IFAC Conference on Advances in PID Control
PID'12
Brescia (Italy), March 28-30, 2012 ThA2.5

Model-Free Adaptive PID Controllers Applied to the Benchmark PID'12

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Abstract: Many tuning strategies for PID control algorithms have been proposed in the literature depending on modeling choice from heuristics, physical laws and/or experimental essays. In this paper, the properties of three model-free adaptive PID control design conceptions, known as direct PID control, are investigated and implemented. The recursive least squares estimator is used to identify the PID controller gains without the knowledge of the plant mathematical model. The tuning strategies of the direct PID controller are applied to the boiler control system proposed by Fernando Morilla (UNED, Spain) as a benchmark problem for the 2012 IFAC Conference on Advances in PID Controllers.

Keywords: PID control, identification, self-tuning, stability, minimum variance, Benchmark PID'12.

1. INTRODUCTION

The concept of adaptation in industrial process control is popular since 1970, with focus to increase control-loop performance in order to reduce the variability of manufactured products. During the last decades a great number of automatic tuning algorithms are being published in the process control literature. The majority of these algorithms, in their original forms, are based on the principle of process identification and control: the process model parameters are estimated and then used in the controller design. When the controller design is based on the process model online estimation, the procedure is usually called indirect adaptive control and its adaptive mechanism can use recursive identification or computational intelligence (Åström and Hägglund, 2006).

In model based control design, the development of the control system requires great simplification of the experimental data in order to derive the process reduced order estimated model (parsimony principle). Differently, direct adaptive control synthesis (also designated as model-free), more experimental data is retained during the control design stage. All simplifications and considerations are specifically to what is important to build the control law and not the plant model (DeKeyser, 2000; Yamamoto and Shah, 2007).

Some characteristics that justify the direct adaptive PID control are: i) since most processes have non-linear behavior, there exists the need of more elaborated control algorithms where the plant mathematical representation (plant identification) should not be applied; ii) the possibility of controller design by directly applying sensor-measured data from the process is attractive, since no restriction is imposed due to plant model simplifications; iii) adaptive control in the self-tuning conception is becoming a conventional method in applications and commercial products; iv) the need of new control algorithms to cope with high-end PLC, DSP, FPGA or within micro-controlled applications, all well established in the PID control industry and capable of handling highly-demanding computer processing, such as the usage of vector and matrix operations to support advanced PID control strategies; v) technological and theoretical enhancements in industrial adaptive PID control are becoming a common practice due to its wide acceptance in many different fields of application (DeKeyser, 2000; Kirecci et al., 2003; Shafiq and Akhtar, 2004; Yamamoto and Shah, 2007).

In this paper, three direct adaptive PID control structures are presented, tuned based on input and output data of the process alone. The first one is based on the synthesis of Model Reference design; the second, on the Generalized Minimum Variance Control, and the third, based on a FIR filter controller. All three methods focused not only on depicting the real design advantages of every tuning technology and the treatment of complex plants, but also to emphasize the real impact on plant operational efficiency. In this sense, this study gets away from the current industrial conservatism mode, or of oscillatory dynamic behavior due to worst case tuning of controllers in order to retain stability within extremely safe margins to face operational condition changes of non-linear plants (Gude et al., 2006).

The three tuning strategies to be studied in this paper are assessed within a simulation environment of a boiler control system proposed by Fernando Morilla (UNED, Spain), known as Benchmark PID'12, adopted as a benchmark problem for the 2012 IFAC Conference on Advances in PID Controllers. Results for the SISO control case are compared to the Benchmark PID'12 controller as the reference for performance evaluation. This means that any reader can reproduce the results to be presented, directly from the Benchmark PID'12 package found at the conference web site: http://pid12.ing.unibs.it/.

2. PID CONTROLLER

The tuning of the PID controller concerns the adequate selection of values for its gains, K_c, K_i, K_d. Many different tuning techniques and surveys have been reported in specialized books, conferences and journals, during the last 60 years, and the results indicates that: i) 30% of installed controllers are being operated in manual mode; ii) 30% of the control-loops are inconsistent (poorly elaborated designs); iii) 25% of the control-loops are tuned at controller's default setup; iv) 30% of the control-loops have problems with equipment issues. One possible reason for the lack of understanding about the process closed-loop dynamics, the PID control algorithm or theoretical/experimental knowledge, is associated to the tuning design procedure being relatively simple for some certain classes of plants and also due to measurement and actuation instrumentation being considerably robust enough to face many problems (Hägglund, 2005; Gude et al., 2006). However, it is statistically shown that many control-loops are poorly tuned and could be optimized with less interference of operators and the production rate increased.

2.1 PID controller structure

The PID controller presents itself under several implementation structures depending of its manufacturer. The parallel structured PID is the most popular and has become very common within modern control systems (Åström and Hägglund, 2006). The parallel digital PID control law is based on the equation given by

$$u(t) = K_c e(t) + K_i \sum_{i=1}^{t} e(i) + K_d \left[e(t) - e(t-1) \right] \qquad (1)$$

where $K_i = K_c T_s / T_i$, $K_d = K_c T_d / T_s$, and T_i, T_d, T_s, are respectively the integral, derivative and sampling time, and $e(t)$ denotes the error signal between a reference sequence $\mathrm{y_r(t)}$ and the process output $\mathrm{y(t)}$. The incremental form of (1) is given by

$$u(t) = u(t-1) + s_0 e(t) + s_1 e(t-1) + s_2 e(t-2) \qquad (2)$$

$$s_0 = K_c + K_i + K_d;\ s_1 = -(K_c + 2K_d);\ s_2 = K_d \qquad (3)$$

By applying a recursive estimator it is possible to identify, online, the parameters s_0, s_1, s_2, from (2). By doing this, the model-free adaptive PID control law is derived using

$$K_c = -(s_1 + 2s_2);\ K_i = s_0 + s_1 + s_2;\ K_d = s_2 \qquad (4)$$

$$\Delta u(t) = (K_c + K_i + K_d) e(t) - (K_c + 2K_d) e(t-1) + K_d e(t-2) \qquad (5)$$

Equations (4) and (5) are adequate for real-time implementation and of easy comprehension from the viewpoint of most operators and engineers (Bobál et al., 2005).

In the PID controller design, the proportional and derivative bands are presented as being multiplied by the error signal. This has an implication in the controller's performance, since abrupt and severe changes in the reference signal, and consequently in the error signal, will produce excessive and sometimes unpractical control actions. This might compromise actuator life-time and the process as a whole. To avoid this in practical applications, the following actions should be considered: i) keep the integral term related to $e(t) = y_r(t) - y(t)$; ii) substitute proportional and derivative terms by $e(t) = -y(t)$. The parallel digital PID is then rewritten in the I+PD form, which is

$$\Delta u(t) = t_0 y_r(t) - s_0 y(t) - s_1 y(t-1) - s_2 y(t-2) \qquad (6)$$

$$t_0 = K_i = s_0 + s_1 + s_2 \qquad (7)$$

with s_0, s_1, s_2, given by (3) and K_c, K_i, K_d, by (4). Observe that not only the implementation equations of the PID and I+PD structures are different, but also the number of estimated parameters. The PID form has 3, and I+PD has 4 parameters.

3. PID DESIGN AND TUNING BY MODEL REFERENCE

The first model-free PID control design to be shown is based on the model reference synthesis (MR) with real-time self-tuning of its control law parameters (DeKeyser, 2000). In this technique, a desired transfer function, which represents the closed-loop dynamic (reference tracking) is defined by the designer. Then the PID controller is tuned in order to make the real closed-loop transfer function to follow the desired one. The controller parameters can be estimated online by the Recursive Least Squares (RLS) or by some other parameter estimation method (Hägglund and Åström, 2000; Bobál et al., 2005). The mathematical model of the process is not necessary to be known, which in practical applications is hard to build and maintain.

In the digital PID control synthesis, the ideal closed-loop dynamic is specified by the designer via a discrete transfer function of the form

$$\frac{y(t)}{y_r(t)} = \frac{y_m(t)}{y_r(t)} = \frac{z^{-d} B_m(z^{-1})}{A_m(z^{-1})} \qquad (8)$$

The selection of polynomials $B_m(z^{-1})$ and $A_m(z^{-1})$ have a fundamental importance on the closed-loop dynamic behaviour of the plant. Selection criteria are available and given in DeKeyser (2000).

For the implementation of the controller, hereinafter designated as PID-MR, from (2) and (8), it is possible to establish the following relation:

$$\left(s_0 + s_1 z^{-1} + s_2 z^{-2}\right)[A_m(z^{-1}) - z^{-d} B_m(z^{-1})] y(t) = (1 - z^{-1}) B_m(z^{-1}) u(t-d) \qquad (9)$$

By defining the auxiliary variables given by

$$y_{abm}(t) = [A_m(z^{-1}) - z^{-d}B_m(z^{-1})]y(t) \tag{10}$$

$$u_{bm}(t) = (1 - z^{-1})B_m(z^{-1})u(t-d) \tag{11}$$

(9) can be expressed as being

$$u_{bm}(t) = \left(s_0 + s_1 z^{-1} + s_2 z^{-2}\right) y_{abm}(t) \tag{12}$$

In order to apply the classical equations of the RLS estimator, (12) can be rewritten in a vector form, which is

$$u_{bm}(t) = \varphi^T(t)\theta(t) \tag{13}$$

where measurements and estimated parameters vectors are now defined as

$$\varphi^T(t) = [y_{abm}(t) \;\; y_{abm}(t-1) \;\; y_{abm}(t-2)] \tag{14}$$

$$\theta^T(t) = [s_0 \;\; s_1 \;\; s_2] \tag{15}$$

In the implementation of the I+PD control structure, (6), (12), measurements and estimated parameters vectors modify to

$$u_{bm}(t) = t_0 y_{am}(t) - \left(s_0 + s_1 z^{-1} + s_2 z^{-2}\right) y_{bm}(t) \tag{16}$$

$$\varphi^T(t) = \left[y_{am}(t) - y_{bm}(t) - y_{bm}(t-1) - y_{bm}(t-2) \right] \tag{17}$$

$$\theta^T(t) = [t_0 \;\; s_0 \;\; s_1 \;\; s_2] \tag{18}$$

$$y_{am}(t) = A_m(z^{-1})y(t); \;\; y_{bm}(t) = B_m(z^{-1})y(t) \tag{19}$$

4. PID DESIGN AND TUNING BY GENERALIZED MINIMUM VARIANCE

The Generalized Minimum Variance controller (GMV) was developed as a modification of the Minimum Variance technique of K. J. Åström and B. Wittenmark for the treatment of complex industrial plants. The cost function of GMV allows additional penalties, or weights, in terms of the process output error and increment of the control signal (Veronesi and Visioli, 2011).

In the design of the incremental GMV controller, plant equations, polynomial identity and cost function, respectively assume the form

$$A(z^{-1})\Delta y(t) = z^{-d}B(z^{-1})\Delta u(t) \tag{20}$$

$$P(z^{-1}) = A(z^{-1})\Delta E(z^{-1}) + z^{-d}S(z^{-1}) \tag{21}$$

$$J = \left[\phi(t+d)\right]^2 = [P(z^{-1})y(t+d) - T(z^{-1})y_r(t) + Q(z^{-1})\Delta u(t)]^2 \tag{22}$$

where d is the discrete delay, Δ = $(1\text{-}z^{-1})$, *ns* = *na*, *ne* = d – 1, with *ns*, *na*, *ne*, corresponding to the order of polynomials $S(z^{-1})$, $A(z^{-1})$, $E(z^{-1})$, respectively. The incremental control law is given by

$$R(z^{-1})\Delta u(t) = T(z^{-1})y_r(t) - S(z^{-1})y(t) \tag{23}$$

$$R(z^{-1}) = B(z^{-1})E(z^{-1}) + Q(z^{-1}) \tag{24}$$

This design, hereinafter designated as PID-GMV, and within the implementation in the context of direct self-tuning control, is based on the definition of a generalized system of output $\phi(t)$, represented as a function of the controller polynomials $R(z^{-1})$, $S(z^{-1})$, $T(z^{-1})$, that is,

$$\begin{aligned}\phi(t) = S(z^{-1})y(t-d) + R(z^{-1})\Delta u(t-d)\\ -T(z^{-1})y_r(t-d)\end{aligned} \tag{25}$$

$$R(z^{-1}) = 1; \;\; S(z^{-1}) = T(z^{-1}) = s_0 + s_1 z^{-1} + s_2 z^{-2} \tag{26}$$

where s_0, s_1, s_2, can be obtained from (2).

In order to adequate (25) to the implementation context of adaptive PID control, with the RLS estimator, it is necessary to regroup some terms, leading to

$$\begin{aligned}\bar{\phi}(t) = \phi(t) - \Delta u(t-d) = \\ S(z^{-1})[y(t-d) - y_r(t-d)] = \varphi^T(t-d)\theta(t)\end{aligned} \tag{27}$$

where the signal $\phi(t)$, measurements and estimated parameters vectors are calculated by the simplified GMV generalized output, with a scalar energy weighting factor q_0,

$$\phi(t) = y(t) - y_r(t-d) + q_0 \Delta u(t-d), \tag{28}$$

and

$$\varphi(t-d) = \begin{bmatrix} y(t-d) - y_r(t-d) \\ y(t-d-1) - y_r(t-d-1) \\ y(t-d-2) - y_r(t-d-2) \end{bmatrix} \tag{29}$$

$$\theta^T(t) = [s_0 \;\; s_1 \;\; s_2] \tag{30}$$

During the practical usage of PID-GMV, the designer must observe some remarks to avoid that the measurements vector elements tend to zero (blow-up phenomena), that may lead to an unstable estimator and, consequently, an unstable closed-loop behavior (Hägglund and Åström, 2000).

The I+PD control structure, in (28), suits better the construction of the measurements and estimated parameters vectors, which are respectively modified to

$$\begin{aligned}\bar{\phi}(t) = \phi(t) - \Delta u(t-d) = S(z^{-1})y(t-d)\\ -T(z^{-1})y_r(t-d) = \varphi^T(t-d)\theta(t)\end{aligned} \tag{31}$$

$$\begin{aligned}\varphi^T(t-d) = \\ [y(t-d) \;\; y(t-d-1) \;\; y(t-d-2) \;\; -y_r(t-d)]\end{aligned} \tag{32}$$

$$\theta^T(t) = [s_0 \;\; s_1 \;\; s_2 \;\; t_0] \tag{33}$$

Similar to this I+PD algorithm based on GMV, Cameron and Seborg (1983) presented an alternative version but in the context of an indirect adaptive controller.

5. PID DESIGN AND TUNING BY FIR FILTER

For an unknown plant, it is assumed that the only available (measurable) information is its input and output. In this way, a discrete dynamic process can be described as

$$y(t) = f\left(y(t-1), y(t-2), \ldots \quad \ldots \tag{34}$$

where $f(.)$ is a general function (there is no restriction on the controlled plant, then it could be linear/non-linear, of minimum/non-minimum phase, stable/unstable or time invariant/variant). The objective is to synthesize a limited control input using a 2-DOF adaptive PID controller based on a self-tuned FIR filter in such a way that the output of the plant would track the reference sequence (Shafiq and Akhtar, 2004), that is:

$$\lim_{t\to\infty} y(t) = y_r(t-L) \tag{35}$$

The digital control law is synthesized from the following equation:

$$\Delta u(t) = F(z^{-1})\left[T(z^{-1})y_r(t) - S(z^{-1})y(t)\right] \tag{36}$$

The $F(z^{-1})$ polynomial in (36), of order *nf*, satisfies a linear relation within the nominal point of operation of the plant, where

$$G_p(z)F(z^{-1}) = z^{-d}\frac{B(z^{-1})}{A(z^{-1})}F(z^{-1}) \cong z^{-L} \tag{37}$$

in which $L \in \mathbb{Z}$, $L \geq d$. It can be said that $F(z^{-1})$ is an inverse linear approximation of the plant, delayed by L samples, and considered as a FIR filter that can be estimated recursively. The plant is locally approximated by a Taylor series expansion applied to (37), leading to

$$\begin{aligned} y(t) = {} & \frac{\partial f}{\partial y(t-1)}y(t-1) + \ldots + \\ & \frac{\partial f}{\partial y(t-na)}y(t-na) + \frac{\partial f}{\partial u(t-d-1)}u(t-d-1) + \\ & \ldots + \frac{\partial f}{\partial u(t-nb)}u(t-nb) + \eta(t) \end{aligned} \tag{38}$$

where *nb* is the order of $B(z^{-1})$ and $\eta(t)$ is representing superior order terms or residue. If $\eta(t)$ is small, then the input and output relation approximates to a linear discrete parametric model relation of the form:

$$G_p(z) = \frac{Y(z)}{U(z)} \cong z^{-d}\frac{B(z^{-1})}{A(z^{-1})} = \frac{z^{-d}\sum_{j=1}^{nb} b_j z^{-j}}{1 + \sum_{j=1}^{na} a_j z^{-j}} \tag{39}$$

However, the parameters a_j, b_j, in (39), are considered to be unknown and the controller will depend solely on the input and output data to build the control law in (36). The polynomial $S(z^{-1})$ needs to satisfy the relation $S(1) = T(1)$ in order to ensure reference tracking. Then, the roots of its characteristic equation, $\Delta + z^{-1}S(z^{-1})$, need to have magnitude of less than 1 (one) in order to guarantee closed-loop stability.

The parallel PID controller by FIR filter is obtained from (36) while assuming the following relations:

$$\Delta u(t) = (f_0 + f_1 z^{-1} + f_2 z^{-2})e_f(t) \tag{40}$$

$$e_f(t) = T(z^{-1})y_r(t) - S(z^{-1})y(t) \tag{41}$$

$$K_c = f_0 - f_2;\;\; K_i = f_0 + f_1 + f_2;\;\; K_d = f_2$$

$$\begin{aligned} u(t) = {} & u(t-1) + (K_c + K_i + K_d)e_f(t) \\ & -(K_c + 2K_d)e_f(t-1) + K_d e_f(t-2) \end{aligned} \tag{42}$$

The RLS estimator applied at a nominal point of operation of the plant, under the assumptions of (34) and (37), results on

$$u(t-L) = F(z^{-1})y(t) = \varphi^T(t)\theta(t) \tag{43}$$

and its measurements and estimated parameters vectors are expressed as

$$\varphi^T(t) = [y(t) \;\; y(t-1) \;\; y(t-2)] \tag{44}$$

$$\theta^T(t) = [f_0 \;\; f_1 \;\; f_2] \tag{45}$$

The PID-FIR structure does not fit in the I+PD design framework, as depicted in (42). As a consequence, it is more likely to produce an aggressive and noisy control signal. One way to avoid such a problem, is to use a low-pass first order filter $u_f(t) = \alpha u(t) + (1-\alpha)u_f(t-1)$, $0 \leq \alpha \leq 1$.

6. SIMULATION RESULTS

The simulation environment adopted matches the one distributed with the Benchmark PID'12 of the 2012 IFAC Conference on Advances in PID Controllers, which includes MATLAB and Simulink files with appropriate startup values for the boiler model and the benchmark PID controller in the SISO case. The major concern within this simulation section, is to evaluate the three control structures presented, with respect to output performance and control effort as compared to the default tuned PID controller. The reader should also keep in mind that there is no need for a deep investigation over the process characteristics within this paper, since it is not the problem to be covered. Then, the process is being considered as a black box and all results are going to be presented simply as input and output signals, being compared to the benchmark case. For a detailed description on the boiler process covered in the Benchmark PID'12, it is well documented at http://pid12.ing.unibs.it/.

The three model-free adaptive digital controllers implemented use a sampling time of 1s and run within the Benchmark PID'12 Simulink diagram as a m-function block. By doing this, the m-function of the controller reads, at every second, the process block output, stores current and past data, to use it recursively for controller parameters estimation and to compute the control law.

6.1 PID-MR simulation

In the Model Reference based controller, the designer needs to specify the closed-loop transfer function as shown in (8). The appropriate selection of desired polynomials is commonly related to the open-loop dynamic behavior of the process. However, as it is going to be shown in this example, a simplistic approach such as a first order description, can give suitable results. Then, considering a continuous first order transfer function with a closed-loop time constant of 100s, its discrete equivalent is

$$\frac{B_m(z^{-1})}{A_m(z^{-1})} = \frac{0.01z^{-1}}{1-0.99z^{-1}} \qquad (46)$$

The selection of the time constant was based on a few trial-and-error tests and was found to cope well with the RLS and the process block initialization. During the first 100s of simulation, the benchmark process starts at its default input and output values, and just after, a load disturbance is sensed at the output. During this initial stage the PID-MR remains dormant, but its RLS is kept estimating.

It was observed that a slow convergence of the RLS estimator was required. Otherwise, when the load disturbance becomes active, the estimator could be already stabilized and no longer updating the controller as it should. Of course a more sophisticated estimation technique could be used instead to face this kind of problem, but a simpler RLS without any type of supervision module is better to evaluate the results and difficulties at this stage of this study (Hägglund and Åström, 2000).

The slow convergence for the RLS estimator was set by selecting a initial covariance matrix, P_{cov}, of small magnitude. $P_{cov} = 10I$ of appropriate dimension was selected. Vectors φ and θ elements values were set close to zero.

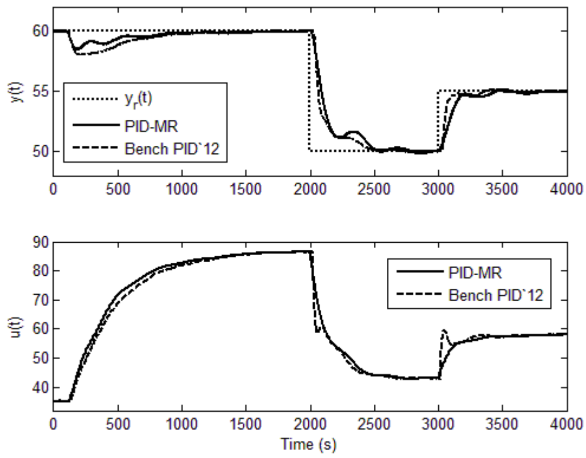

Fig. 1. Process output and PID-MR control signal.

In Figs. 1 and 2, the process output, control signal and PID-MR gains adaptation over time are respectively shown. The PID-MR design handled well the benchmark problem.

6.2 PID-GMV simulation

In the PID-GMV simulation, the same RLS estimator as of the PID-MR case was used. The design is based on the simplest GMV structure of a single tuning parameter, $q_0 = 15$, weighting the energy employed to move the output to the desired reference. The single energy weighting factor was tuned by trial and error, being not too rough on conservatism to avoid the load disturbance, but also restrictive enough to avoid further oscillations observed in Fig. 3 during setpoint changes. The PID-GMV gains adaptation over time are shown in Fig. 4.

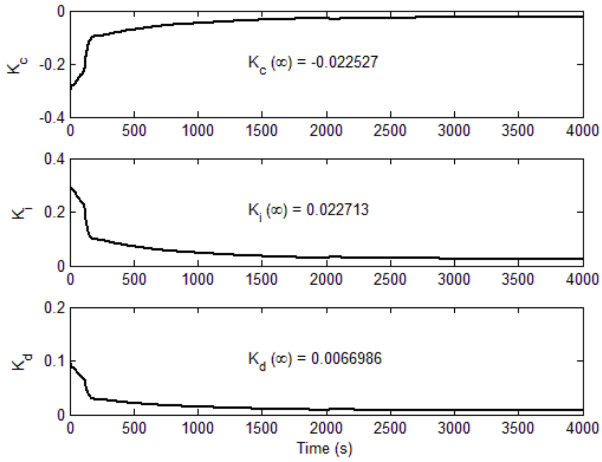

Fig. 2. PID-MR gains adaptation over time.

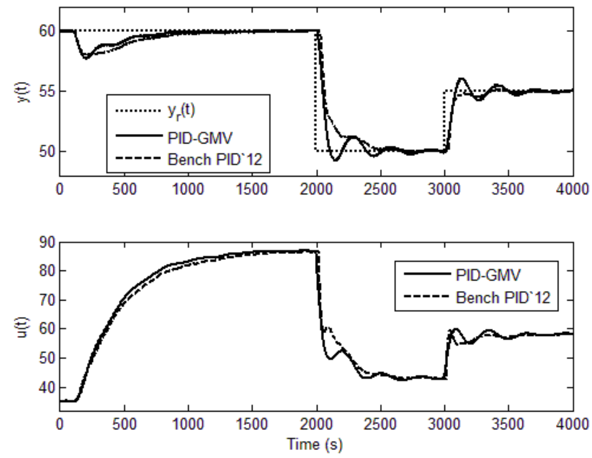

Fig. 3. Process output and PID-GMV control signal.

6.3 PID-FIR simulation

Using the same RLS estimator as in the other two previous simulations, the PID-FIR controller was the one exhibiting the closest output response as compared to the Benchmark PID'12 in the upper plot in Fig. 5. The PID-FIR filter

$S(z^{-1}) = 1 - 0.85z^{-1}$, $T(z^{-1}) = S(1)$, and $L = 1$ were selected to give a satisfactory dynamic behavior in terms of the output response, but even with more conservative parameters, the PID-FIR exhibited a very noisy control signal. See in Fig. 6, that its estimated gains values are very high as compared to Figs. 4 and 2 results. To overcome the problem of the noisy control signal, a low-pass filter with $\alpha = 0.1$ was used.

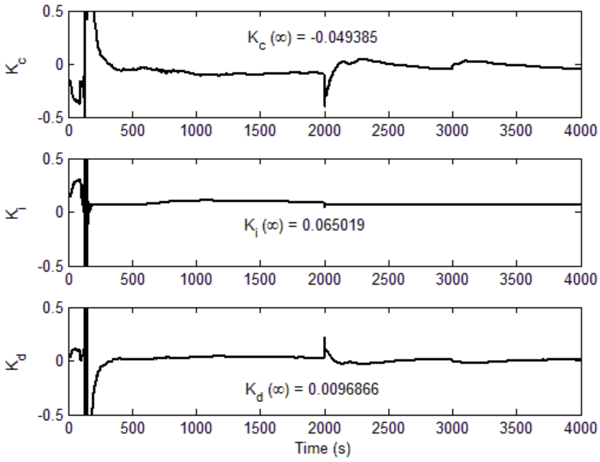

Fig. 4. PID-GMV gains adaptation over time.

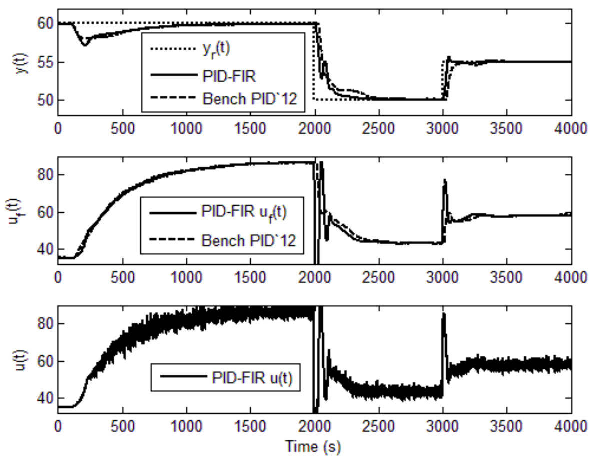

Fig. 5. Process output and PID-FIR control signal.

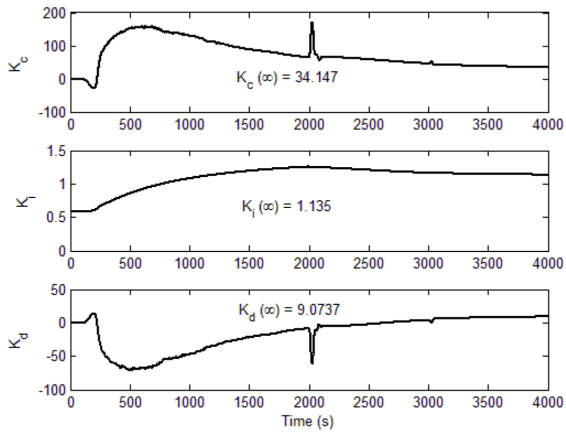

Fig. 6. PID-FIR gains adaptation over time.

7. CONCLUSIONS

The three model-free adaptive controllers reviewed in this work handled well the Benchmark PID’12 with reduced workload of trial and error due to the small number of parameters involved. The PID-MR was found to be the most flexible technique, allowing the best trade-off between robustness and performance within the I+PD framework, whereas the reduced order PID-GMV exhibited difficulties to cope with load disturbance rejection plus oscillation avoidance during setpoint changes. However, it might be possible to improve its performance if a more complex generalized output is used instead of the single tuning parameter form. But this research scope was limited to low maintenance techniques with respect to supervision operators.

And within the ideal PID framework, the PID-FIR exhibited a severe problem of producing a noisy control signal, in spite of some processes not being sensible to it, if the actuator is not designed for such a noisy signal, the PID-FIR technique should be avoided or a low-pass filter must be implemented.

ACKNOWLEDGMENT

This research was supported by the Department of Automation and Systems of the Federal University of Santa Catarina and CNPq.

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