Conclusions

It has been shown that comparing different anti-windup and bumpless transfer methods is no straightforward task. The classical comparison criteria such as process overshoot, settling time, IAE, ISE, etc., are shown to be inadequate. It was proposed that the realisable reference concept could prove to be a suitable criterion for comparison.

It was found that for the PID controller, a smaller integral time constant T_i gives a shorter process settling time after the system leaves saturation and vice versa. The only exception is when using the conditioning technique which generally assures the fastest settling time of the realisable reference.

The area enclosed by the unlimited and limited process response equals zero when no protection against windup is used. The same holds for the area enclosed by the actual and the realisable references. Using such interpretation of the windup phenomenon, it was shown why a suitable anti-windup compensator can reduce the windup effect.

The prediction of the maximum overshoot of the realisable reference was also derived, where the expected overshoot of the realisable reference is higher for smaller integral time constant T_i, and vice versa.

In some cases, even disturbances can cause system saturation. In that event, if the controller is well tuned for reference tracking as well as for disturbance attenuation, the limited process response to disturbances should be quite acceptable when using the conditioning technique. However, when using small values of factor β (in PID controllers), it can happen that limited disturbance rejection is sluggish. In such cases, a disturbance rejection compensator could be used, but not in all cases, as seen in the thesis.

Process noise is relatively common in the process industry (e.g. flow control). It is demonstrated that some anti-windup algorithms are much more sensitive to process noise than others. The conditioning technique appeared relatively inert to process noise.

The anti-windup strategy for multivariable systems should take into account the so-called “deadlock” situation, which does not occur in univariable systems. It is shown that, by using the artificial nonlinearity block in the control scheme, significant improvements of process response can be achieved. The scheme showed superior performance compared to other existing anti-windup algorithms.

PID tuning rules have been developed for more than five decades. Our goal was to find those tuning rules which would only require the open-loop process step-response, and, on the other hand, a demanding tuning criterion (magnitude optimum) was to be achieved. By applying the multiple integrations (moment) method, known from the identification theory, it was found that tuning of the PID controller parameters could be based merely on the process step response. The developed real-time self-tuning algorithm showed excellent performance on several laboratory plants tested.

The author's *contributions* to the fields of anti-windup, bumpless transfer, and PID controller tuning are the following:

- It was shown that classical comparison criteria, like process overshoot, settling time, IAE, ISE, etc., are not appropriate when comparing different anti-windup algorithms. The realisable reference appears as a better criterion.
- It was shown that the process settling time, when limitation occurs and when the conditioning technique is not used, depends highly on the integral time constant T_i. The higher T_i, the longer is the process settling time.
- It was derived that the area enclosed by the unlimited and limited process response equals zero when no anti-windup protection is used. The same holds for the area enclosed by the actual and the realisable reference. Using that interpretation of the windup phenomenon, it was shown that a suitable anti-windup compensator should appropriately increase the area.
- The prediction of the realisable reference overshoot is given for the PID controller. The overshoot is higher when using a shorter integral time constant.
- Achieving good disturbance rejection during the saturation of the process is not such a straightforward task. It was shown that the conditioning technique usually gives quite acceptable process responses, if the controller is well tuned for tracking and disturbance rejection for unlimited conditions. In some cases, the disturbance compensator can be used at the cost of an inferior tracking response.
- It was outlined that some of the anti-windup algorithms are much more sensitive to process noise than others. The conditioning technique appeared relatively inert to process noise.
- The artificial nonlinearity for the multivariable controllers was implemented and tested under the MATLAB-SIMULINK programme package, where numerous experiments on different plants were made.
- Theoretical development of the new tuning approach for the PID controllers was given. It is based on the "magnitude optimum" principle, and with the help of the multiple integration (moment) method, the algorithm only requires the process step response to calculate controller parameters precisely. The calculated PID controller parameters also hold for processes with a time delay.

- A real-time self-tuning algorithm for the PI and PID controller has been developed. Numerous tests on different laboratory plants showed very good tuning results. It was shown that the algorithm is also robust to process noise and nonlinearities.

Appendices

APPENDIX A: Convergence of integral (68)

Consider a stable unlimited and limited closed-loop system. Following (78), the steady-state values of the limited and unlimited process outputs are the same:

$$y_0(\infty) = y_1(\infty) = y_{SS} \ . \tag{A1}$$

Integral (68) can be rewritten as

$$A_y = \int_0^\infty \left[\varepsilon_1(t) - \varepsilon_0(t)\right] dt \ , \tag{A2}$$

where

$$\varepsilon_0(t) = y_{SS} - y_0(t) \ , \tag{A3a}$$

$$\varepsilon_1(t) = y_{SS} - y_1(t) \ . \tag{A3b}$$

The unlimited process output is given by (72). Using the partial fractional expansion and considering the initial conditions of signals $y_0(t)$ and $w(t)$, the unlimited process response on a step change of the reference can be expressed as (c.f. DiStefano et al., (1990)):

$$y_0(t) = y_{SS} + \int_0^t \left[\sum_{i=1}^{r} \sum_{k=1}^{n_i} \frac{\lambda_{ik}}{(k-1)!} \tau^{k-1} e^{-p_i \tau} \right] d\tau + \sum_{i=1}^{r} \sum_{k=1}^{n_i} \frac{\mu_{ik}}{(k-1)!} t^{k-1} e^{-p_i t} \;;\; t \geq 0 \;, \quad \text{(A4)}$$

where $\mathbf{Re}\{p_i\}>0$ for $i=1..r$.

While the system is limited, the process response depends only on the limited variable u^r. Since u^r is limited, and the closed-loop response is assumed to be stable, the process output $y_1(t)$ for $0 \leq t < t_1$ is limited. After $t \geq t_1$, the system returns to a linear regime. For $t \geq t_1$ the closed-loop system response can be viewed as a response to the reference input with initial conditions defined at $t=t_1$:

$$\begin{aligned} y_1(t) = y_{SS} + \int_{t_1}^t \left[\sum_{i=1}^{r} \sum_{k=1}^{n_i} \frac{\lambda_{ik}}{(k-1)!} (\tau - t_1)^{k-1} e^{-p_i(\tau - t_1)} \right] d\tau + \\ + \sum_{i=1}^{r} \sum_{k=1}^{n_i} \frac{\mu'_{ik}}{(k-1)!} (t - t_1)^{k-1} e^{-p_i(t - t_1)} \;;\; t \geq t_1 \end{aligned} \quad . \quad \text{(A5)}$$

The sufficient condition for the convergence of (A2) is that

$$I = \int_0^\infty |\varepsilon_0(t)| dt + \int_0^\infty |\varepsilon_1(t)| dt \quad \text{(A6)}$$

converges.

Sufficient conditions for the convergence of (A6) are

$$\lim_{t \to \infty} |\varepsilon_0(t)| t^\alpha < \infty \;;\; \alpha > 1 \;, \quad \text{(A7a)}$$

$$\lim_{t \to \infty} |\varepsilon_1(t)| t^\alpha < \infty \;;\; \alpha > 1 \;. \quad \text{(A7b)}$$

After multiplying (A4) and (A5) by t^{α}, it follows that (A7a) and (A7b) hold, hence (A2) and (68) do converge.

APPENDIX B: Proof of theorem 1

For the sake of notation distinction, let us rewrite (108) and (110) for AWa and AWb, respectively, namely

$$u_a = K_{1a}(s)(w - y) - K_{2a}(s)v\ , \tag{B1}$$

$$w_a{}^r = w + K_{1a}{}^{-1}(s)(v - u_a)\ . \tag{B2}$$

$$u_b = K_{1b}(s)(w - y) - K_{2b}(s)v\ , \tag{B3}$$

$$w_b{}^r = w + K_{1b}{}^{-1}(s)(v - u_b)\ . \tag{B4}$$

Note that due to the same initial conditions, v (y) when using AWa is the same as v (y) when using AWb at the beginning. If w_a^r and w_b^r are the same, then v (y) when using AWa will be the same as v (y) when using AWb.

From C_1, we can easily find that

$$K_{2b}(s) = \Gamma K_{2a}(s) + \Gamma - I\ . \tag{B5}$$

Combining (B1), (B3) and (B5) yields

$$u_b - v = \Gamma(u_a - v), \tag{B6}$$

This leads to $w_a^r = w_b^r$ which concludes the proof.

APPENDIX C: Proof of theorem 2

From condition C_1, and given that $K_1(s)$ and $K(s)$ are proper, we deduce that $I+K_2(s)$ should be biproper. Hence, $I+K_{2a}(\infty)$ is a non-singular matrix. From (B5), we then have

$$K_{2b}(\infty) = \Gamma\left(I + K_{2a}(\infty)\right) - I\,. \tag{C1}$$

Thus the choice $\Gamma = \left(I + K_{2a}(\infty)\right)^{-1}$ leads to $K_{2b}(\infty) = 0$, i.e. $K_{2b}(s)$ is strictly proper.

APPENDIX D: Explicit solution to the optimal artificial nonlinearity design

In this appendix, we present an explicit solution to the optimal design of AN, the artificial nonlinearity block. For practical applications, it is reasonable to choose the quadratic weighting criterion for minimisation:

$$J\left(w^r - w\right) = \left(w^r - w\right)^T \Lambda\left(w^r - w\right), \tag{D1}$$

where Λ is a diagonal weighting matrix whose elements are all positive. The choice of Λ is dictated by the relative importance of each element of $w^r\!-\!w$ in the criterion. A Kuhn-Tucker multiplier $\mu=[\mu_1,\mu_2,...,,\mu_{2m}]^T \in \mathbf{R}^{2m}$ is introduced to form the following auxiliary Kuhn-Tucker function:

$$J^*\left(w^r - w\right) = \left(w^r - w\right)^T \Lambda\left(w^r - w\right) + \mu^T\left(HD\left(w^r - w\right) + Hu + b\right) \tag{D2}$$

where $H=[H_1^T,H_2^T,...,H_{2m}^T]^T \in \mathbf{R}^{2m\times m}$ and $b=[b_1^T,b_2^T,...,b_{2m}^T]^T \in \mathbf{R}^{2m}$. In equation (D2), $\mu_i=0$ if the ith constraint is active (that is, if $H_iD(w^r\text{-}w)+H_i+b_i<0$), and $\mu_i\geq 0$ if the constraint is active ($H_iD(w^r\text{-}w)+H_i+b_i=0$). From the Kuhn-Tucker theorem, a first-order necessary condition for $w^r\text{-}w$ to be a local minimiser of J (a condition to be satisfied by any optimal AN design) is that

$$\frac{\partial J^*}{\partial w^r} = 2\Lambda\left(w^r - w\right) + D^T H^T \mu = 0 \tag{D3}$$

If the controller output u violates no constraints (i.e. $\mu=0$), the optimal solution is $w^r=w$, in which case $v=u$. If, however, one or more elements of u violate the constraints (i.e. $H_i u + b_i > 0$ for some i), the optimal solution is found by setting $H_i v + b_i = 0$ for the same i. To this end, we partition μ, H and b as follows: $\mu = \left[\mu_0{}^T \quad \mu_+{}^T\right]^T$, $H = \left[H_+{}^T \quad H_0{}^T\right]^T$ and $b = \left[b_+{}^T \quad b_0{}^T\right]^T$, such that $\mu_0=0$ contains the Kuhn-Tucker multipliers associated with the inactive constraints

$$H_+D\left(w^r - w\right) + H_+u + b_+ < 0 \ , \tag{D4}$$

and μ_+>0 are the Kuhn-Tucker multipliers corresponding to the active constraints

$$H_0D\left(w^r - w\right) + H_0u + b_0 = 0 \tag{D5}$$

From (D3) and (D5) we obtain

$$\mu_+ = 2\left(H_0 D \Lambda D^T H_0^{\ T}\right)^{-1}\left(H_0 u + b_0\right) \tag{D6}$$

Note that since H_0 picks out precisely those elements of u which violate the constraints (i.e. $H_0u+b_0>0$), equation (D6) implies $\mu_+>=0$. From (D3) and (D6), we obtain

$$w^r - w = -\Lambda^{-1} D^T H_0^{\ T}\left(H_0 D \Lambda D^T H_0^{\ T}\right)^{-1}\left(H_0 u + b_0\right) \ , \tag{D7}$$

from which equation (119) yields

$$v = -D\Lambda^{-1} D^T H_0^{\ T}\left(H_0 D \Lambda D^T H_0^{\ T}\right)^{-1}\left(H_0 u + b_0\right) + u \ . \tag{D8}$$

It can be readily verified that v computed as in (D8) satisfies

$$H_0 v + b_0 = 0 \ . \tag{D9}$$

If this same v also satisfies

$$H_+ v + b_+ < 0 \ , \tag{D10}$$

then it is the unique optimal solution of the AN design problem. If (D8) does not satisfy (D10), then the corresponding line of v must be modified in order to find the optimal solution. Obtaining the appropriate modification is generally too complex a task for real-time applications. In practice, it is sufficient to apply v computed as in (D8), in which case the nonlinearity N modifies some elements of v so that $H_+ u^r + b_+ \leq 0$. This method can therefore be considered a suboptimal solution to the AN design problem.

APPENDIX E: The tested processes and their parameter values

Process	***Values of parameter***
$G_{P1}(s)=\dfrac{e^{-s}}{(1+sT)}$	*T*=0.1, 0.2, 0.5, 1, 2, 5, 10
$G_{P2}(s)=\dfrac{e^{-s}}{(1+sT)^2}$	*T*=0.1, 0.2, 0.5, 1, 2, 5, 10
$G_{P3}(s)=\dfrac{1}{(1+sT)(1+s)}$	*T*=0.1, 0.2, 0.5, 1, 2, 5, 10
$G_{P4}(s)=\dfrac{1}{(1+sT)^2(1+s)^2}$	*T*=0.1, 0.2, 0.5, 1, 2, 5, 10
$G_{P5}(s)=\dfrac{1}{(1+s)^n}$	*n*=2, 3, 4, 5, 6, 7, 8
$G_{P6}(s)=\dfrac{1}{(1+s)(1+sT)(1+sT^2)(1+sT^3)}$	*T*=0.2, 0.3, 0.4, 0.5, 0.6, 0.7, 0.8
$G_{P7}(s)=\dfrac{(1-sT)}{(1+s)^3}$	*T*=0.1, 0.2, 0.5, 1, 2, 5, 10
$G_{P8}(s)=\dfrac{e^{-s}(1+sT)}{(1+s)^2}$	*T*=0.1, 0.2, 0.3, 0.4, 0.5, 0.7, 1.0
$G_{P9}(s)=\dfrac{1}{(1+s)(1+s(1+j\alpha))(1+s(1-j\alpha))}$	α=0.1, 0.2, 0.3, 0.4, 0.5, 0.7, 1.0

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