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AUTOTUNING PID CONTROL FOR LONG TIME-DELAY PROCESSES

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Abstract: A refined relay feedback identification autotuning PID/PI is proposed in this paper which is capable of controlling long time-delay processes. The process is approximated via FOPDT or SOPDT models, whose parameters are determined through a modified relay feedback identification method. Employing zero-pole cancellation principle, the PID/PI is tuned by the specified amplitude and phase margins, which can guarantee fast response and strong robustness to the closed loop system. Model identification and controller parameter tuning are done on-line without much influence on the normal operation. In addition, this algorithm also has good disturbance rejection capability. *Copyright © 2002 IFAC*

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1 INTRODUCTION

PID controller is widely used in the process control industry because of its relatively simple structure. Lately there have been significant efforts to give it added capabilities by providing facilities for automatic tuning, gain scheduling and adaption (Astrom and Hagglund, 1984; 1995). However, it has some drawbacks. It performs, for example, poorly for processes with long time-delay. Satisfactory system performance can't be achieved for PID strategies, whose parameters, for instance, are tuned by Ziegler-Nichols Method (Z-N), A-H Method (Astrom and Hagglund, 1984), etc. It is thus of interest to refine PID controller to cope with processes with long time-delay and to provide it with some tuning facility. Here long time-delay process refers to a process that satisfies $L/T > 0.5$ where L denotes pure time-delay and T denotes dominant time constant of the system.

A refined relay feedback identification autotuning PID/PI control algorithm is proposed in this paper, which is capable of controlling long time-delay processes. The main contributions of this paper are summarized as follows. The first one is employing a modified relay feedback identification method to obtain the dynamic information of long time-delay processes and introducing a correction coefficient C_{90} of the oscillation amplitude valued within 0.81~1.0. The second one is presenting a PID design method based on amplitude and phase margin and zero-pole cancellation principle, which can guarantee fast response and strong robustness to the closed loop system.

2 MATHEMATICAL MODEL DESCRIPTION

By far the most commonly used model in process control is

$$G_{31}(s)=\frac{K}{Ts+1}e^{-Ls} \tag{1}$$

However, Astrom and Hagglund (1995) emphasize that it is not a representative model and that the conclusions drawn based on it may often be misleading when applied to plants. The other frequently used low-order models are

$$G_{32}(s)=\frac{K}{(Ts+1)^2}e^{-Ls} \tag{2}$$

$$G_4(s)=\frac{K}{(T_1s+1)(T_2s+1)}e^{-Ls}\,,T_1\neq T_2 \tag{3}$$

Simulations (Astrom and Hagglund, 1995) indicate that model (2) and (3) approximate plants better than (1). For convenience of parameter identification on line, now convert model (2) and (3) into

$$G(s)=\frac{1}{as^2+bs+c}e^{-Ls} \tag{4}$$

Wang, et al (1999) point out that the fitting of the Nyquist plot of model (4) to that of the real process is incredibly close over a frequency range important for control performance. Therefore, based on this model, a PID/PI controller designed in terms of the specified amplitude and phase margin can guarantee almost the same amplitude and phase margin, respectively, to the real process as the specified ones.

3 RELAY FEEDBACK IDENTIFICATION

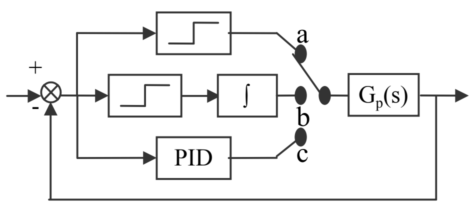


Fig.1 The structure of relay feedback based parameter identification method

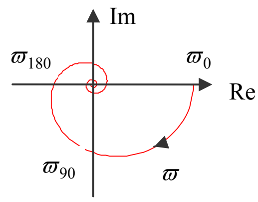


Fig.2 The Nyquist plot of the controlled object

Astrom and Hagglund (1984) and Wang (2000) detail the merits of the relay feedback identification method. In this paper a modified relay feedback structure shown in fig.1 is used to identify model (1) and (4).

In fig.1, the negative reciprocal describing function of relay is

$$-\frac{1}{N(M)}=-\frac{\pi M}{4d} \tag{5}$$

where d is the relay amplitude and M is the amplitude of the relay input (the same as the amplitude of the closed loop output). The oscillation occurs at the point of ϖ_{180} in fig.2. Now turn the switch in fig.1 from position a to b, then there is (Astrom and Hagglund, 1995)

$$N(M)\frac{1}{(j\varpi)}G_p(j\varpi)=-1 \tag{6}$$

Substituting eq. (5) into (6) gives

$$G_p(j\varpi)=-\frac{1}{N(M)}j\varpi=-\frac{\pi M\varpi}{4d}j \tag{7}$$

The corresponding oscillation occurs at the point of ϖ_{90} in fig.2. Eq. (5) and (7) indicate that M is proportional to d. Then M can thus be adjusted by d automatically. So the stable limit cycle amplitude can be kept within the acceptable limits.

A method called, here, Two-Point Method is applied to identify model (1) and (4). For model (1), Wang (2000) and Astrom and Hagglund (1991) employ the information of points ϖ_0 and ϖ_{180} to determine its parameters. However, the information of ϖ_0 and ϖ_{90} is used, instead, to identify model (1) in this paper. As for model (4), a similar method is applied as Wang's (2000), which all make use of the information of points ϖ_{90} and ϖ_{180}. But the relay is preceded the integrator in this paper (see fig.1).

The introduction of the integrator following a relay has two main functions as below: a) Improving the approximating ability of the describing function to nonlinear element. The existence of long time-delay leads to the decrease of the plant's corner frequency and the oscillation frequency will also decrease consequently. When conducting experiments for dynamic characteristic identification by the modified autotuning method, the output of the closed loop system (ie, the input of the relay element) is notably different from a sinusoid wave, which decreases the description precision of describing function greatly. By means of the attenuation characteristic of integral element to high frequency signals, the high harmonics will be attenuated as far as possible when they pass through the integrator. The ratio of the first harmonic will then increase relatively, which improves the approximation of the describing function to nonlinear elements. Hence accurate dynamics can be available. b) Correcting the measured oscillation amplitude. The introduction of the integrator leads to an output of approximate triangular wave in the closed loop when conducting a relay feedback identification experiment. So the input of relay is not a sinusoid signal, which also affects the description precision of describing function. However, the first-harmonic amplitude of the output triangular wave can be easily figured out because of its special waveform and the value of the first-harmonic amplitude is independent of the oscillation frequency. The resulted value can be used as a theoretical basis to correct the measured oscillation amplitude.

Generally (Astrom and Hagglund, 1984; Wang and Shao, 1999; Wang, 2000), the period of the limit cycle oscillation can be easily determined from the times between zero-crossings. The amplitude M may be determined by measuring the peak-to-peak values of the output. For long time-delay processes, however, simulations demonstrate that the resulted value of M often tends to be larger, which must be corrected. Here, let $\hat{a} = C_{90} a$ where C_{90} is a correction coefficient, whose value will vary slightly according to the difference of the real time-delay processes. It can be assigned within 0.81~1.0. Where 0.81 is the ratio between the first-harmonic amplitude of the triangular wave and the triangular wave amplitude (the accurate value is $8/\pi^2$), and 1.0 is corresponding to the triangular wave amplitude. It is suggested $C_{90} = 0.91$ for large time-delay processes and $C_{90} = 1.0$ for little time-delay processes.

3.1 Parameter identification for model (1)

Suppose the frequency characteristics of model (1) at points ϖ_0 and ϖ_{90} equal that of the process $G_P(s)$, respectively. Then

$$\begin{cases} \left|G_{31}(j\varpi_0)\right| =: K_0, \ \left|G_{31}(j\varpi_{90})\right| =: K_{90} \\ \angle G_{31}(j\varpi_0) = 0, \ \angle G_{31}(j\varpi_{90}) = -\pi/2 \end{cases} \tag{8}$$

Substituting eq. (8) to (1) and (7) gives

$$K = K_0, \ L = \frac{1}{\varpi_{90}} \arcsin\left(\frac{\pi a_{90} \varpi_{90}}{4 d_{90} K_0}\right), \ T = \frac{1}{\varpi_{90}} \cot(\varpi_{90} L) \tag{9}$$

In fact, K_0 is static gain of the controlled object. It must be known or can be got from the step response near the operating-point.

3.2 Parameter identification for model (4)

Suppose the frequency characteristics of model (4) at the point of ϖ_{90} and ϖ_{180} equal that of $G_P(s)$, respectively. Then

$$\begin{cases} \left|G(j\varpi_{90})\right| =: k_{90}, \ \left|G(j\varpi_{180})\right| =: k_{180} \\ \angle G(j\varpi_{90}) = -\pi/2, \ \angle G(j\varpi_{180}) = -\pi \end{cases} \tag{10}$$

Substituting eq. (10) to (4), (5) and (7) and conducting some complicated calculations give

$$a = \frac{1}{\varpi_{180}^2 - \varpi_{90}^2} \left(\frac{\sin(\varpi_{90} L)}{k_{90}} + \frac{\cos(\varpi_{180} L)}{k_{180}}\right) \tag{11}$$

$$b = \frac{\sin(\varpi_{180} L)}{\varpi_{180} k_{180}} \ \text{or} \ \frac{\cos(\varpi_{90} L)}{\varpi_{90} k_{90}} \tag{12}$$

$$c = \frac{1}{\varpi_{180}^2 - \varpi_{90}^2} \left(\frac{\varpi_{180}^2 \sin(\varpi_{90} L)}{k_{90}} + \frac{\varpi_{90}^2 \cos(\varpi_{180} L)}{k_{180}}\right) \tag{13}$$

$$\frac{\sin(\varpi_{180} L)}{\cos(\varpi_{90} L)} = \frac{k_{180} \varpi_{180}}{k_{90} \varpi_{90}} \tag{14}$$

Since eq. (14) is a nonlinear equation, Newton-Raphson's method is used to generate a sufficiently accurate solution after a few iterations. Let

$$f(L) = \frac{\sin(\varpi_{180} L)}{\cos(\varpi_{90} L)} - \frac{k_{180} \varpi_{180}}{k_{90} \varpi_{90}} \tag{15}$$

then the Newton-Raphson iterative express can be described as

$$L_{k+1} = L_K - \frac{f(L_k)}{f^{'}(L_k)} \qquad (16)$$

$$f^{'}(L) = \frac{\varpi_{90}\sin(\varpi_{90}L)\sin(\varpi_{180}L) + \varpi_{180}\cos(\varpi_{90}L)\cos(\varpi_{180}L)}{\cos^2(\varpi_{90}L)}$$

where $f^{'}(L)$ is the derivative of $f(L)$. The initial value L_0 of the iterative variable L is very crucial, which has great influence on iteration times and identification precision. From eq. (12), there are $\sin(\varpi_{180}L) > 0$, i.e. $\varpi_{180}L \in (0, \pi)$; $\cos(\varpi_{90}L) > 0$, i.e. $\varpi_{90}L \in (0, \pi/2)$. So the previous sine and cosine functions can be approximated by

$$\sin(\varpi_{180}L) = p(\varpi_{180}L)^2 + q(\varpi_{180}L),\ \varpi_{180}L \in (0, \pi) \qquad (17)$$

$$\cos(\varpi_{90}L) = p(\varpi_{90}L)^2 + r(\varpi_{90}L) + 1,\ \varpi_{90}L \in (0, \pi/2) \qquad (18)$$

where $p \approx -0.3357$, $q \approx 1.1640$, $r \approx -0.1092$. The fittings are exact at the points $\varpi L = 0, \pi/4, \pi/2$. Applying eq. (17) and (18) to (14) obtains

$$0.3357(k_{180}\varpi_{90} - k_{90}\varpi_{180})L^2 + (0.1092k_{180} + 1.1640k_{90})L - k_{180}/\varpi_{90} = 0 \qquad (19)$$

Solving eq. (18) and taking the smaller absolute root yields L_0. Apply it to eq. (16) to obtain the more accurate L by a few iterations. And then substitute the resulted value into eq. (11)~(13) for the other parameters a, b and c of model (4).

4 PID/PI PARAMETER CALCULATION

Here, the zero-pole cancellation principle is applied to determine PID/PI parameters in terms of amplitude and phase margin. The presence of long time-delay leads to tremendously sluggish system response. It is possible to attain a fast response as far as possible in terms of zero-pole cancellation method. Since $e^{-Ls} = \frac{e^{-0.5Ls}}{e^{0.5Ls}} \approx \frac{1-0.5Ls}{1+0.5Ls}$, then $G_{31}(s) = \frac{K}{Ts+1}e^{-Ls} \approx \frac{K}{(Ts+1)}\frac{(1-0.5Ls)}{(1+0.5Ls)}$ In Addition, because of $L >> T$, so it is feasible to cancel the pole of model (1) (for example, $Ts+1$) via the zero of PID/PI controller. Such conclusion can also be drawn for model (2)~(4). At the same time, such a PID/PI controller whose parameters are tuned in terms of amplitude and phase margin can guarantee the closed loop system well stable and strongly robust

For model (4), let PID controller be

$$G_c(s) = K_p(1 + \frac{1}{T_i s} + T_d s) \qquad (20)$$

Determine its parameters so that its zeros cancel all the poles of model (4). That is

$$G_c(s)G(s) = \frac{k_o}{s}e^{-Ls} \qquad (21)$$

where k_o is an unknown parameter whose value will be specified by the amplitude margin A_m and phase margin ϕ_m. From eq. (4), (20) and (21), it follows

$$K_p = k_o b,\ T_i = b/c,\ T_d = a/b \qquad (22)$$

According to the definition of amplitude and phase margin, there is

$$\begin{cases} \arg[G_c(j\varpi_g)G(j\varpi_g)] = -\pi, & |G_c(j\varpi_g)G(j\varpi_g)| = 1/A_m \\ |G_c(j\varpi_p)G(j\varpi_p)| = 1, & \phi_m = \pi + \arg[G_c(j\varpi_p)G(j\varpi_p)] \end{cases} \qquad (23)$$

Substituting eq. (21) into (23) gives

$$\phi_m = \frac{\pi}{2}(1 - \frac{1}{A_m}). \qquad (24)$$

$$k_o = \frac{\pi}{2A_m L} \text{ or } k_o = \frac{1}{L}(\frac{\pi}{2} - \phi_m) \qquad (25)$$

Eq. (24) is a constraint condition of zero-pole cancellation for $G_c(s)G(s)$. Usually A_m values within $2 \sim 5$, then $\phi_m = \pi/4 \sim 2\pi/5$. Fig.3 depicts the relationship between A_m and ϕ_m. Applying eq. (25) to (22) yields

$$K_p = \frac{\pi b}{2A_m L},\ T_i = b/c,\ T_d = a/b \qquad (26)$$

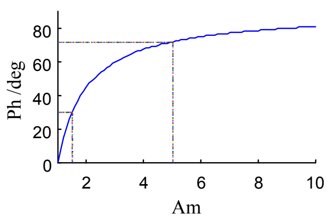


Fig.3 The relationship between A_m and ϕ_m

or

$$K_p = \frac{b}{L}(\frac{\pi}{2} - \phi_m),\ T_i = b/c,\ T_d = a/b \qquad (27)$$

As for model (1), a PI controller $G_c(s) = K_p(1 + \frac{1}{T_i s})$ is designed. Similarly to parameter tuning for model (4) above, there is

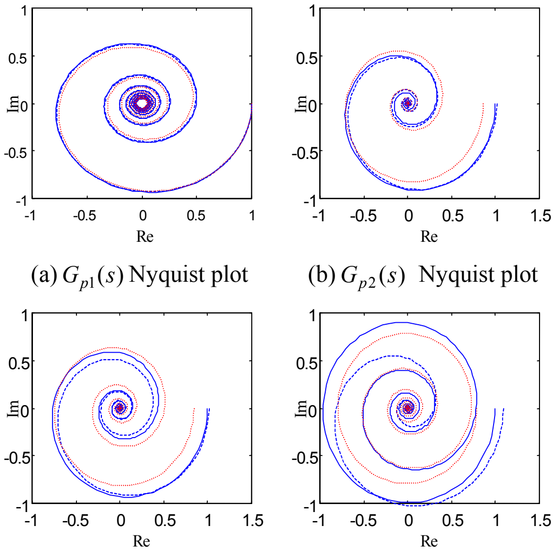
$$K_p = \frac{\pi T}{2A_m KL} \quad or \quad \frac{T}{KL}(\frac{\pi}{2} - \phi_m),\ T_i = T \qquad (28)$$

5 SIMULATIONS

5.1 Simulations for identification precision and control performance

The following four models $G_{p1}(s) = \frac{1}{2s+1}e^{-6s}$, $G_{p2}(s) = \frac{1}{(2s+1)^2}e^{-6s}$, $G_{p3}(s) = \frac{1}{(s+1)(2s+1)}e^{-6s}$ and $G_{p4}(s) = \frac{1}{s^2+1.5s+1}e^{-6s}$ are considered to compare identification precision of the proposed method in this paper with that of Wang (2000). Evidently they all belong to processes with long time-delay. Now approximate them with FOPDT (first-order plus dead time) or SOPDT (second-order plus dead time), respectively. The simulation results are given in the form of tables and graphs. For the sake of margins, only Nyquist plots are presented as shown in fig.4. Apparently, the proposed identification method performs better than Wang's (2000). The improvement of identification precision attributes to the introduction of the correction coefficient C_{90} and the decision of the approximating expresses for time-delay initial value. Compared with Wang's (2000), both are undergone modification so that they are more suitable for identification of long time-delay processes. However, for second oscillatory processes, the identification precision will decrease evidently with the increase of time-delay (see fig.4 (d)).

Simulations are also done for control performance comparisons based on the same four models above. PID/PI controllers are tuned via Z-N method, A-H method, Smith predictive PID and the proposed method in this paper (let $A_m = 3$, then, $\phi_m = \pi/3$),



(a) $G_{p1}(s)$ Nyquist plot (b) $G_{p2}(s)$ Nyquist plot

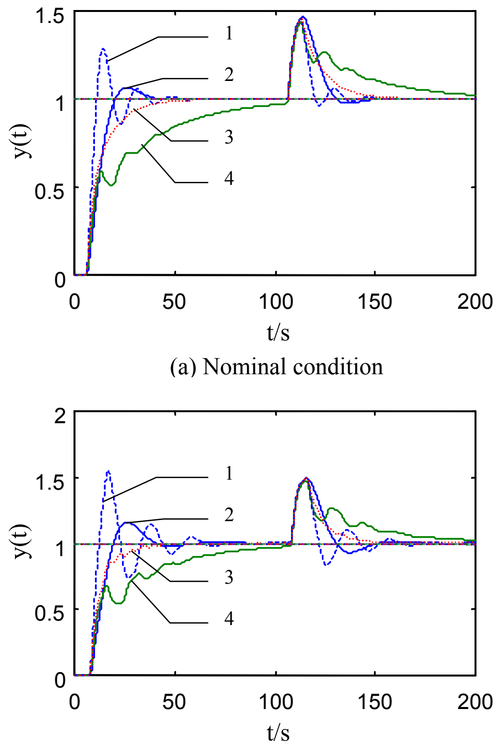
(c) $G_{p3}(s)$ Nyquist plot (d) $G_{p4}(s)$ Nyquist plot

: (solid) the original model

: (dashed) the proposed method

: (dotted) method of Wang (2000)

Fig.4 The resulted Nyquist plots via the proposed method and the method of Wang (2000)



(a) Nominal condition

(b) L increases 30%, and T_1, T_2 decrease 20%

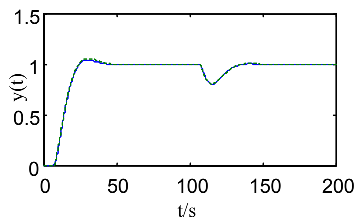
1: ZN-PID; 2: the proposed; 3: ZN-Smith; 4: AH-PID.

Fig.5 The responses of $G_{p3}(s)$ in different controllers

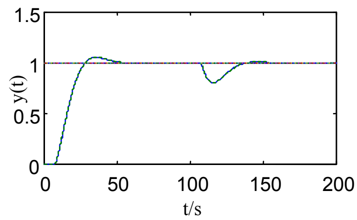
respectively. The PID controller parameters in Smith predictive structure are tuned in terms of Z-N method. The setpoint inputs are unit step signals and the disturbance signals are introduced at t=100s with an amplitude of 50% unit step. It is shown that the proposed method has fast system response and that its disturbance rejection ability and robustness also excel the other three. Fig.5 are response curves based on simulation model $G_{p3}(s)$.

5.2 Simulations for autotuning process

Let the mathematical model of the real plant be $G_{p5}(s) = \frac{1}{(s+1)^4} e^{-6s}$. Firstly, approximating it by model (1) and (4), respectively, gives the following two low-order models $G_{51}(s) = \frac{1}{2.1026s+1} e^{-7.9692s}$ and $G_{52}(s) = \frac{1.0326}{(0.6344s+1)(2.6332s+1)} e^{-6.8708s}$. Then design PID/PI controller (let $A_m = 3$, $\phi_m = \pi/3$) based on $G_{51}(s)$ and $G_{52}(s)$, respectively. And the corresponding controllers are $G_{c1}(s) = 0.1381(1 + \frac{1}{2.1026s})$ and $G_{c2}(s) = 0.2412(1 + \frac{1}{3.2677s} + 0.5113s)$. Finally apply $G_{c1}(s)$, $G_{c2}(s)$ to $G_{p5}(s)$ and $G_{51}(s)$, $G_{52}(s)$, respectively, to observe their control performance. Fig.6 shows the control results. It's easy to see that the response of $G_{p5}(s)$ almost overlaps with that of $G_{51}(s)$ and $G_{52}(s)$ with fast response speed and little



(a) Model (4)



(b) Model (1)

Fig.6 Response of the real process (solid line) and the low-order approximation model (dashed line) under the proposed PID/PI

overshoot. It means that the proposed method of autotuning PID/PI controller in this paper has good identification capability and control performance.

6 CONCLUSION

The proposed control algorithm in this paper has many advantages such as high precision for model identification, explicit parameter expresses for controller tuning, fast response, strong robustness and good disturbance rejection ability. And it's very suitable for controlling such long time-delay processes that require fast response, however, permit the existence of overshoot. At the same time, the method makes fully use of the advantages of the relay feedback identification method, by which model identification and controller parameter tuning can be done on-line without much influence on the normal operation of the real system. Therefore, it is of great value for engineering practice. However, an important problem, bumpless switching between identification on line and PID/PI control, must be tackled before putting into application in industry.

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