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Brief Paper

Obtaining controller parameters for a new Smith predictor using autotuning☆

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Abstract

The paper extends recent work on a modified Smith predictor strategy, which leads to significant improvements in its regulatory capacities for reference inputs and disturbances. High-order or long dead time stable, integrating and unstable plants are modelled as lower-order plant models with a longer time delay. The controllers are designed so that the delay-free component of the output is tuned to be either a first- or second-order response if there are no modelling errors in the assumed plant transfer function. Plant model transfer functions and the controller parameters are estimated using exact analysis from the peak amplitude and frequency of the process output obtained from a single-relay feedback test. Illustrative examples show the simplicity and superiority of the proposed controller design method over previously published approaches both for the setpoint response and for the load disturbance rejection. © 2000 Elsevier Science Ltd. All rights reserved.

Keywords: New Smith predictor; Relay; Autotuning; Stable process; Unstable process; Integrating process

1. Introduction

PID controllers are commonly used in practice in the process control industries. They can be tuned by using rules of thumb (Ziegler & Nichols, 1942) or formulae resulting from analytical design (Åström & Hägglund, 1984). PID controllers tuned using these conventional methods, however, may not provide satisfactory closed-loop responses when the plant considered is either a high-order plant or a plant with a long dead time. Another simple and powerful control technique is the Smith predictor. The controller can be designed as if the system were delay free. However, it was pointed out by Watanabe and Ito (1981) that these regulators cannot reject load disturbance for processes with integration. Subsequently, Åström, Hang and Lim (1994) presented a modified structure for the control of integrator and long dead time processes. This structure decouples the disturbance response from the setpoint response and thereby improves the disturbance response. Later on Mataušek and Micić (1996) proposed a structure similar to that of Åström et al. (1994) but having an additional feedback path from the difference of the plant output and the model output to the reference input for controlling integrating processes. Recently, the limitations of PID controllers controlling resonant, integrating and unstable plants in a conventional feedback structure have been studied (Kwak, Sung & Lee, 1997; Park, Sung & Lee, 1998; Atherton & Majhi, 1999a). These references use an internal feedback loop to convert the integrating or unstable process to an open-loop stable process first and then use a PID controller in the forward loop for improved setpoint and disturbance responses. Majhi and Atherton (1998b) therefore proposed a new Smith structure incorporating a similar inner feedback loop which extends its applicability to resonant, integrating and unstable processes.

Recently, relay feedback automatic tuning of SISO controllers has been studied extensively. Palmor and Blau (1994) developed an autotuning algorithm for the Smith dead time compensator using a first-order plus delay model for some stable plants. The algorithm

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estimates two points on the process Nyquist curve via automatically generated controlled limit cycles. The points are then used in a least-squares procedure to estimate the parameters of the model. Hang, Wang and Cao (1995) have presented methods to autotune and self-tune a modified Smith predictor by relay feedback. Both these works, however, use the describing function approximation to a relay in their analysis, which can result in errors in the estimation of the plant model parameters. Also their results are limited to stable processes only.

In this paper, a simple relay feedback autotuning method is proposed for the new Smith predictor. With prior information on the static gain, a reduced order process model in terms of a first- or second-order dynamics plus dead time (abbreviated as FOPDT and SOPDT, respectively) can be computed and used to autotune the Smith predictor from a single symmetrical relay test. Excellent performance of the autotuned Smith predictor has been substantiated by simulations for stable, integrating and unstable processes.

2. The new Smith predictor structure

The structure of the new Smith predictor (Majhi & Atherton, 1998b) for controlling stable, unstable and integrating processes is shown in Fig. 1. It has three controllers which are designed for different objectives and of the three controllers, G_{c1} in the inner loop is provided to stabilise an unstable or integrating process and modify the pole locations of the transfer function of a stable process. The other two controllers, G_c and G_{c2} are then used to take care of servo-tracking and disturbance rejection, respectively, by considering the inner loop as an open-loop stable process. It is similar to the structure of Mataušek and Micić when $G_{c1} = 0$ and by setting $G_{c1} = G_{c2} = 0$, the standard Smith predictor is obtained.

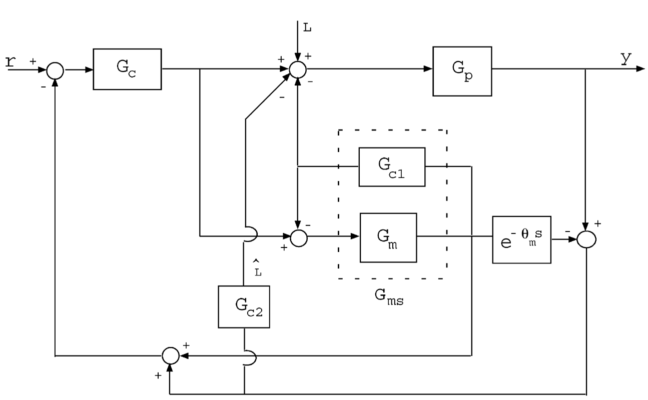

Fig. 1. The new Smith predictor structure.

Let $G_m(s)e^{-\theta_m s}$ and $G_p(s)$ be the transfer functions of the plant model and the plant, respectively. Based on the assumption that the model used perfectly matches the plant dynamics, i.e. $G_m(s)e^{-\theta_m s} = G_p(s)$, the closed-loop response to setpoint and disturbance inputs is given by

$$Y_r(s) = \frac{G_c G_m e^{-\theta_m s}}{1 + G_m(G_c + G_{c1})} = Y'_r(s)e^{-\theta_m s}, \quad (1)$$

$$Y_L(s) = \frac{G_m e^{-\theta_m s}}{1 + G_m(G_c + G_{c1})} \frac{1 + G_m(G_c + G_{c1}) - G_c G_m e^{-\theta_m s}}{1 + G_m G_{c2} e^{-\theta_m s}} = Y'_L(s)e^{-\theta_m s}. \quad (2)$$

3. Estimation of plant model parameters

In principle, from an odd symmetrical limit cycle two unknown parameters can be found, and from an asymmetrical limit cycle four can be found. Also by doing two odd symmetrical limit cycle tests, one without hysteresis and one with, four parameters can be found but this requires extra time for the two tests. Further, estimation of four unknown parameters from an asymmetrical test or two symmetrical tests requires a nonlinear algebraic equation solver to solve the equations which may result in convergence to a false solution if poor initial estimates are used and in an application requires significantly more controller software. Thus since the computations required in the controller become more complicated for other than the single odd symmetrical limit cycle test this is therefore to be preferred if possible. Unfortunately in some instances erroneous results can be obtained using this method if describing function analysis is applied either to estimate the critical point or two plant parameters (Atherton, 1997). Accurate evaluation of parameters is, however, possible assuming the plant model has a known form and there are no measurement errors. This is based on the fact that exact expressions for the limit cycle waveform can be obtained. Using the state space approacch the authors have in recent publications (Atherton & Majhi, 1998a; Majhi and Atherton, 1999b) derived exact relationships to identify an SOPDT model plant transfer function

$$G_p(s) = \frac{k e^{-\theta s}}{(T_1 s \pm 1)(T_2 s + 1)}. \quad (3)$$

The resulting expressions for an odd symmetrical limit cycle for the plant in a feedback loop with an ideal relay are

$$\frac{1}{\alpha_1}\left[1 - \frac{2e^{\alpha_1(T-\theta)}}{1 + e^{\alpha_1 T}}\right] - \frac{1}{\alpha_2}\left[1 - \frac{2e^{\alpha_2(T-\theta)}}{1 + e^{\alpha_2 T}}\right] = 0, \quad (4)$$

where $\alpha_1 = \mp 1/T_1$, $\alpha_2 = -1/T_2$, h is the relay amplitude, A_p is the peak amplitude of the oscillation, and T is the half-period of the oscillation. The expression for the peak amplitude is

$$A_p = \mp kh[1 - 2(R_1^{-\alpha_2/(\alpha_1-\alpha_2)} R_2^{\alpha_1/(\alpha_1-\alpha_2)})], \tag{5}$$

where $R_1 = 1/(1 + e^{\alpha_1 T})$, $R_2 = 1/(1 + e^{\alpha_2 T})$ and $\alpha_1 \neq \alpha_2$. Eqs. (4) and (5) with the R values substituted can be solved from a symmetrical relay test method involving measurement of A_p, T for identification of two unknowns. The following transfer function models which can be obtained from (3) using suitable limiting values for α_1 and α_2 and which may be good approximations for some plants, can also be used with these formulae.

$$G_1(s) = \frac{k e^{-\theta s}}{(T_1 s \pm 1)}, \tag{6}$$

$$G_2(s) = \frac{k e^{-\theta s}}{s}, \tag{7}$$

$$G_3(s) = \frac{k e^{-\theta s}}{s(T_1 s + 1)}. \tag{8}$$

The areas of the process output signal, a_y and input signal, a_u over a half-period of time can be measured, which allows estimation of the steady-state gain k. It has been shown that (Majhi & Atherton, 1999b) if the plant is of the form of (6) then k can be found from the expression

$$\frac{a_y = \int_0^T y(t)\,dt}{a_u = \int_{t_p}^{T+t_p} u(t)\,dt} = k, \tag{9}$$

where t_p is the time at which the peak amplitude appears after a zero crossing. This method unfortunately does not give good results when the plant dynamics differs from Eq. (6) and an alternative approach is required to find k. There are several possibilities which include performing an additional relay test, open-loop step response test (for stable processes), or using data which may have been obtained from the design or from monitoring of the plant input and output signals under normal operating conditions. For $G_1(s)$, setting $\alpha_2 \to -\infty$ and $\alpha_1 = \mp 1/T_1$ in (4) and (5) give

$$T_1 = \frac{\mp T}{\ln\left(\dfrac{1 \mp \mu}{1 \pm \mu}\right)}, \tag{10}$$

$$\theta = \mp T_1 \ln(1 \mp \mu) = T \ln(1 \mp \mu)/\ln\left(\frac{1 \mp \mu}{1 \pm \mu}\right), \tag{11}$$

where the normalised peak amplitude, $\mu = A_p/(kh)$. (10) and (11) show that μ must not be greater than 1 to obtain a solution for T_1 and θ. When the value of $\mu = 1$ is substituted in (11) boundary conditions of $\theta/T_1 \leq 0.693$ for the unstable FOPDT process and $\theta/T_1 \leq \infty$ for the stable FOPDT process are obtained which give the conditions for existence of limit cycles in these cases. Using (4) and (5), the parameters of the integrator with dead time model, $G_2(s)$ are given by

$$k = \frac{A_p}{h\theta}, \quad \theta = T/2. \tag{12}$$

In some situations, $G_2(s)$ may not provide a good model for an integrating process involving a number of poles and zeros. Therefore, the model, $G_3(s)$, which has three parameters may be considered to represent a higher-order integrating plant. It is easy to derive from the expressions given in (4) and (5) setting $\alpha_1 \to 0$ the following nonlinear equations:

$$e^{(T/2-\mu)/T_1} + e^{-(T/2+\mu)/T_1} - 2 = 0, \tag{13}$$

$$(\theta - T/2)/T_1 - e^{(\theta - T/2)/T_1} e^{-\mu/T_1} + 1 = 0 \tag{14}$$

for $T_1 \neq 0$. T_1 can be estimated from (13) from the measurements of T and μ. Then (14) is used to evaluate θ. The nonlinear equations are found to be convex and the NESOLVE routine in MATLAB gives accurate results since they are solved independently.

4. Development of the autotuning formulae

The form of the main PI controller is $G_c = K_p(T_i s + 1)/(T_i s)$. The controllers G_{c1} and G_{c2} take different forms depending upon the assumed order and type of the plant transfer function model. In two cases G_{c1} is taken as an ideal PD controller and the results are given for this idealised case. For practical implementation a time constant filter is used with the derivative term and if this is some 10 times smaller than the derivative time constant, which is normally possible in practice, the results do not change significantly. It is perhaps important to stress here that with the derivative term in the feedback path choice of the filter time constant is based on noise considerations unlike the situation for the derivative term in the forward path where the effect of the derivative 'kick' for a step input also has to be considered. In this section, very simple but straightforward tuning formulae are derived for controllers for the aforementioned plants whose parameters are estimated as described in the previous section from relay feedback measurements.

4.1. For the plant assumed as a stable FOPDT

When the plant transfer function is modelled as the stable FOPDT transfer function $G_1(s)$ so that

$$G_m = \frac{k}{(T_1 s + 1)} \tag{15}$$

using (1) and G_c and letting $T_i = T_1$, the closed-loop transfer function with no delay, $Y'_r(s)$, for setpoint tracking can be written as

$$Y'_r(s) = \frac{1}{(T_1/kK_p)s + 1} = \frac{1}{\lambda s + 1}, \tag{16}$$

where $\lambda > 0$ is an adjustable closed-loop design parameter whose value is typically chosen approximately equal to the time constant of $G_m(s)$, i.e. $\lambda \approx T_1$. This is done since the Smith predictor approach would normally be used for stable plants with time delay greater than the time constant and speeding up the closed-loop response with respect to the open loop normally results in overshoot. The response speed is determined by the parameter λ. The choice of a higher value of λ slows down the system and makes it more robust while small values of λ may cause instability of the modified Smith predictor in the presence of unmodelled dynamics. Further, under perfect matching conditions, the magnitude of the initial control effort is inversely proportional to λ. From (16), one can write

$$kK_p = \frac{T_1}{\lambda}. \tag{17}$$

Since the plant is a stable one, there is no need for the controllers G_{c1} and G_{c2}. Using G_m and G_c the delay free part of (2) can be written

$$Y'_L(s) = \frac{k(\lambda s + 1 - e^{-\theta_m s})}{(\lambda s + 1)^2}. \tag{18}$$

Therefore, the controller is expected to also give approximately a first-order response to a load disturbance input. The design procedure is illustrated below in a sample example.

Example 1. Consider the high-order transfer function (Hang et al., 1995), $G_p = 1/(s+1)^{20}$. From the simulation, a single relay feedback test gives $T = 19.660$ and $A_p = 0.9484$ for an ideal relay amplitude $h = 1$ (In all the simulation studies, A_p is obtained from the measurement of peak-to-peak amplitude of the limit cycle output). This gives $\mu = 0.9484$ and (10) and (11) estimate the FOPDT transfer function model as $G_m e^{-\theta_m s} = e^{-16.0483s}/5.414s + 1$. Using the describing function approach for limit cycle analysis Hang et al. (1995) obtain a SOPDT model as $G_m e^{-\theta_m s} = e^{-11.85s}/(3.60s+1)^2$. The Nyquist plots of the original model and the reduced order models are shown in Fig. 2. They have suggested a PI controller with the parameters $K_p = 0.5$ and $T_i = 3.60$. The proposed method gives $K_p = 1$ and $T_i = 5.414$ choosing $\lambda = 5.414$. The magnitude of the load disturbance is $L = -0.5$. The responses of the closed-loop system for these controller settings are given in Fig. 3. The response by the proposed method is superior to that of Hang et al. (1995) automatically tuned PI controller. The ISE performance index for the proposed method is 3.516 compared to 3.726 obtained by the latter method for the load disturbance response. Since the above two control methods use the same control structures and the same design methods, the control benefit comes from improved modelling of the plant using the exact relay analysis.

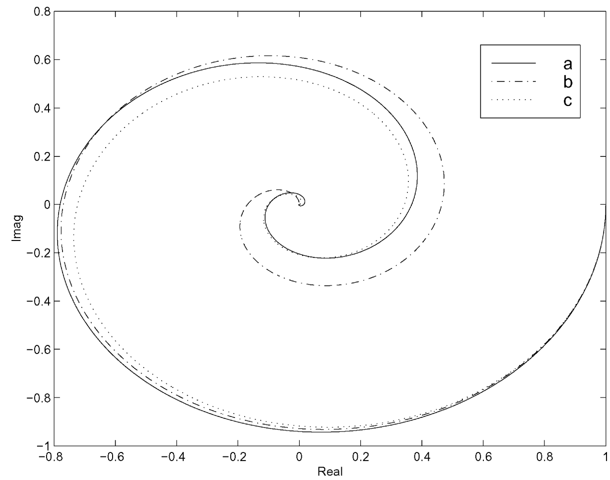

Fig. 2. Nyquist Plots of (a) original plant, (b) proposed reduced model and (c) reduced model of Hang et al.

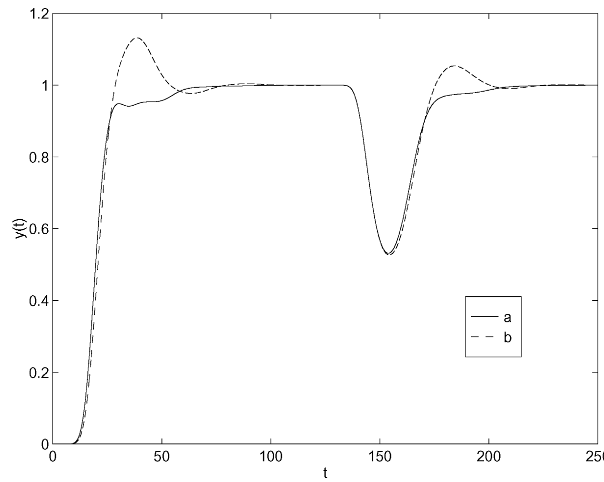

Fig. 3. Responses of Example 1: (a) proposed PI controller and (b) Hang's PI controller.

4.2. *For the plant assumed as an integrator with long dead time*

For the plant transfer function $G_2(s)$, considering the inner-loop controller $G_{c1} = K_f$, the modified process becomes

$$G_{ms} = \frac{k}{(s + kK_f)}. \tag{19}$$

Using (24) and G_c, and letting $T_i = 1/(kK_f)$, the delay free part of (1) becomes

$$Y'_r(s) = \frac{1}{[1/kK_p]s + 1} = \frac{1}{\lambda s + 1}, \quad (20)$$

where λ is the closed-loop parameter. The controller parameter $K_p = 1/(k\lambda)$ can be obtained from a suitable choice of λ while constraining of $kK_f = 1$ gives values of the parameters K_f and T_i. The expression for K_p shows that the initial control effort can be reduced significantly by choosing $1 < \lambda < 5$ which inherently brings robustness to the controller. It is evident from (2) that since the plant is an integrating one, the controller $G_{c2} = K_d$ is necessary for a satisfactory load disturbance rejection. Further, since the roots of the factor $(1 + G_m(G_c + G_{c1}))$ have been placed properly by the earlier design, the controller G_{c2} is designed applying the Nyquist stability criterion to the second factor of the denominator of (2), i.e.

$$1 + G_m G_{c2} e^{-\theta_m s} = 1 + \frac{kK_d e^{-\theta_m s}}{s} = 0. \quad (21)$$

Choosing K_d to give a phase margin, ϕ_m, of 60° (Mataušek and Micić, 1996) gives

$$K_d = \frac{\pi - 2\phi_m}{2k\theta_m} = \frac{0.5236}{k\theta_m}. \quad (22)$$

It has to be noted that ϕ_m is not the phase margin corresponding to the system open-loop transfer function and the choice of $\phi_m = 60°$ provides satisfactory closed-loop system performance both with respect to the set-point response and the load disturbance rejection. The following example illustrates the performance of the proposed autotuning method.

Example 2. Let the integrating process have the parameters $k = 1$ and $\theta = 5$. From a simulation with an ideal relay with $h = 1$, $T = 9.9987$ and $A_p = 4.999$ were measured. This gives from (12) the FOPDT model $G_m(s) = e^{-5s}/s$ to within 10^{-3}. Then the controller parameters are $K_p = 0.5$, $T_i = 1$, $K_f = 1$ and from (27) $K_d = 0.105$ assuming λ to be 2 to decrease the sensitivity of the system to plant parameter perturbations. The magnitude of the load disturbance is $L = -0.1$. The responses of the closed-loop system for these controller settings for the actual plant and with $\pm 10\%$ estimation error in the time delay are given in Fig. 4. The results are similar to the ones obtained by Åström et al. (1994) and Mataušek and Micić (1996) since there is no error in the plant model estimation

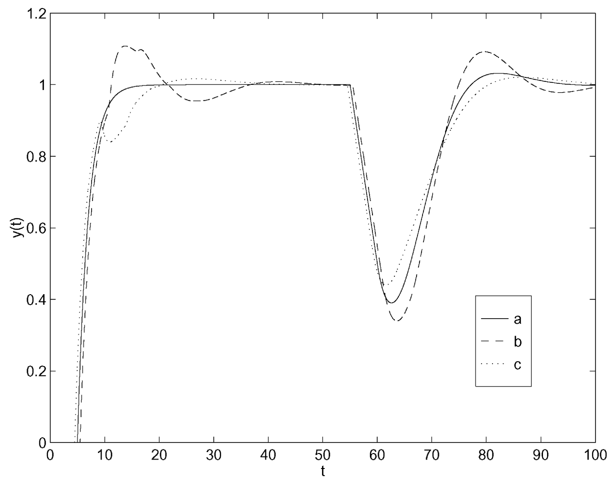

Fig. 4. Responses of Example 2: (a) for nominal $\theta = 5$, (b) for 10% positive perturbation and (c) for 10% negative perturbation in θ.

4.3. For the plant assumed as an integrating SOPDT

When the plant transfer function is modelled as an integrating SOPDT transfer Function $G_3(s)$, then with the inner-loop controller $G_{c1} = K_f(T_f s + 1)$ the modified process becomes

$$G_{ms} = \frac{k}{(s + kK_f)(T_1 s + 1)} \quad (23)$$

assuming $T_f = T_1$. Using G_c, and letting $T_i = T_1$, the delay free part of the closed-loop transfer function for setpoint tracking given in (1) can be written as

$$Y'_r(s) = \frac{1}{(T_i/kK_p)(s^2 + kK_f s + 1)} = \frac{1}{(\lambda s + 1)^2}, \quad (24)$$

where λ is the closed-loop design parameter which is chosen $\lambda \approx T_1$. Then the controller parameters K_p and K_f are obtained from

$$kK_p = T_1/\lambda^2, \quad (25)$$

$$kK_f = 2/\lambda. \quad (26)$$

The controller $G_{c2} = K_d(T_d s + 1)/(T_d/Ns + 1)$ (where $T_d/N \ll 1$ and can be neglected for analysis) is necessary for a satisfactory load disturbance rejection whose design parameters are obtained following similar procedure as discussed earlier. For the typical process

$$1 + G_m G_{c2} e^{-\theta_m s} = 1 + \frac{kK_d e^{-\theta_m s}}{s} = 0 \quad (27)$$

when $T_d = T_1$. Choosing K_d to give a phase margin, ϕ_m, of 60° gives

$$K_d = \frac{\pi - 2\phi_m}{2k\theta_m} = \frac{0.5236}{k\theta_m}. \quad (28)$$

Similarly, load response $Y'_L(s)$ can be obtained like the previous case. This procedure is illustrated by the following example.

Example 3. Consider the integrating process $G_p = e^{-5s}/(s(s+1)^2(3s+1))$. The simulation with an ideal relay with $h = 1$ gives $T = 19.8083$ and $A_p = 7.494$. Then from (13) and (14), the SOPDT model becomes $G_m e^{-\theta_m s} = e^{-6.5672s}/s(3.4945s+1)$. Assuming $\lambda = 3.4945$, then the controller parameters for the integrating SOPDT model become $K_p = 0.2862$, $T_i = 3.4945$, $K_f = 0.5723$ and $T_f = 3.4945$. The controller parameters of G_{c2} become $T_d = 3.4945$ and $K_d = 0.0797$. The magnitude of the load disturbance is $L = -0.1$. The response of the closed-loop system for these controller settings is given in Fig. 5 for the actual plant and also ones with the time delay in error by $\pm 10\%$ for $T_d/N = 0.01$. The performance of the proposed technique is compared with that of Mataušek and Micić (1996) in Fig. 6 and found to be superior because of the modelling, the new Smith predictor structure and the design method used.

4.4. For the plant assumed as an unstable FOPDT

When the plant transfer function is modelled as an unstable FOPDT transfer function, where

$$G_m = \frac{k}{(T_1 s - 1)} \tag{29}$$

considering $G_{c1} = K_f(T_f s + 1)$ and assuming $kK_f = 2$, the modified process due to the addition of the inner loop becomes

$$G_{ms} = \frac{k}{(T_1 + 2T_f)s + 1}. \tag{30}$$

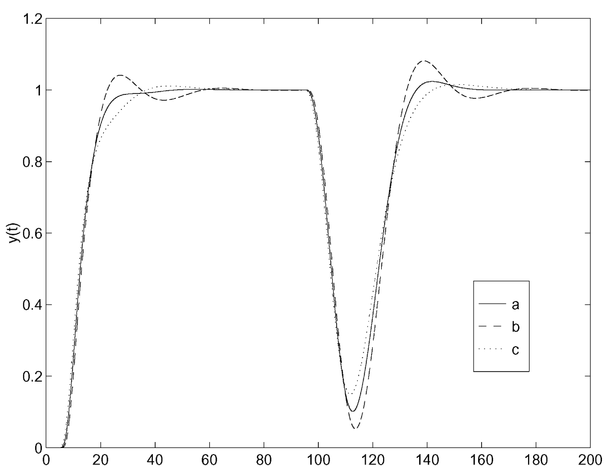

Fig. 5. Step setpoint and disturbance responses for Example 3: (a) for nominal $\theta = 5$, (b) for 10% positive perturbation and (c) for 10% negative perturbation in θ.

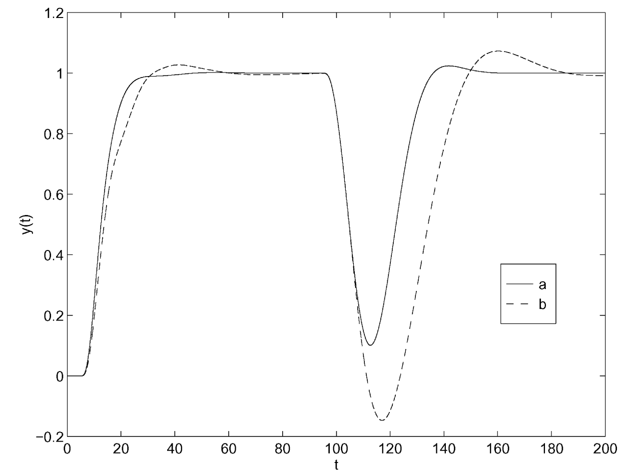

Fig. 6. Step setpoint and disturbance responses for Example 3: (a) proposed method, (b) Mataušek and Micić's method.

Using G_c, and letting $T_i = T_1 + 2T_f$, the delay free closed-loop transfer function for setpoint tracking given in (1) can be written as

$$Y'_r(s) = \frac{1}{[(T_1 + 2T_f)/kK_p]s + 1} = \frac{1}{\lambda s + 1}, \tag{31}$$

where λ is the closed-loop design parameter. Unlike the situation for stable plants the Smith predictor approach would normally be used for plants with time delay less than the time constant and speeding up the closed-loop response with respect to the open loop may be desirable. Assuming $\lambda = T_1 + 2T_f$ which becomes the time constant of the open-loop stable process model G_{ms} given by (30), then the controller parameters are obtained for a chosen value of λ. Further, the controller G_{c1} is helpful in controlling an unstable plant having two poles (one stable and one unstable). The controller $G_{c2} = K_d$, whose prime job is to reject unwanted load disturbances, is designed on the basis of stabilisation of the second part of the characteristic equation of (2)

$$1 + G_m G_{c2} e^{-\theta_m s} = 1 + \frac{kK_d e^{-\theta_m s}}{T_1 s - 1} = 0. \tag{32}$$

De Paor and O'Malley (1989) suggested a proportional controller, K_d, for the stabilisation of an unstable FOPDT process based on the optimum phase margin criterion which gives

$$K_d = \sqrt{\frac{T_1}{\theta_m k^2}} \tag{33}$$

with the constraint $\theta/T_1 < 1$. The delay free part of (2) can be written

$$Y'_L(s) = \frac{k(\lambda s + 1 - e^{-\theta_m s})}{(\lambda s + 1)(\lambda s - 1 + kK_d e^{-\theta_m s})} \tag{34}$$

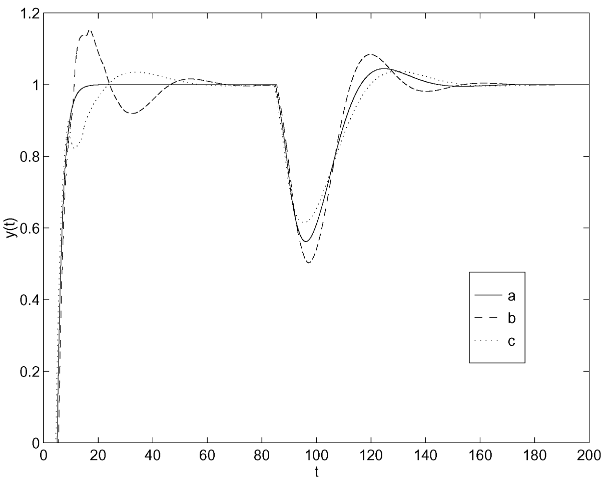

Fig. 7. Responses for Example 4: (a) for nominal $\theta = 5$, (b) for 10% positive perturbation and (c) for 10% negative perturbation in θ.

(34) shows that if $K_d = 0$, the load response $Y'_L(s)$ is unstable. The design procedure is illustrated below in a sample example.

Example 4. Consider an unstable process $G_p(s) = 4e^{-5s}/(10s - 1)$. From the relay feedback test with an ideal relay of unity magnitude, $a_y = 21.8440$, $a_u = 5.4606$, $T = 15.4606$ and $A_p = 2.5948$ were obtained. Using (10)–(11) the estimate for the FOPDT transfer function model of the plant is $4.0003e^{-4.9998s}/(10s - 1)$ and assuming $\lambda = 2$, the controller parameters become $K_p = 0.25$, $T_i = 2$, $K_f = 0.5$ and $T_f = -4.001$. From (33), the controller parameter for the disturbance rejection is $K_d = 0.3535$. A unit step setpoint and a load disturbance $L = -0.1$ are introduced at $t = 0$ and 80, respectively. The response of the closed-loop system for these controller settings is given in Fig. 7 for the plant with time delay equal to the nominal value ($\theta_m = \theta$) and changes of $\pm 10\%$ ($\theta = 5.5$ and $\theta = 4.5$ with $\theta_m = 5$ in both cases).

5. Conclusions

Simple and effective automatic tuning formulae are derived for a new Smith predictor structure assuming low-order model transfer functions with time delay for stable, unstable and integrating processes. Two important advantages of the new scheme are that the process models possess two unknowns namely, the time delay and the time constant which are easily obtainable using a single relay feedback test assuming the process steady-state gain is known and the controller parameters have very simple and straightforward relations with the two unknowns, respectively. The method has been shown to work effectively even when the plant does not have the assumed transfer function forms. The robustness of the proposed controller is apparent from results obtained using incorrect time delay values in the plant model. Illustrative examples have been given to show the simplicity and robustness of the controllers for controlling the class of processes considered.

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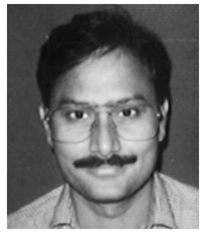

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