Enhanced Automatic Tuning Procedure for Process Control of PI/PID Controllers

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An enhanced automatic tuning procedure developed for process control of PI and PID controllers addresses several potential problems present in current standard autotuners. The proposed enhanced autotuner uses a novel technique based on relay feedback to estimate the process frequency response at two specified phase lags on the Nyquist curve automatically. An iterative procedure then uses these two points to obtain a transfer-function model of the process. Based on this model and a controller-selection scheme, an appropriate controller (PI or PID) is applied to the process automatically. The controller is tuned so that the Nyquist curve of the compensated system is appropriately shaped to satisfy a combined gain and phase-margin type of specification. The effectiveness of this enhanced autotuner is demonstrated both in simulations and in real-time experiments for level control of a coupled-tanks system.

Introduction

Controllers of the proportional-integral-derivative (PID) type are still in widespread use in the process-control industry despite advances and increasing sophistication in the so-called mathematical-control systems theory. This is not surprising since the reliability of these PID controllers has been proved by the decades of success and the wide acceptance they enjoyed among practicing engineers. Far from becoming obsolete, research-and-development efforts for these controllers are currently undergoing a resurgence, with much research effort focused on the automatic tuning of PID controllers (Astrom and Hagglund, 1984; Bristol, 1977; Hagglund and Astrom, 1991; Hess et al., 1987; Ho et al., 1993; Leva, 1993; Schei, 1992). Some of them have already found direct industrial applications (Bristol, 1977; Hagglund and Astrom, 1991; Hess et al., 1987). References to different autotuning approaches can be found in Gawthrop and Nomikos (1990).

Among the techniques currently available, the arguably most frequently quoted one is an ingenious and attractively simple autotuning method due to the work of Astrom and Hagglund (1984). They insert a relay as a feedback controller into the process loop in the arrangement of Figure 1. Under the relay feedback, most processes with dynamics typically encountered in process control will oscillate with a small and controlled amplitude. From these oscillations, the ultimate frequency can be identified. When the *critical point*, that is, the ultimate gain and frequency, is known, it is straightforward to apply the classic Ziegler–Nichols tuning rules (1943) to tune the controller. It is also possible to devise many other design schemes that are based on the knowledge of the critical point. A general approach, proposed by Astrom and Hagglund (1984) (which we shall refer to as the standard Astrom–Hagglund-based method) would be to move the critical point to a desired position on the complex plane through the parameters of the PID controller so that a combined gain and phase margin type of specification is satisfied. This approach is known to work well in many situations in process control, and its advantage is that the specified margin could serve as a measure of robustness, resulting in better system closed-loop performance over those tuned by classical Ziegler–Nichols rules. Nevertheless, there are some potential problems associated with this standard method. First, a relationship between the integral and derivative time constants

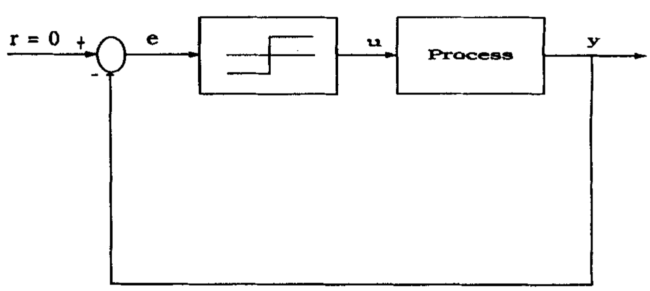

Figure 1. Relay feedback system.

of the PID controller has to be specified, which is difficult without *a-priori* information of the process dynamics Second, the design method cannot be extended to a PI controller, which is necessary in many problems in process control, for example, processes of the first-order dynamics and processes with a long deadtime. Third, the method does not consider how the neighborhood frequency response of the compensated system is affected by the controller parameters. In many cases, the frequency response is not sufficiently attenuated above the critical frequency, and this has a direct adverse effect on the closed-loop performance. To date, none of the various variants (Ho et al., 1993; Leva, 1993) of PID autotuners that rely on the key relay tuning feature of the standard Astrom–Hagglund-based method have satisfactorily addressed these problems.

We present here the design of an enhanced automatic-tuning procedure (autotuner) that overcomes the problems previously discussed; as such, this strategy may be considered as a further enhancement of the well-accepted standard Astrom–Hagglund method. This proposed enhanced strategy shapes the compensated Nyquist curve using two points at process phase lags of $\phi_1 = -180°$ and $\phi_2 = -180° + \phi_m$, where ϕ_m is the equivalent phase margin specification. To estimate the latter point, a novel method incorporating relay-based techniques is presented. The technique is able to automatically track the frequency corresponding to any desired value of ϕ_m between 0° and 180°. An algorithm is provided to convert the frequency response estimated to a transfer-function model; based on this model and a controller-selection scheme, a PI or a PID controller can be automatically applied to the process. The controller is tuned so that the Nyquist curve is appropriately shaped to achieve the desired specification and give improved closed-loop performance over the standard method. The effectiveness of the proposed enhanced strategy is demonstrated in both simulation examples and by a real-time implementation experiment.

Next, we review the standard Astrom–Hagglund-based method of autotuning PID controllers and discuss potential problems associated with it. Then, the enhanced Astrom–Hagglund-based method proposed here is described step by step, including the novel frequency-response estimation technique, the algorithm for frequency response to transfer-function conversion, the controller-selection scheme, and controller tuning algorithms. Some simulation and comparison results are presented, as well as results of real-time experiments on the level control problem of a coupled-tanks apparatus.

Standard Astrom–Hagglund-based Autotuned PID Controller

The key idea in the standard Astrom–Haglund-based method of autotuning the PID controller is to use the relay feedback arrangement of Figure 1. For many classes of process, the relay feedback gives an oscillation with a frequency close to the ultimate process frequency ω_u. The ultimate process gain k_u is approximately given by Astrom and Hagglund (1984) as $k_u = (4u_m/\pi y_m)$, where u_m and y_m are amplitudes of relay and process output, respectively. In the presence of noise, it is advantageous to use a relay with hysteresis (Astrom and Hagglund, 1984) to reduce the sensitivity of the resultant system to noise. The hysteresis level is set according to the noise level. Noise sensitivity is reduced further by averaging measurements of the ultimate frequency and ultimate gain over a few limit cycle oscillations.

When the critical point is known, it is possible to move it to a desired position on the complex plane, $(1/A_m)e^{-j(\pi-\phi_m)}$. This design method is considered as a combination of gain and phase-margin specification. When $A_m = 1$, ϕ_m is the desired phase margin, and conversely, when $\phi_m = 0$, A_m is the desired gain margin. For most purposes in process control applications, the specification of $A_m = 2$ and $\phi_m = \pi/4$ is recommended (Hagglund and Astrom, 1991).

The PID controller is described by

$$G_c(s) = k_c\left(1 + \frac{1}{T_i s} + T_d s\right), \tag{1}$$

where the parameters k_c, T_i, and T_d are chosen so that the critical point is moved to the desired position. The following conditions must be satisfied:

$$|G_c(j\omega_u)G_p(j\omega_u)| = \frac{1}{A_m}, \tag{2}$$

$$\arg G_c(j\omega_u)G_p(j\omega_u) = -\pi + \phi_m, \tag{3}$$

where $G_p(s)$ is the process transfer function. A possible solution is

$$k_c = \frac{k_u}{A_m}\cos\phi_m, \tag{4}$$

$$T_d = \frac{\tan\phi_m + \sqrt{\frac{4}{\alpha} + \tan^2\phi_m}}{2\omega_u}, \tag{5}$$

$$T_i = \alpha T_d \tag{6}$$

where α is chosen arbitrarily.

The standard Astrom–Hagglund-based method just reviewed is known to work well for many industrial processes, and it also has the advantage of simplicity. However, there are some potential problems associated with this standard design method that may have to be considered in certain applications. First, an additional design parameter α has to be specified. In general, this specification requires *a-priori* information on all process dynamics, since the control performances are heavily influenced by the choice of α. If it is too small, the obtained regulator has significant derivative action, and one has to contend with the related problems; conversely, if it is too large, the integral time becomes much higher than needed, which results in a slower control loop. Second, the design is constrained to a PID design, since a PI controller always contributes phase lag and thus would not be able to shift the critical point to a lower phase-lag position in the complex plane. In process control, there are classes of problems where the derivative control action is undesirable. For example, very simple control problems, where the process is approximately a first-order system, can be solved effectively with a PI controller with high gain. Derivative action is of limited use for these problems (Hagglund and Astrom,

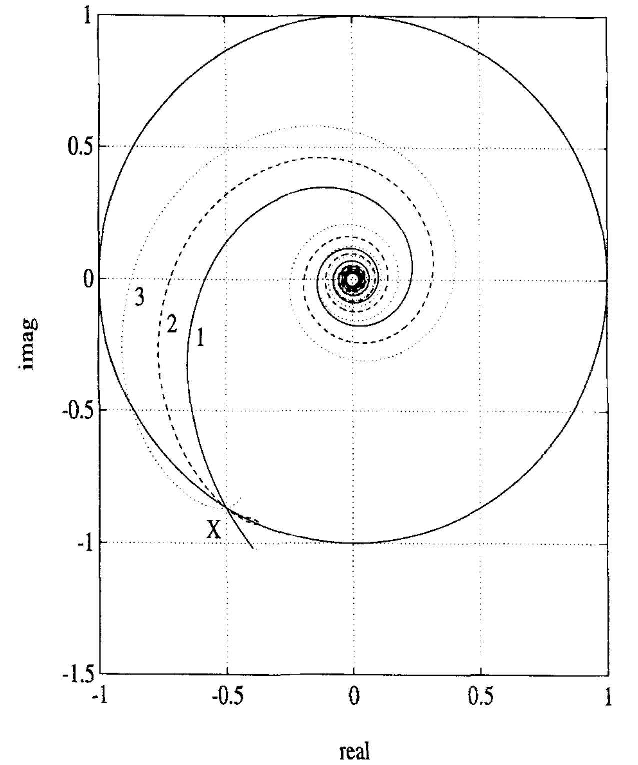

Figure 2. Nyquist curves passing through a common point *X* on the complex plane.

1991). Furthermore, the noise will be amplified by the derivative term. It is also undesirable to use derivative action when the process has a long deadtime (Hagglund and Astrom, 1991). In an automated environment, it is desirable to detect these processes and switch off the derivative action appropriately. This desirable feature is unfortunately not attainable with the standard Astrom–Hagglund-based method of autotuning. Third, the design does not consider the effect of controller parameters on the frequency response of the compensated system. The critical point may be moved to the desired position, but that alone may not necessarily ensure that the Nyquist curve of the compensated system is appropriately shaped, which could result in less than satisfactory control performance. This situation may be illustrated in the examples in Figure 2, where the Nyquist curves of various compensated systems pass through a common point on the complex plane, but the different systems have differing amounts of attenuation at frequencies above that point. These different systems will have significantly different performances, with tighter control from System 1 compared to System 2 or 3. Some additional examples are discussed in the fourth section.

Enhanced Astrom–Hagglund-Based Autotuned PI/PID Controller

The proposed enhanced autotuned PI/PID controller seeks to address the potential problems associated with the standard Astrom–Hagglund-based method of autotuning. In this proposed enhanced strategy, two points on the process Nyquist curve are estimated at specified phase lags of $-\pi$ and $-\pi+\phi_m$ using a novel technique of frequency response estimation. Using the points, the process frequency response is converted to a transfer function model. Based on this model and a controller selection scheme, an appropriate controller (PI or PID) is tuned to achieve the combined gain and phase specification while ensuring that the Nyquist curve of the compensated system is appropriately shaped. All these elements of the proposed enhanced autotuner (referred to as the enhanced Astrom–Hagglund-based method) are methodically described in the following subsections.

Relay-based frequency response estimation

With the simple relay feedback approach, only one point on the process Nyquist curve is determined. It is possible, for example, to cascade a known linear dynamical element to the system in Figure 1 to obtain a frequency other than the ultimate frequency. However, we cannot specify the frequency of interest; it is fixed by the choice of the linear element cascaded and is unknown *a priori*. Besides, the introduction of the linear system affects the amplitude response of the original process, and in the case of a gain reduction, we have a smaller signal-to-noise ratio (SNR), which affects the estimation accuracy adversely.

An interesting variation of the relay feedback system is shown in Figure 3 that can give the frequency at a specified process lag of interest, $-\pi+\phi_m$ $[\phi_m \in [0,\pi)]$, without affecting the amplitude response of the original process. The delay function f is defined such that

$$u(t) = f[v(t)] = v(t-\theta),$$

where $\theta = \phi_m/\omega$ and ω is the oscillating frequency of $v(t)$.

We illustrate this feature from a describing function analysis (Atherton, 1975). Consider the modified relay in the dashed-box of Figure 3, which consists of a relay cascaded to the delay function, f. Corresponding to the reference input, $e(t) = y_m \sin \omega t$, the output of the modified relay can be expanded using the Fourier series and shown to be

$$u(t) = \frac{4u_m}{\pi} \sum_{k=1}^{\infty} \frac{\sin[(2k-1)\omega t - \phi_m]}{2k-1}. \tag{7}$$

Describing function of a nonlinear element is defined as the *complex ratio of the fundamental component of the nonlinear element to the input sinusoid* (Atherton, 1975). Hence, describing function of the modified relay can be computed as

$$N(y_m) = \frac{4u_m}{\pi y_m} e^{-j\phi_m}. \tag{8}$$

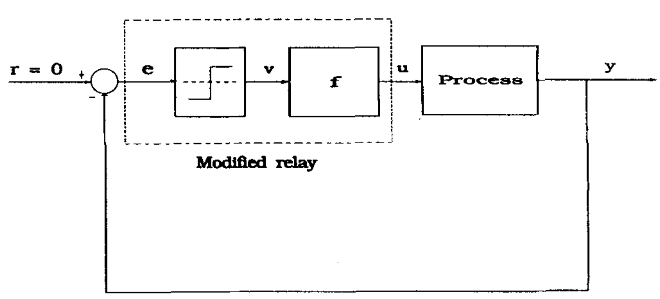

Figure 3. Modified relay feedback system.

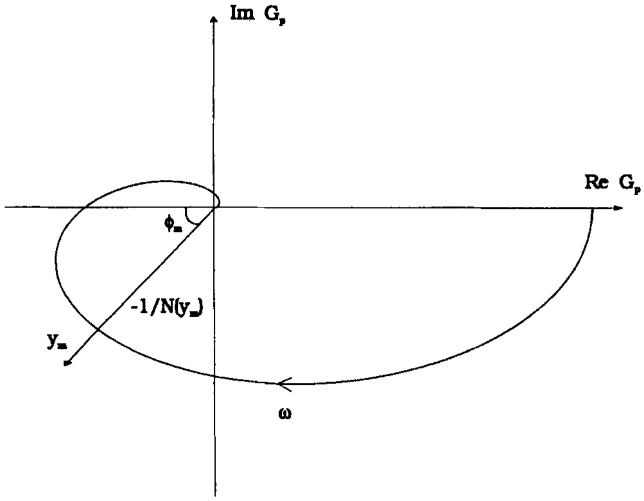

Figure 4. Negative inverse describing a function of the modified relay.

The negative inverse describing function, $-(1/N(y_m)) = (\pi y_m/4u_m)e^{j(-\pi+\phi_m)}$, is thus a straight line segment through the origin, as shown in Figure 4. By the feedback arrangement of Figure 3, it can be shown that the resultant amplitude and frequency of oscillation correspond to the intersection between $-(1/N(y_m))$ and the process Nyquist curve. Hence, we can obtain at the specified phase lag of $-\pi+\phi_m$, the inverse gain (k_ϕ) and the frequency of the process (ω_ϕ) from the output amplitude (y_m) and frequency (ω_{osc}), that is,

$$k_\phi = \frac{4u_m}{\pi y_m}, \tag{9}$$

$$\omega_\phi = \omega_{osc}. \tag{10}$$

With the arrangement of Figure 3, we can thus automatically track a frequency at the specified phase lag $-\pi+\phi_m$ for $\phi_m \in [0,\pi)$ without affecting the amplitude response of the original process. In this respect, it may be viewed as a generalized form of the relay feedback technique that tracks the frequency at the phase lag with $\phi_m = 0$. Besides, like relay feedback, the technique facilitates one-button tuning, a feature that is invaluable for intelligent control applications (Astrom and McAvoy, 1993). In the sequel, a practical implementation of the preceding development is presented to obtain two points from the process Nyquist curve—one at the critical frequency and the other at a specified phase lag, $-\pi+\phi_m$:

- Estimate the ultimate process gain k_u and frequency ω_u with the normal relay feedback.
- With these estimates, an initial guess for the frequency $(\hat{\omega}_\phi)$ at the specified phase of $-\pi+\phi_m$ is $\hat{\omega}_\phi = ((\pi-\phi_m)/\pi)\omega_u$, and we initialize $\theta = \phi_m/\hat{\omega}_\phi$.
- Continue the relay experiment and adapt the delay function, f, to the oscillating frequency ω_{osc}, that is, $\theta = \phi_m/\omega_{osc}$. Extensive simulations performed show that the oscillating frequency converges to the frequency at the specified phase lag ω_ϕ for most processes typically encountered in the industry. Then k_ϕ can be obtained through Eq. 9.

Conversion to a transfer-function model

Most processes in the industry can be adequately modeled by a transfer function of the following form (Seborg et al., 1985; Luyben, 1991):

$$G(s) = \frac{K_p}{(\tau s+1)^n} e^{-sL}. \tag{11}$$

Several authors (Astrom and Hagglund, 1988; Ho et al., 1993) have obtained first-order models with deadtime, for the process from a relay feedback experiment. The transfer-function model, $G(s)$ in Eq. 11 is a more general one that can capture high-order dynamics. However, with four parameters to be identified, two points on the process Nyquist curve are required, obtained in this case through the experiment in the preceding subsection.

In what follows, an algorithm to convert the frequency response obtained in that subsection ($G_p(j\omega_u)$ and $G_p(j\omega_\phi)$) to the transfer function model, Eq. 11, is developed. Equating the gain of the process and model at ω_u and ω_ϕ,

$$|G(j\omega_u)| = \frac{1}{k_u} = \frac{K_p}{\left(\sqrt{1+\tau^2\omega_u^2}\right)^n}, \tag{12}$$

$$|G(j\omega_\phi)| = \frac{1}{k_\phi} = \frac{K_p}{\left(\sqrt{1+\tau^2\omega_\phi^2}\right)^n}. \tag{13}$$

Simplifying Eqs. 12 and 13, we have

$$\left[1+\left(\frac{\omega_\phi}{\omega_u}\right)^2\left((K_pk_u)^{2/n}-1\right)\right]^{n/2} = K_pk_\phi. \tag{14}$$

Equating the phase of the process and model at ω_u and ω_ϕ, we have

$$\arg G(j\omega_u) = -\pi = -n\arctan\tau\omega_u - L\omega_u, \tag{15}$$

$$\arg G(j\omega_\phi) = -\pi+\phi_m = -n\arctan\tau\omega_\phi - L\omega_\phi. \tag{16}$$

A simple algorithm to obtain the parameters in Eq. 11 is outlined below. n_{limit} is a specified upper bound of the process, and for most problems in process control (Seborg et al., 1985; Luyben, 1991), $n_{limit} = 2$ is sufficient. This value of n_{limit} will be applied in our simulations and experiments.

- Iterate from $n = 1$ to $n = n_{limit}$.

  Compute K_p from Eq. 14, τ from an average of the estimates from Eqs. 12 and 13, and L from Eq. 15.

  Compute the cost function derived from Eq. 16,

$$S(n) = |-\pi+\phi_m+n\arctan\tau\omega_\phi+L\omega_\phi|$$

with the earlier values of n, K_p, τ and L.

- At the end of the iteration, a suitable set of model parameters (n, K_p, τ, L) corresponds to $n = n_{min}$, where $S(n_{min}) = \min_n\{S(n)\}$.

Controller design

At this stage, the transfer function model of the process is available. Based on this model, it is possible to decide on an appropriate controller (PI or PID) for the process. Following the discussions in the second section, PI control is applied for the following process categories.

- Processes of the first-order dynamics, $n = 1$.
- Processes with a long deadtime. A suitable measure of the relative magnitude of deadtime is the *normalized deadtime* defined by $\Theta = L/\tilde{\tau}$ (Astrom et al., 1992), where

$$\tilde{\tau} = \tau, \qquad n = 1,$$
$$\tilde{\tau} = 1.5\tau, \qquad n = 2.$$

An empirical threshold to classify a long process deadtime is $\Theta > 0.5$. The algorithms to tune the PI or PID controllers are given in the following subsubsections.

PID Controller. Consider the PID controller described by Eq. 1. First, we move the second estimated Nyquist point $G_p(j\omega_\phi)$ to the desired position specified by the combined gain and phase margins (A_m, ϕ_m). The following conditions must be satisfied:

$$|G_c(j\omega_\phi)G_p(j\omega_\phi)| = \frac{1}{A_m}, \tag{17}$$

$$\arg G_c(j\omega_\phi) = 0. \tag{18}$$

Equations 17 and 18 can be simplified to

$$\left|1 - j\frac{1}{\omega_\phi T_i}\left(1 - \omega_\phi^2 T_i T_d\right)\right| = \frac{k_\phi}{k_c A_m}, \tag{19}$$

$$\arctan \frac{1}{\omega_\phi T_i}\left(1 - \omega_\phi^2 T_i T_d\right) = 0. \tag{20}$$

A simple solution is to choose

$$T_i T_d = \frac{1}{\omega_\phi^2}, \tag{21}$$

$$k_c = \frac{k_\phi}{A_m}. \tag{22}$$

Besides moving the Nyquist point to the desired position, we want to ensure that the Nyquist curve of the compensated system is appropriately shaped for $\omega \in (\omega_\phi, \tilde{\omega}_u]$, where $\tilde{\omega}_u$ is the ultimate frequency of the compensated system, that is, $\arg G_p(j\tilde{\omega}_u)G_c(j\tilde{\omega}_u) = -\pi$.

The technique adopted here is to choose a value of T_i so that the frequency response of the compensated system is attenuated at $\omega = \omega_u$ with attenuation factor, r, that is,

$$|G_p(j\omega_u)G_c(j\omega_u)| = \frac{1}{rA_m}, \qquad 1 < r < \frac{k_u}{k_\phi}, \tag{23}$$

giving

$$T_i = \frac{rk_\phi}{\omega_u\sqrt{k_u^2 - r^2k_\phi^2}}\left[\left(\frac{\omega_u}{\omega_\phi}\right)^2 - 1\right]. \tag{24}$$

Since r relates to the attenuation of the amplitude response, a larger value of r typically implies a more damped response, compared to one with a smaller value of r. Empirical studies show that an initial choice of r at the 90% mark of its allowable range works well in most cases, that is, $r = 1 + 0.9\ ((k_u/k_\phi) - 1)$.

With the value of T_i computed from Eq. 24, we proceed to check whether the frequency response is adequately attenuated for $\omega \in (\omega_\phi, \tilde{\omega}_u]$, that is,

$$|G_c(j\omega)G_p(j\omega)| < \frac{1}{A_m}, \qquad \omega \in (\omega_\phi, \tilde{\omega}_u), \tag{25}$$

where $|G_p(j\omega)|$ can be estimated from the transfer function model, Eq. 11, that is,

$$|G_p(j\omega)| = \frac{1}{\left(\sqrt{1 + (\tau\omega)^2}\right)^n}.$$

In most cases, Eq. 25 is satisfied by the recommended value of the attenuation factor, r. Otherwise, T_i is slowly increased until Eq. 25 is satisfied. The search for T_i will converge since in the limit $T_i \to \infty$, the controller approaches a P controller and Eq. 25 is naturally guaranteed, since the amplitude response of the model in Eq. 11 is strictly decreasing.

PI Controller. The PI controller is described by

$$G_c(s) = k_c\left(1 + \frac{1}{T_i s}\right). \tag{26}$$

Two sufficient conditions to achieve the required margin specification are given in Eqs. 27 and 28.

$$|G_c(j\omega^*)G_p(j\omega^*)| = \frac{1}{A_m}, \tag{27}$$

$$\arg G_c(j\omega^*)G_p(j\omega^*) = -\pi + \phi_m. \tag{28}$$

Condition 25 is trivial in this case, since $|G_c(j\omega)G_p(j\omega)|$ is strictly decreasing for all frequencies for the process model of Eq. 11 and controller of Eq. 26; ω^* is a specified frequency satisfying $(\omega^*/\omega_\phi) = \beta < 1$, since a PI controller contributes phase lag at all frequencies; and β relates to the frequency of the specified point. Hence, a larger value of β implies a higher frequency and a faster response speed. For a process with long deadtime, where controller detuning is typically needed, a smaller value for β is needed. Considering these factors, the empirical design value recommended is $\beta = 0.8$, except in the case of the process having a long deadtime where $\beta = 0.5$ is suitable, that is,

$$\beta = 0.8, \qquad n = 1, \qquad \Theta < 0.5,$$
$$= 0.5, \qquad \qquad \Theta > 0.5.$$

Equations 27 and 28 can then be simplified to

$$T_i = \frac{1}{\beta\omega_\phi \tan[\arg G_p(j\beta\omega_\phi) - \phi_m]}, \tag{29}$$

$$k_c = \frac{\beta T_i \omega_\phi}{A_m |G_p(j\beta\omega_\phi)| \sqrt{(\beta T_i \omega_\phi)^2 + 1}}, \tag{30}$$

where $|G_p(j\beta\omega_\phi)|$ and $\arg G_p(j\beta\omega_\phi)$ can be estimated from Eq. 11, that is,

$$|G_p(j\beta\omega_\phi)| = \frac{1}{\left(\sqrt{1+(\beta\tau\omega_\phi)^2}\right)^n}$$

$$\arg G_p(j\beta\omega_\phi) = -n \arctan \beta\tau\omega_\phi - \beta\omega_\phi L.$$

Simulation Examples

In this section, we present several simulation examples for different kinds of process dynamics to illustrate the applicability and effectiveness of the autotuning method in the previous section. A comparison with the control system tuned by the standard Astrom–Hagglund-based method is also provided to highlight the improved performance achieved with the enhanced method. White noise has been deliberately introduced into the simulation to test the performance of the experiments under more realistic conditions. A hysteresis level of one relay amplitude is used in these examples to reduce the sensitivity of the measurements to noise.

Example 1: First-Order Process. Consider a process of the first-order $G_p(s) = (1/(s+1))e^{-0.2s}$. The recommended default specification (Hagglund and Astrom, 1991) of $A_m = 2$ and $\phi_m = \pi/4$ is used. After the relay-based frequency response estimation, the process model is fitted as $G(s) = (0.95/(0.83s+1))e^{-0.21s}$.

Since $\Theta = 0.25$, the PI controller is selected with $\beta = 0.8$. By Eqs. 29 and 30, the PI parameters are computed as $(k_c, T_i) = (1.63, 0.88)$ and the PI controller design is thus completed. This autotuning process is shown in Figure 5. The first part from $t = 0$ to $t = 8$ exhibits the relay-based experiment. The PI controller is commissioned at $t = 10$. A setpoint change occurs at $t = 12$, and a load response at $t = 20$. The response is clearly shown in Figure 5. The corresponding tuning and closed-loop performance for a PID controller using the standard Astrom–Hagglund-based method is shown in the same figure, showing that improved performance is achieved with the proposed enhanced method. The frequency response of the compensated system is shown in Figure 6 for both methods. From the figure, it may be seen that for this example, the Nyquist curve of the compensated system tuned with the standard Astrom–Hagglund-based method is not adequately attenuated at high frequencies, resulting in significant noise amplification, while the proposed enhanced Astrom–Hagglund-based method corrects this deficiency.

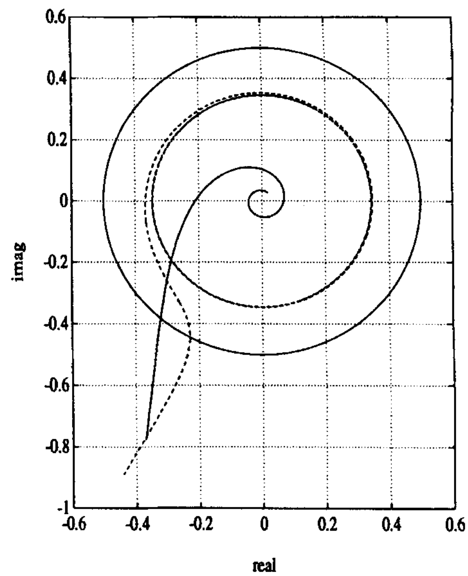

Figure 6. Frequency response of compensated system.
—— enhanced method; - - - standard method.

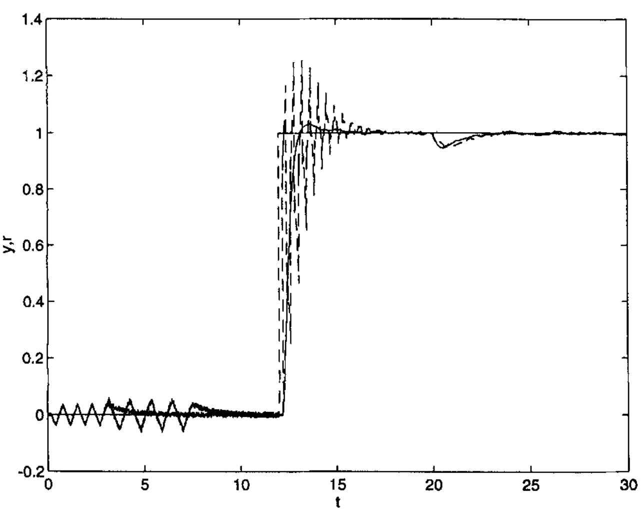

Figure 5. Autotuning session and closed-loop performance for a first-order process.
—— enhanced method; - - - standard method.

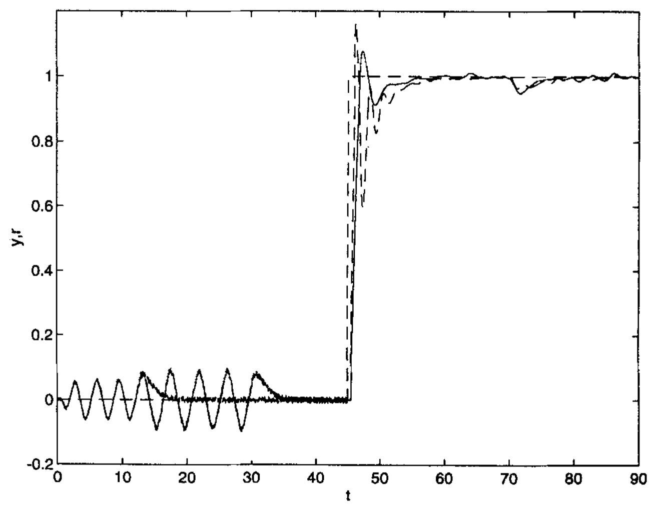

Figure 7. Autotuning session and closed-loop performance for a second-order process.
—— enhanced method; - - - standard method.

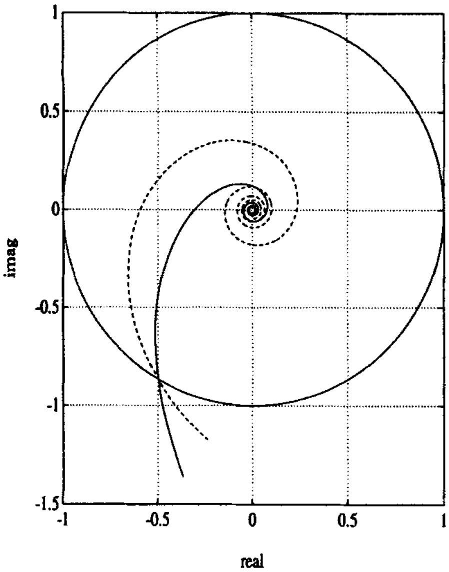

Figure 8. Frequency response of compensated system.
—— enhanced method; - - - standard method.

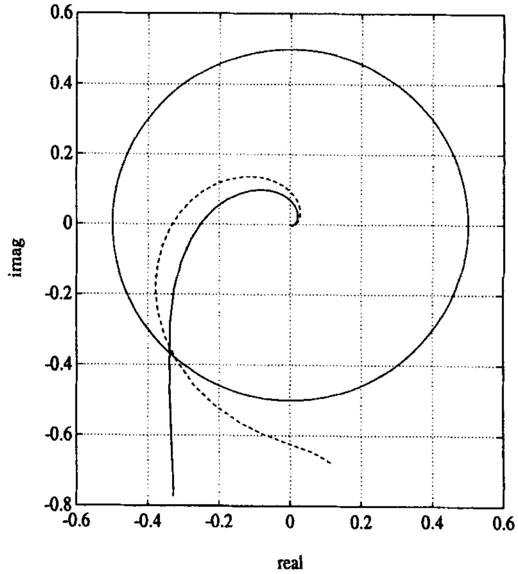

Figure 10. Frequency response of compensated system.
—— enhanced method; - - - standard method.

Example 2: Second-Order Process. Consider a process described by $G_p(s) = (1/(s+1)^2)e^{-0.5s}$. The specification of $A_m = 1$ and $\phi_m = \pi/3$ is chosen. After the relay-based frequency response estimation, the process model is fitted as $G(s) = (1.05/(0.98s+1)^2)e^{-0.52s}$. The normalized deadtime is given by $\Theta = 0.35$, and hence, the PID controller is selected. The autotuning and closed-loop performance is shown in Figure 7, and the frequency response of the compensated system in Figure 8. The improved closed-loop performance obtained by using the proposed enhanced method with lower overshoot and faster settling time is evident from these figures.

Example 3: High-Order Process. Consider a process described by $G_p(s) = (1/(s+1)^5)e^{-0.5s}$. The default specification of $A_m = 2$ and $\phi_m = \pi/4$ is chosen. After the relay-based frequency response estimation, the process model is fitted as $G(s) = (1.12/(1.91s+1)^2)e^{-2.10s}$. The normalized deadtime is given by $\Theta = 0.73$, and hence, the PI controller is chosen with $\beta = 0.5$. The autotuning and closed-loop performance are shown in Figure 9, and the frequency response of the compensated system is shown in Figure 10. Again, the improved closed-loop performance obtained by using the proposed enhanced method is evident from these figures.

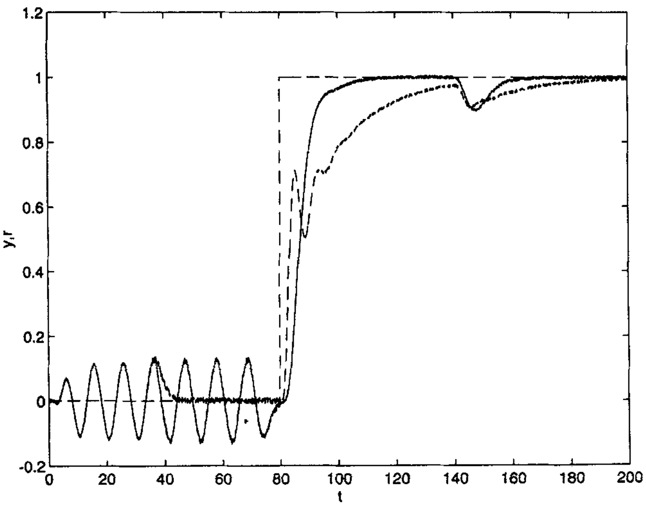

Figure 9. Autotuning session and closed-loop performance for a high-order process.
—— enhanced method; - - - standard method.

Real-Time Implementation Example

The proposed enhanced autotuner described earlier has been applied to experiments on a coupled tank system. As shown in Figure 11, the process consists of two square tanks, Tank 1 and 2, coupled to each other through an orifice at the bottom of the tank wall. The inflow (control input) is supplied by a variable speed pump that pumps water from a reservoir into Tank 1. The orifice between Tank 1 and 2 allows the water to flow into Tank 2. In the experiments, we

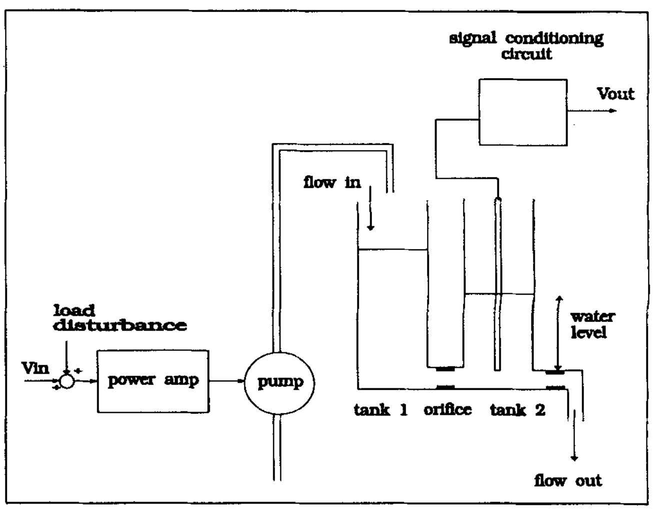

Figure 11. Experiment setup of coupled-tanks system.

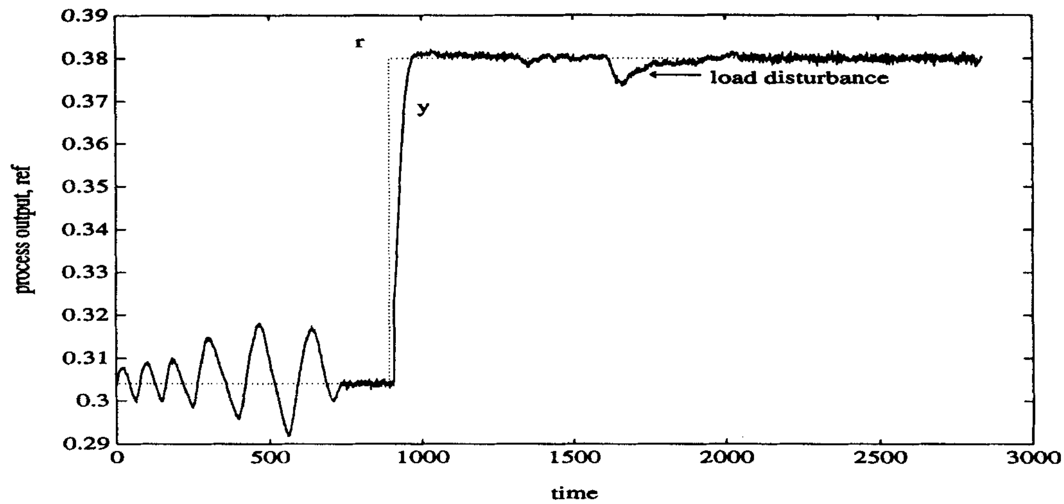

Figure 12. Autotuning and closed-loop performance for the enhanced PID autotuner implemented on a coupled-tanks system.

are interested in controlling the process with the voltage to drive the pump as input, and the water level in Tank 2 as process output. This coupled tank pilot process has process dynamics that represent the many fluid-level control problems faced in the process-control industry.

Results of the real-time experiments are in Figure 12. The default specification of $A_m = 2$ and $\phi_m = \pi/4$ is chosen. From $t = 0$ to $t = 750$, the relay-based frequency response estimation is run, and the transfer function model is obtained as $G(s) = [0.84/(76s+1)^2]e^{-27.6s}$. The normalized deadtime is given by $\Theta = 0.24$, and hence, the PID controller is selected.

Upon completion of the autotuning, a set-point change is made at $t = 900$. In addition, the experiments also included a 10% load disturbance in the process at $t = 1{,}600$. It can be seen from Figure 12 that the proposed enhanced autotuner attained tight set-point response and also good closed-loop regulation of load disturbances.

Conclusions

This article has presented the design of an enhanced autotuner for PI and PID controllers that require minimal *a-priori* process information. There are three main contributions here. First, a novel technique of frequency response estimation is developed that allows automatic estimation at specified phase lags. Two points on the process Nyquist curve are extracted using the technique and an algorithm is developed to convert the frequency response to a rational transfer function. Second, the autotuner is capable of self-diagnosis—from the process model, it can automatically apply an appropriate process control strategy (PI or PID). Third, algorithms are developed for tuning these controllers so that the compensated Nyquist curve is appropriately shaped to achieve a gain and phase margin type of specification. Simulation results as well as a real-time implementation example have been presented to demonstrate the practical utility of the autotuner and the good performance achieved.

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