Classical and Fuzzy Approaches to 2–DOF Control Solutions for BLDC–m Drives

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Abstract. This chapter gives two–degree–of–freedom (2–DOF) speed control solutions for brushless Direct Current motor (BLDC–m) drives with focus on design methodologies. A classical 2–DOF structure, 2–DOF proportional-integral (PI) and proportional–integral–derivative (PID) structures and 2–DOF fuzzy control solutions are presented and approaches regarding the methods are highlighted. A case study concerning a BLDC–m drive with variable moment of inertia is presented. Comparative studies based on digital simulation results are included to exemplify the design methods.

Keywords: Speed control, 2–DOF control, brushless direct current motor, PID control.

1 Introduction

The research results obtained in mechatronics systems during the last decade have been focused on setting theoretical foundations and, based on it, on enlarging the application domains. The studies are oriented to assessing the quality of motion control systems and on disclosing the insurmountable performance limitations inherent in the mechanical structures. Both accuracy and robustness are essential characteristics in high performance motion applications with variable moment of inertia (VMI), [1]. As result, robust control solutions for servomechanisms have been

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proposed to realize effective disturbance suppression in the presence of stability degree constraints [2], [3].

The main advantages of using two–degree-of-freedom (2–DOF) control solutions concern simultaneous good feedback properties, reference tracking and disturbance rejection. One drawback of 2–DOF controllers is that the overshoot reduction is paid by slower set–point responses [3], [4], [5]. So, the design of 2–DOF controllers represents a multi-objective problem. With this regard, Miklosovic and Gao offer in [3] a robust 2–DOF control design technique that extends the concepts of active disturbance rejection control.

Several 2–DOF control structures (CSs) have been proposed during the last decades. They are characterized by different combinations of the inclusion in the feed-back loop the reference part and of the disturbance part requirements. Even the classical design has different approaches [6], [7], [8], [9], [10]. For the development of classical 2–DOF PI(D) CSs for low order plants, Araki and Taguchi present in [6] similarities between 2–DOF control structures and one–degree–of–freedom (1–DOF) controllers (mainly PI(D) ones) extended with input filters (reference and feedback).

Alternative approaches to the design of 2–DOF proportional–integral (PI) and proportional–integral–derivative (PID) control solutions are the ESO–m [11] and the 2p–ESO–m design methods [12]. These approaches are recommended mainly for applications with VMI. The computer–aided design of various types of 2–DOF controllers based on algebraic methods is analyzed in [12], [14], [13].

The fuzzy logic technique can be also inserted in the 2–DOF CSs, and several approaches are suggested for 2–DOF fuzzy control (FC) structures. A 2–DOF controller consisting of a one-step-ahead fuzzy pre-filtering in the feed–forward loop and a PI–fuzzy controller in the feedback loop dedicated to the foot trajectory tracking control is discussed in [15] and [16], where self–tuning and model reference adaptive 2–DOF PID–fuzzy controllers are presented. A new framework for the design of generic 2–DOF linear and fuzzy controllers dedicated to plants with integral components and nonlinearities is proposed and applied in [17] and [18]. In addition, the variability of the plant parameters needs sometime also the permanent adaptation of the control algorithm or of its parameters as shown in [13], [15] and [16].

The presented research results are based mainly on easy accessible references. They are focused on two classes of 2–DOF controllers, the classical 2–DOF (and its PI(D) representation) and 2–DOF fuzzy controllers. The applications involve a class of mechatronics systems.

The chapter is structured as follows. Section 2 presents a brief overview on the classical 2–DOF controller approach based on [14] and [19], on 2–DOF–PI(D) equivalent structures providing the foundation for discussion and comparison for the design methodology, and on the basic structures for 2–DOF fuzzy controllers derived from a PI(D) approach. Section 3 gives a short description of the plant, i.e., a servo system application built around a brushless Direct Current Motor

(BLDC–m) with VMI, with inner control (current) and usable models for design. Section 4 discusses some aspects regarding speed control solutions with 2–DOF controllers, experimental scenarios and simulation results. Section 5 is dedicated to the concluding remarks.

2 Classical Structures of 2–DOF and 2–DOF PI(D) Controllers

2.1 Basic Structure and Polynomial Design of 2–DOF Controllers

The 2–DOF CS in its classical (discrete or continuous) form [4], [19], uses two distinct controllers (Fig. 1): the reference controller $T(z)/R(z)$, where $T(z)$ is the reference filter, and the feedback controller $S(z)/R(z)$. The polynomial $R(z)$ is the common part which include mainly the integral components. The classical design of the unknown polynomials $T(z)$, $S(z)$ and $R(z)$ is known as polynomial design problem based on solving the polynomial Diophantine equation [5], [14] under different particularities in treating the constrains and causality (degree) conditions for the polynomials; such conditions are exemplified in [4] and [5]. Unlike the 1–DOF CS, in case of the 2–DOF controller the enlisted attributes can be separately adjusted without influencing one another.

In the usual discrete form the plant is characterized by the pulse transfer function (t.f.) calculated from the continuous model

$$P(z) = (1 - z^{-1})Z\{\frac{P(s)}{s}\} = \frac{B(z)}{A(z)}, \tag{1}$$

where the plant parameters can be time variable. The servo performances are given by a reference model in the form $\frac{B_m(z)}{A_m(z)}$ with an additional condition for the zero control error. Finally, the t.f. of the CS obtains the form

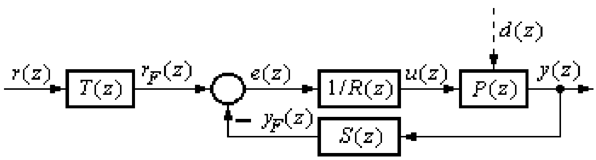

Fig. 1 Structure of classical 2–DOF controller and control structure with the $R(z)$ component placed in the loop

$$\frac{T(z)}{A(z)R'(z)+B^-(z)S(z)} = \frac{B'_m(z)}{A_m(z)} \cdot \frac{A_o(z)}{A_o(z)},$$
$$T(z) = B'_m(z)A_o(z), \tag{2}$$
$$\mathrm{A}(z)\mathrm{R}'(z) + \mathrm{B}^-(z)\mathrm{S}(z) = \mathrm{A_m}(z)\mathrm{A_o}(z).$$

The last equation in (2) is a Diophantine equation over the ring of polynomials, and its solutions are the coefficients of the polynomials $T(z)$, $R(z)$ and $S(z)$ [12].

2.2 PID Controllers. 2–DOF Controller Interpretation

For 1–DOF controllers, the CS performance can be improved using several particular controller structures with non-homogenous dynamics with respect to the two inputs [5] as shown in Fig. 2. Each controller block can be characterized by its own t.f.s. The presented approach permits also an easy 2–DOF interpretation of the design as discussed in [7] for classical PI(D) controllers and in [8] for fuzzy controllers.

Three such 2–DOF CSs are presented in Fig. 3 and referred to as [17] the reference input filter structure, Fig. 3 (a), the feed-forward structure, Fig. 3 (b), and the feedback structure, Fig. 3 (c). The connections between 2–DOF and extended with input filters of 1-DOF controller structures are synthesized in Table 1, where: P – proportional, D – derivative, I – integral, L1(2) – first (second) order lag filter. The choice of a certain representation of the controller depends on the structure of the available controller and on the adopted algorithmic design method and the result of this design.

Using the approach proposed in [6], the main PI(D) controller component – or/and $C^*(s)$ - defined in Fig. 3 (a) to (c) are characterized by the following t.f.s:

$$C^*(s) = \begin{cases} \frac{k_c(1+T_F s)}{s} & \text{PIcontroller} \\ \frac{k_c(1+T_F s)(1+T'_c s)}{s[1+(T_d/N)s]} & \text{PIDcontroller} \end{cases}, \tag{3}$$

with a possible additional block $C_F(s)$ which accelerates the effect of the reference input in the control signal:

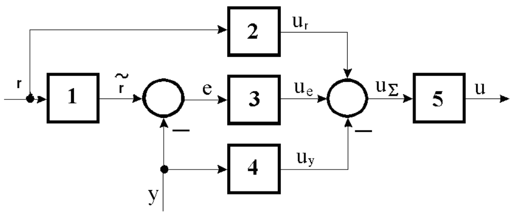

Fig. 2 Typical controller structures and particular forms of the modules

Table 1 Connections between 2–DOF controllers and extended 1–DOF controller structures

Fig. 3 (a)		$F(s)$	–	$F(s)C(s)$	$C(s)$	Remarks
Fig. 3 (b)		–	$C_F(s)$	$C(s)–C_F(s)$	$C(s)$	-
Fig. 3 (c)		–	$C_F(s)$	$C^*(s)$	$C^*(s)+C_F(s)$	-
α_1	α_2	–	–	(ref. channel)	(feedback)	
0	0	1	0	PID	PID	1-DOF controller
0	1	PDL2	DL1	PI	PID	1-DOF with non-homogenous behavior
1	0	PD2L2	P	PID-L1	PID	
1	1	PL2	PDL2	I	PID	
α_1	α_2	PID controller with pre-filtering (2-DOF controller)				

$$C_F(s) = \begin{cases} k_c(T_c - T_F) & \text{PIcontroller} \\ \frac{k_c(T_c-T_F)(1+T_c' s)}{1+(T_d/N)s} & \text{PIDcontroller} \end{cases}$$

The digital implementation can be supported by the classical informational diagram presented in detail for example in [5]. The bump-less switching between two control algorithms (c.a.s) – connected to linearized plant models and referred as c.a. (1) and c.a. (2) – needs a permanent modification of the tuning parameters and the reconsideration of the past values in the control algorithms.

2.3 2–DOF Takagi–Sugeno Fuzzy Control Structures for PID Controllers

The main advantage of the classical Takagi–Sugeno (TS) fuzzy control (FC) structure concerns an easy modeling of the nonlinearities. They also do not need the special bump-less circuit. Extended 2–DOF FC structures can be defined on the ba-

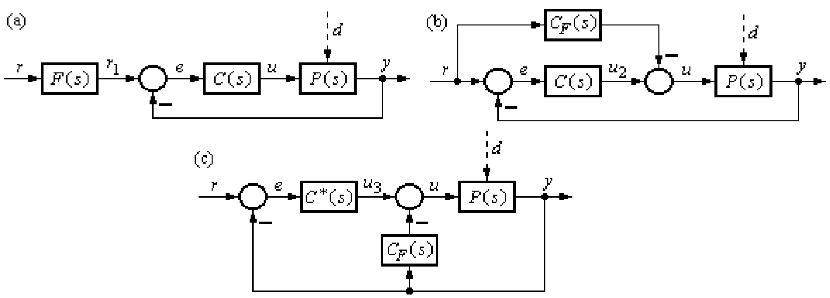

Fig. 3 Structures for 2–DOF controllers as extensions of 1–DOF controllers

sis of the structures given in Figs. 2 and 3; for example, [5] and [17] offer 2–DOF FCs defined around TS fuzzy blocks FB–T_c implemented in terms of Figs. 4 to 7.

The development of the extended 2–DOF PI–FC starts with the definition and development of the classical PI block with the t.f.

$$G^{\tau}(s) = \frac{k_c}{s}(1 + s\tau),$$

with $\tau \geq 0$.

The main PI(D) block is fuzzified in the set-point filter structure or in the feed-forward structure, and the transfer function $C(s)$ (or $C^*(s)$) is expressed in (3). The fuzzification of the generic PI block with the t.f. $G^{\tau}(s)$ leads to the fuzzy block FB–τ; it is accepted that the continuous-time linear block with the t.f. $G^{\tau}(s)$ has the control error e as input and the control signal u as output (other variants are also possible). The structure of the block FB–τ is presented in Fig. 8, where FB is the TS fuzzy block without dynamics, with:

$$\Delta e(k) = e(k) - e(k-1)$$

is the increment of control error,

$$\Delta u(k) = u(k) - u(k-1)$$

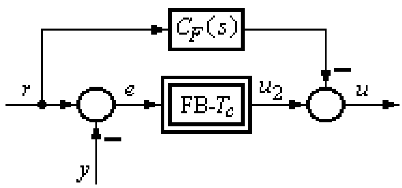

Fig. 4 Structure of feed-forward 2–DOF PI-fuzzy controller

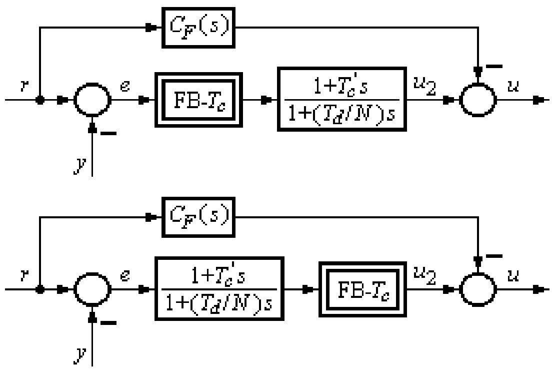

Fig. 5 Structures of feed-forward 2–DOF PID–fuzzy controllers

is the increment of control signal, and k is the index of the current sampling interval because the block FB–τ is implemented as a digital controller.

The fuzzification in the block FC is based on the input membership functions illustrated in Fig. 9, which can be applied for the TS fuzzy block FB–τ. For a low-cost implementation of the 2–DOF fuzzy controllers initially three input membership functions are defined. Fig. 9 points out the tuning parameters, B_e and $B_{\Delta e}$.

More membership functions can be defined for nonlinear plants and high performance specifications. The analysis of and design of the fuzzy controllers should account for the necessary nonlinear scaling factors of the input and output variables of the block FB which must be inserted in the plant. Accepting the sampling period T_s, Tustin's method can be applied to discretize the continuous-time linear generic PI block with the t.f. $G^{\tau}(s)$. This results in the following recurrent equation of the incremental digital generic PI block and its parameters:

$$\Delta u(k) = K_P\,[\Delta\, e(k) + \mu\, e(k)],\;\; K_P = k_c\,(\tau - T_s/2),\;\; \mu = 2T_s/(2\tau - T_s).$$

The complete rule base of the TS fuzzy block FB–τ is given by

$$\begin{aligned}
&\text{Rule 1: IF } e(k) \text{ IS N AND } \Delta e(k) \text{ IS P THEN } \Delta u(k) = K_P^1[\Delta e(k) + \mu^1 e(k)],\\
&\text{Rule 2: IF } e(k) \text{ IS ZE AND } \Delta e(k) \text{ IS P THEN } \Delta u(k) = K_P^2[\Delta e(k) + \mu^2 e(k)],\\
&\text{Rule 3: IF } e(k) \text{ IS P AND } \Delta e(k) \text{ IS P THEN } \Delta u(k) = K_P^3[\Delta e(k) + \mu^3 e(k)],\\
&\text{Rule 4: IF } e(k) \text{ IS N AND } \Delta e(k) \text{ IS ZE THEN } \Delta u(k) = K_P^4[\Delta e(k) + \mu^4 e(k)],\\
&\text{Rule 5: IF } e(k) \text{ IS ZE AND } \Delta e(k) \text{ IS ZE THEN } \Delta u(k) = K_P^5[\Delta e(k) + \mu^5 e(k)],\\
&\text{Rule 6: IF } e(k) \text{ IS P AND } \Delta e(k) \text{ IS ZE THEN } \Delta u(k) = K_P^6[\Delta e(k) + \mu^6 e(k)],\\
&\text{Rule 7: IF } e(k) \text{ IS N AND } \Delta e(k) \text{ IS N THEN } \Delta u(k) = K_P^7[\Delta e(k) + \mu^7 e(k)],\\
&\text{Rule 8: IF } e(k) \text{ IS ZE AND } \Delta e(k) \text{ IS N THEN } \Delta u(k) = K_P^8[\Delta e(k) + \mu^8 e(k)],\\
&\text{Rule 9: IF } e(k) \text{ IS P AND } \Delta e(k) \text{ IS N THEN } \Delta u(k) = K_P^9[\Delta e(k) + \mu^9 e(k)].
\end{aligned}$$

This rule base shows, by the additional upper indices in the rule consequents, that the TS fuzzy block FB–τ can be obtained from the separate tuning of nine linear blocks FB–τ. Therefore the TS fuzzy block FB–τ exhibits like a bump-less interpolator of nine separately tuned linear PI blocks defined in accordance with (9). The SUM and PROD operators are used in the inference engine of the TS fuzzy block FB–τ, and the weighted average method is used in the defuzzification. The modal equivalence

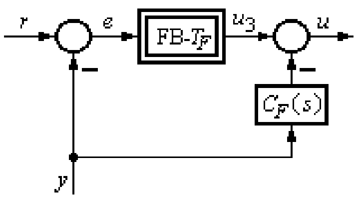

Fig. 6 Structure of feedback 2–DOF PI–fuzzy controller

principle is applied to guarantee the quasi–PI behavior of the fuzzy block FB–τ. This results in the useful tuning conditions

$$B_{\Delta e} = \mu\, B_e,\ \ B_{\Delta u} = K_P\, \mu\, B_e. \tag{4}$$

The tuning conditions are applied in the TS fuzzy block FB–τ. The restructuring of the controller structure allows: accounting for the experience design with PI and PID controllers; the easy introduction of additional specific facilities specific to PI(D) controllers (output or inner limitations, anti windup reset referred to as AWR, smoothing the transition from one algorithm to another one), and the conversion of a PI, PID controller into a 2–DOF controller and vice versa. Such conversion relations are given in [17] and [18].

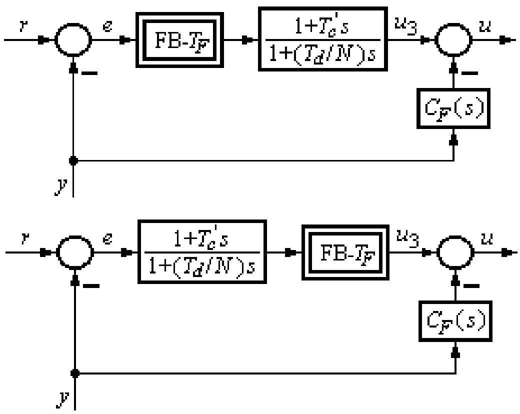

Fig. 7 Structures of feedback 2–DOF PID–fuzzy controllers

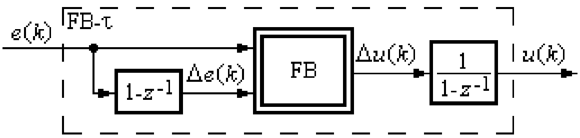

Fig. 8 Structure of FB–τ

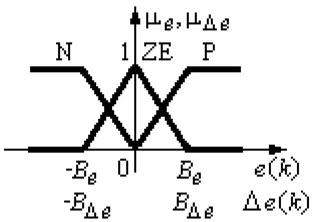

Fig. 9 Input membership functions of FB–τ

The improvement of CS performance can be ensured by the proper choice of the parameter $B_e > 0$ in (4). This can be assisted by the stability analysis accompanied by useful tools specific to the analysis and modeling of fuzzy control systems [20], [21], [22], [23], [24], [25], [26], [27], [28], [29].

3 Mathematical Modeling of Plant as BLDC–m Drive

The controlled plant is represented by a BLDC–m drive with internal current control loop. In the symmetrical operating mode the mathematical models (MMs) of classical DC motors and of BLDC–ms are very close. This fact leads to some similarities in the development of control solutions, relative simple control structure and cheap implementation of the control algorithms [30], [31]. In case of BLDC–ms the current switch is obtained by specialized converters whose commutation time is determined by the position of the rotor, determined either by position sensors or by sensor-less techniques. The major advantages of BLDC–ms are lifespan, high efficiency, very good torque-speed characteristics, and quiet operation. It is accepted that in case of vector control value of the current trends to be zero and the speed control is achieved through the current. Moreover, it is assumed the excitation flux is constant; and the nonlinear effects due to different constructive elements are neglected.

The matrix form of the main equations of the MM of a BLDC–m is presented in [30]. The electromagnetic torque m_e is used in the movement equation

$$m_e = J_e \frac{d}{dt}\omega_r + k_f \omega_r + m_{Load},$$

where m_{Load} is the load torque (i.e., a time variable load disturbance input); the moment of inertia of the driven mechanism J_{mech} can be constant or time-variable:

$$J_e(t) = J_{BLDC} + J_{mech}(t).$$

Two case studies will be considered in the next section. The first case study, with time-variable reference input (including regions characterized by $r(t)$ = const), is representative for defining the adopted controller design methods (the classical 2–DOF and the case with 2–DOF PI controller and feed-forward filter). The main CS performance indices and some simulation results are synthesized in [31].

The second case, to be treated in this chapter, considers the application (the plant) as a simulated plant for a winding process with VMI and constant linear speed, $v_t(t)$ = const according to Fig. 10, where the reference input $r(t)$ for the angular speed $\omega(t)$ must be correlated with the modification of the working roll radius $r_r(t)$. In this context, the CS should ensure the modification of reference input $r(t)$ and the tuning and retuning of the controller parameters as well.

To treat the first aspect, the following condition must be fulfilled, and the measurement of $r_r(t)$ enables the continuous modification of the reference input $r(t)$:

$$v_t(t) = \text{const} \Rightarrow r(t) = k/r_r(t)\ \ (a),\ \ J_e(t) = \frac{1}{2}\rho\pi r_r^4 t\ \ (b). \tag{5}$$

In the design step the inner loop can be characterized by linearized equivalent second order benchmark-type t.f. connected to representative operating points [7]. Such MMs, expressed as benchmark-type t.f.s, are:

- * in the speed control applications:

$$H_P(s) = \frac{k_P}{(1+sT_\Sigma)(1+sT_1)},\ \ T_1 >> T_\Sigma, \tag{6}$$

$$H_P(s) = \frac{k_P}{(1+sT_\Sigma)(1+sT_1)(1+sT_2)},\ \ T_1 > T_2 >> T_\Sigma, \tag{7}$$

- * in the position control applications:

$$H_P(s) = \frac{k_P}{s(1+sT_\Sigma)}, \tag{8}$$

$$H_P(s) = \frac{k_P}{s(1+sT_\Sigma)(1+sT_1)},\ \ T_1 >> T_\Sigma, \tag{9}$$

with time-variable $T_1 = f(J_e(t))$.

The main PI(D) controllers can be tuned by the Extended Symmetrical Optimum method (ESO–m) [11] which can improve the CS performance. The main advantage consist ins fact that only one design parameter (referred to as β) must be adopted. Useful design diagrams concerning the choice of the parameter β and the tuning relations are given in [11]:

$$k_c = \frac{1}{k_P\beta\sqrt{\beta}T_\Sigma^2},\ \ T_c = \beta\ T_\Sigma,\ \ T_c' = T_1. \tag{10}$$

The CS performance indices can be improved further by introducing the reference filter of first-order (a) or of second-order (b), with the t.f.s:

$$F(s) = \frac{1}{1+\beta\ T_\Sigma s}\ \ (a)\ \ or\ \ F(s) = \frac{1+(\beta-\sqrt{\beta})T_\Sigma s+\beta\ T_\Sigma s^2}{(1+\beta\ T_\Sigma s)(1+T_\Sigma s)}\ \ (b). \tag{11}$$

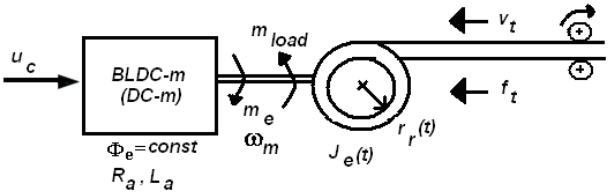

Fig. 10 Block diagram of controlled plant for winding process with VMI

Since $r(t)$ is time-variable and the inner loop can be characterized by a simplified, benchmark-type model, the main (speed) controller is a PID one, extended or not with an integral (I) component. The open–loop CS has $k_0 = k_P k_c$, $T_1 >> T_\Sigma$, k_P and T_1 – time–variable, T_1 must be compensated (applying the pole–zero compensation technique, $T_c' = T_1$), and the gain k_0 must be maintained constant using the permanent recalculation of k_c according to (10).

The considered BLDC–m–based servo system with VMI is characterized by the following parameters: $p = 2$, $R_a = 1\ \Omega$, $L_a = 0.02$ H, $V_{DC} = 220$ V, $J_{e0} = 0.005$ $kg \cdot m^2$. The inner loop, which contains an on–off controller, ensures a second-order (with lag) behavior of the plant. The controlled parameters of the BLDC–m, θ and lpm, were set to ensure that the motor can operate at any desired speed within the range $0 \leq \omega \leq 314\,\mathrm{s}^{-1}$. The functional diagram of a speed controlled BLDC–m drive is presented in Fig. 11.

The internal loop (block diagram) contains the PWM inverter, the current controllers (on–off–type controller with hysteresis and also the current sensors). The main loop contains the actual speed controller and the incremental speed (position) sensor [32], [33], [34], [35]. The phase selection block ensures the proper switching of the phases and the initialization as well.

In the design step of the speed controller the inner control loop characterized by the equivalent second–order benchmark–type t.f. (6) with $k_P = 40$, $T_1 = 0.03$ s and $T_\Sigma = 0.015$. Accepting the variation range of the equivalent moment of inertia $_e(t)$ as relatively small, the controller is tuned using the linearized models (6)–(9). In the first step the classical PI(D) speed controller is designed and tuned using the ESO–m for $\beta = 12$ (or in its extended form of double parameterized (2p–SO–m) form [12]). Finally, due to the fact that the reference input is permanently variable, the reference filter (11) (b) is applied. So the design becomes as a classic case of a

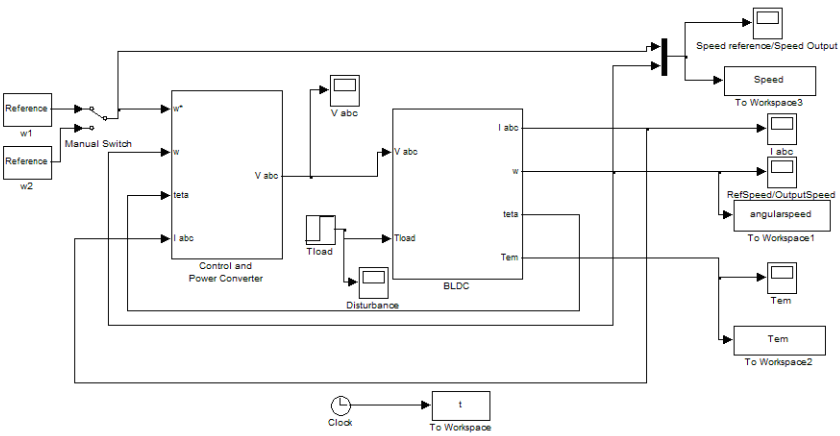

Fig. 11 Functional diagram of a speed controlled BLDC–m drive

2–DOF PI(D) controller, Fig. 3 (a). Accepting a high value for β, which ensures a great value for the nominal phase margin, the controller's parameters are calculated for the average value of $J_e(t)$, without parameter adaptation and with AWR.

4 Experimental Scenarios and Simulation Results

This chapter investigates analytical structures of two-input single-output (TISO) Takagi–Sugeno fuzzy PI(D) controllers versus conventional PI(D) control and variable gain control. Generally, to design and tune PI(D) fuzzy controllers, the continuous–time PI controllers are discretized resulting in the incremental versions of the quasi–continuous digital PI controllers with input/output integration.

The higher operating speed can be accounted for the components of the trapezoidal speed curve, resulting in an average speed equal to the movement speed. The selected structures of 2–DOF PI(D) CSs and of 2–DOF fuzzy PI(D) CSs are presented in detail in Fig. 12 (a), (b) and (c).

In the first experimental scenario the reference input contains (in chronological order) an acceleration part (0 – 1.0 s), a part with constant velocity (1.0 – 1.5 s), a deceleration part (1.5 – 3.0 s), a part with constant velocity (3.0 – 3.5 s), a part characterized by a step disturbance input applied at 3.1 s, and finally a part of deceleration until a stop is reached (3.5 – 4.5 s). This chapter presents the results of the tests conducted with the presented CSs in a first step through simulation for a driving system with BLDC–m with constant moment of inertia.

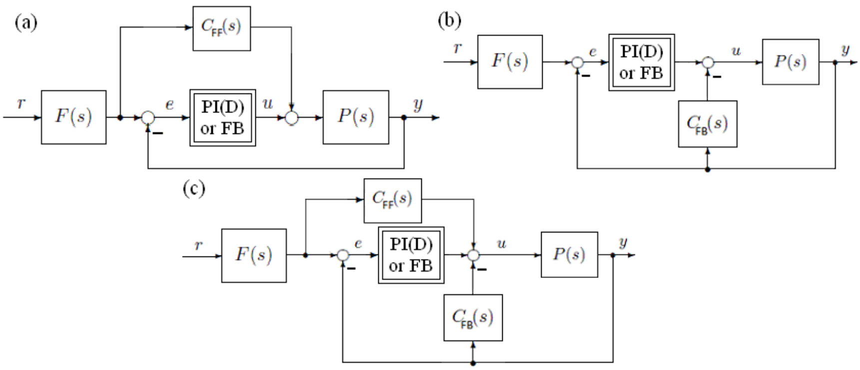

Fig. 12 (a): structures of feed-forward-set-point filter (FF–SP) 2–DOF PI(D)/fuzzy PI(D) CSs, (b): structures of feedback–set–point filter (FB–SP) 2–DOF PI(D)/fuzzy PI(D) CSs, (c): structures of feed–forward–feedback–set–point filter (FF–FB–SP) 2–DOF PI(D)/fuzzy PI(D) CSs

4.1 Feed–Forward–Set–Point Filter Structures

The main simulation results are synthesised in Fig. 13. Fig. 13 (a) illustrates for the simulation scenario the system's output, the angular speed $\omega(t)$ of BLDC–m drives in case of the FF–SP structure. The output is almost the same for both cases, the FF–SP–2–DOF PI(D) controller and the FF–SP–2–DOF fuzzy PI(D) controllers. Fig. 13 (b) details the output around the portion between 0.95 and 1.25 seconds sustaining that the FF–SP structure with 2–DOF fuzzy PI(D) controller ensures better behavior. The control error versus time is presented in Fig. 13 (c); it can be seen that in the tracking phase the control error for FF–SP structure with 2–DOF fuzzy controller is less than the control error for the FF–SP structure with 2–DOF PI(D) controller. For both cases if the angular speed is constant, the control error reaches zero (due to the presence of the I component in controller's structure).

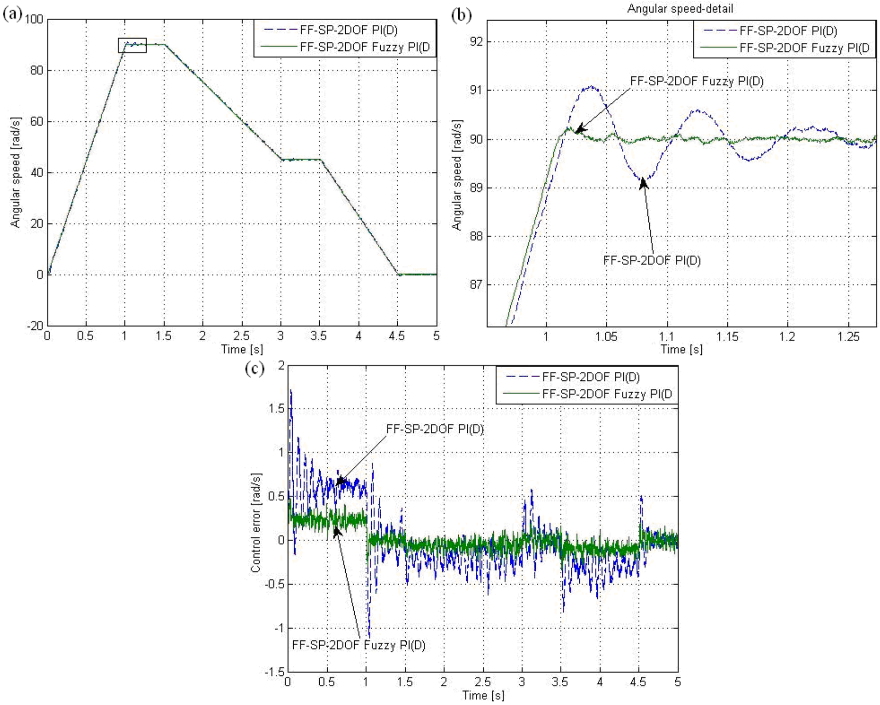

Fig. 13 Simulation results for the FF–SP structures

4.2 Feedback–Set–Point Filter Structures

For the same simulation scenario applied to the FB–SP structure, the system performance indices can be computed using the system responses given in Fig. 14 (a), (b) and (c). For the FB–SP structure with 2–DOF fuzzy PI(D) controller and for the FF–SP structure with 2–DOF PI(D) controller the angular speed are almost the same as suggestively illustrated in Fig. 14 (a).

The results presented in Fig. 13 and 14 show that no discontinuities in the variation of $v(t)$ are observed. During the winding regime the output of the controller remains within the limitations.

According to Fig. 14 (b), the overshoot for the FB–SP structure with 2–DOF PI(D) controller is less than the overshoot for the FF–SP structure with 2–DOF PI(D) controller. In comparison with the FB–SP structure, the FF–SP structure has a more oscillatory character. Finally, Fig. 14 (c) presents the evolution of the control error for the FB–SP structure with 2–DOF PI(D) controller and for the CS with 2–DOF fuzzy PI(D) controller.

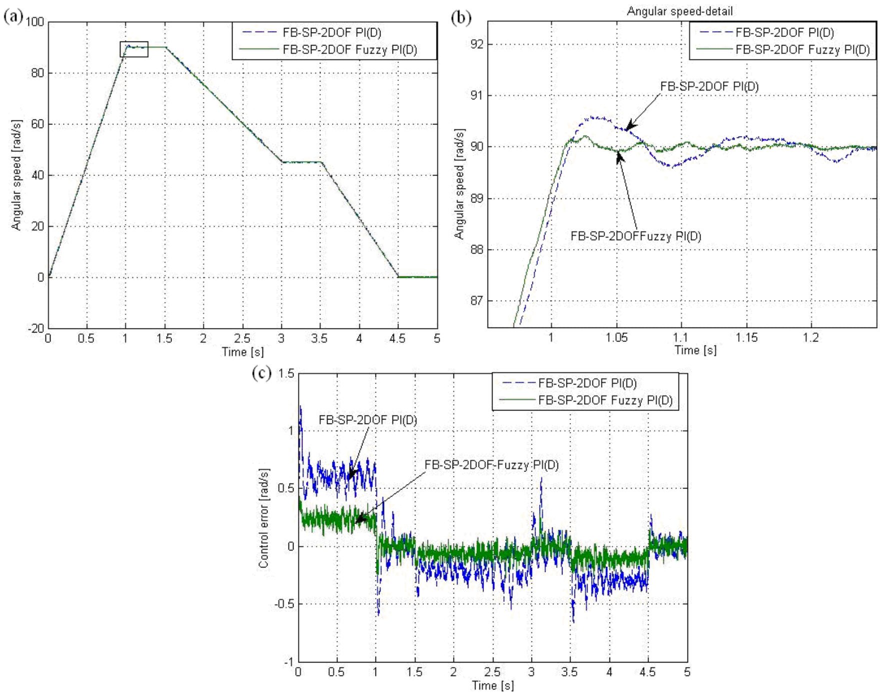

Fig. 14 Simulation results for the FB–SP structures

4.3 Feed–Forward–Feedback–Set–Point Filter Structures

The same simulation scenario is considered for both FF–FB–SP structures. Fig. 15 (a) offers the evolution of the angular speed. Fig. 15 (b) offers some details concerning the angular speed around 1.0 - 1.5 seconds, and it points out that the FF–FB–SP structure with 2–DOF fuzzy controller performs better in comparison with the FF–FB–SP structure with 2–DOF PI(D) controller. Fig. 15 (c) illustrates the control error versus time.

The analysis of these simulations results shows the main conclusion which states that the FF–FB–SP structure with 2–DOF fuzzy PI(D) controller has the best transient behaviors. Therefore, in this case the FF–FB–SP structure with 2–DOF fuzzy PI(D) controller will be adopted for the next experiments (simulations) focused on the case of BLDC–m drive with VMI.

Fig. 16 (a) to (d) synthesizes the main simulation results for fixed controller's parameters according to (10). Since the reference input is permanently variable, the reference filter (11) (b) is applied. The drum radius is calculated according to (5) (a), and the moment of inertia $J_e(t)$ is calculated according to (5) (b). The simulation scenario illustrated in Fig. 16 consists of the start regime, and the angular velocity modification to ensure the desired linear speed v_r, which needs a proper modification of reference input, corresponding to increasing the radius $r_r(t)$ and to the variation of the moment of inertia.

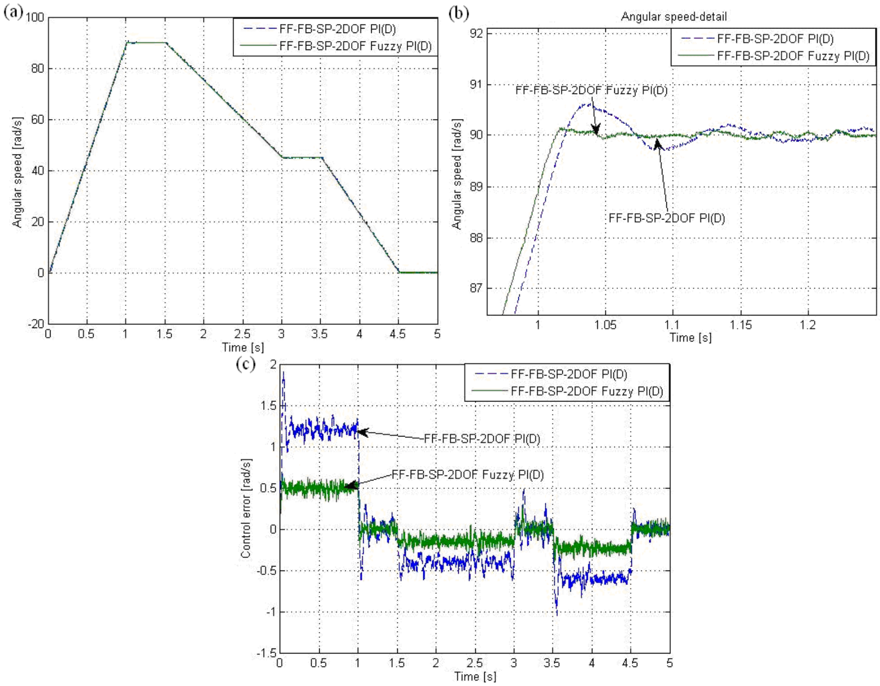

Fig. 15 Simulation results for the FF–FB–SP structures

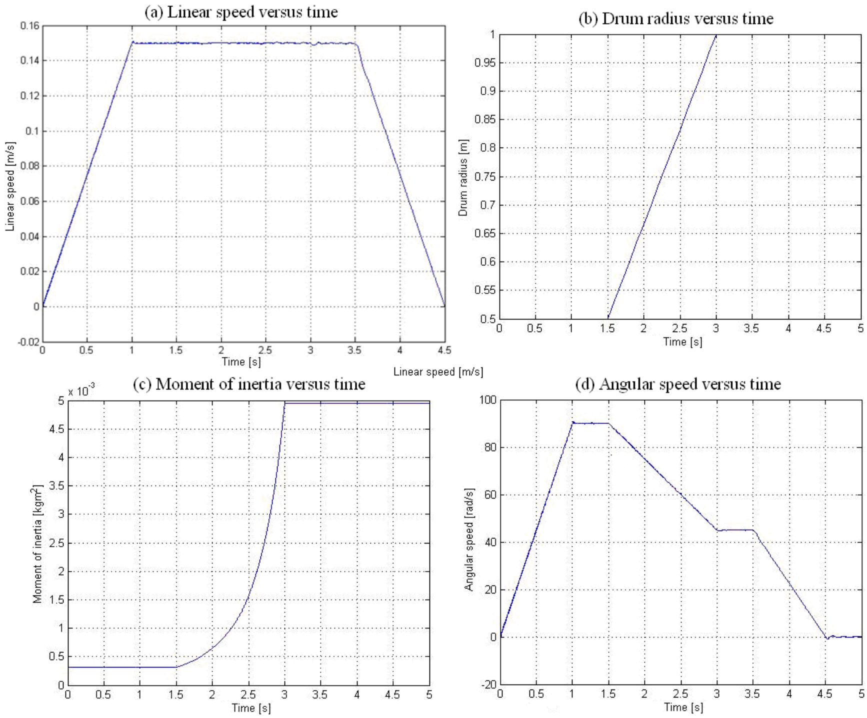

Fig. 16 Simulation results for the FF–FB–SP structure with VMI.

5 Conclusions

This chapter has presented control solutions and methods based on extensions of 2–DOF PI(D) control structures and derived Takagi–Sugeno 2–DOF fuzzy controllers, focused on three basic structures: the FF–SP–, the FB–SP– and FF–FB–SP– structures. The application is a BLDC–m based servo system driving system. The integral element specific to the 2–DOF controllers is included in the forward channel of the control loop.

The advantage of fuzzy logic is the ability to tune certain variables easily by varying the linguistic rules or input variables. The main feature of TS fuzzy models is the expression of the local dynamics of each fuzzy rule by linear system models, and this has been employed in our control solutions.

The proposed controller structures are tuned by the straightforward adaptation of the tuning relations given in the literature. The choice of different representations depends on the structure of the controller, the methods used in controller design and tuning and the final form of the t.f.

Due to the nonlinearities in the plant, the fuzzy control solutions are more advantageous in comparison with other BLDC–m control solutions reported by the state–of–the–art. The proposed controller structures can be implemented relatively easily in quasi continuous digital version by using well–known approaches [4], [5], [7].

The application related to a BLDC–m drive system with VMI confirms the applicability of the methods. Other aspects of interest for future research include sliding mode control and state observers with disturbance observation.

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