estimated value $x(t_i+h)$ is used as the initial value of the predictor which will deliver a "continuous" estimate of the state $x(t+h)$. The initial condition of the model is periodically resetted to the value of the state of the system.

The elements of the control scheme are next described in detail.

Resetting Predictor: During the time interval $[t_k - h, t_k)$ we compute the value of the integral

$$\int_{t_k-h}^{t_k} e^{A(t-\tau)} B u(\tau) d\tau$$

as a solution to the differential equation

$$\dot{z}(t) = Az(t) + Bu(t)$$

with zero initial conditions at time $t_k - h$. At time t_k, this integral, along with the measured value of the state of the system allows the computation of

$$x_c(t_k + h) = e^{Ah} x(t_k) + \int_{t_k-h}^{t_k} e^{A(t_k-\tau)} B u(\tau) d\tau \quad (4)$$

which is an estimate of $x(t_k + h)$ computed at time t_k. Next, at time $t_k, k = 1, 2, ...$ the predictor is initialized with the computed value $x_c(t_k + h)$. The prediction $x_p(t)$ of the state $x(t + h)$ is then

$$x_p(t) = e^{A(t-t_k)} x_c(t_k + h) + \int_{t_k}^{t} e^{A(t-\tau)} B u(\tau) d\tau,$$
$$t \in [t_k, t_{k+1}). \quad (5)$$

Finally, substituting (4) into (5) leads to

$$x_p(t) = e^{A(t-t_k+h)} x(t_k) + \int_{t_k-h}^{t} e^{A(t-\tau)} B u(\tau) d\tau,$$
$$t \in [t_k, t_{k+1}). \quad (6)$$

Resetting Model: At time $t_k, k = 1, 2, ...,$ the model is initialized with the computed value $x(t_k)$. The state $x_m(t)$ is given by

$$x_m(t) = e^{A(t-t_k)}x_m(t_k) + \int_{t_k}^{t} e^{A(t-\tau)}Bu(\tau - h)d\tau, \quad (7)$$

Error feed into the controller

$$e(t) = Cx_p(t) + y(t) - Cx_m(t) + r(t). \quad (8)$$

It is possible to perform a robustness analysis with respect to parameter and delay mismatch, and with respect to bounded disturbances of the CRSP, as the one done in Mondié *et al.* (2001b) for the Resetting Process Model Control. However, due to the complexity of the control law, the difficulty of the analysis is increased. Moreover, the analysis of Mondié *et al.* (2001b) is based on the assumption that the resetting time is greater than the delay. This hypothesis is restrictive in particular for large delays, because, between resetting the controller acts only on the predictor, and the state is not feedback.

4. DISCRETE RESETTING SMITH PREDICTOR (DRSP)

The periodic resetting introduced above reminds somehow the period of discrete time controller. Moreover there is a practical interest for digitally implemented controllers. These facts are strong motivation for developing a discrete implemented version of the RSP.

4.1 *Discrete Resetting predictor*

For the discretization of the scheme, the sampling period is selected so that the design time delay is a multiple over the integers of the sampling period T. It is well known that in the discrete framework the sampling period must de small enough so that it meets the requirements inherent to the discretization process. Since the sampling period is a design parameter it can be chosen so that it fulfills both requirements, namely, $nT = h$, where n is an integer and where T is smaller than the maximum allowed value for an exact discretization.

The DRSP is obtained by choosing the resetting time of the previous section so that it coincides with the sampling time. A zero order hold is used in the control action.

Consider the expression (6) at time $t = kT, k = 0, 1, 2, ...$, then

$$x_p(kT) = e^{A\cdot nT}x(kT) + \int_{kT-nT}^{kT} e^{A(kT-\tau)}Bu(\tau)d\tau.$$

Defining $kT - \tau = \zeta$, it follows that

$$x_p(kT) = e^{AnT}x(kT) + \int_{0}^{nT} e^{A\cdot\zeta}Bu(kT - \zeta)d\zeta.$$

Because of the zero order hold, $u(kT-\zeta)$ is piecewise constant over each intersampling interval, hence

$$x_p(kT) = e^{AnT}x(kT) + \sum_{j=1}^{n} \int_{T\cdot(j-1)}^{T\cdot j} e^{A\cdot\zeta}Bd\zeta \cdot u(kT - jT). \quad (9)$$

Now, observe that defining $nT + \zeta = \sigma$; $nT - T + \zeta = \sigma$; $nT - 2T + \zeta = \sigma$; ...; $\zeta = \sigma$, respectively in the following equations

$$\int_{nT-T}^{nT} e^{A\cdot\zeta}Bd\zeta = e^{A\cdot(nT-T)} \int_{0}^{T} e^{A\cdot\sigma}Bd\sigma,$$

$$\int_{(nT-2T)}^{nT-T} e^{A\cdot\zeta}Bd\zeta = e^{A\cdot(nT-2T)} \int_{0}^{T} e^{A\cdot\sigma}Bd\sigma,$$

$$\vdots$$

$$\int_{0}^{T} e^{A\cdot\zeta}Bd\zeta = \int_{0}^{T} e^{A\cdot\sigma}Bd\sigma,$$

then, defining z as the backward shift operator, (9) can be written as

$$x_p(kT + nT) = e^{A\cdot nT}x(kT) + \Phi(z)\cdot u(kT) \quad (10)$$

where

$$\Phi(z) = \phi^{n-1}\Gamma\cdot z^n + \cdots + \phi\Gamma\cdot z^2 + \Gamma\cdot z^1,$$

and

$$\phi = e^{AT}\ ; \qquad \Gamma = \int_{0}^{T} e^{A\sigma}Bd\sigma. \quad (11)$$

Again, due to the delayed nature of the input, the prediction at the sampling time $kT + nT$, depends only on the value of $x(kT)$ and of values of the input at kT and previous sampling instants. Hence this prediction is available at time kT.

Remark 1. It is possible to derive the expression for the predictor by considering the discretization of system (1) (Astrom and Wittenmark, 1997). An interpretation of $\Phi(z)$, that will be useful in the sequel, is in terms of the model transfer function with no delay $P(z^{-1}) := (z^{-1}I - \phi)^{-1}\Gamma$. Indeed one can see in a straightforward manner that

$$\Phi(z) = P(z) - \phi^n \cdot z^n \cdot P(z) \quad (12)$$

4.2 *Discrete Resetting Smith Predictor*

Now, if we apply the same resetting process as in the continuous case, at each sampling instant, we are able to describe in detail the elements of the Discrete Resetting Smith Predictor.

Predictor: The resetting predictor is the one developed above, namely

$$x_p(kT) = \phi^n x(kT) + \sum_{j=1}^{n} \phi^{j-1}\Gamma u(kT - jT). \quad (13)$$

Model error: In this case, due to the fact that the resetting occurs at each sampling time instant, we observe that

$$x(kT) - x_m(kT) = 0, \quad (14)$$

hence the model error is forced to zero at each sampling time and there is no need to compute it.

Error feed into the controller: It is given by $e(kT) = r(kT) - Cx_p(kT) + y(kT) - Cx_m(kT)]$

and because of (14) it reduces to

$$e(kT) = r(kT) - Cx_p(kT) \quad (15)$$

Indeed, we are now in a discrete framework. In the following, let us consider the control law $u(z) = G(z)e(z)$, and let then define the discrete transfer function without delay, of the model employed in the design $P(z) = (z^{-1}I - \phi)^{-1}\Gamma$ and of the process $P^*(z) = (z^{-1}I - \phi^*)^{-1}\Gamma^*$, the delay of the process z^{n^*} and the delay employed in the design z^n. The overall control scheme is then:

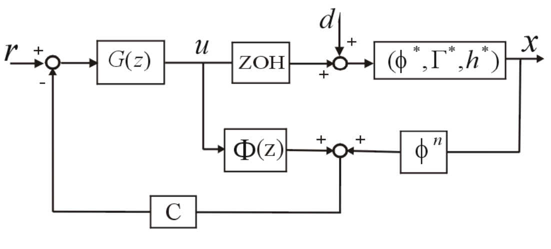

Fig. 3 Discrete resetting Smith Predictor

Remark 2. The resulting scheme is equivalent to the one that would result from the discretization of the Resetting Process Model presented in Mondié *et al.* (2001b). This shows indeed the closeness of the two approaches.

5. STABILITY AND ROBUSTNESS OF THE DRSP

An additional advantage of the DRSP, compared to the CRSP introduced in the first part of this work, is that it is possible to use in a straightforward manner the machinery available for the discrete time approach to perform the stability and robustness analysis of the scheme. For simplicity, we restrict our attention to the single input, single output case in the rest of the paper.

5.1 *Stability analysis*

The reference to output transfer function is $y(z)/r(z) = N_r(z)/D_r(z)$ where

$$N_r(z^{-1}) = G(z)C^*P^*(z)z^{n^*} \quad (16)$$
$$D_r(z^{-1}) = 1 + G(z)C\Phi(z) + G(z)C\phi^n P^*(z)z^{n^*}$$

Substitution of the expression (12) for $\Phi(z)$ gives:

$$N_r(z^{-1}) = G(z)C^*P^*(z)z^{n^*}$$
$$D_r(z^{-1}) = 1 + G(z)C\{P(z) - \phi^n P(z)z^n\} + G(z)C\phi^n P^*(z)z^{n^*}$$

When there is no mismatch in the parameters and in the delay of the process and of the design model, this transfer function simplifies to

$$\frac{y(z)}{r(z)} = \frac{G(z)CP(z)}{1 + G(z)CP(z)}z^n.$$

Then if the controller $G(z)$ is designed so that the closed loop system with no delay has a stable closed loop polynomial, it follows that the closed loop resulting from for the Scheme of Figure 3 has the same characteristic polynomial, hence it is stable.

The disturbance to output transfer function is $y(z)/d(z) = N_d(z)/D_d(z)$ where

$$N_d(z^{-1}) = C^*\{1 + G(z)C\Phi(z)\}P^*(z)z^{n^*}$$
$$D_d(z^{-1}) = 1 + G(z)C\Phi(z) + G(z)C\phi^n P^*(z)z^{n^*} \quad (17)$$

Again, the unstable poles of the numerator are canceled with those of the denominator, and if (16) is stable then (17) is also stable.

Remark 3. If $\Phi(z)$ is realized as the right hand side of (12) the transfer function between the output and the disturbance when there is no mismatch,

$$\frac{y(z)}{d(z)} = \left\{1 - \frac{G(z)C\phi^n P(z)z^n}{1 + G(z)CP(z)}\right\}CP(z)\cdot z^n$$

is always unstable, because the unstable poles of the term $CP(z)$ outside the brackets of the numerator remain. Indeed, although $\Phi(z)$ and $P(z) - \phi^n \cdot P(z)\cdot z^n$ are equal from an algebraic point of view, there is indeed a fundamental difference in the implantation.

5.2 *Robustness Analysis*

It is now possible to give necessary conditions for robust stability based on the principle of variation.

Theorem 1. Consider the discretizaton of the process (1) in closed loop with the control law (15), (13) and assume that the controller is designed so that under ideal circumstances, the closed loop is stable. Assume also that the real process and design model differ. Then, if the condition

$$\left|G(z)C\phi^n\{P^*(z)z^{(n^*-n)} - P(z)\}z^n\right| < |1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^n| \quad (18)$$

holds for all z on the unit circle, then the closed loop remains stable.

Proof. Observe that the stability region in the z-plane is the outside the unit circle. The controller $G(z)$ is designed so that it stabilizes the process, then the closed loop

$$1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^n \qquad (19)$$

has no roots inside the unit circle. It follows from the above condition and from Rouché's theorem that (19) and

$$\begin{aligned} &1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^n \\ &+G(z^{-1})C\phi^n\{P^*(z)z^{(n^*-n)} - P(z)\}z^n \\ =&\,1 + G(z)C\Phi(z) + G(z)C\phi^n P^*(z)z^{n^*}, \end{aligned}$$

the closed loop characteristic equation when uncertainty is present, have no roots inside the unitary circle, hence it is stable. ■

Corollary 2. If there is no mismatch in the delay, then the sufficient condition (18) reduces to

$$\begin{aligned} &|G(z)C\phi^n\{P^*(z) - P(z)\}| \qquad (20) \\ &< |1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^n| \end{aligned}$$

Proof. Substituting $n = n^*$ into (18) leads to

$$\begin{aligned} &|G(z)C\phi^n\{P^*(z) - P(z)\}|\,|z^n| \\ &< |1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)\cdot z^n|, \end{aligned}$$

on the unit circle $|z^{-n}| = 1$ and (20) follows ■

Theorem 3. When there is no mismatch in the parameters, then if

$$\begin{aligned} &|G(z)C\phi^n P(z)| \qquad (21) \\ &< \frac{1}{2}|1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^n| \end{aligned}$$

holds for z on the unit circle, then the closed loop is stable.

Proof. Let $\Delta n = n - n^*$. It follows from (21) and from the fact that $\left|z^{n^*}\right| = 1$ and $\left|1 - z^{\Delta n}\right| \leq 2$ that

$$\begin{aligned} &|G(z)C\phi^n P(z)|\left|z^{n^*}\right|\left|z^{\Delta n} - 1\right| \\ &< \left|1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^{-n}\right| \end{aligned}$$

and that

$$\begin{aligned} &\left|G(z)C\phi^n P(z)z^{n^*}(1 - z^{\Delta n})\right| \\ &< |1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^n|. \end{aligned}$$

The arguments used above imply that

$$1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)z^{n^*} \qquad (22)$$

has no roots inside the unit circle, hence the closed loop is stable for all delay. ■

Theorem 4. When there is no mismatch in the parameters, if there exists some z on the unit circle so that,

$$W(z) \geq \frac{1}{2},$$

where

$$W(z) = \frac{|G(z)C\phi^n P(z)|}{|1 + G(z)C\Phi(z) + G(z)C\phi^n P(z)\cdot z^n|}.$$

Let

$$\omega_m = \sup\{\omega_i : W(e^{-j\omega T}) < \frac{1}{2}, \forall\omega \in (0, \omega_i]\}$$

then the closed loop is stable for $\Delta n \in [0, \delta_n)$ where δ_n is the closest integer smaller than $\hat{\alpha}$ defined as

$$\hat{\alpha} = \min\{\inf_{\omega\in(0,\omega_m)} \frac{2}{\omega}\arcsin\frac{1}{2W(e^{-j\omega T})}, n^*\}.$$

Proof. Let $z = e^{-sT}$ and let us associate to the characteristic equation (22)

$$1 + G(e^{-sT})C\Phi(e^{-sT}) + G(e^{-sT})C\phi^n P(e^{-sT})e^{sTn^*} = 0$$

where n^* is assumed to be a real variable. Substitute $\alpha = n - n^*$, rearranging terms and taking modules give

$$\begin{aligned} &\left|G(e^{-sT})C\phi^n P(e^{-sT})\right|\left|e^{-sTn^*}\right|\left|(1 - e^{-sT\alpha})\right| \\ =&\left|1 + G(e^{-sT})C\Phi(e^{-sT}) + G(e^{-sT})C\phi^n P(e^{-sT})e^{-sTn}\right|. \end{aligned}$$

For s on the imaginary axis $\left|e^{-j\omega Tn^*}\right| = 1$ and $\left|e^{-j\omega T\Delta n} - 1\right| = 2\left|\sin(\frac{\omega T\alpha}{2})\right|$. The above is then

$$\begin{aligned} &\left|G(e^{-j\omega T})C\phi^n P(e^{-j\omega T})\right| 2\left|\sin(\frac{\omega T\alpha}{2})\right| \\ =&\left|1 + G(e^{-j\omega T})C\Phi(e^{-j\omega T}) + G(e^{-j\omega T})C\phi^n P(e^{-j\omega T})e^{-nj\omega T}\right|. \end{aligned}$$

We can conclude that if there exist ω and α (continuous) such that this equation holds, a crossing of the imaginary axis occurs in the s-plane. Then $\hat{\alpha}$ defined above is a solution to this equation. Moreover for $\Delta n \leq \delta_n$, the closest integer smaller than $\hat{\alpha}$, no crossing of the unit circle occurs in the z-plane, and the closed loop (22) is stable. ■

6. ILLUSTRATIVE EXAMPLE

The scalar system $\dot{x}(t) = x(t) + u(t-1)$ introduced by Manitius and Olbrot (1979) has been extensively studied in the literature.

As it is well known this unstable process in closed loop with a SP is unstable, even when there is no mismatch.

In the simulation results that follows, the design parameters are $a = 1$, $b = 1$ and $h = 1$. When the system parameters are known, we use $a^* = 1$, $b^* = 1$ and $h^* = 1$. When the system parameters

are uncertain, unless otherwise indicated, we assume that $a^* = 1.1$, $b^* = 1.1$ and $h^* = 1.1$.
As shown on Figures 4 and 5 the CRSP and the DRSP stabilize the plant and they are robust. The higher frequency of the resetting in the DRSP allows indeed a softer response and control law.
The stability and robustness of the scheme are indeed crucial, but from a practical point of view, the reference following and disturbance rejection are also important properties. Next, a simple modification of the original scheme, consisting of an adaptive gain allows zero steady state error and disturbance rejection. This situation is illustrated in Figure 6 for a unit step reference, and a disturbance of a magnitude of 50% of the reference occurring at time $20s$.

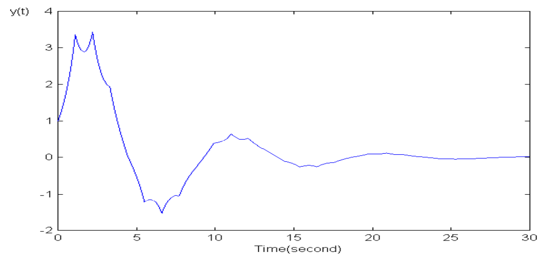

Fig. 4: CRSP uncertain parameters ($b^* = 1$)

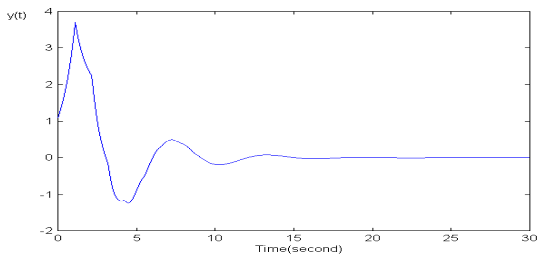

Fig. 5: DRSP uncertain parameters.

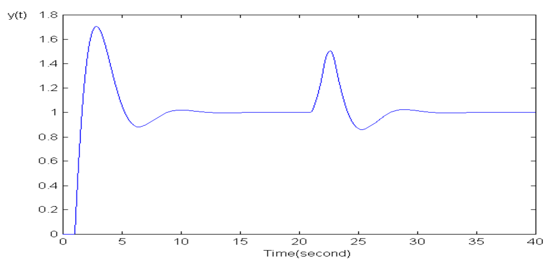

Fig. 6: DRSP with adaptive gain

7. CONCLUDING REMARKS

The resetting technique introduced in Mondié *et al.* (2001b) can be extended successfully to the Smith Predictor in order to allow the stabilization of unstable systems with delay in the input or in the output. The discrete version of this scheme leads to the Discrete Resetting Smith Predictor. This discrete scheme allows a simple proof of the stability of the scheme. The robustness with respect to parameter and delay mismatch is established. An important feature of the Discrete Resetting Smith Predictor is that, as a Smith Predictor, its design is done in two steps, namely, the design of the controller ignoring the delay, and the straightforward construction of an equivalent controller for the delayed system. Moreover, a discrete time version of this controller is of certain interest in practical applications.

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