On the Use of Iterative Feedback Tuning Algorithms in Fuzzy Control System Design

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Abstract: The paper deals with an overview on the possibility to use Iterative Feedback Tuning algorithms in the design of a class of fuzzy control systems employing Mamdani-type PI-fuzzy controllers. The presentation is focused on the two-degree-of-freedom fuzzy control system structure. The design method is validated by real-time experimental results for fuzzy controlled nonlinear DC drive-type laboratory equipment.

Keywords: Iterative Feedback Tuning, PI-fuzzy controllers, experiments

1 Introduction

PI and PID controllers are widely used in more then 80% of industrial applications worldwide due to the good control system (CS) performance they offer [1]. Since the main tasks in control, the achievement of good CS performance in reference input tracking and the regulation in the presence of disturbance inputs, are difficult to be accomplished by means of PI and PID controllers in one-degree-of-freedom structures, an alternative is to develop two-degree-of-freedom (2-DOF) controllers which have advantages over the one-degree-of-freedom ones [2, 3]. However, the main drawback of 2-DOF control structures in the linear case is that although they ensure regulation, the reduction of the overshoot is paid by slower responses with respect to the modification of reference input. The presentation in the paper will be concentrated on the PI controller case.

Another solution with to ensure good CS performance in the conditions of complex, even ill-conditioned, plants is fuzzy control. The development of fuzzy control systems (FCSs) is usually performed by heuristic means, incorporating human skills, with the drawback in the lack of general-purpose design methods. A major problem, which follows from this way to design fuzzy controllers (FCs) is the analysis of several properties of the FCS including stability, controllability, parametric sensitivity and robustness [4, 5].

If low cost automation solutions are required then systematic design methods devoted to relatively simple FCs. One approach is to design firstly 2-DOF PI controllers for the plants characterized by simplified linearized models. Then, transfer of results from the linear case to the fuzzy one resulting in 2-DOF PI-fuzzy controllers (PI-FCs) is done in terms of the modal equivalence principle [6] accepting the well acknowledged equivalence in certain conditions between FCSs and linear / linearized CSs [7].

Iterative Feedback Tuning (IFT) [8, 9] is a gradient-based approach, based on input-output data recorded from the closed-loop system. The CS performance indices are specified through certain cost functions (c.f.s). Optimizing such functions usually requires iterative gradient-based minimization, implemented as IFT algorithms, observing that the c.f.s can be complicated functions of the plant and of the disturbances dynamics. The key feature of IFT is that the closed-loop experimental data are used to compute the estimated gradient of the c.f. Several experiments are performed at each iteration and the updated controller parameters are obtained based on the input-output data collected from the system.

In this context, the aim of combining IFT algorithms with fuzzy control by the transfer of results from the linear case to the fuzzy one is to obtain new and attractive low cost fuzzy control solutions ensuring FCS performance enhancement.

The paper is organized as follows. An overview on the IFT algorithms used in tuning the linear 2-DOF PI controllers is presented in Section 2. Then, Section 3 is focused on a new design method for a class of Mamdani-type two-degree-of-freedom PI-fuzzy controllers (2-DOF PI-FCs). Section 4 is dedicated to the validation of the method by applying it in a case study regarding the speed control of a nonlinear DC drive-type laboratory equipment, and Section 5 ends the paper with concluding remarks.

2 Overview on Iterative Feedback Tuning Algorithms

Two versions of equivalent control system structures can be used in case of 2-DOF control structures to ensure either the simultaneous tuning of controller parameters [10] or their separate tuning for each of the controller blocks [11]. The first version will be presented as follows, with the nomenclature according to Fig. 1(a), where the two controller blocks are characterized by the transfer functions $C_r(s)$ and $C_y(s)$. The basic details in the two-degree-of-freedom control structure case are presented in [12].

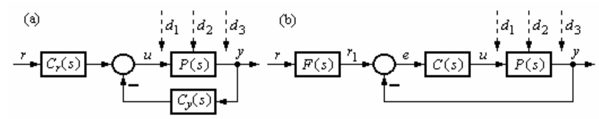

Figure 1

2-DOF control system structure used in IFT (a), and used feedforward filter (b)

The IFT method consists of the steps A) … E) to be presented as follows in relation with the 2-DOF PI controller with the structure presented in Fig. 1(b), where: $C(s)$ – transfer function of the basic PI controller:

$$C(s) = k_c(1 + sT_i)/s = k_C[1 + 1/(sT_i)],\ k_C = T_i k_c, \tag{1}$$

with k_C – controller gain and T_i – integral time constant, $F(s)$ – transfer function of the feedforward filter:

$$F(s) = 1/(1 + T_i s), \tag{2}$$

r – reference input, y – controlled output, $e = r_1 - y$ – control error, u – control signal, r_1 – output of block $F(s)$ (filtered reference input), d_1, d_2, d_3 – load disturbance input scenarios, with the general disturbance input $d \in \{d_1, d_2, d_3\}$, and te connections between the controller blocks in Fig. 1(a) and (b) are:

$$C_r(s) = C(s)F(s),\ C_y(s) = C(s). \tag{3}$$

In these conditions, the steps of the IFT approach are:

A) A controller, of desired complexity, which stabilizes the system, has to be chosen. A discrete form of the controller is needed. The parameterization of the controller is such that the transfer functions $C_r(s, \boldsymbol{\rho})$ and $C_y(s, \boldsymbol{\rho})$ are differentiable with respect to its parameters, $\boldsymbol{\rho}$ being the parameters vector.

In order highlight the controller tuning parameters, the parameters vector $\boldsymbol{\rho}$ has been added as additional input variable to the transfer function. This nomenclature will be used in the sequel in both continuous- and discrete-time not only for the transfer functions but also to the variables regarding the plant (the control signal u and the controlled output y).

B) A reference model must be chosen, prescribing the desired CS behaviour observed in y. This model is typically chosen with first- or second-order dynamics and, for the sake of simplicity and better CS performance, it can be also chosen to be without dynamics, having the transfer function equal to the unity.

C) The general expression of the c.f. J is proposed in (4) in the framework of this optimization problem:

$$\boldsymbol{\rho}^* = \arg\min_{\boldsymbol{\rho}\in SD} J(\boldsymbol{\rho}),\ J(\boldsymbol{\rho}) = \frac{1}{2N}\cdot\sum_{k=1}^{N}\{[L_y(q^{-1})\,\delta y(k,\boldsymbol{\rho})]^2 + \\ +\lambda\,[L_u(q^{-1})u(k,\boldsymbol{\rho})]^2\}, \tag{4}$$

where: N – length of each experiment, L_y, L_u – weighting filters, introduced to emphasize certain frequency regions, λ – weighting constant, δy – output error, the difference between the actual output (y) and the desired output (y_d):

$$\delta y = y - y_d\,. \tag{5}$$

D) The update law must be set by which the next set of parameters will be computed. This law corresponds usually to a Gauss-Newton scheme of type (6), other versions being also used to avoid the computation of second-order derivatives:

$$\boldsymbol{\rho}^{i+1} = \boldsymbol{\rho}^i - \gamma^i(\mathbf{R}^i)^{-1}est\left[\frac{\partial J}{\partial\boldsymbol{\rho}}(\boldsymbol{\rho}^i)\right], \tag{6}$$

where: i – index of the current iteration, $est[x]$ – estimate (generally) of the variable x, $\gamma^i > 0$ – parameter to determine the step size.

E) The regular matrix $\mathbf{R}^i$ in (6) is a positive definite matrix, usually the Hessian of $J(\boldsymbol{\rho})$:

$$\mathbf{R}^i = \frac{1}{N}\sum_{k=1}^{N}\left(est\left[\frac{\partial y}{\partial\boldsymbol{\rho}}(k,\boldsymbol{\rho}^i)\right]est\left[\frac{\partial y}{\partial\boldsymbol{\rho}}(k,\boldsymbol{\rho}^i)\right]^T + \lambda\,est\left[\frac{\partial u}{\partial\boldsymbol{\rho}}(k,\boldsymbol{\rho}^i)\right]\cdot \\ \cdot est\left[\frac{\partial u}{\partial\boldsymbol{\rho}}(k,\boldsymbol{\rho}^i)\right]^T\right). \tag{7}$$

Choosing the identity matrix for $\mathbf{R}^i$ ensures the negative direction of the gradient, but it is recommended to compute $\mathbf{R}^i$ by a quasi-Newton method or as the Hessian of the c.f.

IFT algorithms are used to implement the process of solving the optimization problem (4), where several additional constraints can be imposed regarding the plant or the closed-loop system. One necessary constraint concerns the stability of the closed-loop system, and SD in (4) stands for stability domain. In addition, the expression of the c.f. can be modified by adding quadratic terms with the output sensitivity functions defined in the time domain, accordingly weighted to reduce the sensitivity of the CS with respect to the parametric variations of the controlled plant similar to the presentation in [13] for FCSs.

IFT algorithms are iterative, in case of 2-DOF PI controllers considered here this corresponds to three real-time experiments performed with the CS, the first and the third one referred to as normal ones and the second one referred to as the gradient one. The normal experiments are characterized by the reference input fed to the CS while in case of the gradient experiment the role of reference input fed to the CS is played by the control error in the first experiment. The input-output data recorded from these three experiments are employed to compute the estimated gradients of the controlled output and of the control signal required in computing the estimated gradient of the c.f. $J(\boldsymbol{\rho})$.

The IFT algorithms considered here, contain the steps 1 … 8 to obtain the next set of parameters:

Step 1 The three experiments are done and the input-output data (u_1, y_1), (u_2, y_2) and (u_3, y_3) are recorded.

Step 2 The output of the reference model is generated, y_d, and the output error δy is computed by (5).

Step 3 The estimated gradient of the output is computed based on the data recorded form the real-time experiments. Before applying this approximation, the sensitivity function S and the complementary sensitivity function T must be expressed in discrete-time, with the following definitions according to the CS structure illustrated in Fig. 1(a):

$$\begin{aligned} S(q^{-1},\boldsymbol{\rho}) &= 1/[1+P(q^{-1})C_y(q^{-1},\boldsymbol{\rho})], \\ T(q^{-1},\boldsymbol{\rho}) &= P(q^{-1})C_r(q^{-1},\boldsymbol{\rho})/[1+P(q^{-1})C_y(q^{-1},\boldsymbol{\rho})]. \end{aligned} \tag{8}$$

The analytical expression of the gradient of δy is obtained using (8) and taking the derivatives with respect to $\boldsymbol{\rho}$ [10]:

$$\begin{aligned} est\left[\frac{\partial \delta y}{\partial \rho}(k,\boldsymbol{\rho})\right] = \frac{1}{C_r(q^{-1},\boldsymbol{\rho})}\cdot\left[\frac{\partial C_r}{\partial \boldsymbol{\rho}}(q^{-1},\boldsymbol{\rho})y_3(k,\boldsymbol{\rho}) - \right. \\ \left. - \frac{\partial C_y}{\partial \boldsymbol{\rho}}(q^{-1},\boldsymbol{\rho})y_2(k,\boldsymbol{\rho})\right]. \end{aligned} \tag{9}$$

Step 4 The control signal is a perfect realization of the control signal in the first experiment:

$$u = u_1 \,. \tag{10}$$

Step 5 The estimated gradient of the control signal u is computed using (8), (10) and taking the derivatives with respect to $\boldsymbol{\rho}$:

$$est\left[\frac{\partial u}{\partial \rho}(k,\boldsymbol{\rho})\right]=\frac{1}{C_r(q^{-1},\boldsymbol{\rho})}\cdot\left[\frac{\partial C_r}{\partial \boldsymbol{\rho}}(q^{-1},\boldsymbol{\rho})u_3(k,\boldsymbol{\rho})-\right.$$
$$\left.-\frac{\partial C_y}{\partial \boldsymbol{\rho}}(q^{-1},\boldsymbol{\rho})u_2(k,\boldsymbol{\rho})\right]. \quad (11)$$

Step 6 The c.f. $J(\boldsymbol{\rho})$ is computed according to (4) with the estimated and its estimated gradient results as follows:

$$est\left[\frac{\partial J}{\partial \boldsymbol{\rho}}(\boldsymbol{\rho})\right]=\frac{1}{N}\sum_{k=1}^{N}\{L_y(q^{-1})\delta y(k,\boldsymbol{\rho})est\left[\frac{\partial \delta y}{\partial \boldsymbol{\rho}}(k,\boldsymbol{\rho})\right]+$$
$$+\lambda L_u(q^{-1})u(k,\boldsymbol{\rho})est\left[\frac{\partial u}{\partial \boldsymbol{\rho}}(k,\boldsymbol{\rho})\right]\}, \quad (12)$$

with the values of the estimated gradients obtained as steps 3 and 5.

Step 7 The matrix $\mathbf{R}^i$ is computed in terms of (7).

Step 8 The next set of parameters is obtained by a Gauss-Newton scheme according to (6).

3 Design Method for Mamdani-type Two-degree-of-freedom PI-fuzzy Controllers

The structure of the 2-DOF linear considered here consists of fuzzifying the basic linear PI controller with the transfer function $C(s)$ in Fig. 1(b). The 2-DOF PI-FC, with the structure presented in Fig. 2, represents a discrete-time controller involving a basic fuzzy controller without dynamics (B-FC). The dynamics is inserted by the numerical differentiation of the control error e_k expressed as the increment of control error, $\Delta e_k=e_k-e_{k-1}$, and by the numerical integration of the increment of control signal, Δu_k.

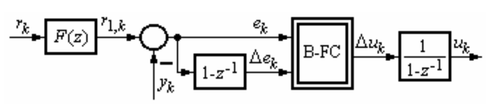

Figure 2
Structure of two-degree-of-freedom PI-fuzzy controller

The block B-FC is a nonlinear two inputs-single output system, which includes among its nonlinearities the scaling of inputs and output as part of its fuzzification module. The fuzzification is solved in terms of the regularly distributed input and output membership functions presented in Fig. 3. Other distributions of the membership functions can modify in a desired way the controller nonlinearities.

The inference engine in B-FC employs Mamdani's MAX-MIN compositional rule of inference assisted by the rule base presented in Table 1, and the centre of gravity method for singletons is used for defuzzification.

The design method for this class of Mamdani-type 2-DOF PI-FCs consists of the following design steps:

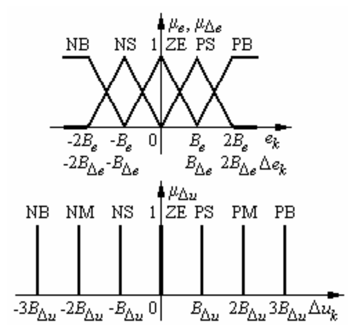

Figure 3

Membership functions of B-FC in Fig. 2

Table 1

Decision table of B-FC

Δe_k	e_k				
	NB	NS	ZE	PS	PB
PB	ZE	PS	PM	PB	PB
PS	NS	ZE	PS	PM	PB
ZE	NM	NS	ZE	PS	PM
NS	NB	NM	NS	ZE	PS
NB	NB	NB	NM	NS	ZE

Step A Design the continuous-time 2-DOF PI controller by a specific method to the linear case depending on the class of considered controlled plants (in simplified mathematical models for the design) and on the desired / imposed CS performance indices.

Step B Set the value of the sampling period T_s chosen in accordance with the requirements of quasi-continuous digital control, express the discrete-time equation of the-point filter $F(z)$, express the discrete-time equation or equations of the digital PI controller $C(z)$ in its incremental version:

$$\Delta u_k = K_P \cdot \Delta e_k + K_I \cdot e_k = K_P (\Delta e_k + \alpha \cdot e_k), \quad (13)$$

and compute the parameters $\{K_P, K_I, \alpha\}$, expressed in (14) when Tustin's method is applied:

$$K_P = k_C[1 - T_s/(2T_i)],\ K_I = k_C T_s / T_i,\ \alpha = K_I / K_P = 2T_s/(2T_i - T_s)\,. \tag{14}$$

Step C Apply the modal equivalence principle to obtain two of fuzzy controller parameters, $B_{\Delta e}$ and $B_{\Delta u}$:

$$B_{\Delta e} = \alpha B_e, B_{\Delta u} = K_I B_e, \tag{15}$$

where the third parameter, B_e, represents designer's option.

4 Case Study

A case study is considered to validate the new design method dedicated to the two-degree-of-freedom PI-fuzzy controllers proposed in the previous Section. The case study is focused on a fuzzy controller design for the class of plants with the transfer function $P(s)$ characterizing the simplified mathematical models used in servo system control as part of mechatronic systems and of embedded systems:

$$P(s) = k_P/[s(1+T_\Sigma s)], \tag{16}$$

where k_P is the controlled plant gain and T_Σ is the small time constant or an equivalent time constant as sum of parasitic time constants.

One solution to control this class of plants is represented by PI control [1]. A simple and efficient way to tune the parameters of the 2-DOF PI controller controlling the plant (16) is represented by the Extended Symmetrical Optimum (ESO) method [14], characterized by only one design parameter, β. The choice of the parameter β within the domain $1 < \beta < 20$, leads to the modification of the CS performance indices (σ_1 – overshoot, $\hat{t}_r = t_r / T_\Sigma$ – normalized rise time, $\hat{t}_s = t_s / T_\Sigma$ – normalized settling time defined in the unit step modification of r, φ_m – phase margin) according to designer's option and to a compromise to these performance indices using the diagrams presented in Fig. 4 in the situation without feedforward filter. The presence of the feedforward filter with the transfer function $F(s)$ improves the CS performance indices.

The PI tuning conditions, specific to the ESO method, are:

$$k_c = 1/(\beta\sqrt{\beta}T_\Sigma^2 k_P),\ T_i = \beta T_\Sigma\,. \tag{17}$$

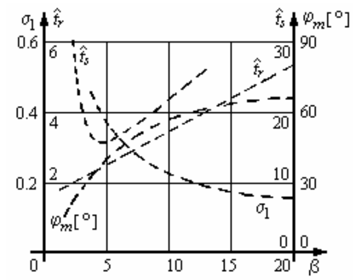

Figure 4
Control system performance indices versus β in the situation without feedforward filter

These tuning conditions highlight the presence of only design parameter, β. This simplifies the application of the IFT algorithms in Section 2 because the parameters vector becomes a scalar:

$$\boldsymbol{\rho} = \beta \,. \tag{18}$$

The experimental setup consists of speed control of a nonlinear laboratory DC drive (AMIRA DR300). The DC motor is loaded using a current controlled DC generator, mounted on the same shaft, and the drive has built-in analog current controllers for both DC machines having rated speed equal to 3000 rpm, rated power equal to 30 W, and rated current equal to 2 A. The speed control of the DC motor is digitally implemented using an A/D-D/A converter card. The speed sensors are a tacho generator and an additional incremental rotary encoder mounted at the free drive-shaft. The block diagram of the hardware station is presented in Fig. 5.

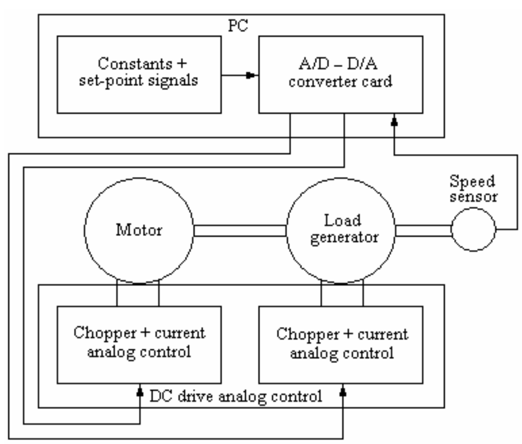

Figure 5
Block diagram of hardware station

The mathematical model of the plant can be well approximated by the transfer function $P(s)$ in (16), with $k_P = 4900$ and $T_\Sigma = 0.035$ s.

The development method proposed in the previous Section is applied, and for the sake of simplicity only the main parameter values are presented. The method starts with the choice of the initial value of the design parameter β = 6. Then, a version of IFT algorithm presented in Section 2 is applied in the condition of the following c.f., J:

$$J = \frac{1}{2N} \sum_{k=1}^{N} (\delta y^2(k)) . \tag{19}$$

The following "optimal" values of the PI-FC tuning parameters have been obtained after six iterations: $B_e = 0.3$, $B_{\Delta e} = 0.03$, $B_{\Delta u} = 0.0021$, for $\beta^* = 5.76$.

Part of the real-time experimental results – the variations of r and y versus time – are presented in Fig. 6 and Fig. 7 for the linear CS (with 2-DOF PI controller) and for the fuzzy CS (with 2-DOF PI-FC) in the conditions:

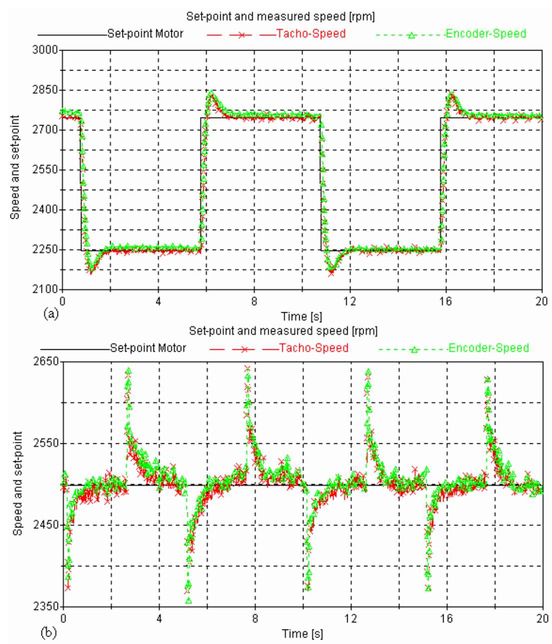

Figure 6

Control system response with 2-DOF PI controller

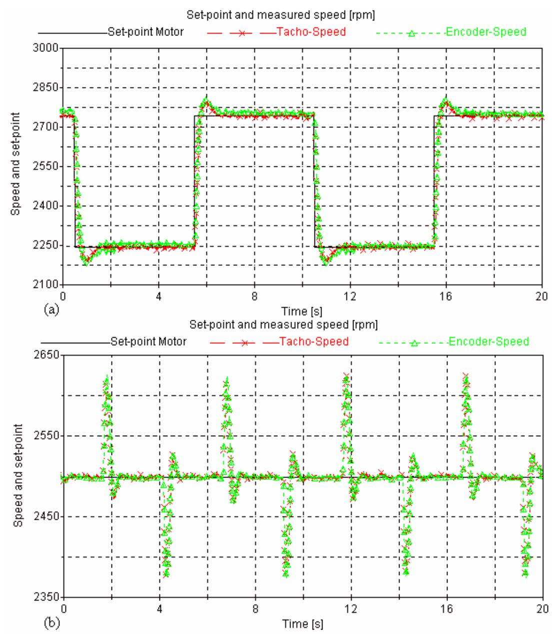

Figure 7

Control system response with 2-DOF PI-fuzzy controller

- without load in Fig. 6(a) and Fig. 7(a),
- 5 s period of 10% rated $d = d_2$ type load and $r = 2500$ rpm in Fig. 6(b) and Fig. 7(b).

Conclusions

The paper presents aspects concerning the use Iterative Feedback Tuning algorithms in the design of a class of fuzzy control systems employing Mamdani-type PI-fuzzy controllers.

It is proposed a new design method for the PI-fuzzy controllers validated by real-time experiments related fuzzy control solutions dedicated to a class of plants applied in servo systems as part of mechatronics systems and of embedded systems.

The design method illustrates the potential of IFT employed in connection with fuzzy control in complex plants.

Future research will focus on the on-line implementation of IFT algorithms to with control other laboratory equipment with discrete-event systems including robots and manufacturing systems [15, 16, 17].

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