Auto-tuning and adaptive tuning of PID controllers

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Refresher article

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Introduction

PID algorithms are by far the most common way used tot introduce feedback in a control loop, and they constitute the basis of stand-alone controllers, Direct Digital Control systems or hierarchical Distributed digital Control Systems. These algorithms, enhanced with remote set point, override and ratio strategies, can be combined with logic and arithmetic function blocks, selectors and sequencing machines (Programmable Logic Controllers, for example), resulting in powerful systems suited for the control of continuous or batch industrial processes.

The ideal PID algorithm is defined by the equations:

$$e(t) = r(t) - y(t) \tag{1.a}$$

$$u(t) = K\left(1 + \frac{1}{T_I s} + T_D s\right) e(t) \tag{1.b}$$

where "s" is the derivative operator, r(.) is the set point, y(.) is the process output, e(.) is the error signal, u(.) is the control signal and K (proportional gain), T_I (reset time) and T_D (derivative time) are the tuning (or design) parameters.

The most commonly used industrial versions are the non-interacting algorithm:

$$u(t) = K\left[\left(1 + \frac{1}{T_I s}\right)e(t) - \frac{T_D s}{1 + \alpha T_D s}\, y(t)\right] \tag{2}$$

and the interacting one:

$$u(t) = K\left[1 + \frac{1}{T_I s}\right]\left[r(t) - \frac{1 + T_D s}{1 + \alpha T_D s}\, y(t)\right] \tag{3}$$

Note that a low pass filter with a coefficient α ($0.1 \leqslant \alpha \leqslant 0.3$) is incorporated to the derivative term and also that this term is computed from the process output signal only.

PID algorithms implemented in a digital form correspond to discretized versions of such equations. Considering expression (2) and using the "z^{-1}" backward shift operator, the discrete PID can be written as:

$$u(t) = \frac{t_0 + t_1 z^{-1} + t_2 z^{-2}}{(1 - z^{-1})(1 + \gamma z^{-1})} r(t) - \frac{s_0 + s_1 z^{-1} + s_2 z^{-2}}{(1 - z^{-1})(1 + \gamma z^{-1})} y(t) \tag{4}$$

where the coefficients t_i, s_i and γ are related to the sampling time utilized and to the parameters K, T_I, T_D and α present in the continuous PID.

Although PID controllers are simple and well known, to tune them is sometimes a time-consuming task which can even be very difficult if the process has complex behaviour due to non-linearities, time varying characteristics, high order dynamics, or dominant delay. So PID controllers are often poorly tuned and the derivative term is frequently switched off. As a solution for the mentioned problems, some commercial PID controllers with a self-tuning capability have appeared on the market during the last years.

The objectives of introducing such a self-tuning capability are twofold: firstly, to provide systematic, standard and automatic procedures for obtaining the tuning parameters, which is generally classified as auto-tuning; secondly to automatically update these parameters when the process characteristics or its environment are changing, in order to ensure the desired performance, which is known as adaptive tuning.

This paper is intended to review the main ideas used for deriving self-tuning PID controllers. The different strategies that have been proposed in the literature may be characterized by the methods adopted:

i) to extract the knowledge about the process behaviour (dynamics);

ii) to determine the PID tuning parameters by using the knowledge obtained in i).

These two aspects will be considered for the strategies presented in the following text.

This subject is discussed in more detail in reference [1], which provides a survey of the literature published on self-tuning PID controllers up to now.

Auto-tuning

In this strategy, the tuning parameters are determined automatically on demand by the user and, once obtained, the parameters remain unchanged. This method can also be repeated in the whole range of process operation conditions, in order to build a table with the tuning parameters, so that the best parameter values are later recovered depending on the process operating point. The use of such a table is known as gain scheduling strategy, and it introduces a higher level of adaptation in the controller.

The simplest way of auto-tuning is to obtain automatically the open loop process reaction curve, which leads to the approximated first order model with delay given by:

$$G_P(s) = \frac{K_P\, e^{-\tau s}}{1 + Ts} \tag{5}$$

with the coefficients K_P (static gain), T(time constant) and τ (time delay). The tuning parameters are then computed by use of well established tuning rules such as those of Ziegler-Nichols for example. This approach is adopted in the products of Yokogawa, Eurotherm and Honeywell, and is used as a pre-tuning aid in the self-tuning controllers of Leeds & Northrup, Turnbull Control Systems and Foxboro.

The model given by (5) can also be derived from the process closed loop response when a step is added to the set point or the controller output. In this case, the time evolution of the control input and process output signals are recorded in order to compute some "characteristic areas" which are then used to find the coefficients K_P, T and τ in (5).

Another strategy for auto-tuning, used in the products of Satt Control Instruments and Fisher Controls, is shown in *Figure 1*. This strategy and others are described in the excellent reference [2].

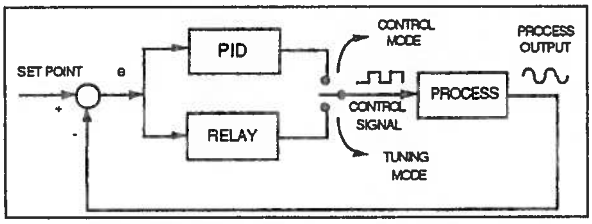


Fig. 1 : Relay method for auto-tuning.

The relay, with suitable values of amplitude $\pm\delta$ and hysteresis $\pm\varepsilon$, is introduced in the feedback loop and leads the process to a state of controlled oscillation. The process output oscillation amplitude σ and period T_O are measured, and the following process model in the frequency domain is obtained:

$$G_P(j\omega) = r_P\, e^{\,j(\pi + \Phi_P)} \quad \text{at} \quad \omega = \frac{2\pi}{T_O} \tag{6.a,b}$$

$$r_p = \frac{\pi\,\sigma}{4\,\delta} \quad ; \Phi_p = \tan^{-1}\left(\frac{\varepsilon}{\sqrt{\sigma^2 - \varepsilon^2}}\right) \tag{6.c,d}$$

As $0 \leqslant \varnothing_p \leqslant \pi/2$, this model corresponds to a point of the process Nyquist curve in the 3 rd quadrant, whose imaginary part depends only on the ε/δ ratio. If $\Phi_p = 0$ ($\varepsilon \approx 0$), we have the intersection between the Nyquist curve and the negative real axis, where the process is found in its stability limit; this intersection, known as the critical point and characterized by the ultimate gain and ultimate period (K_{CR}, T_{CR}), corresponds to

$r_p = 1/\, K_{CR}$ and $T_O = T_{CR}$

Several methods can be used to determine the PID tuning parameters. Let the ideal PID controller, described by equations (1.a,b), be rewritten in a frequential formulation:

$$G_{PID}(j\omega) = r_{PID}\, e^{\,j\,\Phi_{PID}} \tag{7.a}$$

$$r_{PID} = K\sqrt{1 + \left(\omega T_D - \frac{1}{\omega T_I}\right)^2} \tag{7.b}$$

$$\Phi_{PID} = \tan^{-1}\left(\omega T_D - \frac{1}{\omega T_I}\right) \tag{7.c}$$

Introducing the relation $T_D = \mu T_I$ (usually $\mu = 1/4$), the PID parameters can be obtained as the solution of:

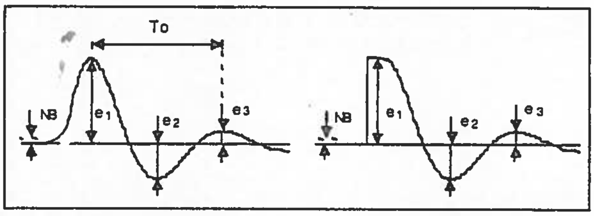
$$r_{PID} = \frac{r_N}{r_p} \quad \text{and} \quad \Phi_{PID} = \Phi_N - \Phi_P \tag{8.a,b}$$

where Φ_N and r_N are chosen such that the phase margin or the amplitude margin is specified. For instance, if $\Phi_N = 0$ and $r_N = 1/AM$ then an amplitude margin AM is specified, whereas $\Phi_N = \Phi_M$ and $r_N = 1$ corresponds to the specification of a phase margin Φ_M. The frequential Ziegler-Nichols tuning rules can be interpreted as the case where $r_P = 1/K_{CR}$, $\Phi_p = 0$, $r_N = 0.66$ and $\Phi_N = 25°$; modified Ziegler-Nicols rules for the case $r_P = 1/K_{CR}$, $\Phi_p = 0$, $r_N = 0.5$ and $\Phi_N = 45°$, as well as other methods to determine the PID tuning parameters that need the determination of more than one point in the Nyquist curve, are also presented in [2].

Pattern recognition

In the pattern recognition approach, the tuning parameters are adjusted when a set point change or a load disturbance occurs. In these cases, the process is characterized by its closed loop response: the error signal e(.) between the set point and the process output is always monitored and compared with a noise band NB that accounts for the noise present in the process output, as shown in *Figure 2.* When |e(.)| exceeds NB, the algorithm searches the values of three successive peaks, e_1,e_2 and e_3 and of the natural oscillation period T_O. The following variables are then computed:

$$\text{overshoot} = \frac{|e_2|}{|e_1|} \quad \text{damping} = \frac{|e_3| + |e_2|}{|e_1| + |e_2|} \tag{9.a,b}$$



(a) for load disturbances (b) for set point changes
Fig. 2 Pattern recognition for a typical error signal.

The tuning parameters are adjusted so that these computed overshoot and damping approach the desired values specified by the user. This adjustment procedure is implemented by means of rules that are drawn from the reasoning followed by experienced operators and engineers when faced with the same problem. However, the variables defined by (9.a,b) are interdependent and, besides this approximately the same values of damping and overshoot but with different oscillation periods can be obtained for several combinations of the tuning parameters. So, other conditions must then be introduced in the adjustment procedure with the purpose of obtaining a unique set of tuning parameters. These conditions are stated through the ratios T_I/T_O and T_D/T_O between the integral and derivative times and the measured natural period. These ratios define respectively the lag and lead angles of the PID controller; from (7.c), the resulting phase angle is given by:

$$\Phi_{PID} = \tan^{-1}\left(\frac{2\pi T_D}{T_O} - \frac{T_0}{2\pi T_I}\right) \tag{10}$$

The value of Φ_{PID} gives some indirect indication of the PID and closed loop performances.

The pattern recognition strategy described above was first presented in the Foxboro product [3]. The rules that adjust the tuning parameters are implemented following the expert systems ideas; the ratios T_I/T_O and T_D/T_O are a function of the τ/T ratio (delay / time constant) and the noise level of the process. A pre-tuning facility that automatically obtains the open loop process reaction curve is used to approximate the model given by (5) and also to determine the noise band NB; then, based on this information, the values of the T_I/T_O and T_D/T_O ratios are chosen and initial values for the tuning parameters are computed.

Other methods based on this approach have been proposed in the literature. For example, the process closed loop response can also be characterized by using different variables like the rise and settling times or ISE and related criteria. Another possibility is to incorporate some confidence indexes in the pattern measurements and to the tuning parameters adjustment rules; these indexes introduce a sort of qualitative reasoning in the adjustment procedure.

Parameter estimation based methods

A very important class of adaptive controllers, usually described as self-tuning controllers, is composed of two loops that are executed at each iteration. One loop comprises a recursive parameter estimation algorithm that identifies a process model; the estimated parameters are then used to adjust the coefficients of a linear control law that constitutes the other loop.

If the mentioned control law has a PID structure, one can obtain a self-tuning PID controller based on parameter estimation, as shown in *Figure 3*. In this case, it is possible to update the tuning parameters continuously.

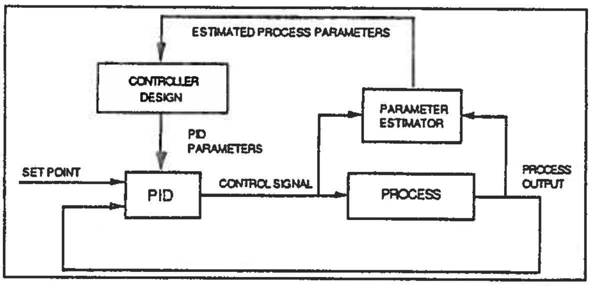


Fig. 3: Self-tuning PID Controller based on parameter estimation.

Several strategies can be considered to implement the "controller design" box in the diagram of Figure 3. In the following, we will present some strategies where the user specifies the desired closed loop system transfer function $G^*_{CL}(.)$.

For instance, consider that the PID controller can be represented as the transfer function $G_{PID}(.)$ given by:

$$\frac{u(t)}{e(t)} = G_{PID}(z^{-1}) = \frac{s_0 + s_1 z^{-1} + s_2 z^{-2}}{(1 - z^{-1})(1 + \gamma z^{-1})} \tag{11}$$

This discrete formulation can be obtained for example by making $t_i = s_i$ in equation (4), which corresponds to the continuous algorithm in (2) but with e(.) also used in the derivative action.

Assume also that the process can be approximated through a 2^{nd} order model defined as the discrete transfer function $G_p(.)$ given by:

$$\frac{y(t)}{u(t)} = G_p(z^{-1}) = \frac{B(z^{-1})}{A(z^{-1})} z^{-d} \tag{12.a}$$

where $d \geqslant 0$ is the process time delay expressed as an integer number of the sampling period h, and the polynomials A(.) and B(.) are:

$$B(z^{-1}) = b_1 z^{-1} + b_2 z^{-2} \tag{12.b}$$

$$A(z^{-1}) = 1 + a_1 z^{-1} + a_2 z^{-2} \tag{12.c}$$

The expression for the closed loop transfer function $G_{CL}(.)$ is then:

$$\frac{y(t)}{r(t)} = G_{CL}(z^{-1}) = \frac{G_{PID}(z^{-1})\, G_p(z^{-1})}{1 + G_{PID}(z^{-1})\, G_p(z^{-1})} \tag{13}$$

Considering that the user has specified the desired closed loop transfer function $G^*_{CL}(.)$, from equation (13) we arrive at:

$$G_{PID}(z^{-1}) = \frac{1}{G_p(z^{-1})} \quad \frac{G^*_{CL}(z^{-1})}{1 - G^*_{CL}(z^{-1})} \tag{14}$$

In order to obtain a PID controller, certain constraints must be imposed on the tranfer functions $G_p(.)$ and $G^*_{CL}(.)$. The first one implies that $d = 0$ in the process model (12.a), which means that the sampling period adopted is greater than the process time delay.

In the product of Leeds & Northrup [4], $G^*_{CL}(.)$ is specified according to the method proposed by Dahlin, which consists in choosing a 1st order model for the closed loop system. As $d = 0$ is assumed, we have:

$$G^*_{CL}(z^{-1}) = \frac{1 - e^{-(h/T^*)}}{1 - e^{-(h/T^*)} z^{-1}}\, z^{-1} \tag{15}$$

where T* is the desired time constant for an exponential closed loop response when a step is applied in the set point. The PID controller, according to (11), is given by:

$$u(t) = p_0 \frac{A(z^{-1})}{(1 - z^{-1})\,(1 + \gamma z^{-1})}\, e(t) \tag{16}$$

with $\gamma = b_2/b_1$ and $p_o = (1 - e^{-h/T^*})/b_1$. The coefficients in (16) are then converted to the tuning parameters $K, T_I\ T_D$ and α.

In other developments, $b_2 = 0$ in (12.b) and $\gamma = 0$ in (11) are assumed; this last condition arises when $\alpha = 0$ in the continuous PID, meaning that the derivative action has no filter. In this case, one of the available strategies consists in computing p_o in (16) so as to satisfy a pre-fixed phase margin [5]. There are also certain methods that include a Smith predictor in the PID control law, in order to compensate for a process delay $d > 0$.

However, the above mentioned strategies have some drawbacks. One of these drawbacks is that they can be applied only to stable processes because of the cancellation of the A(.) polynominal; note that occasionally, as in the case of the Leeds & Northrup controller, also the B(.) polynomial must be stable. Another problem is the difficulty to translate the coefficients in equation (16) to the corresponding K, T_I, T_D and α PID parameters.

The closed loop transfer function $G_{CL}(.)$ can be derived following a somewhat different formulation from that based on equation (13). Starting with the PID controller defined by equation (4) and the process model given by (12.a,b,c) with the delay $d = 0$, one obtains:

$$\frac{y(t)}{r(t)} = G_{CL}(z^{-1}) = \tag{17}$$

$$\frac{(t_0 + t_1 z^{-1} + t_2 z^{-2})\, B(z^{-1})}{(1 - z^{-1})(1 + \gamma z^{-1})\, A(z^{-1}) + B(z^{-1})\,(s_0 + s_1 z^{-1} + s_2 z^{-2})}$$

In this way, the user will specify the poles of the desired closed loop transfer function $G^*_{CL}(.)$. Equating the 4th order characteristic equation in (17) to the P(.) polynomial that contains the desired closed loop poles, the following equation arises:

$$P(z^{-1}) = (1 - z^{-1})\,(1 + \gamma z^{-1})\, A(z^{-1}) + B(z^{-1})\,(s_0 + s_1 z^{-1} + s_2 z^{-2}) \tag{18.a}$$

Let the P(.) polynomial be defined as:

$$P(z^{-1}) = (1 + n_1 z^{-1})\,(1 + n_2 z^{-1})\,(1 + p_1 z^{-1} + p_2 z^{-2}) \tag{18.b}$$

with $p_1 = -2\, e^{(\omega \xi h)} \cos(\omega h \sqrt{1 - \xi^2})$ and $p_2 = e^{-2\omega\xi h}$. The coefficients p_1 and p_2 introduce the two dominant closed loop poles with relative damping ξ and natural frequency ω, where the values of ξ and ω can be computed from the desired overshoot and damping specifications. Different approaches can be considered with respect to the specification of the other poles, ($-n_1$ and $-n_2$). They can be:

i) placed at the origin, ($n_1 = n_2 = 0$);
ii) placed somewhat far from the dominant poles, ($n_1 = n_2 = -e^{-8\omega h}$);
iii) used to cancel one or more zeros introduced by the controller, (roots of T(.) in the numerator of equation (17));
iv) used to cancel the (stable) process zero (root of B(.) in the numerator of (17))
v) used tot obtain a trade-off between the control signal and the process output signal variances.

Provided that $(1 - z^{-1})$ A(.) and B(.) have no common factors, from (18.a,b) we obtain a system of equations with a unique solution that gives the PID coefficients s_i and γ. Due to the PID structure, the t_i coefficients cannot be chosen arbitrarily but are determined as a function of the s_i and γ coefficients [6,7].

It is also possible to derive a pole placement algorithm where the PID coefficients are directly estimated [7]. Another approach is to combine the PID controller with a Smith predictor in order to compensate for a process delay $d > 0$. A drawback of those strategies lies in the difficulty to translate the obtained s_i, t_i and γ coefficients to the K, T_I, T_D and α parameters.

Optimization

In this strategy, the PID tuning parameters are obtained as those that minimise a given criterium, expressed as a loss function J. The most used loss functions are defined by the equations:

$$J_1 = \int_0^T e^2(t)\, dt \qquad (19)$$

$$J_2 = \int_0^T e^{\lambda t}\, e^2(t)\, dt \qquad (20)$$

$$J_3 = \int_0^T [\lambda\, e^2(t) + \beta\, (u(t) - u(t-1))^2]\, dt \qquad (21)$$

where e(.) and u(.) are respectively the error and control signals.
J_1 corresponds to the widely adopted ISE criterium, while J_2 can be interpreted as a weighted ISE. In J_3, the weighting factors λ and β make possible to get a trade-off between the process output variance and the control effort.

The optimization problem is solved through the following steps:

i) evaluation of the loss function J for a given horizon T,
ii) adjustment of the tuning parameters by using numerical optimization methods, in such a direction that J is minimized,
iii) if still necessary, return to i).

In step i) above, the starting assumptions and the approximations adopted during the development of a specific strategy will dictate the way that the J criterium will be evaluated and, as a consequence, the rate of tuning parameters adjustment. So, it is possible to obtain auto-tuning or adaptive tuning strategies. For the adaptive strategies, the parameters can be updated either at each sampling time ($T = h$) or after several sampling times ($T \gg h$); in this last case, the parameters adjustment procedure runs in parallel with the controller computations.

The criterium J can be evaluated by directly using the measurements of y(.) and $u(t) - u(t-1)$ or from a process model which can be identified on-line or is previously determined. When the process model is used, it serves to generate the u(.) and y(.) values through a simulation for the evaluation of J; another possibility is to rewrite J in terms of the closed loop transfer function derived from the process model.

In step ii) above, different optimization methods can be used. Examples are the conjugated gradient, minimum search gradient, stochastic approximation, direct search and hill climbing algorithms.

A characteristic of the optimization based strategies is that no constraints are imposed on the order or the delay of the process model, but on the other hand, as the convergence of numerical methods is usually slow, the time required for tuning the PID controller may be rather long.

In the off-line optimization strategy implemented in a product of Sensycon [8], the considered loss function J is that of the equation (21) with $\lambda = 1$ and $\beta = \nu K_p^2$, where K_p is the process static gain. The process model given by (12.a) is identified recursively and then used for deriving the closed loop transfer function. A new expression for the loss function is then obtained from the closed loop transfer function and, in the optimization procedure that runs in parallel with the controller, this rewritten J is evaluated by either applying the Parseval's theorem or computing the u(.) and y(.) values for a step change in the set point. Next, the controller coefficients are adjusted through a hill climbing method. Once the new values have been determined, the PID coefficients in (11) are updated, where $\gamma = 0$ is used.

Other methods and further developments

A number of other strategies have been proposed in the literature for deriving selftuning PID controllers, within a wide variety of different approaches.

In the product of Turnbull Control Systems for example, described in [2], the process model given by equations (12.a,b.c) is identified recursively; a short sampling period h is used, so that an estimated coefficient $b_0 \neq 0$ appears in the B(.) polynomial. The estimated discrete model is then converted to the corresponding continuous case and the PID tuning parameters are obtained through a frequency based design method that imposes a phase margin of 60°.

Considering the "controller design" and the "parameter estimation" boxes in Figure 3, several different methods can be used for their implementation, see for instance [9-10].

It is also possible to adapt the tuning parameters through empirically established equations or by using the fuzzy sets theory, as in reference [11].

Different auto-tuning and adaptive tuning methods can be combined with an intelligent supervisor that is placed in a higher hierarchical level; this supervision level, conceived according to the expert systems techniques, decides and coordinates the application of the available methods, resulting in an expert tuner for PID controllers as proposed in [12]. Another research area in the self-tuning PID domain is the development of connectionist (or neuromorphic) tuners, where some of the usual linear relations in auto-tuning and adaptive tuning strategies are substituted by neural networks and the principles of neuro-engineering are exploited.

Conclusion

The development of self-tuning PID controllers is still a new and promising research area, with direct application to industrial processes. Autotuners for PID controllers provide systematic and automatic procedures for obtaining the tuning parameters; they can be used on existing plants or to reduce the commissioning time for new installations. It is also possible to derive adaptive tuners so that the PID parameters are adjusted in order to maintain the desired closed loop performance, even when the process characteristics are changing.

In this paper, we have reviewed the most commonly used strategies for auto-tuning and adaptive tuning of PID controllers. We have also briefly described some of the commercially available PID controllers that incorporate self-tuning capabilities.

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