

Time-domain based design and analysis of new PID tuning rules

G.K.I.Mann, B.-G.Hu and R.G.Gosine

Abstract: The time domain PID analysis includes three types of first-order plus time delay (FOPTD) models: (a) zero or negligible time delay (b) low to medium long time delay and (c) very long time delay. The first part of the analysis proves that the optimum PID controller for plants having negligible time delay is a PI controller, and the corresponding PI terms based on the actuator's capacity and set-point overshoot are explicitly derived. For low to medium time delay problems, a new PID tuning scheme is then developed. The proposed tuning rule is capable of accommodating the actuator's saturation and therefore has the ability to select an optimum PID controller. By using a separate time response analysis, a new PI tuning scheme for large normalised time delay is then derived. Numerical studies are made for higher-order processes having monotonic open-loop characteristics. The performance is compared with other commonly available tuning rules. With new tuning rules, better performance is observed and the rules have the capability to cover time delays ranging from zero to any higher value.

1 Introduction

As a result of extensive investigations to devise ways of choosing optimum controller settings for the PID controllers, Ziegler and Nichols [1, 2] showed how they could be estimated using open and closed-loop tests on the plant. The method is referred to as ZN rules. The ZN settings usually experience excessive overshoot of the plant response and also the method cannot be used to tune plants that have a relatively longer normalised time delay or NTD (ratio of process time delay to time constant). With the ease of computations, the numerical optimisation and curve fitting techniques have later become significant in devising formulae for PI and PID parameters. The error integral criteria are the most common for such optimisations [3–7]. Most of these tuning schemes are valid only for limited range of NTD problems. More recently Khan and Lehman [8] have used extensive simulations and data fittings to obtain PI tuning formulae, which can satisfy the NTD ranging from 0.2 to 20. In the curve-fitting approaches the PID parameters are arbitrarily expressed as functions of process terms, typically in terms of NTD and first-order time constant. Hang *et al.* [9] have critically examined the ZN settings and, by introducing an additional variable (set-point weighting), the excessive overshoot in the original ZN settings have been reduced while preserving

the same load disturbance characteristics. In development of the former refined ZN (RZN) rules, the ZN PI settings have been completely revised to cope with relatively long time delay problems. The set-point adjustment has the ability to control the transient response at a desired speed. By having a variable set-point weighting, the RZN method has been further refined [10]. By allowing the set-point weighting to vary during transient, the improved RZN method has been able to recover the response speed. Using numerical simulations, empirical formulae have been developed for determining the adaptive set-point values. The RZN method is quite effective and shows excellent performance for short NTD problems. However, it has limitations to use for processes having NTD greater than one.

Frequency domain analysis and development of PID tuning are reported [11–14]. These methods have the advantages of obtaining different PID parameters by selecting desired points in the Nyquist curve [11, 12] or by using user specified phase and gain margins [13, 14]. The later is called as GPM tuning. The internal model control (IMC) approach (time-domain based study) to design PID parameters is shown in [15]. The IMC design simplifies the PID settings to a single parameter, which is directly related to the proportional gain and therefore the response speed. Therefore, IMC design has the advantage of obtaining PID parameters to accommodate the actuator saturation [16]. Also, the PID formula for zero time delay simplifies to a pole zero cancellation PI controller. Due to the first-order Padé approximation of the exponent term, the applicability of the IMC method is also limited to relatively short time delay problems. Hang *et al.* [17] have presented a comparative study of the IMC and GPM approaches and concluded the IMC design has a lower flexibility in terms of robustness.

Based on this literature review, we address three main issues, which are related to PID tuning. The first is related to PID control of processes modelled with zero or negligible time delays. The numerical optimisation in [4]

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suggests that for zero time delay processes the best selection is a PI controller. While the majority of tuning rules such as ZN, RZN, GPM and error integral optimised methods are not applicable, the IMC design simplifies the PID settings to a pole zero cancellation PI controller for zero time delay plants. However, a time domain-based closed form mathematical analysis of PID controlled response for zero time delay has not so far been reported in the literature. Therefore, in this paper we first analyse FOPTD systems with zero time delay and the PID controller settings are analytically deduced. The second issue is related to PID parameter selection to accommodate the maximum capacity or gain of the actuator while avoiding the hazard of integral wind-up. The applicability of tight PID control based on NTD and the normalised process gain was discussed in [18]. The importance of the PID design to limit the overshoot of controller signal has been argued at length [16, 19]. However, the optimum PID controller selection based on the actuator limitations hasn't so far been adequately addressed. We extend the analysis to deduce new PID tuning scheme applicable for short and medium long time delay ($0 < \text{NTD} < 2$). The PID controller gains are selected based on the actuator's saturation. The practical limits for PI and PID control are also deduced. The third issue is related to plants having large NTD (> 2). To accommodate them, a new PI tuning scheme is analytically derived and the tuning is based on user defined two points in the transient response curve. This scheme can be used for any higher value NTD. Therefore this paper intends to provide a complete analysis of PID tuning for first-order plant models covering the complete range of NTD.

1.1 Controller/process specifications

This paper aims at developing PID tuning schemes for the class of problems that have monotonic open-loop characteristics except for the initial time period when such a process can be approximated to a FOPTD model. Let a FOPTD with time delay (t_d), time constant (T) and steady state gain (k) is given by the transfer function

$$G(s) = \frac{k \exp(-t_d s)}{Ts + 1}$$

The experimental identification of the three terms using many techniques are well described in [20]. The PID controller with gain (k_c), derivative time (T_d) and integral time (T_i) is given by

$$u(t) = k_c \left(e(t) + T_d \frac{de(t)}{dt} + \frac{1}{T_i} \int_0^t e(t) dt \right) \quad (1)$$

The feedback error $e(t)$ = reference signal $r(t)$ – response signal $y(t)$. In the practical PID controller with a derivative filter, the controller output is given by

$$u(t) = k_c \left(e(t) - T_d \frac{dy_f(t)}{dt} + \frac{1}{T_i} \int_0^t e(t) dt \right) \quad (2)$$

$$(T_d/N) \frac{dy_f(t)}{dt} = y(t) - y_f(t)$$

The N is an arbitrary number associated with the derivative filter. For $N > 10$ the same PID parameter values obtained from (1) can be implemented with the derivative filter without any significant difference [7]. The other practical forms generally implemented in commercial controllers can be obtained from eqn. 2 as described in [20]. For this work, the design of a PID controller is considered as a

process of determining three unknown gain parameters given by

$$u(t) = k_p e(t) + k_D \frac{de(t)}{dt} + k_I \int_0^t e(t) dt \quad (3)$$

where the PID gain terms are, respectively, given by: $k_p = k_c$, $k_I = k_p/T_i$ and $k_D = k_c T_d$. The remainder of this paper consists three parts of analysis followed by simulation examples. Without loss of generality, the PID methodology in this paper is described with respect to unit step response and zero initial conditions are assumed. With a suitable error transformation, the general step response can be transformed to unit step case with least effort.

2 Analysis I: For zero or negligible NTD problems

The Laplace form of eqn. 3 can be expressed with the initial error signal $e(0)$ by

$$U(s) = k_p E(s) + k_D (sE(s) - e(0)) + k_I E(s)/s$$

where for closed-loop feedback control $E(s) = R(s) - Y(s)$. The cascade closed-loop PID controller system with no external disturbance, the plant output in the Laplace form is given by

$$Y(s) = U(s)G(s)$$

For a unit step input, $R(s) = 1/s$ and $e(0) = 1$. Substituting $t_d = 0$ for $G(s)$ the output can be simplified to

$$Y(s) = \frac{1}{(K_3 + T)s^2 + (K_1 + 1)s + K_2} \left(K_1 + \frac{K_2}{s} \right) \quad (4)$$

where the normalised PID gain terms are defined as, $K_1 = k k_p$, $K_2 = k k_I$ and $K_3 = k k_D$. Our objective in this exercise is to relate the PID parameters to the closed-loop response behaviour. Therefore, expressions for the rise time and overshoot (or undershoot) of eqn. 4 had been deduced and the derivation was based on the nature and positions of the closed-loop poles in the s -plane. The final results are shown below.

2.1 Rise time and peak overshoot in the transient response

Case I: The closed-loop poles are real and distinct

By examining the real closed-loop poles, a general relationship between the normalised gains was established and is given by

$$K_2 = \frac{K_1}{(K_3 + T)} \left(1 + \beta \frac{(K_1 - 1)^2}{4K_1} \right) \quad (5)$$

where, β is a positive real number and its range is constrained to be within $0 \leq \beta < 1$ for the closed-loop poles in eqn. 4 to be real and distinct. Also, within this range and when $K_1 > 1$, the peak overshoot is always positive. Case II covers $\beta = 1$, which corresponds to equal roots. Also, it can be easily proved that, when $\beta = 0$, the peak overshoot of the response also becomes zero.

Case I-a: $K_1 > 1$ and $0 < \beta < 1$

The rise time (T_r) based on 0–100% response and peak overshoot (M) are given by

$$T_r = \frac{(K_3 + T)}{(K_1 - 1)\sqrt{1 - \beta}} \ln \left(\frac{1 + \sqrt{1 - \beta}}{1 - \sqrt{1 - \beta}} \right) \quad (6)$$

$$M = \frac{(1 + \sqrt{1 - \beta})}{(\gamma - \sqrt{1 - \beta})} \times \left(\frac{(\gamma - \sqrt{1 - \beta})(1 - \sqrt{1 - \beta})}{(\gamma + \sqrt{1 - \beta})(1 + \sqrt{1 - \beta})} \right)^{(\gamma + \sqrt{1 - \beta}/2\sqrt{1 - \beta})} \quad (7)$$

respectively, where $\gamma = (K_1 + 1)/(K_1 - 1)$.

Case I-b: $\beta = 0$

In this case the overshoot, $M = 0$ and T_r based on 10–90% response is given by

$$T_r = ((K_3 + T)/(K_1 + 1)) \ln 9 \quad (8)$$

Case II: Closed-loop poles are real and equal

This case refers to the gain relationship of eqn. 5 with $\beta = 1$.

Case II-a: $K_1 > 1$

The T_r based on 0–100% response and M are given by

$$T_r = 2(K_3 + T)/(K_1 - 1) \quad (9)$$

$$M = \frac{(K_1 - 1)}{(K_1 + 1)} \exp(-2K_1/(K_1 - 1)) \quad (10)$$

From eqn. 10 it is clear, when $K_1 < 1$ the system shows stable overdamped response.

Case II-b: $K_1 = 1$

The critically damped response has zero overshoot and the rise time based on 10–90% is given by

$$T_r = (K_3 + T) \ln 9 \quad (11)$$

Case III: The closed-loop poles are complex with negative real parts

This case is realised when the relative damping factor ζ of the closed-loop system is chosen within $0 < \zeta < 1$, while satisfying the normalised gain relationship given by

$$K_2 = \frac{(K_1 + 1)^2}{4\zeta^2(K_3 + T)} \quad (12)$$

The peak overshoot of the under damped response can be shown as

$$M = \sqrt{(1 + P^2)(1 - \zeta^2)} \exp\left(\frac{-\zeta(\theta + \phi)}{\sqrt{1 - \zeta^2}}\right) \quad (13)$$

where

$$P = \frac{\zeta}{\sqrt{1 - \zeta^2}} \frac{(K_1 - 1)}{(K_1 + 1)}, \quad \phi = \tan^{-1} \frac{\sqrt{1 - \zeta^2}}{\zeta}$$

and

$$\theta = \begin{cases} \tan^{-1}(1/P) & \text{for } K_1 \geq 1 \\ \pi - \tan^{-1}(1/|P|) & \text{for } K_1 < 1 \end{cases}$$

It can also be shown that M is always positive when ζ is within 0 and 1. Therefore, the rise time based on 0–100% can be shown as

$$T_r = \frac{\zeta(K_3 + T)\theta}{(K_1 + 1)\sqrt{1 - \zeta^2}} \quad (14)$$

2.2 Optimal tuning law for processes having zero time delay

By observing the rise time given for all the cases above, we can see that the addition of the derivative term, which corresponds to K_3 in eqns. 6, 8, 9, 11, and 14 slows down the transient response. Also the overshoot in all cases can be controlled by the normalised proportional gain while choosing the integral gain satisfying the gain relation, eqn. 5 or 12, corresponds to each case. Therefore, we can conclude that, for optimum design of PID controller for any FOPTD system with zero lag-time, the derivative gain should be zero. The theoretical model with the PI controller has an infinite gain margin and the system can be operated with any value of a controller gain. The upper saturation level of the actuator gain can determine the maximum gain of the controller. For unit step response the maximum controller signal u_{max} is given as follows.

Case A: when closed-loop poles are real or are complex with $K_1(2\zeta - 1) \geq 1$

$$u_{max} = u(0) = k_p$$

Case B: closed-loop poles are complex with $K_1(2\zeta - 1) < 1$

The peak controller signal corresponding to time

$$t_p = \frac{1}{(K_1 + 1)} \frac{2\zeta T}{\sqrt{1 - \zeta^2}} \times \tan^{-1} \left[\frac{\left((K_1 + 1)^2 - (2K_1\zeta)^2 \right) \sqrt{1 - \zeta^2}}{4K_1^2(1 - \zeta^2) - (K_1 - 1)^2} \right]$$

and

$$u_{max} = k_p e(t_p) + k_i \int_0^{t_p} e(t) dt$$

This can be simplified to as

$$u_{max} = \frac{1}{k} \left(\frac{K_1 + 1}{2\zeta} - \frac{2K_1\zeta}{K_1 + 1} \right) \exp\left(-\frac{(K_1 + 1)}{2\zeta T} t_p\right) - \frac{1}{k}$$

The above conditions hold only if $K_1 > 1$. With 5% or less overshoot, case B occurs when the normalised proportional gain is closer to 1 and also u_{max} in this case is not significantly greater than k_p . Therefore, in most cases the proportional gain k_p can have any value as high as the actuator's upper limit of saturation. This condition is implicitly stated in the IMC design [15]. The remaining integral gain can be selected by choosing a desired level of overshoot of response. Let M_d denote the desired peak overshoot level. The optimal tuning law can be stated as follows.

- With zero time delay, PI controller is optimum where $K_3 = 0$, and therefore $k_D = 0$.
- Select the proportional gain based on the actuator saturation. If the actuator's upper limit of saturation is U_u , select $(k_p)_{max}$ by assigning $u_{max} = U_u$. This allows the fastest rise time.
- If the $K_1 \leq 1$, select the necessary relative damping for the given M_d from eqn. 13 and thus compute K_2 using eqn. 12.
- If $K_1 > 1$, compute the peak overshoot using eqn. 10 by assuming closed-loop poles are real and equal. If the computed value is greater than M_d , select β for the given M_d from eqn. 7 and then compute K_2 using eqn. 5. Otherwise, choose relative damping and compute as in (c).

3 Analysis II: For short and medium NTD problems

The second part of the analysis considers FOPTD systems satisfying the range $0 < \text{NTD} < 2$. The exact time-domain analysis of $G(s)$ with the time exponent is a complex mathematical task, owing to the nonlinear exponential term in the transfer function. The Padé approximation [15] or truncated time series approximation [21] of the exponent term results in losing significant poles that exist at distances from the imaginary axis of the s -plane.

3.1 Ultimate gain and ultimate frequency

The definition of ultimate gain and ultimate period refers to the continuous oscillation of the closed-loop response with constant amplitude when the process is controlled only through a proportional controller. The closed-loop characteristic equation of a proportional controlled FOPTD system with zero derivative and integral actions is given by

$$Ts + 1 + K_c \exp(-st_d) = 0$$

where the normalised gain $K_c = kk_c$. The ultimate normalised gain (K_u) and frequency at which the inner locus cuts the imaginary axis are given by

$$K_u = \sqrt{1 + T^2 \omega_u^2} \quad \text{and} \quad \omega_u t_d - \sin^{-1}(1/K_u) = \pi/2$$

By simplifying above, the FOPTD terms can be expressed in terms of ultimate terms as

$$\left. \begin{aligned} T &= \frac{t_u}{2\pi} \sqrt{k^2 k_u^2 - 1} \\ t_d &= \frac{t_u}{2\pi} \left(\frac{\pi}{2} + \sin^{-1} \left(\frac{1}{k k_u} \right) \right) \end{aligned} \right\} \quad (15)$$

where k_u and t_u are the ultimate gain and ultimate period of the FOPTD system, respectively. The method of obtaining approximate estimates k_u and t_u through closed-loop relay experiment is described [20]. By estimating the process gain k by an open-loop test and by using eqn. 15, estimates for T and t_d can be made.

3.2 PI and PID tuning analysis

The limiting values of the gain terms very much depend on the actuator's upper level of saturation. Also, the allowable overshoot of the controller signal is limited in most industrial problems [19]. Therefore, in this exercise overshoot is selected as the main performance criterion.

To simplify the analysis, we assume some of the results given in Section 2.1. It has been observed that, when overshoot control parameter β is set to zero, the response of a zero NTD process has zero overshoot. Therefore we force $\beta = 0$ and the gain relationship of eqn. 5 is simplified to

$$K_2 = K_1/(K_3 + T) \quad (16)$$

3.2.1 PI tuning: For PI controller set $K_3 = 0$. Eqn. 16 further simplified to $K_2 = K_1/T$. The closed-loop characteristic equation with the PI controller then simplifies to

$$(Ts + 1) \left(s + \frac{K_1}{T} \exp(-st_d) \right) = 0$$

The roots of the inner root locus are at $-1/T$ and $-1/t_d$. The first root corresponds to cancellation of the dominant process pole by the PI controller. With this PI setting it can be seen that the theoretical response during the second delay period would be a straight line. The second root corresponds to the break point gain (K_{1b}), and is given by

$$K_{1b} = (T/t_d) \exp(-1) = 0.368(T/t_d)$$

The break point gain exhibits the critical damping condition of the closed-loop system. From this analysis we have seen that the normalised proportional gain is always a function of the scaled time constant (T/t_d). Therefore, it is now reasonable to write a general expression for K_1 as

$$K_1 = \rho \tau_d$$

where the scaled time constant is defined as the reciprocal of NTD, $\tau_d = T/t_d$.

The coefficient ρ is termed as proportional weighting. A value for ρ can be selected to generate the required performance characteristics. In the ZN step response method, this term is fixed and assigned 0.9 for PI controller and 1.2 for the PID controller. With the above PI settings, the plant response for unit step input has been further analysed. For convenience we define a dimensionless scaled time as $\tau = t/t_d$. For three initial scaled time periods the response expressions are shown below and to avoid the overshoot during each period, the limiting values for ρ are also shown.

$$(i) \quad 0 \leq \tau \leq 1 \quad y(\tau) = 0 \quad (17a)$$

$$(ii) \quad 1 \leq \tau \leq 2 \quad y(\tau) = \rho(\tau - 1). \quad \text{For } y(2) \leq 1, \rho \leq 1 \quad (17b)$$

$$(iii) \quad 2 \leq \tau \leq 3 \quad y(\tau) = \rho(\tau - 1) - \frac{1}{2} \rho^2 (\tau - 2)^2. \quad \text{For } y(3) \leq 1, \rho \leq 0.586 \quad (17c)$$

Fig. 1 shows responses for three PI settings that correspond to three values of ρ . From this analysis it is clear now that, if the closed-loop gain is less than the break point gain or $\rho \leq 0.368$, the PI controller would be sufficient. The addition of derivative controller within this range makes the response more sluggish and will take a longer duration to reach the set-point value. Practical control problem can allow a small overshoot to ensure faster rise time and also a

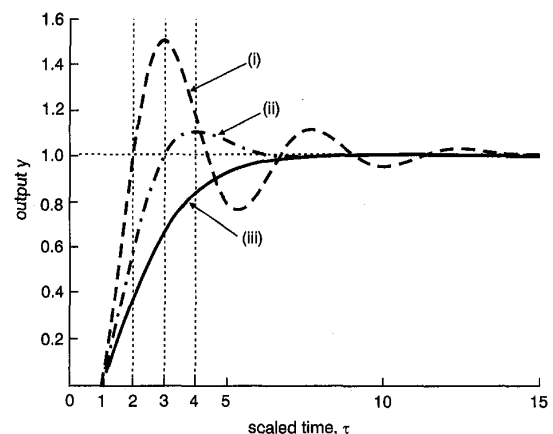


Fig. 1 Effect of proportional weighting in unit-step PI control

- (i) $\rho = 1$
- (ii) $\rho = 0.586$
- (iii) $\rho = 0.368$

faster settling time within an allowable error tolerance. The simulation results showed when $\rho \leq 0.51$, the overshoot $M < 5\%$. This corresponds to about 50% of the set-point response within second time delay period.

In summary, the pole zero cancellation PI controller is

$$\left. \begin{aligned} k_P &= \rho \frac{T}{k t_d} \\ k_I &= \rho \frac{1}{k t_d} \end{aligned} \right\} \quad (18)$$

3.2.2 PID tuning: The tight PID controller can be allowed only when the actuator gain allows the proportional weighting to exceed 0.51 (refer to the next Section). The derivative action is imposed to minimise the undesirable overshoot while achieving a faster rise, which is limited in the above PI design. During the first unit of scaled time period, $0 \leq \tau \leq 1$, error is uniform and its derivatives are zero. Therefore, only proportional and integral control actions are active. With the above PI settings and when $\rho = 1$, the response reaches its target value during this period and overshoot is inevitable (Fig. 1). Hence the maximum limit for proportional weighting is set as $\rho_{max} = 1$. Next, assume the normalised derivative gain can be represented by, $K_3 = \alpha T$ and α is termed as derivative weighting. Using the same gain relationship in eqn. 16, the first-order differential equation showing the time response within the scaled time $n \leq \tau \leq n+1$ can be expressed as

$$\tau_d \frac{dy(\tau)}{d\tau} + y(\tau) = \rho \tau_d e(\tau - n) + \alpha \tau_d \frac{de(\tau - n)}{d\tau} + \frac{\rho}{(\alpha + 1)} \left[\int_n^\tau e(\tau - n) d\tau + \sum_{j=0}^{n-1} \int_j^{j+1} e(\tau - j) d\tau \right]$$

The time response is therefore, $y(\tau) = f(\tau_d, \rho, \alpha)$. This suggests that, when the proportional weighting, ρ is fixed, the response pattern can now be controlled by α alone. The over-weighting of damping through α lowers the gain margin and at the maximum limit of α the system may become unstable. As an example, when $\alpha = 1$, the gain margin drops below 1 dB. For these PID settings, we can define a safety range for α as, $0 \leq \alpha < 1$. Also, from the above equation we can infer that, for optimum control based on a given cost function, α is a function of ρ only. Using time-response simulations when $\rho = 1$, α is set at 0.4. Similarly, for the other limit, when $\rho = 0.50$, α is set at 0.1. The PID settings derived by [5], which corresponds to optimum integral of the absolute value of the error (IAE) have been carefully analysed. The equivalent α and ρ computed for the PID values in [5] have shown an approximately linear relation to each other. Therefore, within the range of $0.50 \leq \rho \leq 1$, the relationship between two weightings is assumed to be linear and fixed as

$$\beta = 0.6\rho - 0.2 \quad (19)$$

The analysis up to now has simplified the PID settings to a single unknown variable (ρ). This term can be adjusted until a desirable overshoot is achieved. Since with the simplified PID settings the response is only a function of τ_d and ρ , a single relationship for proportional weighting can be obtained in terms of the scaled time constant τ_d . A simulation experiment was performed for different τ_d values and the proportional weighting was adjusted in each simulation to retain 5% or less overshoot. The plot of ρ against τ_d is shown in Fig. 2. Since the derivative

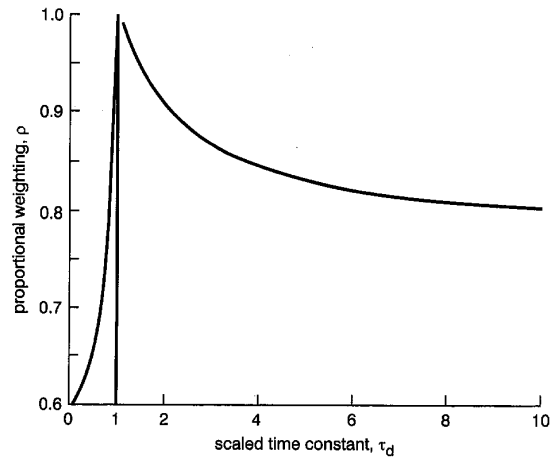


Fig. 2 Proportional weighting for 5% or less overshoot

weighting has been set proportional to the proportional weighting, a low value of ρ means that the amount of damping needed by PID is also low. From this variation we can conclude that, when the normalised time delay is very small (or $\tau_d \gg 1$) the PI control is sufficient. This is in agreement with our previous results on Analysis I. Also, when the normalised dead time is sufficiently large (or $\tau_d \ll 1$), the amount of damping requirement reduces. This is mainly due to the existence of many closed-loop poles near the imaginary axis, where the effect of zero addition by the derivative term is insignificant to change the response characteristics. To derive a PID tuning formula, least square curves have been fitted to the curve shown in Fig. 2 and the expressions for the proportional weighting to provide 5% or less overshoot are obtained.

For relatively short time delay problems $\tau_d > 1$

$$\rho_{PID} = 0.770 + 0.245(\tau_d)^{-0.854} \quad (20a)$$

For relatively long time delay problems $\tau_d \leq 1$

$$\rho_{PID} = 0.603 + 0.275(\tau_d)^{2.4} \quad (20b)$$

Using eqn. 19 the PID parameters are finally expressed using process terms and proportional weighting as shown below. Recommended range: $0.51 \leq \rho \leq 1$

$$\left. \begin{aligned} k_P &= \frac{\rho}{k} \left(\frac{T}{t_d} \right) \\ k_D &= \frac{\alpha T}{k} = \frac{(0.6\rho - 0.2)}{k} T \\ k_I &= \frac{k_P}{k k_D + T} = \frac{\rho}{k(0.6\rho + 0.8)} \left(\frac{1}{t_d} \right) \end{aligned} \right\} \quad (21)$$

A suitable value for ρ is obtained from eqn. 20. The phase and gain margins drawn for this PID tuning law are shown in Fig. 3 and observed they are, respectively, above 2 and 60°. This is in agreement with most of existing tuning techniques as described in [13].

3.2.3 Optimum controller selection for short and medium long NTD problems: In set-point control of any FOPTD system, the integral action causes the control signal to rise monotonically during the initial period. The wind-up problem can be avoided only if the actuator has a higher capacity than the maximum PID controller signal required for the desired settings. With the above PI or PID settings, the maximum controller signal during the transient

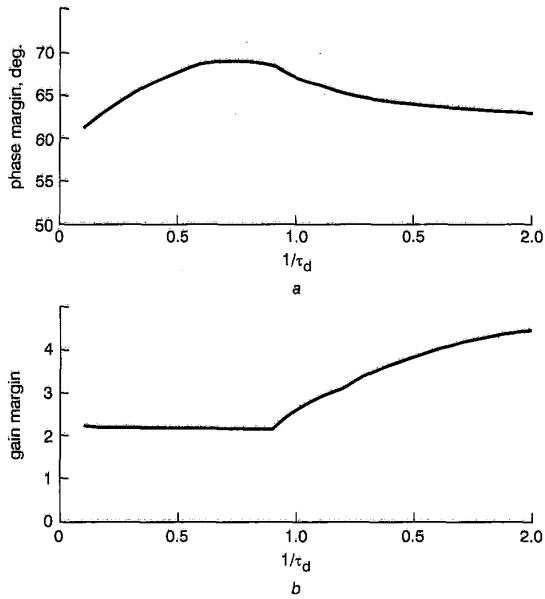


Fig. 3 Gain and phase margin under new PID settings against normalised time delay

response of long time delay processes ($\text{NTD} > 1$) always shows less than 10% overshoot and has an extremely low bandwidth. In such circumstances an actuator with little extra capacity can drive a system with no danger of integral wind-up. Thus we only consider short time delay processes. From the above analysis we have seen that tight PID control can be allowed only when the proportional weighting is chosen at a higher level. Therefore, by knowing the maximum permissible proportional weighting, we can determine whether tight PID control can be allowed or not.

With the PI settings in eqn. 18 and using time response eqn. 17, the maximum controller signal can be determined. There are two cases.

Case A: When $\rho\tau_d \geq 1$ $\tau_p = 1$

Case B: When $\rho\tau_d < 1$ $\tau_p = 1 + (1 - \rho\tau_d)/\rho$

where τ_p is the time at which the PI controllers signal reaches its maximum. By equating the maximum controller signal to the upper saturation margin of the actuator (i.e. $u_{\max} = U_u$) the maximum allowable proportional weighting (ρ_a) is obtained as follows.

$$\text{When } \rho\tau_d \leq 1, \quad \rho_a = kU_u/(1 + \tau_d) \quad (22a)$$

$$\text{When } \rho\tau_d > 1, \quad \rho_a = \left[(1 + \tau_d^2(2U_u - 1))^{1/2} - 1 \right] / \tau_d^2 \quad (22b)$$

For no integral wind-up in PI settings, choose $\rho < \rho_a$. If the optimum proportional weighting is ρ_{PI} , select $\rho = \min(\rho_a, \rho_{PI})$. Based on 5% or less overshoot criterion, $\rho_{PI} = 0.51$.

With the new PID settings it can be shown that the maximum controller signal always occurs at the end of the first dead time period (i.e. $\tau_p = 1$). If there is no saturation limit, the controller signal reaches the maximum value when $t = \tau_d$ and then falls down due to the addition of derivative action. Using the new PID tuning rules, a condition for the maximum PID signal can be expressed by

$$\rho\tau_d/k + \rho/k(0.6\rho + 0.8) \leq U_u \quad (23)$$

To find the limit, consider the equal condition of the above equation and solve the quadratic expression given by

$$0.6\tau_d\rho_b^2 + (0.8\tau_d - 0.6kU_u + 1)\rho_b - 0.8kU_u = 0 \quad (24)$$

As $k > 0$ and $U_u > 0$, eqn. 24 has only one positive solution for ρ_b . To avoid any integral wind-up during PID control, select $\rho = \min(\rho_b, \rho_{PID})$. The analysis above shows that the requirement of a PID controller arises only when the system can allow higher gain, which is governed by the saturation limits. To select the controller, we propose the following rule.

If ($\rho_a \leq \rho_{PI}$), the optimum selection is a PI controller. Otherwise, the selection should be a PID controller.

For very small NTD problems, the limiting proportional weighting (ρ_a) is always small (eqn. 22a) and PI controller would be the optimum. This agrees with our results of Section 2.

4 Analysis III: For large NTD problems (two-point PI design)

In the previous analysis it was observed that, when the normalised time delay becomes larger, the required proportional weighting to retain 5% or less overshoot is reduced to the minimum and the derivative action becomes ineffective. The process response with PI and PID controller becomes almost identical. This phenomenon has been observed in the IMC-PID design [15]. The process dynamic differs, mainly due to the existence of closed-loop poles closer to the imaginary axis of the s -plane. Therefore, we can conclude that, for larger NTD processes, the PI controller is sufficient. The pole zero cancellation PI controller shows an extremely low bandwidth. Therefore, in this Section we derive separate PI tuning formulas for such processes where $T \ll \tau_d$ (or $\tau_d \ll 1$).

It is observed from simulations that, for larger NTD processes, the PI tuning described in Section 3.2.1 overestimates the integral action and gives a lower estimate of proportional weighting. Larger dead time processes require a slower rate of integration to avoid any integral wind-up. Therefore, the normalised integral gain can be assumed to be inversely proportional to the dead time. For this class of problems, the PI gains are redefined as

$$\left. \begin{aligned} K_1 &= \rho_l \tau_d \\ K_2 &= \frac{\gamma_l}{t_d} \end{aligned} \right\} \quad (25)$$

where ρ_l and γ_l are the redefined proportional and integral weightings, respectively, allocated for the PI gains. In the pole zero cancellation PI controller the two weightings are equal to each other and the maximum value permitted for the proportional weighting is one. Since the time delay is sufficiently greater than the process time constant, we can set $\gamma_l < \rho_l$ and ρ_l can be allowed to be greater than one. The first-order differential equation showing the time response within the scaled time $n \leq \tau \leq n+1$ can now be described as

$$\tau_d \frac{dy(\tau)}{d\tau} + y(\tau) = \rho_l \tau_d e^{(\tau - n)} + \gamma_l \left[\int_n^\tau e^{(\tau - n)} d\tau + \sum_{j=0}^{n-1} \int_j^{j+1} e^{(\tau - j)} d\tau \right]$$

The step response solution for the first three delay periods has been obtained and the final expressions are given below. The equal conditions of the response equations

correspond to boundary conditions of the past and future responses.

$$(i) \quad 0 \leq \tau \leq 1 \quad y(\tau) = 0$$

$$(ii) \quad 1 \leq \tau \leq 2 \quad y(\tau) = \rho_l \tau_d + \gamma_l((\tau - 1) - \tau_d) - \tau_d(\rho_l - \gamma_l) \exp\left(\frac{-(\tau - 1)}{\tau_d}\right) \quad (26a)$$

$$(iii) \quad 2 \leq \tau \leq 3 \quad y(\tau) = y_1(\tau) + y_2(\tau) \quad (26b)$$

where

$$y_1(\tau) = [(\rho_l - \gamma_l)\tau_d(\tau - 2) + \tau_d(\tau_d(\rho_l - 2\gamma_l) - \exp(-1/\tau_d))](\rho_l - \gamma_l) \exp\left(\frac{-(\tau - 2)}{\tau_d}\right)$$

$$y_2(\tau) = (\rho_l - \gamma_l)(\tau_d - \rho_l + 2\gamma_l) + \gamma_l \left(1 - 2(\rho_l - \gamma_l)\tau_d - \gamma_l \frac{(\tau - 2)}{2}\right)(\tau - 2)$$

When $\tau_d \ll 1$, the exponent terms in eqn. 26 can be assumed to be negligible and the responses can be approximately expressed as,

$$\text{for } 1 \leq \tau \leq 2, \quad y(\tau) \approx \rho_l \tau_d + \gamma_l((\tau - 1) - \tau_d) \quad (27a)$$

$$\text{and for } 2 \leq \tau \leq 3, \quad y(\tau) \approx y_2(\tau) \quad (27b)$$

It can be observed from the above expressions that the response during the second period of scaled time given by eqn. 27a has a monotonically increasing function during the valid period of time. This implies that it is impossible for the PI controlled response to reach its steady state before the time $\tau=2$. This is the clear limitation of PI controller performance when it is used for long dead time process [22]. With this limitation, we can allow the system to accelerate as much as possible during the second delay period and control the overshoot in the third period. Therefore, two target points are first defined corresponding to the two response periods. Let us assume the response level to be reached at the end of second time duration $y(2)=y_a$. We set $y_a < 1$ to avoid any excessive overshoot of response. By substituting this condition to eqn. 27a we can obtain

$$\rho_l = (y_a - \gamma_l(1 - \tau_d))/\tau_d \quad (28)$$

To achieve fast settling of the response, we can allow the response during the third delay period to reach the peak of the overall response. Let the expected maximum response height of the unit step response be $y_m (\geq 1)$. By substituting the time given by $dy_2(\tau)/d\tau=0$ to eqn. 27b, we can obtain an approximate estimate of y_m

$$y_m = 1/2 + \gamma_l + \tau_d(\rho_l - \gamma_l)(\rho_l \tau_d - 1)$$

By simplifying the above, the integral weighting can now be expressed as

$$\gamma_l = \frac{1}{(1 - \tau_d)} \left[2(y_m - 1) - y_m \tau_d + (\tau_d(y_m^2 \tau_d - 4y_a + 2) + 2 - 4(y_m - y_a))^{1/2} \right] \quad (29)$$

Using predefined two target points and using eqns. 28 and 29 we can now determine the two weightings ρ_l and γ_l to estimate the PI terms. The error of the approximation is dominant only when the NTD is medium long in which case y_a will be an overestimate and y_m will be an under-

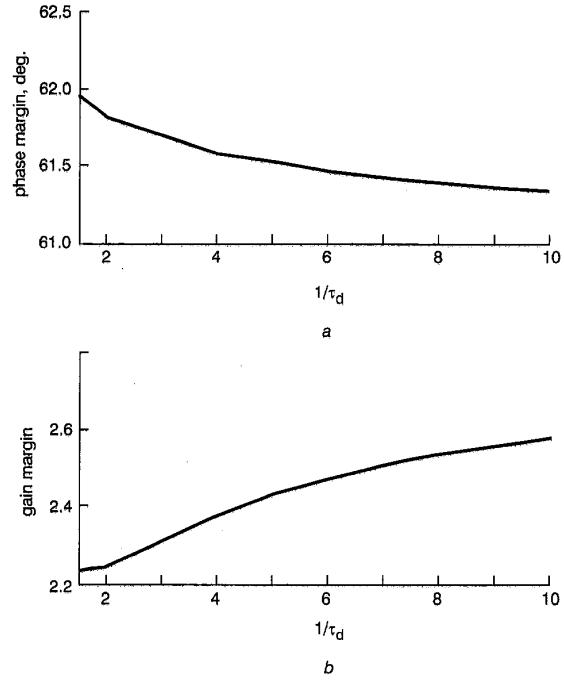


Fig. 4 Gain and phase margins under new two-point PI design against normalised time delay, $y_a = 0.8$ and $y_m = 1.02$

estimate. For larger time delay problems ($NTD > 6$), the above estimations can very accurately be used. The numerical simulations provide the following safe limits for the target points assuming the allowable overshoot is about 5%. Valid range for normalised delay: $\tau_d < 1$ or $t_d > T$

$1 < (t_d/T) < 2$	$y_a = 0.6, y_m = 1.02$
$2 < (t_d/T) < 4$	$y_a = 0.7, y_m = 1.02$
$4 < (t_d/T) < 6$	$y_a = 0.8, y_m = 1.02$
$6 \leq (t_d/T)$	Accuracy is sufficient to predict the two points.

The proposed PI setting is easy to understand. Irrespective of the magnitude of time delay, the user can select two points by relating to the desired set-point response. Since the response is slow for long time delays, the integral wind-up with the above settings would not be a serious problem. With about 10% extra capacity of an actuator, the response with 5% or less overshoot can be easily accommodated with no integral wind-up. The gain and phase margins computed for these settings while seeking $y_a = 0.8$ and $y_m = 1.02$ is shown in Fig. 4. If a higher gain or phase margin is sought, the two target points can be changed appropriately.

5 Simulation examples

In this Section some simulation results are demonstrated. All the simulations were carried out using PID controller architecture with the derivative filter as given in eqn. 2 and without loss of generality $N=10$ is used throughout. The approximated process parameters are evaluated either by using a relay experiment or from a plant open-loop step test. The critical gain evaluation by the relay experiment for negligible time delay or very long time delay gives rather an erroneous estimate. Therefore, those types are estimated by using the open-loop step response method as

shown [23]. During each simulation a constant 50% load disturbance is added.

Example 1: A third-order system with zero time delay
A third-order transfer function model is chosen.

$$G(s) = \frac{2}{(5s+1)(0.5s+1)(0.2s+1)}$$

The open-loop step response test yields $k=2$ and $T=5.72$ s and the time delay component is negligible. For unit step response, the steady-state controller gain required is 0.5. Let the actuator saturation limits be given by $[0 \ 4]$ units. Using the results of Section 2.1, the optimum PID controller is PI. Using the tuning law in section 2.2, $u_{max} < 4$. Select $k_p = 3$ or $K_1 = 6$. By assuming the maximum expected overshoot to be 2% ($M_d = 0.02$), we obtain $k_i = 0.6086$. Therefore, the PID parameters are $k_c = 3$ and $T_i = 4.93$. The response curve for this PI setting is shown in Fig. 5. The response has achieved its fastest rise without exceeding the saturation limits, but has shown somewhat poor load disturbance characteristics. By sufficiently increasing the integral gain, it is possible to obtain better load disturbance, but at the expense of poor transient response.

Example 2: A second-order system with a relatively short time delay

This example is taken from the reference [7]. The transfer function is given by

$$G(s) = \frac{\exp(-0.5s)}{(s+1)^2}$$

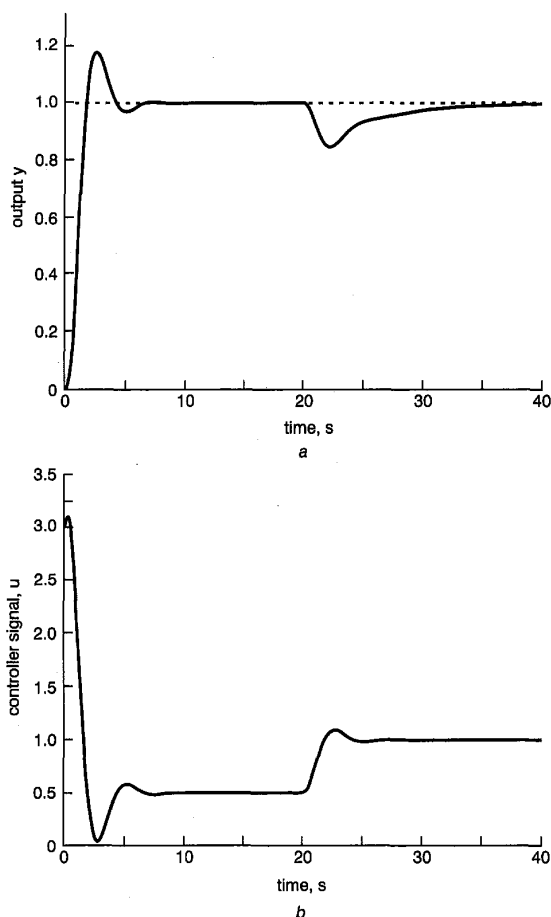


Fig. 5 Step-response of Example 1

The closed-loop relay experiment yields $k_u = 3.45$ and $t_u = 3.32$ s. Using eqn. 24 the process estimates are $T = 1.746$ s and $t_d = 0.985$ s. The process gain $k = 1$. The design is related to a short normalised time delay problem ($t_d/T < 1$ or $\tau_d > 1$). According to our design criteria, the optimum selection is either a PI or PID controller, which depends on the actuator saturation. The proportional weighting from eqn. 20a is 0.92 and the estimated PID parameters using eqn. 21 are $k_p = 1.631$, $k_i = 0.691$, and $k_D = 0.615$ (or $k_c = 1.631$, $T_i = 1.77$, and $T_d = 0.377$). Next, the actuator saturation limits of $[0 \ 1.6]$ are imposed where $U_u = 1.6$. This corresponds to 60% extra capacity at steady-state values. With this limit, the above proportional weighting value violates eqn. 23 where there is a possibility of integral wind-up. Using eqn. 22 the limiting proportional weighting for PI design is 0.58. Since this value is higher than the limiting value corresponding to a PI controller, we can allow tight control through a PID controller. The limiting proportional weighting calculated from eqn. 24 implies $\rho \leq 0.608$. Using the PID formula given in eqn. 32, the new PID terms are given by, $k_p = 1.078$, $k_i = 0.530$, and $k_D = 0.288$ (or $k_c = 1.078$, $T_i = 2.034$, and $T_d = 0.267$). For comparison, the RZN PID settings and Zhung and Atherton's ITSE PID (ZA-PID) settings are also tested. With the estimated process terms above, the RZN-PID terms or $k_c = 2.071$, $T_i = 1.66$, and $T_d = 0.415$ and the set-point weighting = 0.626 and ZA-PID terms; $k_p = 1.76$, $T_i = 2.01$, and $T_d = 0.415$ have been obtained.

The simulation results are shown in Fig. 6. The proposed method shows acceptable step response performance and similar to ZA performance. The RZN tuning generally shows better transient response as well as better load disturbance properties for short time delay problems. This is mainly due to the fact that RZN is a four-parameter controller and the method is specifically designed for short time delay problems. However, the transient response characteristics of the proposed method has shown acceptable performance. With actuator saturation, the method requires lowering the gains to accommodate the limiting requirements and therefore shows poor load disturbance characteristics compared to other two methods. It can be seen from Fig. 6 that the proposed method has satisfied the limiting conditions of the actuator gain and therefore the response has not been affected by the integral wind-up. However, by having a low set-point weighting, it is possible to avoid saturation with RZN settings as well. The controller signals based on the other two designs have reached the upper saturation. This example illustrates the flexibility of the PID design for accommodating the actuator saturation. The IMC-PID design has shown similar results as proposed design and therefore, it has been excluded from the diagram.

Example 3: Second-order plant having a normalised time delay closer to one

This example was chosen from the reference [14] and the transfer function is

$$G(s) = \frac{\exp(-1.0s)}{(s+1)(0.5s+1)}$$

The closed-loop relay experiment yields $k_u = 2.137$ and $t_u = 4.1$ s. The estimates are $k = 1$, $T = 1.232$ s and $t_d = 1.343$ s. The normalised time delay is therefore, 1.09. Using the proposed PID formula, $\rho = 0.827$, the gain terms are: $k_p = 0.759$, $k_i = 0.475$ and $k_D = 0.365$ (or $k_c = 0.759$, $T_i = 1.60$, and $T_d = 0.481$). For comparison, the Zhung and Atherton ITSE PID settings and the PID settings based on

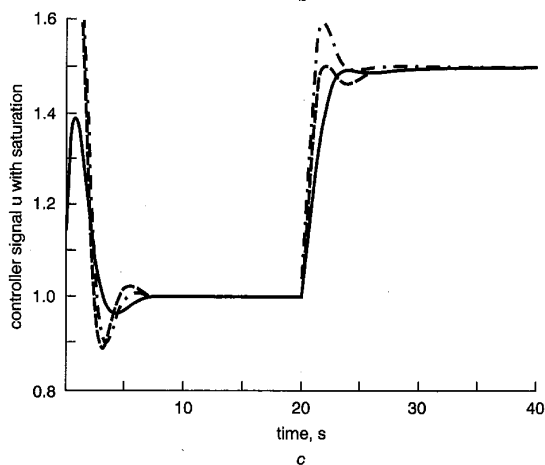
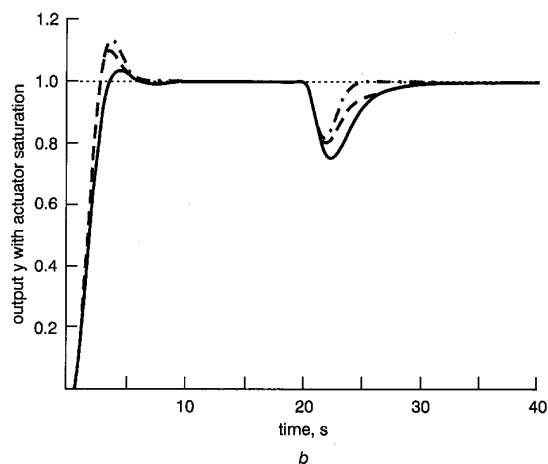
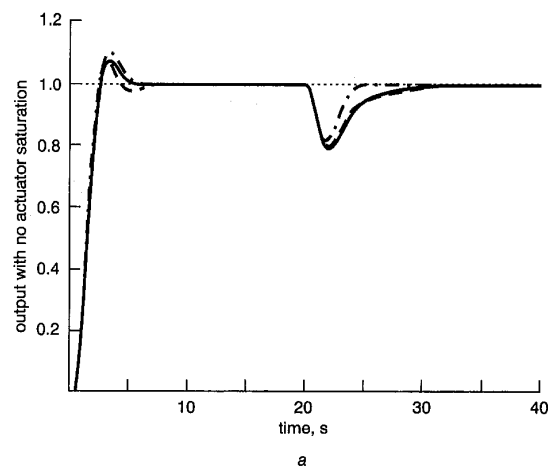


Fig. 6 Step-response of Example 2

— proposed PID
 - - - RZN-PID
 . . . ZA-PID

gain and phase margin specifications (GPM-PID) are also used. The ZA-PID settings are: $k_c = 1.07$, $T_i = 2.06$, and $T_d = 0.46$. The GPM-PID settings corresponding to a gain margin of 3 and a phase margin of 60° are $k_c = 0.78$, $T_i = 1.50$, and $T_d = 0.33$. The response curves are shown in Fig. 7. The proposed PID setting shows minimum set-point overshoot and satisfactory load disturbance characteristics compared with both ZA-PID and GPM-PID methods. The RZN PID setting is inadequate to cover the

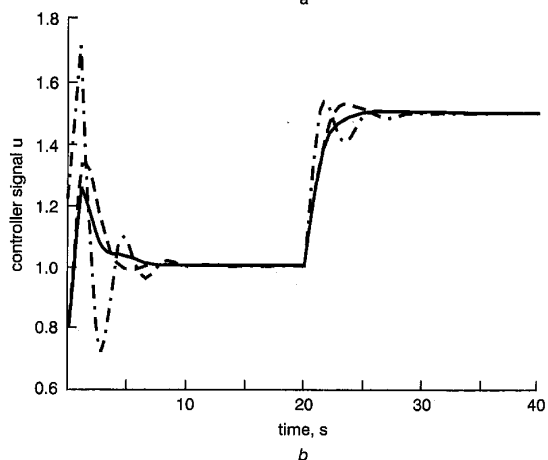
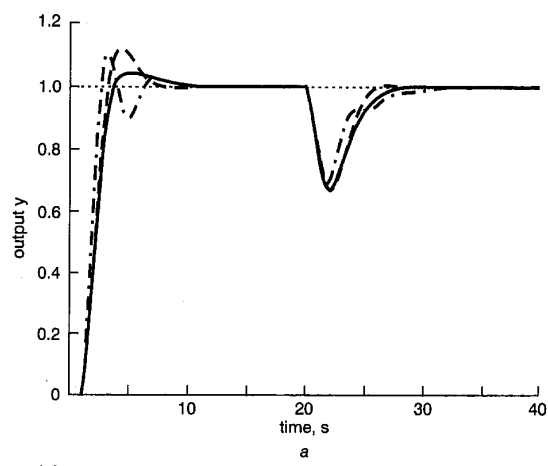


Fig. 7 Step-response of Example 3

— proposed PID
 - - - ZA-PID
 . . . GPM-PID

particular example, and with the closed estimated values the response has shown quite an oscillatory response.

Example 4: Higher-order plant with long normalised time delay

This example was chosen from the reference [22]. The transfer function of the model is given by

$$G(s) = \frac{\exp(-t_d s)}{(s+1)(0.5s+1)(0.5s+1)(0.125s+1)}$$

and for simulations two cases are considered.

Case I: $t_d = 4$ s.

The closed-loop relay experiment yields, $k_u = 1.302$ and $t_u = 11.46$ s. The estimates are $k = 1$, $T = 1.521$ s and $t_d = 4.462$ s. The normalised time delay is therefore 2.934. Using the proposed PID rules, $\rho = 0.6238$ and the gain terms are: $k_p = 0.213$, $k_i = 0.119$ and $k_D = 0.265$ (or $k_c = 0.213$, $T_i = 1.79$, and $T_d = 1.244$). Next, we have used the two-point PI design (TP-PI). Using the recommended values given in Section 3.3, the two points are decided as $y_a = 0.7$, $y_m = 1.02$. Using the expressions (eqns. 28 and 29) the PI weightings are $\rho_i = 0.9058$ and $\gamma_i = 0.133$. Using eqn. 25 the PI terms are $k_p = 0.3088$ and $k_i = 0.133$ (or $k_c = 0.309$ and $T_i = 2.32$). For comparison, the IMC-PI and PI settings developed for large normalised time delay processes by Khan and Lehman, (KL-PI) [8] are also simulated. The IMC-PI terms corresponding to

($\varepsilon/t_d = 1.7$ are $k_c = 0.495$ and $T_i = 3.76$. The KL-PI parameters are $k_c = 0.319$ and $T_i = 2.56$. The unit step response curves and the variation of the manipulated signal are shown in Fig. 8.

Case II: $t_d = 10$ s

For better approximation, the parameters are estimated by the open-loop step response. The estimates are $k = 1$, $T = 1.5$ s and $t_d = 10.5$ s. Based on the new two-point PI design, two designs are evaluated. With $y_a = 0.75$, $y_m = 1$ the PI gains are $k_p = 0.2281$ and $k_i = 0.0580$ (or $k_c = 0.228$ and $T_i = 3.93$) and with $y_a = 0.8$, $y_m = 1.02$ the PI gains are $k_p = 0.2516$ and $k_i = 0.0609$. Again, the KL-PI setting is compared and the corresponding PI gains are $k_c = 0.266$ and $T_i = 4.59$. The unit step response curves are shown in Fig. 9.

It can be seen from Example 4-I that the proportional weighting allowed for PID settings is low. As a result, the performance based on the proposed two-point PI setting is superior to the proposed PID controller. This proves that the PI controller is sufficient for controlling larger normalised time delay processes. Figs. 8 and 9 show that, when the normalised time delay is large, the signal overshoot is minimal and therefore, a proper PID design giving no excessive overshoot of response automatically satisfies the actuator limitations. The IMC-PI design employed

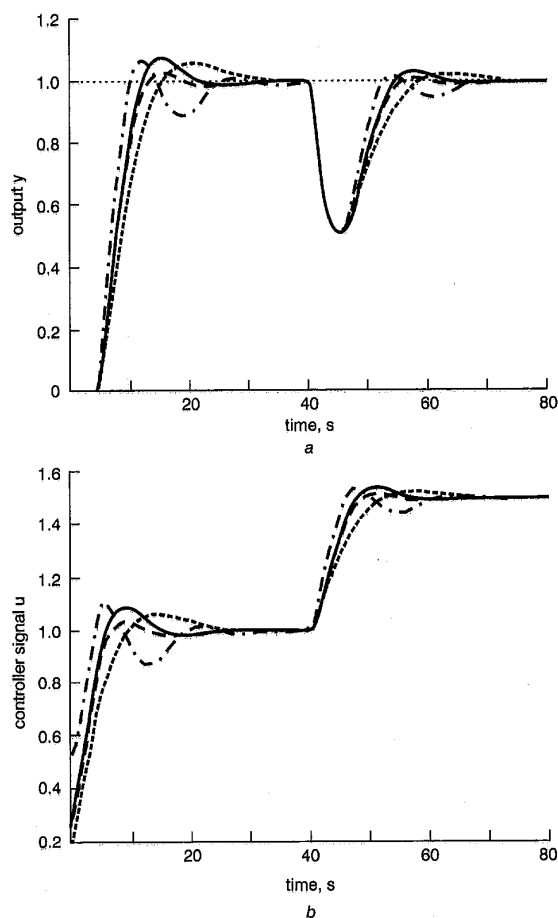


Fig. 8 Step-response of Example 4-I
 — proposed TP-PI ($y_a = 0.7$, $y_m = 1.02$)
 --- proposed PID
 ... KL-PI
 -.- IMC-PI

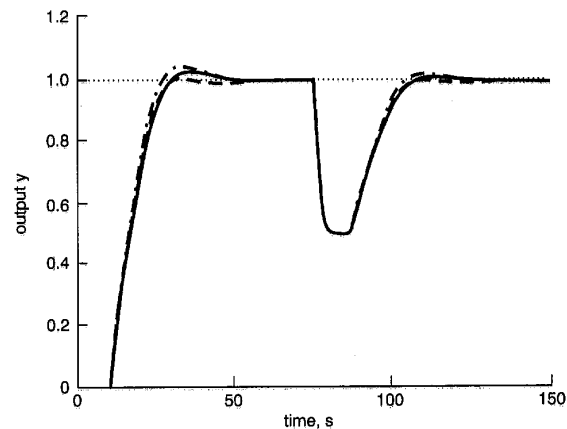


Fig. 9 Step-response of Example 4-II
 — proposed TP-PI ($y_a = 0.75$, $y_m = 1$)
 --- proposed TP-PI ($y_a = 0.8$, $y_m = 1.02$)
 ... KL-PI

the Padé approximation, which resulted in poor PI settings and therefore exhibited the poorest response characteristics. Though KL-PI settings have shown satisfactory performance with zero overshoot, the proposed method has more flexibility to select PI parameters for user specified response behaviour. The two designs in Example 4-II show how one could adjust the overshoot by simply lowering the anticipated two points in the response curve. The results show better performance of TP-PI settings in both step response and disturbance rejection. The two-point PI design analysis in Section 3.3 has shown the best that a PID controller can achieve. Further improvement to the transient response requires employment of different control algorithms, such as the predictive PI controller (PI(P) controller) in [22]. Any increase in the rise time by raising the first expected point y_a would produce rather an excessive overshoot of the response. The other tuning methods, such as RZN-PI, ZA-PI and GPM-PI are unable to provide solutions to very long time delay problems.

6 Summary

This paper has presented a new time response based design methodology for PID controllers. Based on the magnitude of NTD, three types of tuning rules have been developed to cover the time delay ranging from zero to any higher value. The PID tuning rules for zero dead time processes have been analytically obtained. It has been shown that the derivative action is detrimental to those plants having negligible or large normalised time delay. The tight PID control can be applied for plants having low to medium normalised time delay. The new design technique has the flexibility to accommodate the actuator saturation to avoid the danger of integral wind-up in the transient response. Based on the actuator's upper limit of saturation, a selection of PI or PID controllers for such plants have been described. For large normalised time delay plants, a new PI tuning scheme based on user defined two-points of the time response curve has been derived. The analysis has also shown the existing limitations of PI control with respect to performance of transient response.

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