A NEW METHOD OF TUNING THREE TERM CONTROLLERS FOR DEAD-TIME PROCESSES WITH A NEGATIVE/POSITIVE ZERO

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Abstract: The use of the Pseudo-Derivative Feedback (PDF) structure in the control of stable or unstable dead-time processes with a negative or a positive zero is investigated. A simple direct synthesis method for tuning the PDF controller is presented. Moreover, a modification of the proposed method, which ensures its applicability in the case of large overshoot response processes with dead time, is also presented. The PDF control structure and the proposed tuning method ensure smooth closed-loop response to set-point changes, fast regulatory control and sufficient robustness against parametric uncertainty. Simulation results show that, in most cases, the proposed method is as efficient as the best of the most recent PID controller tuning methods for dead-time processes with negative/positive zeros, while its simplicity in deriving the controller settings is a plus point over existing PID controller tuning formulae.

1 INTRODUCTION

In the process industry, stable second order dead-time models as well as second order dead-time models with one right-half-plane pole are frequently used to adequately describe numerous processes for the purpose of designing controllers. However, these types of process models are inadequate in the case where a process controlled variable encounters two (or more) competing dynamic effects with different time constants from the same manipulated variable (Waller and Nygardas, 1975). This composite dynamics results to a process behaviour that exhibits an inverse response or a large overshoot response. Inverse response or large overshoot response is portrayed by the presence of one (or an odd number of) positive or negative zeros, respectively, in the process models, and they can cause, together with the process dead-times, serious problems in designing and tuning simple controllers for the process under consideration.

Inverse response second order dead-time process models (SODT-IR) are used to represent the dynamics of several chemical processes (like level control loops in distillation columns and temperature control loops in chemical reactors), as well as the dynamics of PWM based DC-DC boost converters in industrial electronics. In the extant literature, there is a number of studies regarding the design and tuning of three-term controllers for SOPD-IR processes (Waller and Nygardas, 1975; Scali and Rachid, 1998; Zhang *et al*, 2000; Luyben, 2000; Chien *et al,* 2003; Padma Sree and Chidambaram, 2004; Chen *et al,* 2005; Chen *et al,* 2006). In particular, Waller and Nygardas (1975) presented an

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empirical tuning of PID controllers based on the Ziegler-Nichols method for SOPDT-IR processes. In Scali and Rachid (1998) and Zhang *et al* (2000), analytical design methods based on the Internal Model Control framework and the H_∞ control theory, have been proposed for inverse response processes without time delay. In Luyben (2000), an empirical method that gives large overshoot and oscillatory response has been proposed to design PI controllers for SODT-IR processes. In Chien *et al* (2003), a direct synthesis tuning method is presented to tune PID controllers for both under-damped and over-damped SODT-IR processes. In Chen *et al* (2005), an analytical PID controller design for SOD-IR processes is derived based on conventional unity feedback control. In Chen *et al* (2006), an analytical design scheme based on IMC theory has been proposed to control SODT-IR processes. Finally, in Padma Sree and Chidambaram (2004), a method of tuning set-point weighted PID controllers for unstable SODT processes with a positive or a negative zero is presented. This method is based on appropriately equating coefficients of like powers of s in the numerator and the denominator of the closed-loop transfer function.

In contrast, controller tuning for large overshoot response dead-time processes have received less attention in the past, although they used to model several physical phenomena, like blending processes, mixing processes in distillation columns and temperature of heat exchangers (see Chien *et al* (2003), for details). In Chang *et al* (1997) a tuning method of controllers in first order lead-lag form has been proposed for such processes. Furthermore in Chien *et al* (2003), a direct synthesis tuning method is presented in order to tune PID controllers for both under-damped and over-damped large overshoot response processes.

The present paper investigates some aspects of the controller configuration proposed by Phelan (1978), and called the "pseudo-derivative feedback controller" (PDF), which is put forward here as an alternative means of tuning three-term controllers for stable or unstable dead time processes with a negative or positive zero. The aim of the paper is to propose a set of tuning rules for the PDF controller when it is applied to such processes. The proposed method is a direct synthesis tuning method and it is based on the manipulation of the closed loop transfer function through appropriate approximations of the dead-time term in the denominator of the closed loop transfer function as well as appropriate selection of the derivative gain, in order to obtain a second order dead-time closed-loop system. On the basis of this method the settings of the PDF controller are obtained in terms of two adjustable parameters, one of which can further be appropriately selected in order to achieve a desired damping ratio for the closed-loop system, while the other is free to designer and can be selected in order to enhance the obtained regulatory control performance. Moreover, an appropriate modification of the proposed method, that makes it applicable in the case of large overshoot response processes with dead time, is also presented. For assessment of the effectiveness of the proposed tuning method and in order to provide a comparison with existing tuning methods, a series of simulation examples are presented. Simulation results verify that the PDF control structure and the proposed direct synthesis tuning method ensure smooth closed-loop response to set-point changes, fast regulatory control and sufficient robustness in case of model mismatch.

2 THE PSEUDO-DERIVATIVE FEEDBACK CONTROLLER

The Pseudo-Derivative Feedback (PDF) controller has first been proposed by Phelan (1978), and its general feedback configuration is shown in Figure 1. The transfer function $G_{CL}(s)$ of the closed loop system is given by

$$G_{CL}(s) = \frac{K_I G_P(s)}{s + \left(K_{D,n-1}s^n + \ldots + K_{D,1}s^2 + K_{D,0}s + K_I\right)G_P(s)} \quad (1)$$

The PDF controller is essentially a variation of the conventional PID controller. In contrast to the PID controller, the PDF controller does not contribute to closed-loop zeros, and hence it is expected that it will not render worst the overshoot of the closed-loop response. The two configurations differ in the way they react to set-point changes (as it can be easily checked, they are equivalent for load or disturbance changes). The PID controller often has an abrupt response to a step change because the step is amplified and transmitted directly to the feedback control element and downstream blocks. This can induce a significant overshoot in the response that is unrelated to the closed loop system damping. For this reason, it is a common practice to ramp or filter the set-point. The PDF structure avoids this because naturally ramps the controller effort, since it internalizes the pre-filter that one would apply to cancel any closed-loop zeros introduced in the PI/PID control configuration.

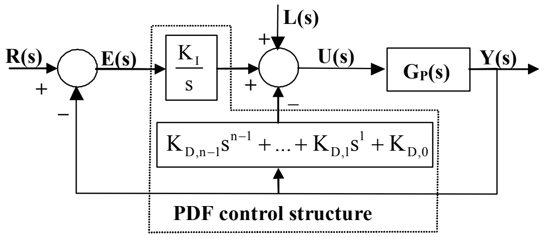

Figure 1: The general PDF control structure.

In the present paper, we focus our attention on the specific form of the general PDF control structure which contains proportional as well as a single derivative action in the feedback path (i.e. $K_{D,i}=0$ for $i=2,\ldots,n-1$ and $K_P\neq 0$, $K_{D,1}\neq 0$). We call this feedback scheme, the PD-1F control structure, in contradistinction with the PDF controller without derivative action (i.e. the controller with $K_{D,1}=0$), which is designated as the PD-0F controller. We shall next analyze its performance, in the case where the system under control is a second order process with both dead-time and a minimum or a non-minimum phase zero, which can be described by the following general transfer function model

$$G_P(s)=K\left(ps+1\right)\exp(-ds)/\left[\left(\tau_1 s+1\right)\left(\tau_2 s+q\right)\right] \quad (2)$$

where

$$p=\begin{cases}-\tau_z & \text{in the case of a positive zero}\\ \tau_z & \text{in the case of a negative zero}\end{cases},\ \tau_z>0$$

and q=1, in the case of a stable process or q=-1, in the case of an unstable process, and where, K, d, τ_z, τ_1 and τ_2, are the process gain, the dead-time, the zero's time constant and the process time constants, respectively.

To this end, observe that, equation (1), in the case of a PD-1F controller and for process models of the form (2), takes the form

$$G_{CL}(s)=\frac{KK_I\left[\left(ps+1\right)/\left(\tau_1 s+1\right)\right]\exp(-ds)}{s\left(\tau_2 s+q\right)+\left(KK_d s^2+KK_P s+KK_I\right)\dfrac{(ps+1)}{(\tau_1 s+1)}\exp(-ds)} \quad (3)$$

Relation (3) will be the starting point for the development of the tuning method that will be presented in the sequel.

3 A SIMPLE TUNING METHOD

In order to systematically present the proposed tuning method, observe that by making use of the approximation

$$\left(ps+1\right)/\left(\tau_1 s+1\right)\approx 1-(\tau_1-p)s \quad (4)$$

in the numerator of (3), and observing that

$$\exp(-ds)=\frac{\exp\left[-\left(1-\alpha\right)ds\right]}{\exp(\alpha ds)}$$

for some $\alpha\in\Re$, we obtain

$$G_{CL}(s)=\frac{\left[1-(\tau_1-p)s\right]\exp(-ds)}{s\left(\dfrac{\tau_2}{KK_I}s+\dfrac{q}{KK_I}\right)+\left(\dfrac{K_d}{K_I}s^2+\dfrac{K_P}{K_I}s+1\right)\dfrac{(ps+1)\exp\left[-(1-\alpha)ds\right]}{(\tau_1 s+1)\exp(\alpha ds)}} \quad (5)$$

Next, using the approximations,

$$\left(ps+1\right)\exp\left[-(1-\alpha)ds\right]=\left[p-(1-\alpha)d\right]s+1$$
$$\left(\tau_1 s+1\right)\exp(\alpha ds)=\left(\tau_1+\alpha d\right)s+1$$

in (5), we obtain

$$G_{CL}(s)=\frac{\left[1-\left(\tau_1-p\right)s\right]\exp(-ds)}{s\left(\dfrac{\tau_2}{KK_I}s+\dfrac{q}{KK_I}\right)+\left(\dfrac{K_d}{K_I}s^2+\dfrac{K_P}{K_I}s+1\right)\left[\dfrac{\left[p-\left(1-\alpha\right)d\right]s+1}{\left(\tau_1+\alpha d\right)s+1}\right]} \quad (6)$$

Relation (6) may further be written as

$$G_{CL}(s)=\frac{\left[1-\left(\tau_1-p\right)s\right]\exp(-ds)}{s\left(\dfrac{\tau_2}{KK_I}s+\dfrac{q}{KK_I}\right)+P(s)} \quad (7)$$

where

$$P(s)=\left[\frac{K_d}{\left(\tau_1+\alpha d\right)K_I}s+\left(\frac{K_P}{\left(\tau_1+\alpha d\right)K_I}-\frac{K_d}{\left(\tau_1+\alpha d\right)^2 K_I}\right)+Q(s)\right]\times\left[\left[p-\left(1-\alpha\right)d\right]s+1\right]$$

$$Q(s)=\frac{1-\dfrac{K_P}{\left(\tau_1+\alpha d\right)K_I}+\dfrac{K_d}{\left(\tau_1+\alpha d\right)^2 K_I}}{\left(\tau_1+\alpha d\right)s+1}$$

Observe now that by selecting

$$K_d = (\tau_1 + \alpha d)K_P - (\tau_1 + \alpha d)^2 K_I \quad (8)$$

we obtain Q(s)=0 and

$$P(s) = \left[\left(\frac{K_p}{K_I} - \tau_1 - \alpha d\right)s + 1\right]\left[\left[p - (1-\alpha)d\right]s + 1\right]$$

Therefore, relation (7) yields

$$G_{CL}(s) = \frac{\left[1 - (\tau_1 - p)s\right]\exp(-ds)}{s\left(\frac{\tau_2}{KK_I}s + \frac{q}{KK_I}\right) + \left[\left(\frac{K_p}{K_I} - \tau_1 - \alpha d\right)s + 1\right]\left[\left[p - (1-\alpha)d\right]s + 1\right]} \quad (9)$$

which can further be written in the form

$$G_{CL}(s) = \frac{\left[1 - (\tau_1 - p)s\right]\exp(-ds)}{\lambda^2 s^2 + 2\zeta\lambda s + 1} \quad (10)$$

$$\lambda = \sqrt{\frac{\tau_2}{KK_I} - \left(\frac{K_p}{K_I} - \tau_1 - \alpha d\right)\left[(1-\alpha)d - p\right]} \quad (11)$$

$$\zeta = \frac{\frac{K_p}{K_I} - d - \tau_1 + p + \frac{q}{KK_I}}{2\sqrt{\frac{\tau_2}{KK_I} - \left(\frac{K_p}{K_I} - \tau_1 - \alpha d\right)\left[(1-\alpha)d - p\right]}} \quad (12)$$

The Routh stability criterion about (10) yields

$$K_P > (d + \tau_1 - p)K_I - \frac{q}{K} \quad (13)$$

and

$$K_P < (\tau_1 + \alpha d)K_I + \frac{\tau_2}{K\left[(1-\alpha)d - p\right]} \quad (14)$$

Therefore, as for K_P one can choose the middle value of the range given by inequalities (13) and (14). That is

$$K_P = \frac{\left[2\tau_1 + (1+\alpha)d - p\right]K_I + \frac{\tau_2 - q\left[(1-\alpha)d - p\right]}{K\left[(1-\alpha)d - p\right]}}{2} \quad (15)$$

Then, from (15), we obtain

$$\beta \triangleq \frac{K_P}{K_I} = \frac{\left[2\tau_1 + (1+\alpha)d - p\right] + \frac{\tau_2 - q\left[(1-\alpha)d - p\right]}{K_I K\left[(1-\alpha)d - p\right]}}{2} \quad (16)$$

which yields,

$$K_I = \frac{\tau_2 - q\left[(1-\alpha)d - p\right]}{K\left[(1-\alpha)d - p\right]\left[2\beta - 2\tau_1 - (1+\alpha)d + p\right]} \quad (17)$$

Therefore,

$$K_P = \frac{\beta\left[\tau_2 - q\left[(1-\alpha)d - p\right]\right]}{K\left[(1-\alpha)d - p\right]\left[2\beta - 2\tau_1 - (1+\alpha)d + p\right]} \quad (18)$$

$$K_d = \frac{\left[\beta(\tau_1 + \alpha d) - (\tau_1 + \alpha d)^2\right]\left[\tau_2 - q\left[(1-\alpha)d - p\right]\right]}{K\left[(1-\alpha)d - p\right]\left[2\beta - 2\tau_1 - (1+\alpha)d + p\right]} \quad (19)$$

Clearly, relations (17)-(19) provide the settings of the desired PD-1F controller as functions of two adjustable parameters α and β, which must be selected in order to guarantee positive controller settings (in the case where the process parameters take positive values), as well as to fulfil inequalities (13) and (14). For a pre-specified value of $\alpha \in \Re$, parameter β can be selected in order to assign a specific damping ratio ζ_{des} of the closed-loop system. Indeed, using relations (12), (17) and the definition of β, and after some trivial algebra, one can resort the following quadratic equation with regard to β,

$$A_2\beta^2 + A_1\beta + A_0 = 0 \quad (20)$$

$$\begin{aligned} A_1 = {} & 4\zeta_{des}^2\left[\frac{2}{T_2 - q\left[(1-\alpha)d - p\right]} - 1\right]\left[(1-\alpha)d - p\right] \\ & - 2\left[1 + \frac{2q\left[(1-\alpha)d - p\right]}{T_2 - q\left[(1-\alpha)d - p\right]}\right] \\ & \times\left[d + \tau_1 - q + \frac{q\left[(1-\alpha)d - p\right]\left[2\tau_1 + (1+\alpha)d - p\right]}{T_2 - q\left[(1-\alpha)d - p\right]}\right] \end{aligned} \quad (21)$$

$$A_2 = \left[1 + \frac{2q\left[(1-\alpha)d - p\right]}{T_2 - q\left[(1-\alpha)d - p\right]}\right]^2 \quad (22)$$

$$\begin{aligned} A_0 = {} & \left[d + \tau_1 - q + \frac{q\left[(1-\alpha)d - p\right]\left[2\tau_1 + (1+\alpha)d - p\right]}{\tau_2 - q\left[(1-\alpha)d - p\right]}\right]^2 \\ & - 4\zeta_{des}^2\left[\frac{\tau_2\left[2\tau_1 + (1+\alpha)d - p\right]}{\tau_2 - q\left[(1-\alpha)d - p\right]} - \tau_1 - \alpha d\right]\left[(1-\alpha)d - p\right] \end{aligned} \quad (23)$$

Then, β is chosen as the maximum real root of (20)

Clearly, the method presented above is applicable when $p=\tau_Z$ or $-\tau_Z$ and q=1 or -1. However, extensive simulations show that, in the case where $\tau_Z >> 0$ (i.e. in the case of large overshoot processes), the method provides controller settings that renders the closed-loop unstable or marginally stable. This is due to the swings of the controller

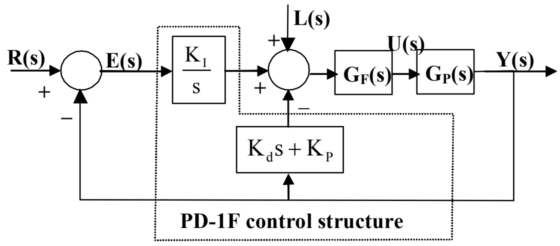

Figure 2: PD-1F control structure in the case of large overshoot response processes.

output induced by the excessive derivative action. One way to avoid this problem is to filter the controller output using a first order filter of the form (see Figure 2).

$$G_F(s) = 1/\left(\tau_F s + 1\right) \tag{24}$$

while calculating the PD-1F controller settings as suggested by relations (17)-(19). The time constant of the filter can be selected as $\tau_F = \tau_Z$.

Alternatively, one can select the controller settings according to the following method, which is a modification of the method resulting in the settings given by relations (17)-(19): In the case where the filter of the form (24) is introduced in the control loop, relation (3) is modified as

$$G_{CL}(s) = \frac{KK_I\left[\frac{ps+1}{(\tau_F s+1)(\tau_I s+1)}\right]\exp(-ds)}{s(\tau_2 s+q)+\left(KK_d s^2+KK_P s+KK_I\right)\left[\frac{ps+1}{(\tau_F s+1)(\tau_I s+1)}\right]\exp(-ds)}$$

Then, making use of the approximations

$$\left(ps+1\right)/\left(\tau_F s+1\right)\left(\tau_I s+1\right) \approx 1-(\tau_I-\bar{p})s$$

$$\exp(-ds) = \frac{\exp\left[-\left(1-\alpha\right)ds\right]}{\exp(\alpha ds)}$$

$$\left(\frac{ps+1}{\tau_F s+1}\right)\exp\left[-(1-\alpha)ds\right] = \left[\bar{p}-(1-\alpha)d\right]s+1$$

$$\left(\tau_I s+1\right)\exp(\alpha ds) = \left(\tau_I+\alpha d\right)s+1$$

where, $\bar{p} = p - \tau_F$, we finally obtain

$$G_{CL}(s) = \frac{\left[1-\left(\tau_I-\bar{p}\right)s\right]\exp(-ds)}{s\left(\frac{\tau_2}{KK_I}s+\frac{q}{KK_I}\right)+\left[\left(\frac{K_P}{K_I}-\tau_I-\alpha d\right)s+1\right]\left[\left[\bar{p}-(1-\alpha)d\right]s+1\right]} \tag{25}$$

It is now obvious that relation (25) is similar to relation (9) when p is replaced by $\bar{p} = p - \tau_F$. Therefore, following an argument similar to that used above to produce relations (17)-(19), we may easily conclude that, in the present case

$$K_I = \frac{\tau_2 - q\left[(1-\alpha)d-\bar{p}\right]}{K\left[(1-\alpha)d-\bar{p}\right]\left[2\beta-2\tau_I-\left(1+\alpha\right)d+\bar{p}\right]}$$

$$K_P = \frac{\beta\left[\tau_2 - q\left[(1-\alpha)d-\bar{p}\right]\right]}{K\left[(1-\alpha)d-\bar{p}\right]\left[2\beta-2\tau_I-\left(1+\alpha\right)d+\bar{p}\right]}$$

$$K_d = \frac{\left[\beta(\tau_I+\alpha d)-(\tau_I+\alpha d)^2\right]\left[\tau_2 - q\left[(1-\alpha)d-\bar{p}\right]\right]}{K\left[(1-\alpha)d-\bar{p}\right]\left[2\beta-2\tau_I-\left(1+\alpha\right)d+\bar{p}\right]}$$

Now, it only remains to select the filter time constant. A suitable choice of τ_F, is $\tau_F=\tau_Z$. With, this selection, the PD-1F controller settings in the case of large overshoot processes are obtained as suggested by the relations

$$K_I = \frac{\tau_2 - q\left[(1-\alpha)d\right]}{K\left[(1-\alpha)d\right]\left[2\beta-2\tau_I-\left(1+\alpha\right)d\right]} \tag{26}$$

$$K_P = \frac{\beta\left[\tau_2 - q\left[(1-\alpha)d\right]\right]}{K\left[(1-\alpha)d\right]\left[2\beta-2\tau_I-\left(1+\alpha\right)d\right]} \tag{27}$$

$$K_d = \frac{\left[\beta(\tau_I+\alpha d)-(\tau_I+\alpha d)^2\right]\left[\tau_2 - q\left[(1-\alpha)d\right]\right]}{K\left[(1-\alpha)d\right]\left[2\beta-2\tau_I-\left(1+\alpha\right)d\right]} \tag{28}$$

4 SIMULATION RESULTS

For assessment of the effectiveness of the proposed tuning methods and in order to provide a comparison with existing tuning methods, a series of simulation examples are carried out for different dead-time processes.

4.1 Inverse Response Processes with Two Stable Poles

Consider the typical inverse response process with K=1, τ_1=1, τ_2=1, d=0.8, p=-0.5, q=1. Applying the proposed method with α=0.6 and ξ_{des}= 0.8225, yields β=2.15. The PD-1F controller settings are then obtained as K_I=0.4221, K_P=0.9076 and K_d=0.4186. The settings of the series form PID controller with filtered derivative, tuned according the method proposed by Chien *et al* (2003), are K_C=0.3367, τ_I=1, τ_D=1, while the low-pass filter parameter takes the value a=0.1 and the inverse of the cyclic frequency of the desired critically damped closed-loop system takes the value τ_{cl}= 0.8348. The settings of the conventional PID controller that is tuned according to the method reported in Chen *et al* (2006), are K_C=0.71, τ_I=2, τ_D=0.5. Figure 3 illustrates the comparison of the servo-responses as well as of the regulatory control responses obtained by the proposed method and by the methods reported in Chien *et al* (2003) and Chen *et al* (2006), in the case of nominal process parameters. In case of set-point tracking, the proposed method provides a

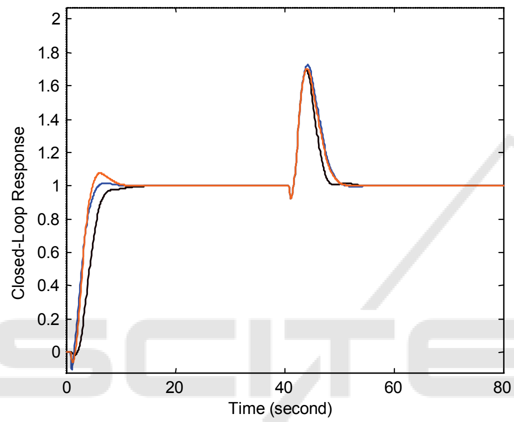

Figure 3: Servo-responses and regulatory control responses for the system $G_P(s)=(-0.5s+1)\exp(-0.8s)/(s+1)$, in case of nominal process parameters. Black line: Proposed Method. Orange line: Method in Chien *et al* (2003). Blue line: Method in Chen *et al* (2006).

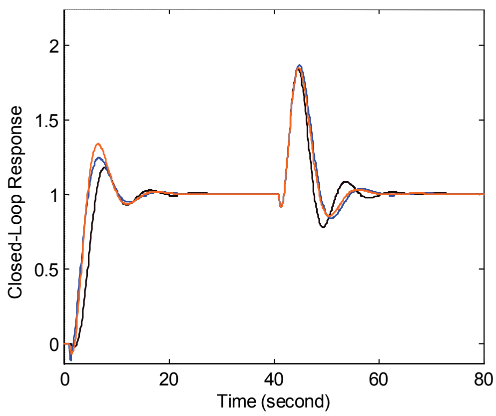

Figure 4: Servo-responses and regulatory control responses for the system $G_P(s)=(-0.5s+1)\exp(-0.8s)/(s+1)$, in case of a +20% mismatch in all process parameters. Other legend as in Figure 3.

slightly more sluggish response as compared to the abovementioned PID tuning methods, while the initial jump obtained by our method is smaller. In the case of regulatory control, our method gives a better response in terms of maximum error, while the settling time is comparable to that obtained by the methods in Chien *et al* (2003) and Chen *et al* (2006).

A comparison in terms of the ISE criterion, in the case of regulatory control, gives the values 1.2002 for the proposed method, while for the methods in Chien *et al* (2003) and Chen *et al* (2006), we obtain ISE=1.5782 and ISE=1.4425, respectively. The respective IAE values for the methods under comparison are obtained as 2.443, 3.058 and 2.8941. Figure 4 shows the comparisons of the servo-responses and of the regulatory control responses in the case where a simultaneous +20% uncertainty in all process parameters is assumed. The responses obtained by the proposed method are better in terms of overshoot, maximum error and initial jump, while the settling time is similar to that of the responses obtained by the PID controllers tuned according to the methods by Chien *et al* (2003) and Chen *et al* (2006). The ISE values, in case of regulatory control, are 1.9514, for the proposed method, 2.3765 for the method of Chien *et al* (2003) and 2.1673, for the method of Chen *et al* (2006). The respective IAE values are 3.8221, 4.2492 and 3.9183.

As already mentioned, for a pre-specified value of adjustable parameter α, parameter β is directly related to the damping ration ζ of the second order approximation (10) of the closed-loop system. In

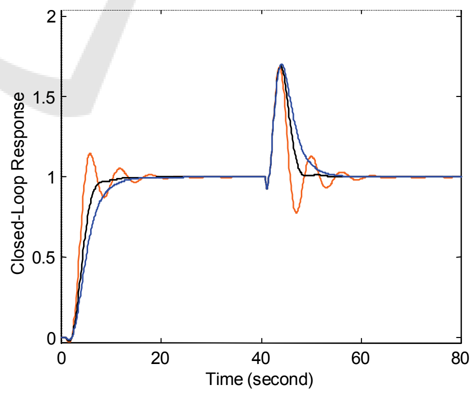

Figure 5: Servo-responses and regulatory control responses for the system $G_P(s)=(-0.5s+1)\exp(-0.8s)/(s+1)$, in case of nominal process parameters, for α=0.6 and for three values of β. Orange line: β=2.05; Black line: β=2.15;. Blue line: β=2.25.

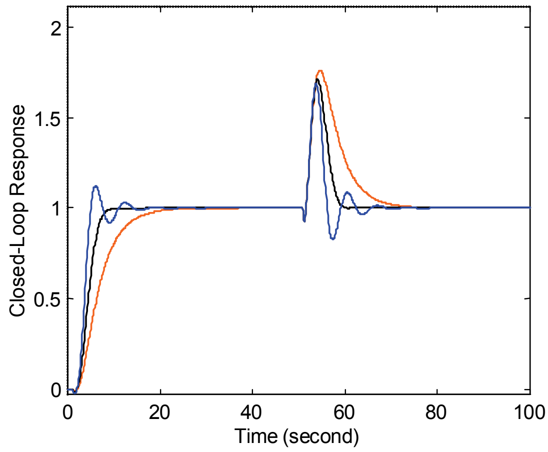

Figure 6: Servo-responses and regulatory control responses for the system $G_P(s)$=(-0.5s+1)exp(-0.8s)/(s+1), in case of nominal process parameters, for β=2 and for three values of α. Orange line: α=0.45; Black line: α=0.5;. Blue line: α=0.55.

particular, as shown in Figure 5, β increases when ζ is increased. This of course results to a more conservative PD-1F controller. Therefore, a greater value of β, renders the closed-loop system more robust. Parameter α has an inverse effect on the closed-loop system robustness: For a pre-specified value of the parameter β, an increase of the parameter α, leads to a less robust but faster closed-loop system, as illustrated in Figure 6.

4.2 Control of a Continuous Stirred Tank Reactor

Let us consider the transfer function model of a CSTR reported in Padma Sree and Chidambaram (2004), and having the form

$$G_P(s) = \frac{-2.07(0.1507s+1)}{2.85s^2+2.31s-1}\exp(-0.3s) = \frac{-2.07(0.1507s+1)}{(0.8905s+1)(3.2005s-1)}\exp(-0.3s)$$

The process has one dominant unstable pole and one stable pole, at s=0.3125 and s= -1.123, respectively, as well as a stable zero -6.6357. Here, K=-2.07, τ_1=0.8905, τ_2= 3.2005, d=0.3, p=0.1507, q=-1. Application of the proposed method with α=-0.5, β=2.5, yields the PD-1F controller settings K_P=-4.3862, K_I=-1.7545, K_d=-2.2859. The settings of the set-point weighted PID controller tuned according to the method reported in Padma Sree and Chidambaram (2004) are, K_C=-0.7205, τ_I=39.7228, τ_D=0.1494, while the tuning parameter used in the above mentioned paper, as well as the set-point weight b, take the values 0.15 and 0.3275, respectively. Figure 7 illustrates the servo-responses obtained by the two controllers. Figure 8 shows the comparison of the regulatory control responses for a negative unit step load change. Obviously, the PD-1F controller tuned according to the proposed method provides a considerably better performance, particularly in the case of regulatory control, where the response obtained by the controller tuned according to the method in Padma Sree and Chidambaram (2004) is practically unacceptable.

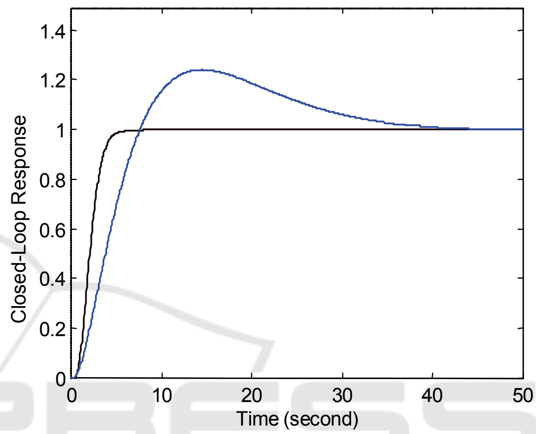

Figure 7: Closed-loop servo-responses of the CSTR model. Black line: Proposed method. Blue line: Set-point weighted PID controller tuned according to the method proposed by Padma Sree and Chidambaram (2004).

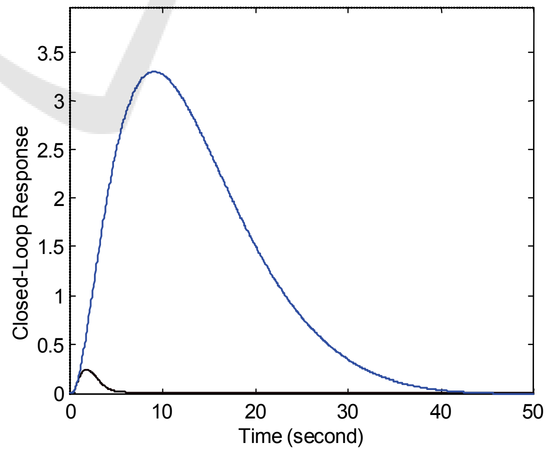

Figure 8: Regulatory control responses of the CSTR. Other legend as in Figure 7.

4.3 Second Order Unstable Process with a Positive Zero

Consider the process with K=1, τ_1=2.07, τ_2=5, d=0.939, p=-1, q=-1. The process has a stable pole, an unstable pole and a strong non-minimum phase zero. To the authors' best knowledge, controller design for second order processes with one or two righ-half-plane poles and a right-half-plane zero has not yet been addressed in the literature. Application of the proposed method, with α=0.3 and β=25, yields the PD-1F controller settings K_I=0.0920, K_P=2.3012, K_d=4.9027. The process model is next approximated as $G_P(s) = \exp(-1.939s)/\left[(2.07s+1)(5s-1)\right]$, i.e. the negative numerator time constant has been approximated as a time delay term of the form exp(-s). This is reasonable since an inverse response has a deteriorating effect on control similar to that of a time delay. We next apply the method reported in Lee *et al* (2000), in order to design a PID controller with first order set-point filter for the given process, on the basis of the approximated model. Application of the method reported in Lee *et al* (2000), with the IMC parameter λ=6.25, yields the PID controller settings K_C=1.9570, τ_I=34.9614 and τ_D=2.4889. Figure 9 illustrates the comparison of the servo-responses and the regulatory control responses for a unit step set-point change at t=0 sec and an inverse unit step load change at t=75 sec. It is seen that the proposed method results in an improved load disturbance response as compared to the method in Lee *et al* (2000), while the set-point responses are similar, with comparable settling times.

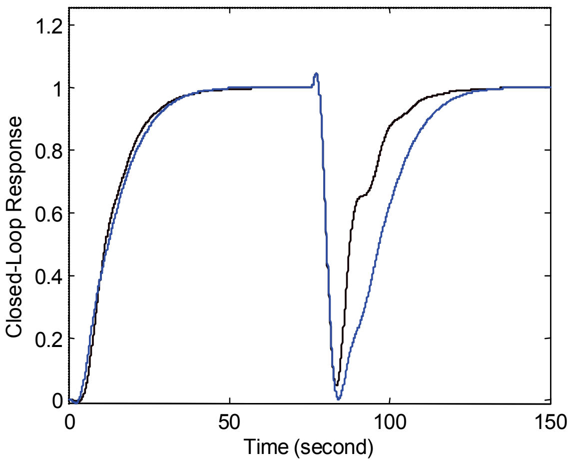

Figure 9: Servo-responses and regulatory control responses for the system G(s)=(-s+1)exp(-0.939s) / [(2.07s+1)(5s-1)]. Black line: Proposed method; Blue line: Method in Lee *et al* (2000).

4.4 Stable Second Order Unstable Process with a Positive Zero

Consider the process model of the form (2), with K=1, τ_1=2, τ_2=1, d=1, p=0.3, q=1. Application of the proposed method with α=0.4, β=3, yields K_I=1.0358, K_P=3.1073, K_d=1.6340. The settings of the series form PID controller with filtered derivative, tuned according the method proposed by Chien *et al* (2003) are K_C=1.0355, τ_I=2, τ_D=1, while the low-pass filter parameter takes the value τ_F=0.3 and the inverse of the cyclic frequency of the desired critically damped closed-loop system takes the value τ_{cl}= $d/\sqrt{2}$ = 0.5457. Figure 10 illustrates the comparison of the servo-responses as well as of the regulatory control responses obtained by the proposed method and by the method reported in Chien *et al* (2003). In the regulatory control case our method gives a considerably better response, whereas, although our method provides a smooth response, the method in Chien *et al* (2003) is better in the case of set-point tracking.

Let us now consider the case of a large overshoot process with K=1, τ_1=2, τ_2=1, d=1.2, p=5, q=1. Evaluating relations (17)-(19), while assuming α=0.2, β=3, yields the PD-1F controller settings K_I=0.2208, K_P=0.6624, K_d= 0.3759. Application of the above controller yields an unacceptable oscillatory response, as shown in Figure 11. Let us try, another design by evaluating relations (17)-(19) in the case where we select a=0.6, β=3. This yields K_I=0.1951, K_P=0.5853, K_d= 0.1486, i.e. a more conservative controller. The obtained servo-response

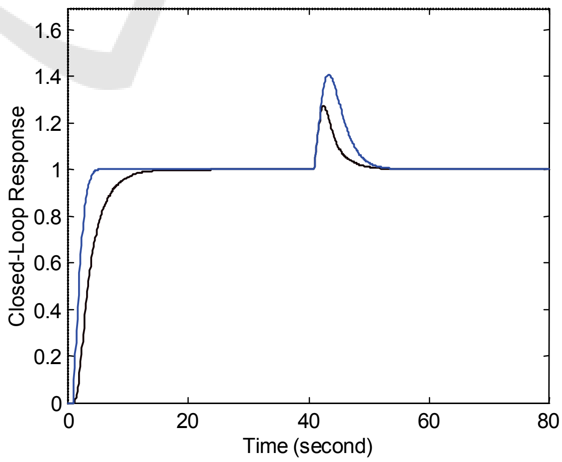

Figure 10: Servo-responses and regulatory control responses for the system $G_P(s)$=(0.3s+1)exp(-0.8s) /(2s+1)(s+1). Black line: Proposed method. Blue line: Method in Chien *et al* (2003).

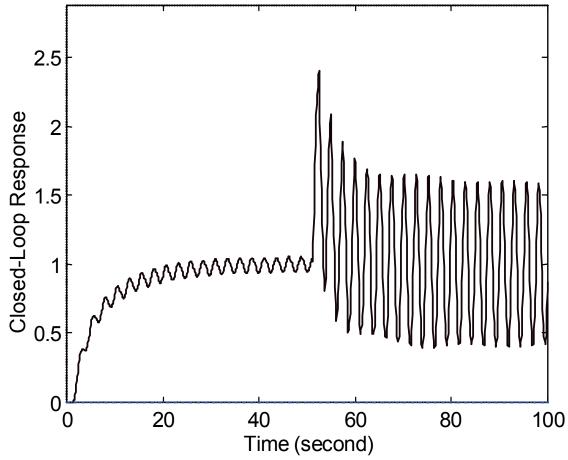

Figure 11: Closed-loop servo-response and regulatory control response of the system G(s)=(5s+1)exp(-1.2s) /(2s+1)(s+1), in the case of he PD-1F controller with parameters K_I=0.2208, K_P=0.6624, K_d= 0.3759.

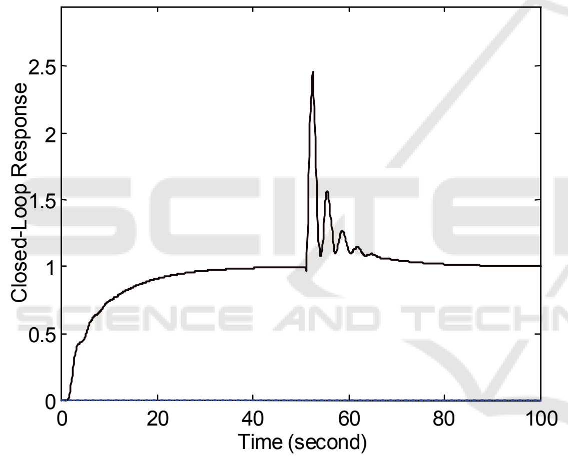

Figure 12: Closed-loop servo-response and regulatory control response of the system G(s)=(5s+1)exp(-1.2s) /(2s+1)(s+1), in the case of he PD-1F controller with parameters K_I=0.1951, K_P=0.5853, K_d= 0.1486.

and regulatory control responses are given in Figure 12. In the later case, the servo-response is quite smooth while the regulatory control response is less oscillatory. However, the robustness of the closed-loop system is marginal, and a small parameter mismatch can readily lead to instability.

Let us now consider filtering the output of the PD-1F controller that is designed for the case where α=0.2, β=3, with settings K_I=0.2208, K_P=0.6624, K_d= 0.3759, by a filter of the form (24), where τ_F=5. Moreover, let us design a PD-1F controller with filtered output as suggested by relations (26)-(28), with α=-0.2, β=2, τ_F=5. In this case the controller

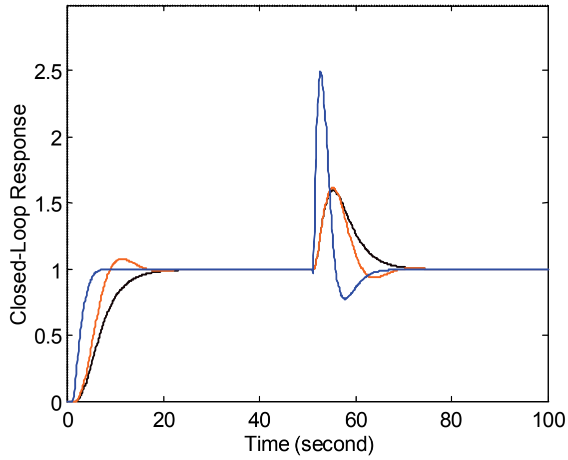

Figure 13: Closed-loop servo-response and regulatory control responses of the system G(s)=(5s+1)exp(-1.2s) /(2s+1)(s+1). Black line: PD-1F controller with filtered output tuned according to relations (17)-(19); Orange line: PD-1F controller with filtered output tuned according to relations (26)-(28); Blue line: Series form PID controller with filtered derivative tuned according to the method in Chien *et al* (2003).

settings are K_I=0.3183, K_P=0.6366, K_d= 0.1344. Figure 13 shows the obtained servo-responses and regulatory control responses for both designs, together with the respective responses obtained by a series PID controller with filtered derivative, designed according the method reported in Chien *et al* (2003). It is seen that, in the regulatory control case our method gives a considerably better response, whereas, although our methods provide smooth responses, the method in Chien *et al* (2003) is better in the case of set-point tracking. A comparison in terms of ISE in the case of regulatory control gives the ISE values 1.8326 and 1.5892, for the proposed methods and 4.5754 for the method in Chien *et al* (2003). The respective IAE values are 4.5287, 3.7058 and 4.9453.

5 CONCLUSIONS

A new direct synthesis method of tuning the PDF controller for stable or unstable dead-time processes with a negative or a positive zero has been presented. The proposed tuning method ensures smooth closed-loop response to set-point changes, fast regulatory control and sufficient robustness against parametric uncertainty. Numerical simulation examples verify the advantages of the proposed method over known PID controller tuning methods for the classes of dead-time processes under study. Extension of the proposed tuning method in the case of frequency domain specifications of the closed-loop system in terms of gain and phase margins is currently under investigation.

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