MODEL BASED DESIGN METHODS FOR SPEED CONTROL APPLICATIONS

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Zsuzsa Preitl, Eng.

Superviser: Univ. Prof. Dr. Eng. Radu-Emil Precup

Opponents: Univ. Prof. Dr. Eng. Sergiu Caraman
Univ. Prof. Dr. Eng. Clement Feştilă
Univ. Prof. Dr. Eng. Toma-Leonida Dragomir

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România, 300159 Timişoara, Bd. Republicii 9,
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Preface

This thesis represents a synthesis of the research results of the author in the time period 2002-2007 in the frame of PhD studies at the "Politehnica" University of Timişoara. During this period I have the chance to be a member of different research groups from "Politehnica" University of Timişoara as well as from Budapest University of Technology and Economics, Hungary.

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To my parents.

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Cuvinte cheie:
Structuri şi algoritmi de reglare automată, proiectare bazată pe model, reglare GPC, reglare Fuzzy, reglarea turaţiei, sisteme de acţionare, hidrogeneratoare

Rezumat:
Teza de doctorat este orientată spre dezvoltarea de noi soluţii de reglare şi metode de proiectare a regulatoarelor destinate reglării turaţiei. Clasa de procese vizate sunt: - sisteme de acţionare electrică; - hidrogeneratoare. Metodele de reglare propuse sunt: Metoda dublei parametrizări a criteriului Optimului Simetric (2p-SO-m). Reglare GPC in cascada cu buclă interna minmax. Reglare Fuzzy cu regulator Takagi-Sugeno (TS-FC) cu patru intrari si două ieşiri, proiectat pe principiul minimizării funcţiilor de sensitivitate. Reglare Fuzzy cu regulator Mamdani dezvoltat in domeniul delta.
Soluţiile propuse au fost testate prin simulare pe modele de procese cu date numerice reale si pe modele de laborator.

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Index of Abbreviations

Abbreviation	Meaning
DOF	Degree of Freedom
1-DOF	One Degree of Freedom
2-DOF	Two-Degrees of Freedom
CS	Control System, Control Structure
CCS	Cascade Control Structure (System, solution)
MIMO	Multi-Input Multi Output (system)
SISO	Single-Input Single Output (system)
CAD	Computer Aided Design
MBC	Model Based Control
IMC	Internal Model Control,
MPC	Model-Predictive-Control
GPC	General Predictive Control
DB	Dead-Beat Control
FC	Fuzzy control (controller)
TS-FC	Takagi-Sugeno Fuzzy Controller
t.f.	transfer function
f.r.f.	frequency response function
F-r, F-y	Input filters regarding to reference channel or to feedback channel
MM	Mathematical Model
EM	Electric Machine
DC-m	Direct Current motor
BLDC-m	Brushless Direct Current motor
NEDC	New European Driving Cycle (test cycle)
EG	Electric Generator
MO-m	Modulus Optimum method
SO-m	Symmetrical Optimum method
E-SO-m	Extended Symmetrical Optimum method
2p-SO-m	double parameterization (2p) of the Symmetrical Optimum method (SO-m)
HT	Hydraulic-turbine
HG	Hydrogenerator
SG	Synchronous Generator
PS	Power System
HTPS, PsS	Hydro-turbine and Penstock (penstock system)
HTG	Hydro-Turbine and Generator
C	Controller
P	Plant
DL1	Derivative with first order Lag type (filter, subsystem, model)
PL1	Proportional with first order Lag type (filter, subsystem, model)
PL2	Proportional with second order Lag (subsystem, model)
PL3	Proportional with third order Lag
PI(D)	Controller type: P-proportional, I-integrative, D-derivative
PDL1	Proportional Derivative with first order Lag

ISE	Integration of Square Error cost function
MCARE	Modified Control Algebraic Riccatti Equation
EHS	Electro-Hydraulic System
EHC	Electro-Hydraulic Converter
SVD	Slide-Valve Distributor
MSM	Main Servo-Motor
R, S, T	Polynomials in 2-DOF representation (RST- structure)
RST-representation	Polynomial representation of 2-DOF controller
LQ	Linear Quadratic (optimization method)
NFS	Non-minimum phase systems
FC-S	Fuzzy Control System
B-FC	nonlinear fuzzy-block B-FC, Takagi-Sugeno fuzzy system
PI-C-r	PI-controller with optimized parameters regarding the CS reference
PI-C-d	PI-controller with optimized parameters regarding the CS load disturbance
2-DOF FC	Two Degree of Freedom Fuzzy Controller
RST	Two-degree-of-freedom (RST) polynomials
Q-C	quasi-continuous
p-t.f.	pseudo-transfer function
PI(PID)-FC	quasi-PI (PID) fuzzy controller
PI-FC-OI	quasi-PI fuzzy controller with output integration
PI-FC-II	quasi-PI fuzzy controller with input integration
RB	Rule base
MF	Membership Function
LTs	Linguistic Terms
LVs	Linguistic Variables
ZOH	Zero-Order-Hold element/block

Index of notations

Notation	Meaning
$H_{x,y}(s)$	transfer function (t.f.), where $_{x,y}$ – dedicated indices
$H_{x,y}(j\omega)$	frequency function (f.r.f.)
$A(s), B(s)$; $P(s), Q(s)$	Polynomials in a rational t.f. form: - for the plant; - for the controller
$\underline{A}, \underline{B}, \underline{C}, \underline{L}$	matrices in a state-feedback MM (underlining can be omitted)
$S(s)$	sensitivity function
$T(s)$	complementary sensitivity function
$G(s)$, $N(s)$, $M(s)$, $Q(s)$, $X(s)$, $Y(s)$	Rational forms, polynomial representation (parameterization) (see Youla parameterization)
$L(s), H_0(s)$	the open loop transfer function
$H_r(s)$	Closed loop t.f. regarding to the reference input
$H_{d1}(s), H_{d2}(s)$	Closed loop t.f. regarding to the disturbance input
$L_0(s)$, $H_{r0}(s)$, $S_0(s)$	Optimized (with index 0) expression for the mentioned t.f.
$H_c(s), C(s)$	t.f. of the controller
$k_c, k_C; T_c, T_c'$ T_f, T_i, T_d	Controller parameters
$H_p(s), P(s)$	t.f. of the plant
T, T_1, T_2, T_k, τ	time constants (in general, of a plant, of a subsystem, ...; indices can be associated)) [sec]
τ	(also) the delta-transformation zero [sec]
T_{Σ},	equivalent time constant (sum of small time constants) [sec]
T_m	time delay, dead-time constant; also mechanical time constant [sec]
$M_{r(p)}(\omega) = \lvert H_r(j\omega)\rvert$	the magnitude function of the f.r.f. regarding to the reference signal
$M_{d1,d2}(j\omega) = \lvert H_{d1,d2}(j\omega)\rvert$	the magnitude function of the f.r.f. regarding to the disturbance signal
$Z\{\ \}$	symbol for the Z transform
$r(t)$	reference signal
$u(t), y(t), \underline{x}(t)$	general notations for system input, output and state
$e(t)$	control error (the error signal)
u, u_c	control signal, command from the controller; some particular notations are also used (for example u_{CE}, $u_{C\omega}$ ...)
$d(t)$, $d_x(t)$	disturbance (index x can be associated)
$y(t)$	measured output
$z(t)$	controlled output
k_M	measurement equipments' gain (with a supplementary index)
k_{awr}	gain of Anti-Windup-Reset block

T_A , T_E	time constant of an actuator (A, E) [sec]
T_a	electrical time constant [sec]
u_a	armature voltage [V]
k_a, k_A	actuator gain
u, u_c	voltage, command voltage [V]
L_a	Inductance [H]
R_a	Resistance [Ω]
i, i_a	current, field current [A]
e , e_m	(counter) electromotive voltage [V]
k_e	electromotive voltage coefficient [V/rad/sec]
k_e	current-torque coefficient [Nm/A]
k_f	friction coefficient [Nm/rad/sec]
ω	(angular) speed, [rad/sec], [sec^{-1}]
ω_v	the speed of the drive shaft and wheel[rad/sec], [sec^{-1}]
J_m	moment of inertia of the motor [kg m^2]
J_{veh}	moment of inertia of the vehicle reduced to the motor axis [kg m^2]
J_w	moment of inertia of the two driven wheels reduced to motor axis (converted) [kg m^2]
J_{tot}	total moment of inertia of the plant [kg m^2]
M_a , m_a, Δm_a	active torque [Nm]
M_s , m_s, Δm_s	load torque (the notation M_d or M_{load} will be also used) [Nm]
M_f	friction torque [Nm]
w_r, F_d , A_a, C_d, M_d ,	parameters in vehicle dynamics (Part I, rel.(2.2-1))
v	linear velocity of vehicle;. [m/sec]
m_{tot}	the total mass of the vehicle (lower and an upper limit, $m_{tot, min}$ and $m_{tot, max}$.) [kg]
g	gravity acceleration, g =9.81 [m/sec^2]
r, w	the wheel radius (in the first application) [m]
A_d	frontal area of vehicle[m^2]
C_d	air drag coefficient
C_r	rolling resistance coefficient
ρ	air / water density[kg/m^3]
f_r	drive ratio
P	power, (generally, in particular mechanical or electrical power) [W]
η	efficiency
H , Δh	the water-fall (in the second application) [m], [p.u.]
Q, Δq,	Water flow [m^3/sec]
$y(t)$, $\Delta y(t)$	position of the electro-hydraulic actuator (part I) [m]
p_G	active power (of the generator) [W]
q_G	reactive power (of the generator) [VAr]
u_G	armature voltage (of the generator) [V]
T_w , T_L	the water time constant, the reflection time constant[sec]
a_m	the network self-control coefficient
$g_{\ddot{o}}$	electro-hydraulic converter's gain Part I, fig.3.2.3
σ_1	overshoot of a CS
t_1	first settling time

t_s	settling time
γ_n	static coefficient
φ_m, φ_r	phase margin (phase reserve);
ω_c	crossover frequency
$M_{p\,max}=\max\lvert T(j\omega)\rvert$	maximum magnitude of the frequency response
$M_{s\,max}=\max\lvert S(j\omega)\rvert$	maximum value of the loop sensitivity function
$t_{s(d1,d2)}$	the settling time regarding to the disturbance
$m=T_\Sigma/T_1$	specific parameter in 2p-SO-method
β	specific parameter in ESO-m and 2p-SO-methods
-20 (-40) dB/dec.	the slope of the Bode diagram
φ	the set of all bounded rational forms with real coefficients
J	Integral Cost Function
N_1 , N_2	limits of the prediction horizon,
N_u	the control horizon
$\hat{y}(t+j\mid t)$	the j-step ahead prediction of the output
$r(t+j)$	the future reference trajectory
$\delta(j)$, $\lambda(j)$	weighting sequences
q^{-1}	the shift operator
K, K_u, K_d,	Feedback gain matrix and its components
ρ , γ	design parameter in Modified Control Algebraic Riccatti Equation
$K_p^{r(d)}, K_I^{r(d)}$	Discretized value for parameters of the PI-C-(r, d) controllers
h, T_e	sampling period
S_e, $S_{\Delta e}$, $S_{\Delta r}$, S_s	Parameters which characterize membership functions
$\Delta r_k=r_k-r_{k-1}$	increment for the reference input
$\Delta e_k=e_k-e_{k-1}$	increment for the error signal
$\Delta u_k=u_k-u_{k-1}$	increment for the control signal
$a_1,a_2,a_3,a_4,$	Parameters for computing Δu_k. in TS-FC (part III-chapter 4)
S_s	Parameter for computing s_{k-1} in TS-FC (part III-chapter 4)
ZE, PS, PM, PB, NS, NM, NB	The names of the membership function
$F(\gamma)=T\{f(t)\}$	generalized delta transform of a function:
γ or δ	the variable associated to the delta operator
$H(\gamma)$	the delta t.f.
$H_{C\,PI}(\gamma)$, $H_{C\,PI}(\gamma)$	t.f.s for the delta PI and discreet PI controllers
$H_{C\,DB}(\gamma)$	t.f.s for the delta DB controllers
$B^+(\gamma),B^-(\gamma)$	Decomposition of $B^+(\gamma)$ into cancelling zeros $B^+(\gamma)$ and non-cancelling zeros $B^-(\gamma)$
$C_{SM}(\gamma)$, $C_{SM}(z)$	Controller with included Smith-predictor (in δ and z domain)
$H_m(z)=P_m(z)$ $A_m(z), B_m(z)$	The reference model's t.f. and its polynomials
$A_0(z)$	Observer polynomial
$\partial\{S.R,T\}$	Degree of polynomials
$<\Delta r_k, \Delta^2 r_k>$	the phase plane for two variable
B_e, $B_{\Delta e}$, $B_{\Delta u}$	tuning parameters (in FC-s)

Part I. Introduction. Controlled applications

"You see things, and you say: 'Why?' But I dream things that never were and I say 'Why not?'" (George Bernard Shaw)

1. A short overview of the Thesis

1.1. Presentation of the Thesis

The PhD thesis is oriented on developing new control design methods and control structures dedicated to speed control in two domains: electrical driving systems and hydrogenerators (HG). The thesis is finalized by presenting the design methods and control structures, and refers to two applications.

Some words about the title. In this thesis the term of **Model Based design** is used to refer to control solutions where the model of the plant is connected strictly to the controller design and sometimes directly to the control algorithm.

Generally, the term of **Model Based Control** (MBC) is used to refer to control solutions that explicitly embed a plant model in the control algorithm. In particular algorithms, such as Internal Model Control, Model Predictive Control, Inferential Control (control using secondary measurements) and Smith predictor based-control, includes in the algorithm the designer's knowledge about the plant in form of a model and they are generically included in the category of Model Based Control systems.

Even the classical task of design (tuning) a PID controller is based on model (model based), much better control system performances can be obtained if the model of the plant is actually obtained and used [I-93].

The current interest in the research topic is reflected by the large amount of publications in the domain: Journals (Automatica, IEEE a.o.), Congresses (IFAC Congresses) Conferences (Control Design, Control Applications of Optimization a.o.), Symposia and Workshops, research reports, PhD theses. I had the chance to take part on some of them with presented and published papers (see the reference list).

This thesis has five parts that are addressing control problems focused on development of two control applications in speed control. The presented thesis has a length of 192 pages and is based on 207 cited works. Out of these I contributed to elaborating 22 papers out of which 12 as first or single author, and the rest as member of the research team. The own paper reference list include also the Diploma Work (2002), the Master Thesis (2003) and the three PhD Reports (2006, 2007). The references are marked as follows: with a general numbering (column 1) from [1] to [207] and referred in each Part under a Part reference number; for example reference [6] is referred in parts I, II and IV under numbers [I-6], [II-13], [IV-12], (see the reference list).

Part I, entitled **Introduction. Controlled applications,** presents first (chapter 1) a synthesis of the contributions and the Acknowledgements, followed by the mathematical modelling of the applications (chapters 2 and 3). The applications are an electrical driving system for an electric vehicle and control solutions for speed control for a hydro generator.

Part II, entitled **Controller design for ensuring good reference tracking and disturbance rejection performances using PID controllers,** presents a new development method for PID controllers. The current interest in the presented design method is due to the fact that over 90% of industrial applications are PID controllers, so any improvement of the design methods is welcome and easily accepted in industry [I-4]. The new method is based on a double parameterization in frequency domain, in order to obtain good performance of the system both in reference tracking and load disturbance rejection. The application is focused on an electrical driving system for an electric vehicle, and uses a cascade control solution (CCS). An appendix presents details about the connection between one Degree Of Freedom (1-DOF) and two Degrees Of Freedom (2-DOF) controllers and a design method for 2-DOF controllers.

The third part, entitled **New Speed Control solutions for hydrogenerators,** deals with two control solutions for speed control for a hydrogenerator (HG) with medium water fall, using combined control schemes and development strategies:

- The first proposed solution is based on a cascade control structure (CCS) with an internal minimax controller (to reject internally located deterministic disturbances) and a main General Predictive Control (GPC) loop (to reject external stochastic disturbances induced by the power system (PS) [I-76], [I-77];
- The second structure is a fuzzy control (FC) solution dedicated to the speed control of hydro-turbine-generators (HTG) [III-15]. In the first phase based on the model of the plant a conventional PI controllers are developed ensuring desired maximum values for both the sensitivity function and the complementary sensitivity function in frequency domain. Further, by accepting the approximate equivalence between fuzzy controllers (FC) and the linear ones (in certain conditions), a design method for a four inputs-two outputs Takagi-Sugeno Fuzzy Controller (TS-FC) is given.

Based on research results in delta domain defined by Middleton, R.H. and Goodwin, G.C. [IV-1], [IV-2] and synthesized in an unified system-theory approach [IV-3], the Part IV entitled **Development of Fuzzy Controllers in delta domain,** presents a delta approach in design of fuzzy control solutions [IV-21]. The new design method resulting in Mamdani PI-fuzzy controllers is validated by real-time experiments in controlling a servo system with nonlinearities and by digital simulation.

Part V, entitled **Contributions** synthesizes the contributions and possible further research directions and topics.

The **Appendices** treat research details which are not included in the main parts II, III and IV of the thesis, but are in strong connection with them.

1.2. Contributions of the Thesis. A short overview

The contributions of the thesis are spread in each part of it. Table 1.2-1 synthesizes them. The contributions are highlighted in detail in the end of each part and finaly, in Part V of the thesis.

Table 1.2-1 Contributions of the thesis (a short overview)

Part	Chapter	Paragraph	Contributions	Papers
0	1	2	3	4
I	2.	2	A synthesis regarding modelling of an electric driving system used in electrical traction with DC motors and BLDC motors, oriented towards CS design	[I-19], [I-20]
	3.	3	A synthesis of the Mathematical Models (MMs) of subsystems appearing in the structure of a speed control of a HG. The models are oriented towards controller development	[I-89], [I-88] (3rd PhD report) (2nd PhD report)
II	2	2.2 2.3	A short overview on optimal design methods based on Modulus Optimum criteria, detailing the MO method, SO method and ESO method	[I-87] (1st PhD report), [II-95]
	3	3.1 3.2 3.3	A novel controller design method based on a double parameterization of the optimality conditions specific for the SO method. Comparative simulations allow a good view of the cases when the method proves to be efficient	[I-6], [II-21], [II-95], [I-87] (1st PhD report)
	3	3.4	A Youla parameterization approach of the MO-m, ESO-m and 2p-SO-method	[II-61], [II-95]
	Appendix 1		2-DOF approach for PI and PID controllers and a design method which can easy applied in practice	[II-70], [I-77], [IV-35]
III	2.	2.2	A synthesis upon recent research results for cascade control structures (CCS) based on different design methods	[I-88], [I-89] (2nd PhD report) (3rd PhD report)
	3	3.2-3.5	A new two-stage CCS with an internal minimax state controller dedicated for rejecting internally located deterministic disturbances and a main GPC loop. The use of the GPC controller under IMC representation based on the GPC's polynomial RST structure has the advantage of easy implementation. The control solution is applied to the speed control of HGs	[III-26],[I-88] (2nd PhD report) [I-88] (3rd PhD report) [II-17]
	4	4.2-4.4	A new FC development solution dedicated to the speed control of HG. The contribution contains a four inputs-two outputs TS-FC, developed by starting with the design of two sets of conventional PI controllers ensuring	[III-15],[I-89] (3rd PhD report)

			the desired maximum values of the sensitivity function and of the complementary sensitivity function in the frequency domain	
	Appendix 2		The IMC equivalent of GPC structure was derived. From an applicative point of view the problem of constraint handling of the control signal with different AWR measures for quick leaving of the saturation zone is also dealt with for the GPC and IMC structures	[III-33], [I-88] (2nd PhD report) [I-89] (3rd PhD report)
IV	2	2.1	A short synthesis upon the advantages of using the δ (delta) transform at the implementation of digital control algorithms	[IV-6], [IV-9]
	2	2.3.2	A study regarding controller design in δ domain: - PI, PID controller design based on MO-, SO- and 2p-SO-methods, highlighting the advantages of the implementation; sensitivity analysis - DB controller design	[IV-5], [IV-6], [IV-9]
	2	2.3.3 2.3.4	IMC-based Smith predictor for plants with dead time. A new approach to control system design based on the IMC in delta domain, with a mixed representation of the plant model within the IMC controller (hybrid architecture). The method is based on the dual representation in delta and Z discrete domain of the plant	[IV-7], [IV-8], [IV-16], [I-88] (2nd PhD report)
	3	3.3 3.4	Based on the delta representation of the systems, a new design method dedicated to a Mamdani type PI FC for benchmark-type plants model is proposed. The design is based on the simplicity and transparency of the method having in view a relatively simple implementation.	[IV-21], [I-89] (3rd PhD report)
	Appendix 3		A method is presented for the development of 2-DOF FCs. The method is easy to understand and to implement as CAD development.	[IV-22], [IV-35], [IV-45], [IV-46], [IV-47]

1.3. Acknowledgements

This thesis represents a synthesis of the research results of the author in the time period 2002-2007 in the frame of PhD studies at the "Politehnica" University of Timişoara. During this period I have the chance to be a member of different research groups from "Politehnica" University of Timişoara as well as from Budapest University of Technology and Economics, Hungary.

I am grateful for the chance for being involved in many different areas of Control Engineering which had a big influence on my professional development, even if much of these results are not part of the thesis. I am grateful for all the people I had contact with, for helping me to get here.

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Last but not least, I would like to warmly thank my Parents for all.

2. Speed control of an electric driving system

2.1. General aspects

The speed control and positioning applications are present in all industrial and domestic areas; the permanent development of control techniques and technologies ensures new solutions. From the multitude of driving system applications the thesis focuses on those where the reference is continuously changing, permanent load is acting and system parameters can change, for example due to changes of the moment of inertia. This happens for example in case of electrical traction systems.

An electric driving system can be represented through a relatively simple block diagram, characterized by reduced number of parameters and variables. For electrical traction systems the different subsystems have different dynamics. The electric machines used in such applications are mainly DC-motors, Asynchronous motors and Synchronous motors. The according Mathematical Models (MM) can be of different complexity [I-10], [I-22], [I-24]. The block diagram of an electric drive is depictured in figure 2.1-1. The "load" is specific to each application affecting the load torque, $m_s(t)$; for example, [I-13] [I-14], [I-21], [I-29], [I-30], [I-42]. The indicated papers treat the driving system of a control oriented point of view.

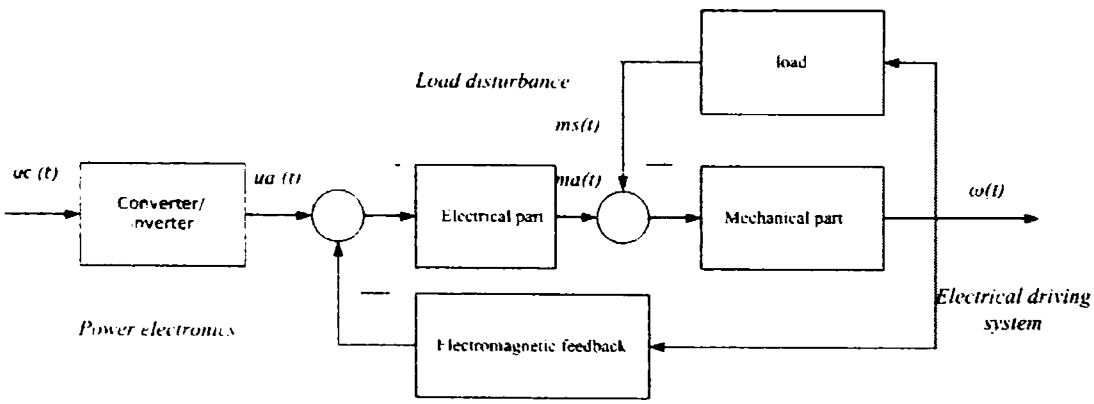


Fig. 2.1-1.The structure of an electromechanical driving system

The testing of a driving system is a standard procedure using constant (persistent) and non-persistent inputs (reference and disturbance). The disturbances can occur due to internal or external factors [I-1]-[I-4], [I-24]. Two categories are taken into account here:

- Load disturbances, d(t) whose rejection is task of the control system;
- Changes in parameters due to internal or external causes.

The thesis analyses the interaction {electrical driving motor - load} for a hybrid electrical vehicle application using a relatively simple model for the driving system, widely accepted in practice.

2.2. Modelling of an electric traction system

2.2.1. General structure of the electric traction system

The traction for an electric vehicle consists in the electric driving system [I-14], [I-17], [I-22]. For these vehicles the energy sources can be various [I-14], [I-17]:

- Pure electrical sources, based on batteries;
- Hybrid primary energy sources with different structure and components.

The main part (the drive engine) is an electric motor (EM) which drives the wheels; it can be a DC-motor (DC-m), with brushes or brushless (with permanent magnets). Both variants are accompanied by dedicated control and power electronics. The electric machine can work as a motor (traction) or as a generator (during the regenerative braking regime).

In case of vehicles with hybrid primary energy sources, the electrical energy for the EM can be delivered (for example) by the battery and by an electric generator (EG) [I-13], [I-14], [I-31], [I-32], [I-33].

The functional block diagram of an electrical driving system as part of an electric vehicle is presented in figure 2.2-1.

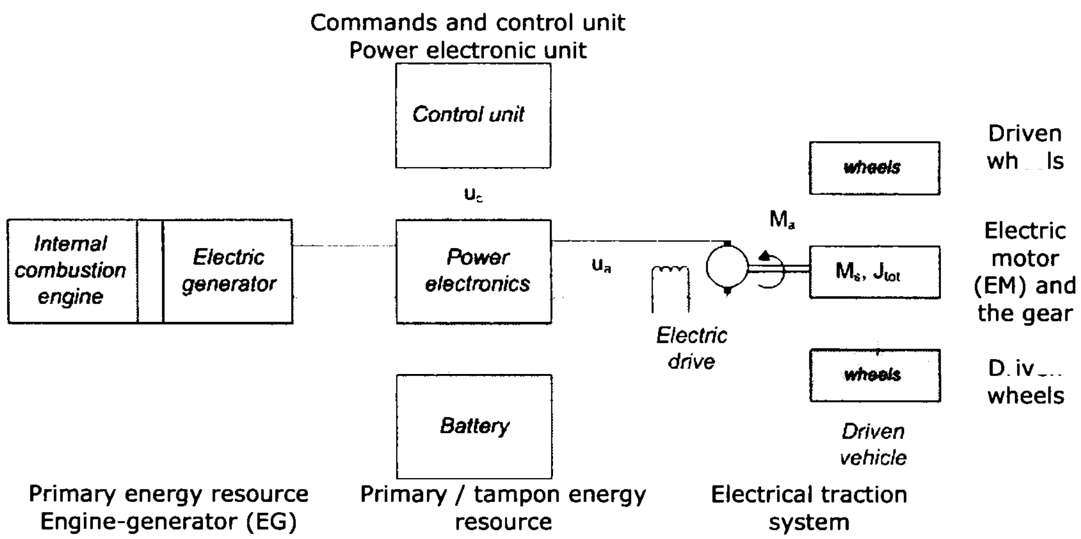


Fig. 2.2-1. Basic diagram of an electric vehicle

The main components of such a traction system are:

- the EM, a DC-m, which drives the wheels and whose control is dealt with in part II chapter 4;
- the energy resources' which deliver electrical energy for the EM: - the battery as a buffer for the energy and the electric generator;
- the power electronics and the control unit.

In hybrid applications, the EG is in rigid connection with the internal combustion engine.

An attractive alternative for the DC motors in case of electric vehicle driving systems are the Brushless DC machines (BLDC-m) [I-14], [I-25], [I-30], [I-31], [I-32], [I-33], [I-34]. They can function both in motor and generator regimes. It must be mentioned that the BLDC-m is in fact the combination of a permanently excited synchronous motor and a frequency inverter, where the inverter „replaces" the converter of a classical DC motor [I-17], [I-33], [I-34] [I-35], [I-43]. BLDCs with inverter are mainly used in high performance electric drives with variable speed, where these values largely outrun the nominal rotation velocity.

The advantages and disadvantages of each machine can be found in the literature [I-14], [I-31], [I-33]. In the considered application a DC-m is used, the control structure and the controller will be designed to the model of a separately excited DC motor. This consideration does not restrict the application, since the BLDC motor can be represented through an equivalent electric scheme as the separately excited DC-m (see paragraph B, [I-25], [I-39], [I-40], [I-43], [I-44]).

The testing of the model through simulation requires a given reference that must be followed. Such reference signals, consisting in a pre-defined "time-vehicle velocity" scheme, are called drive-cycles. An example for drive cycles is the New European Drive Cycle (NEDC), figure 2.2-2 [I-14].

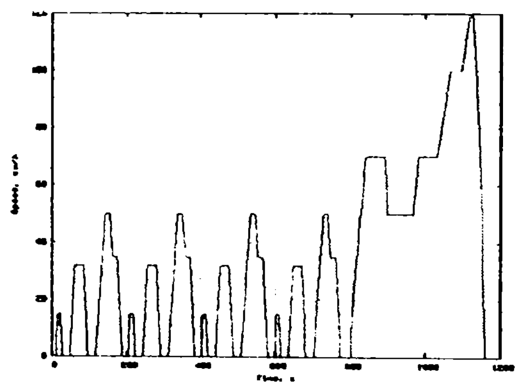


Fig.2.2-2. Test Driving Cycle: the NEDC; time [s] on horizontal and velocity [km/h] on vertical axis

2.2.2. Simplified models of the traction system

Regarding the system in figure 2.2-1, the electric drive and the "load" will be modelled separately. The modelling will refer to a separately excited DC motor drive. Some remarks will also be given regarding the BLDC-m.

A. Electrical drive with DC-m

• **Vehicle dynamics.** The basic relations which describe the driven system consist of the simple longitudinal dynamics of the vehicle [I-41], [I-42], [I-43]. According to Newton's second law, they are as follows:

$$\omega(t) = \frac{f_r}{w_r} F_d(t)$$

$$M_s(t) = \frac{w_r}{f_r} F_d(t) \qquad (2.2\text{-}1)$$

$$F_d(t) = m\dot{v}(t) + \frac{1}{2}\rho v^2(t) A_d C_d + m g (C_r + \sin(\gamma(t)))$$

where ω is the angular velocity and M_s is the torque required from the EM (the load). The velocity v(t) is given in the specified drive cycles, the acceleration ($\dot{v}(t)$) can be simply calculated from it; $\gamma(t)$ is the ramp angle, which will be considered for simplicity equal zero (0^0). The parameters are definied in Part II chapter 4 with concrete numerical values.

The vehicle is modelled considering rolling, hill climbing resistance and aerodynamic drag. This equation gives the required drive force, from which the drive moment can be determined considering wheel radius and velocity.

- **Driving system with DC-m** [I-23], [I-45]. The hypotheses accepted at modelling imply that in normal regimes the DC-m works in the linear domain where the flux (current) is constant in value. An eventual change in the excitation regime will modify the basic model, but a linearization in the new working point results in the basic situation.

The basic equations that characterize the functionality of the system are given in (2.2-2):

$$\begin{gathered} T_A \dot{u}_a + u_a = k_A u_c \\ L_a di_a / dt + R_a i_a = u_a - e \\ T_a = L_a / R_a \quad , \quad e = k_e \omega \\ M_a = k_m i_a \\ J_{tot} \dot{\omega} = M_a - M_s - M_f \\ J_{tot} = J_m + J_{veh} + J_w \end{gathered} \tag{2.2-2}$$

where the following notations were used: T_A – time constant of the actuator (power electronics) [sec], u_a – armature voltage [V], k_a – actuator gain, u_c – command voltage from controller [V], L_a – inductance [H], T_a – electrical time constant, i_a – field current [A], e – counter electromotive voltage [V], k_e – coefficient [V/rad/sec], ω – rotor speed [rad/sec], J_{tot} – total moment of inertia of the plant [kg m^2], M_a – active torque [Nm], M_s – load torque [Nm] (the notation M_{load} will be also used), M_f – friction torque [Nm], J_m – moment of inertia of the DC-m [kg m^2], J_{veh} – moment of inertia of the vehicle reduced to the motor axis [kg m^2], J_w – moment of inertia of the two driven wheels reduced to motor axis [kg m^2].

Accepting that the total inertia of the system can change with max 25% regarded to the basic value $J_{tot\,0}$, wich corresponds to the vehicle without passengers, it results:

$$J_{tot} = J_{tot\,0} + \Delta J_t \qquad \text{with} \qquad \Delta J_t \le 0.25\, J_{tot\,0} \tag{2.2-3}$$

Based on (2.2-2) a block diagram of system can be built, se figure 2.2-3. The load torque M_s is generated by the vehicle dynamics.

Accepting for simplicity that T_A can be neglected, the derived (linearised) state-space equations of the DC-m are (2.2-4):

$$\begin{aligned} \dot{i}_a &= -\frac{R_a}{L_a} i_a + \frac{k_e}{L_a}\omega + k_A u_c \\ \dot{\omega} &= \frac{k_m}{J_{tot}} i_a - \frac{k_f}{J_{tot}}\omega - \frac{1}{J_{tot}} M_s \end{aligned} \quad , \quad \begin{aligned} i_a &= i_a \\ \omega &= \omega \end{aligned} \tag{2.2-4}$$

Based on the block diagram given in fig.2.2-3, the DC-ms' t.f.s can be computed $\{H_{\omega,\omega c}(s), H_{\omega,ms}(s), H_{ia,uc}(s), H_{ia,ms}(s)\}$.

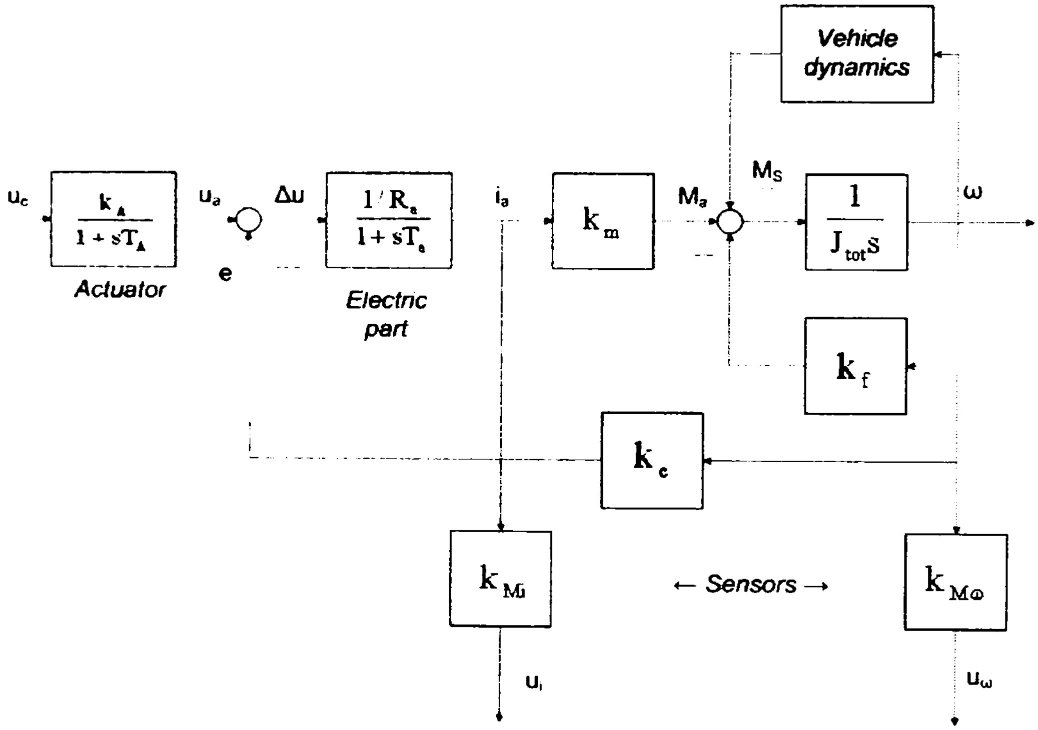


Fig.2.2-3. Block diagram of a separately excited DC-motor (in figure e -> em)

For two particular cases $k_f \neq 0$ and $k_f = 0$ (an approximation widely accepted in practice) the t.f. regarded to the control signal $H_{\omega,uc}(s)$ is explicited:

- the case $k_f \neq 0$ and T_A neglected:

$$H_{\omega,uc}(s) = \frac{\omega(s)}{u_c(s)} = k_A \frac{1/k_e}{(1+\frac{R_a k_f}{k_m k_e}) + s[T_a \frac{R_a k_f}{k_m k_e} + T_m] + s^2 T_a T_m} \quad \text{where (2.2-5)}$$

$$T_m = \frac{J_{tot} R_a}{k_m k_e}$$ - is the mechanical time constant of the plant.

- the case $k_f = 0$, (T_A is not neglected):

$$H_{\omega,uc}(s) = \frac{k_A}{1+sT_A} \frac{1/k_e}{1+sT_m + s^2 T_a T_m} \quad (2.2\text{-}6)$$

In case of electric traction applications $T_m \gg T_a$ and (2.2-6) can be rewritten as

$$H_{\omega,uc}(s) \gg \frac{k_A}{1+sT_A} \frac{1/k_e}{(1+sT_a)(1+sT_m)} \quad (2.2\text{-}7)$$

This form corresponds to a second order with lag benchmark type model. Numerical values for the application are given in Part II, chapter 4.

B. Electrical drive with BLDC-m

The alternative to DC-m is the drive with BLDC-m with concentrated coils and bipolar supply (functioning in reversible regime) [I-2], [I-39], [I-40]. Such drives can be modelled with a MM equivalent with a DC-m and represented through block diagrams like figure 2.2-4 ([I-25], [I-30], [I-40], [I-45]).

- **Vehicle dynamics.** The mechanical part of the plant is the same as the case of DC-m and with dynamics given by relations (2.2-1).
- **Driving system with BLDC-m**. In practice the most common BLDC-m is characterized by three phase coils and associated drive electronics [I-25], [I-39], [I-40], [I-45]. The MM of the BLDC-m can be determined based on its block diagram. Papers [I-25], [I-36], [I-30], [I-37], [I-40], [I-45] present the equivalence from a mathematical modeling point of view between the BLDC-m and DC-m drive systems.

The modelling of the other functional blocks of an electric vehicle is not subject of this thesis.

2.2.3. Operating regimes

The speed control for an electric traction system (application considered in Part II, chapter 4) must satisfy multiple requirements:

- Depending of the traffic conditions, the reference of the system is permanently changing (see for example the NEDC in figure 2.2-2); the velocity of the vehicle is correlated with the big time constants of the plant and the motor power;
- According to (2.2-1) the load disturbance is permanently present and changing, depending on the speed, traffic and weather conditions.
- According to (2.2-3), due to possible changes of the vehicle mass, the equivalent moment of inertia will change and through this the large time constant of the plant.

From these points of view the control and design method presented in Part II chapter 3 proves to be of actual interest.

- The aims of the CS applied to the DC-m can be grouped as follows:
  - To ensure good reference signal tracking (speed) with small settling time and small overshoot (good transients and zero-steady-state error at v=const. velocity).
  - To ensure load disturbance rejection due to modifications in the driving conditions.
  - To show reduced sensitivity to changes in the total inertia of the system:

$$J_{tot} = J_{tot\,0} + \Delta J_t \quad \text{with } \Delta J_t \le 0.25\, J_{tot\,0} \qquad (2.2\text{-}6)$$

- Control solutions adopted in the thesis. CCSs are among the simplest multivariable control schemes. In spite of their simplicity, the CCS can substantially improve the dynamics (see chapter 3). Based on this, two CCS are adopted in the thesis, both having two control loops:
  - One internal control loop of the current, consisting in a PI controller and Anti-Windup-Reset (AWR) measure.
  - One external control loop of rotor speed ω [rad/sec] with a PI controller in two variants: a classical one and a more complex CS where in the outer loop a correction block was added.

2.3. Conclusions

The mathematical modeling was aimed to obtain a relatively simple MM and block diagram based on linear(ised) dependences, which can be easily used for controller design. The model facilitates the application of the new design method proposed in chapter 3 of Part II. The basic relations that describe the vehicle dynamics consist of the simple longitudinal dynamics of the vehicle according to Newton's second law. The model allows CS performance analysis especially regarding modifications of J_{tot}.

3. Speed control of a hydrogenerator

3.1. General aspects

The plant consisting in a hydro-power plant (power generating system) [I-47]-[I-50] is presented in figure 3.1-1 (a) (using [I-58], with authors' acceptance). The plant consist in a hydroelectric dam, a penstock system, the servo system which controls the water flow through the turbine (acting the wicket gate), the turbine and the synchronous generator connected to the power system (PS). The plant is a MIMO system with interconnections between each input and each output, figure 3.1-1 (b).

In normal functioning regime the plant can be considered with minor (without) interconnections (decoupled) and so each transfer channel can be modeled independently [I-58].

A simple formula for approximating electric power produced in a hydroelectric plant is:

$$P = \eta \rho g H Q \tag{3.1-1}$$

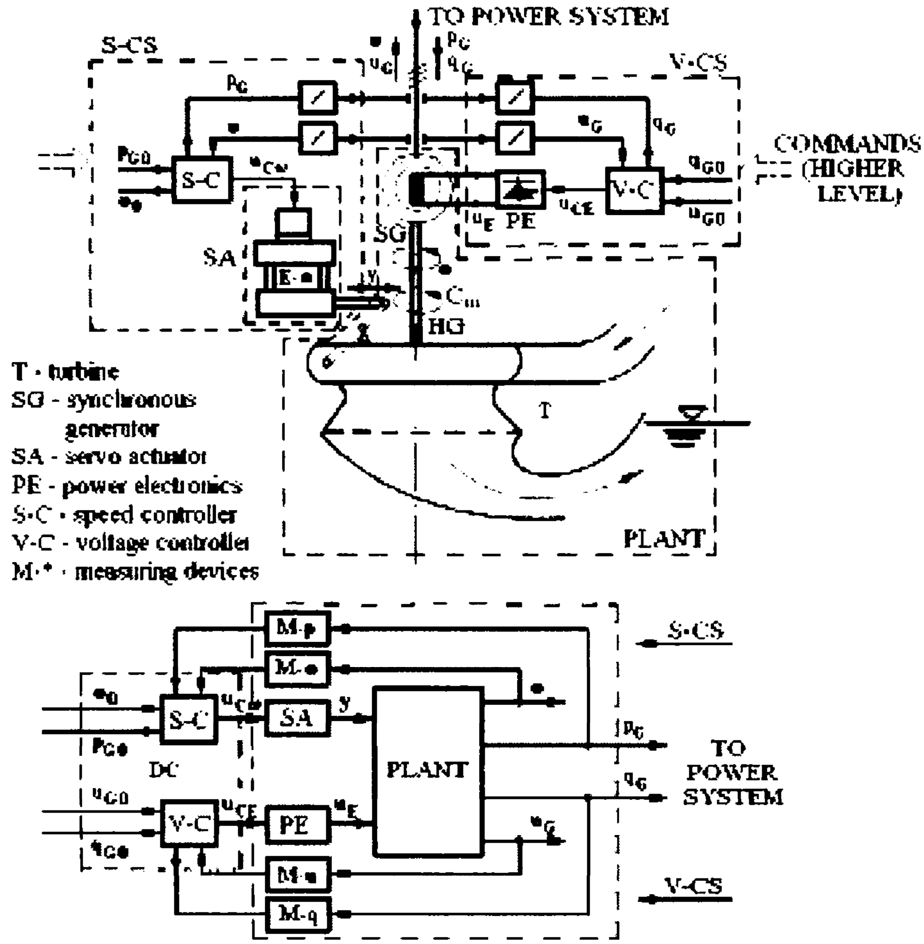
where P [W] power in watts, η the efficiency, ρ [kg/m^3] the water density [kg/m^3], H [m] the water-fall (the height), Q [m^3/sec] the flow rate, g=9.81 [m/sec^2] the gravity.

Two to basic control systems acting upon the HGs depicted in figure 3.1-1 (b), are [I-91]:

(1) The speed control of HG which ensures the active power transfer (p_G) from the HG to the PS controlling the frequency (speed of SG) in the PS, [I-61], channel $u_{C\omega} \rightarrow \omega$; in frame of this PhD thesis mainly this system will be approached. p_G represents the disturbance.

(2) The voltage control of SG which ensures the reactive power (q_G) transfer between the SG and PS, [I-61], channel $u_{CE} \rightarrow u_G$; q_G represents the disturbance.

Because of the important role the governors play in the control of the active power transfer to electrical PSs, a great deal of research effort has been to investigate the effects of CS, governor structure and parameter settings on the overall system performance: mainly stability and transient behavior after load disturbances [I-54].



(a)

(1)

(b)

(2)

Fig.3.1-1. Block scheme of a power generating system: (a) Main scheme; (b) Block diagrams regarding the CSs' structures

3.2. Modelling of the power generating systems' blocks

The part of the process which will be used for modeling in purpose of speed control is presented in 3.2-1 [I-48]-[I-51]. Under simplifying conditions [I-48], [I-50] the diagram can be divided into separate subsystems. The assumption that the angular speed ω represents the controlled output is valid in the conditions of a large power SG connected to a weak PS or of a SG connected to a local load (insulated regime) [I-50].

So, the active power plays the role of load disturbance:

$$P(t) = m_s(t)\omega(t) \tag{3.2-1}$$

and is presented in figure 3.2-1 by the load torque $m_s(t)$.

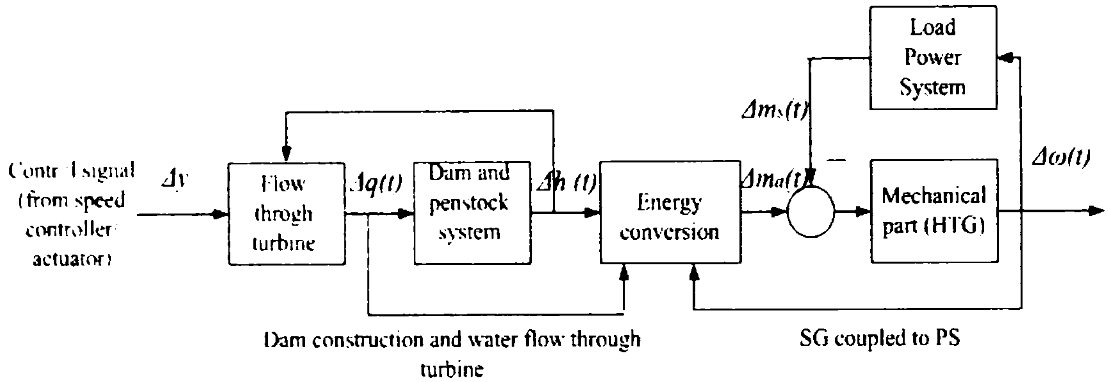


Fig.3.2-1. The part of the process which will be used for modeling in purpose of speed control

The changes in water flow through the turbine represent a second category of disturbances which must be taken into account by designing the control system.

For the development phase of the CS, the modelling of the subsystems and plant is treated thoroughly, from simple to very detailed models [I-47] – [I-53] and [I-58], [I-59] and [I-63] (for the actuator).

By taking into account the reference outlining that simplified MMs for the plant ensure within some limits acceptable conclusions from the point of view of primary control quality verification [I-51], the thesis presents in this chapter, simplified MMs of the main subsystems which occurs in the block diagram; they are frequently called in the design of control structures and in the first phase of their testing. The procedure is recommended in [I-51], [I-52], [I-53] and used in several references with satisfactory reported results ([I-48] - [I-57]).

3.2.1. Simplified models for the hydraulic subsystem and the synchronous generator coupled to the power system

Based on papers [I-47]-[I-59], table 3.2-1 presents a synthesis of the most frequently used t.f.s. the given models are relatively simple and are based on figure 3.2-2, where three subsystems can be delimited:

- the hydraulic subsystem,
- the synchronous generator and the load subsystem (particularly the power system).
- the actuator (electro hydraulic servosystem).

A. **The hydraulic subsystem**

The subsystem - known also as dam construction (including the penstock) and turbine – converts the energy of a hydronenergetic reservoir in mechanical energy necessary for acting the SG, and it is characterized by (3.2-5) or (3.2-6), where T_w - represent the water time constant, T_L- the reflexion time constant. In [I-50], [I-51] these parameters appear with values depending on the steady state operating point and on the characteristics of the hydraulic turbine.

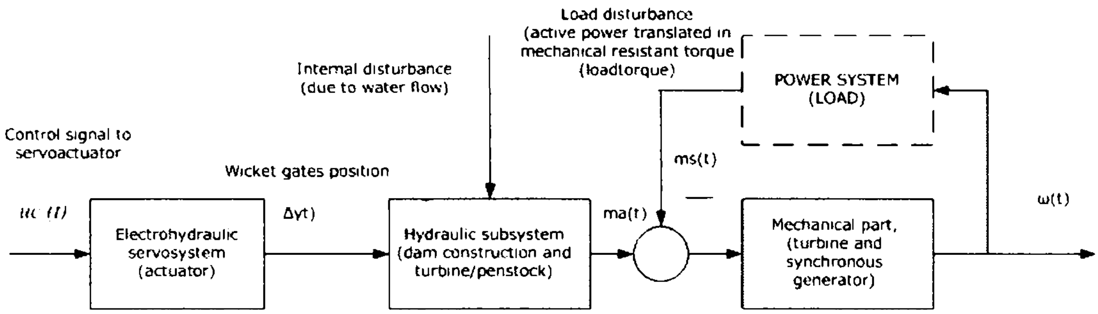


Fig.3.2-2 The simplified structure of the plant

***B.* The synchronous generator coupled to the power system**

The detailed mathematical model of the SG used in PS transient regime studies can be replaced also by reduced order models [I-48]-[I-53]. Based on the following simplifying assumption:

- linearization in the vicinity of a steady state operating point,
- giving up the supratransient phenomena,
- order reduction,

linear(ised) MM with different degrees of complexity can be accepted. A linearised model in form of first order with lag (PL1) (3.2-7) is mainly accepted for the development of the speed controller, [I-48]-[I-57]: T_m- represents the mechanical time constant of the HG, a_m - is the network self-control coefficient and represents a measure of the degree of connection of the SG to the PS. The value of a_m depends of the steady-state operating point; usually [I-48] - [I-53]:

- $a_m=0$ for idle running of the HG (pre-synchronizing regime);
- $0< a_m<1.3$ for the HG connected to the PS; small values a_m correspond to a HG operating in insulated regime on a local idle.

The value of a_m increases when the degree of connection to the PS increases.

3.2.2. Mathematical modeling of the servosystem (the actuator)

Based on [I-59], [I-59] and neglecting the minor nonlinearities in the servosystem, a linearised block diagram can be constructed, see figure 3.2-3. Using the notations from [I-58], the following state model is presented:

$$\dot{x}_1 = \frac{k_A g_0}{T_1} e_c$$
$$\dot{x}_2 = \frac{1}{T_2} x_1 \qquad (3.2\text{-}2)$$
$$\Delta y = x_2 \quad , \quad x_{m1} = k_{M1} x_1 \quad , \quad x_{m2} = k_{M2} x_2$$

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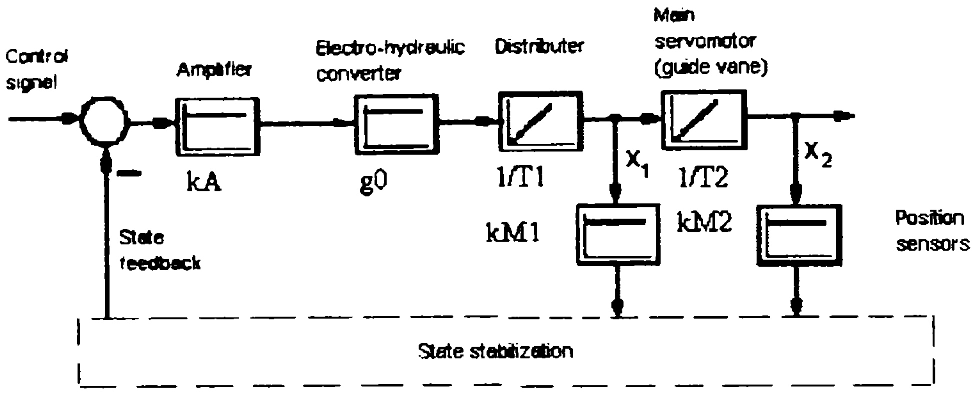


Fig.3.2-3. Linearised block diagram for the electro hydraulic servosystem (after [I-58])

The servosystem can be stabilized using different control design methods:

- pole placement,
- modern control systems which take into account also the disturbance acting on this level.

Part III of the thesis presents a new approach regarding the stabilization of the subsystem, the proposed CS is based on a special design of inner stabilizing and outer speed control loops. The disturbances acting on the plant are located in two places [I-51]:

- disturbance acting upon the actuator due to the water flow [I-48],[I-57];
- load disturbance due to the changes in power demand; they have an oscillatory non-persistent behaviour.

The disturbances will be approached in Part III of the thesis.

Table 3.2-1. Most frequently used t.f.s. for the plants' subsystems

No	Block	Mathematical Model	Remarks
1	The servosystem $H_{\Delta m \Delta y}(s) = \dfrac{\Delta y(s)}{\Delta u_c(s)}$	$\dot{x}_1 = \dfrac{k_A g_0}{T_1} e_c,\ \dot{x}_2 = \dfrac{1}{T_2} x_1,\ \Delta y = x_2$ (3.2-2)	Model for the unstabilised servosystem
		$H_S(s) = \dfrac{k_S}{1 + sT_s}$ (3.2-3)	PL1 type model for the stabilized servosystem
		$H_S(s) = \dfrac{k_S}{1 + 2\varsigma T_s s + T_s^2 s^2}$ (3.2-4)	PL2 type model for the stabilized servosystem
2	Penstock-turbine subsystem (hydraulic part) $H_{\Delta m \Delta y}(s) = \dfrac{\Delta m(s)}{\Delta y(s)}$	$\dfrac{1 - sT_w}{1 + sT_w/2}$ (3.2-5) T_w- the water time constant [I-51]	First order nonminimum-phase model
		$\dfrac{1 - sT_w + T_w^2 s^2}{1 + sT_w/2 + T_w^2 s^2}$ (a) or	Second order nonminimum-phase model

		$\frac{1 - sT_w + (2T_L/\pi)^2 s^2}{1 + sT_w/2 + (2T_L/\pi)^2 s^2}$ (b) (3.2-6) T_L- the reflection time constant [I-51]	
3	Turbine-SG $H_{\Delta m \Delta y}(s) = \frac{\Delta\omega(s)}{\Delta m(s)}$	$\frac{1}{\alpha_m + sT_m}$ (3.2-7) α_m - the network self-control coefficient	PL1 type model $0 < \alpha_m < 1.3$

Remark: The situation $\alpha_m = 0$ represents a special case which can be solved by frequency domain design.

The non minimum-phase systems (NFS) represent complex processes where it is often needed to use advanced control structures to ensure good CS performance.

The basic speed CS is presented in Fig. 3.2-4, where: r – reference input (speed setpoint), y – controlled output (SG speed), e=r−y – control error, u – control signal, d_1 –disturbance input, d_2 – load disturbance (sum of time-varying contributions from the PS).

The wicket gate acting servo, playing the role of the actuator, is not detailed in the CS structure because the actuator is usually a local control system that ensures fast dynamics included in the Hydro-turbine and Penstock (HTPS) dynamics, and deals also with the disturbance rejection acting at this level.

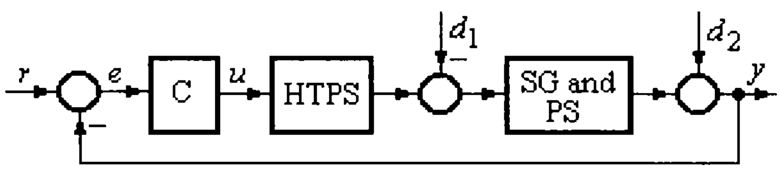


Fig.3.2-4. Speed control system structure

3.3. Conclusions

Based on the cited literature mathematical models for the HG system (with detailed subsystems) were presented, with the aim of later speed control design. The simplified mathematical model presented in table 3.2-1, are standardized and recommended by IEEE Working Group in IEEE report [I-53] and IEEE Committee in IEEE report [I-54].

Part II. Controller design for ensuring good reference tracking and disturbance rejection performances using PID controllers

"The PID controller can be said to be 'the bread and the butter' of the control engineering" (K.J. Åström [II-2])

Many optimal control synthesis methods result in controllers of high order related to the order of the plant (see for example in parragraph 2.3, the general case of the Symmetrical Optimum method (SO-m)). Applying the small time constant theorem, this order can be reduced reasonably. So, for many applications, for example the electric driving applications, the t.f. of the plant can be accepted in the design phase as a benchmark-type MM.

The speed control for an electric traction system (application considered in Part II, chapter 4) must satisfy multiple requirements:

- To ensure good reference signal tracking (speed) with small settling time and small overshoot (good transients and zero-steady-state error at v=const. velocity).
- To ensure load disturbance rejection due to modifications in the driving conditions.
- To show reduced sensitivity to changes in the total inertia of the system.

Some of the reference papers which deal with PI PID controller design ([II-2], [II-3], [II-9], [II-36], [II-78], [II-79]) highlight the fact that optimization methods based on Modulus criteria in frequency domain are still attractive. From this chategory the SO-m [II-24], [II-25] is frequently appealed in literature [II-2], [II-4], [II-9], [II-30], [II-38], [II-39].

In accordance with the application treated in the thesis – a driving system for an electric vehicle – this Part introduces a new design method based on a generalization of the SO-m using two parameterizations:

- One regarded to the criterion,
- Another regarded to the plant model.

1. PI, PID and two-degree-of-freedom (2-DOF) controllers

The number of degrees of freedom (DOF) of a controller is defined by the number of t.f.-s of parameters can be independently adjusted [II-1]. The design of a Control Structure (CS) is a multi-objective problem; in this context, by increasing the number of degrees of freedom from one degree of freedom (1-DOF) to two degrees of freedom (2-DOF), better system performances can be achieved. The main tasks in developing a CS are the following [II-2] - [II-6]:

- Design of a controller that should bring the process' behaviour asymptotically to the desired values determined by the reference;
- Design of a controller that should ensure the process lowest possible deviation from the desired behavior caused by disturbances (load disturbances);
- Design of a robust controller that ensures adequate behaviours when changes in the system parameters occur:
  - modifications in the values of the plant's parameters;
  - limitation in the capacity of the system to introduce the necessary energy to control the plants (systems with limited capacity).

The popularity of PI(D) controllers is due to the fact that despite of its simple structure, it ensures some important control functions such as: feedback, ability to eliminate steady state error through integral action and can anticipate the future through derivative action. Since the number of tuning parameters of a PI(D) controller is relatively low, the tuning techniques can be brought to transparent forms that are easy to use [II-6].

The future of PI(D) control and 2-DOF control [II-2], [II-8] - [II-14] consists in integrating this into advanced control strategies: Smith-predictor, Internal Model Control (IMC), nonlinear 2-DOF extension, fuzzy and neuro-fuzzy, adaptive PI(D), robust PI(D) control, gain scheduling, where the experience in using traditional controllers can be extended and the performance improved.

1.1 PI, PID and 2-DOF controllers. Plant models

The structures of PI-PID controllers can be grouped into four categories [II-3], [II-6], Table 1.1-1.

Table 1.1-1. The structures of PI-PID controllers

No	Usual denomination	Transfer function (t.f.) $H_c(s)$	Parameters	Rel.
0	1	2	3	4
1.	Classical ideal PI(D) controller structure (parallel realization / non-interacting structure)	$k_C(1+\frac{1}{sT_i}+sT_d)$ or $k_p+\frac{1}{s}k_i+sk_d$	k_p, k_i, k_d and k_C, T_i, T_d with $k_p=k_C, k_i=\frac{k_C}{T_i}$, $k_d=k_C T_d$	(1.1-1)
2.	Classical "serial" PI(D) controller structure, (cascade structure or interacting structure	$\frac{k_c}{s}(1+sT_{c1})(1+sT_{c2})$ $k_C(1+\frac{1}{sT_i})\frac{1+sT_d}{1+sT_f}$;	k_c, T_{c1}, T_{c2} k_C, T_i, T_d, T_f; $T_d=nT_f$ $k_c=k_C/T_i$, n>1	(1.1-2) (1.1-3)
3.	Non-interacting PI(D) controller structure (non-homogenous): - PI regard. Ref. - PID regard. feed-back	$u(s)=k_C(1+\frac{1}{sT_i})[e(t)-\frac{sT_d}{1+sT_d}y(s)]$ $u(s)=k_C\frac{1}{sT_i}[e(t)-\frac{sT_d}{1+sT_d}y(s)]$	See rel. (1.1-2), (1.1-3)	(1.1-4) (a) (1.1-4) (b)

4.	Two-degree-of-freedom (2-DOF) controllers	$u(s) = C_T(s)r(s) - C_S(s)y(s)$	C_T - the reference controller, C_S - the feedback controller,	(1.1-5)

Remarks: 1. The structures for controllers with non-homogenous information processing are detailed in [II-16]-[II-20]; they offer various sorts of t.f. regarded to the controller inputs.

2. For low order plants the 2-DOF controller can be regarded to a non-interacting - PI(D) controller; the parameters of $C_T(s)$ and $C_T(s)$ can be expressed as function of the PI(D) controller parameters (for example in [II-17] and appendix 2).

In order to avoid difficulties due to contradictory results obtained from design according to reference tracking and disturbance rejection, different "optimal" - or in special cases, "optimum-like" - tuning techniques can be adopted. Optimum-like tuning techniques are considered techniques which are derived from "optimisation relations" but which do not fulfil entirely the requirements imposed by this. Such techniques were developed by the author/in cooperation and will be presented in this work. Namely:

- The 2-ESO-m method, based on a double parameterization of the Symmetrical Optimum method (SO-m).
- A CAD development of SISO 2-DOF controllers. For low order benchmark-type plants relations between 2-DOF and PI(D) controller are done.

The developed tuning technique can be extended also in fuzzy-control domain [II-16], [II-19], [II-20] with various application-domain. Several analysis techniques can be regarded to these tuning techniques.

The exact model of the plant is very often difficult to be determined exactly; that is why PI(D) or 2-DOF control design methods and the performance verification of the CS are both based on benchmark type models. Some of these models are synthesized in [II-6] - [II-8] in form of:

- Low order lag systems without or with dead time;
- Multiple equal pole model;
- Right half plane zero model;
- Fast and slow models etc..

In case of linear continuous SISO systems the t.f. of the plant will be noted as:

$$H_p(s) = \frac{B(s)}{A(s)} \quad \text{or} \quad H_p(s) = \frac{B(s)}{A(s)} e^{-sT_m} \qquad (1.1\text{-}6)$$

In discrete time, the pulse t.f. of the process from the continuous model will be used:

$$H_p(z) = (1 - z^{-1})Z\{\frac{H_p(s)}{s}\} \quad , \quad H_p(z) = \frac{B(z)}{A(z)} \qquad (1.1\text{-}7)$$

These models can be extended by supplementary particularities of the process: process nonlinearity, place of disturbance, time varying parameters [II-1], [II-4], [II-6], [II-7], [II-8], [II-2].

1.2. The structure of the Part II

Chapter 2 starts with a synthetic presentation of the aims of optimization techniques in frequency domain. As a particular case of optimization in frequency domain for low order plants and PI (PID) controllers variants of the Symmetrical

Optimum method (SO-m) are shortly summarized: - the basic variant of Symmetrical Optimum Method (given by Kessler) [II-24]; - the version of SO-m given by Voda&Landau's (KVL- relations) [II-38], [II-39]; - a parameterization of SO-m: the Extended Symmetrical Optimum method (ESO-m).

Based on this results, for the special case of plants without real integral components and dominant time-constant(s) – which is also the case of electrical traction systems - in chapter 3 a double parameterization of the SO-m (marked 2p-SO-m) was introduced by the author (papers [II-20] - [II-23], [II-64]). The tuning method was applied under different forms in some different application papers (see references [II-18], [II-16], [II-20], [II-49], [II-53]). Designing the controller based on this approach there can be ensured:

- Use of pre-calculated (crisp) tuning relations;
- The possibility of improving load disturbance rejection for some specific cases: $T_1>>T_\Sigma$ and $T_1> T_2>>T_\Sigma$.
- The possibility of improving the phase margin of the CS, reducing its sensitivity and increasing its robustness
- The possibility of using both types of controllers: - with homogenous and with non-homogenous structure regarded to the inputs;
- The possibility of improving reference signal tracking by using reference filters (F-r) with parameters that can be easily fixed.

Finally, a Youla parameterization of the 2p-SO-method [II-60], [II-61] is presented.

Chapter 4 presents an application where, the advatages of the desidn method are combined with the advantages of the CCS.

Appendix 1 based on paper [II-70] treate a design method for 2-DOF controllers and comparison with PID controller. The equivalence between a 2-DOF controller and the conventional 1-DOF (PI, PID) controllers with reference filters are analysed.

2. PI, PID controller design techniques in frequency domain. Modulus Optimum based methods. The Modulus Optimum and the Symmetrical Optimum methods

In order to avoid difficulties due to contradictory results obtained from design according to reference tracking and disturbance rejection, different "optimum" - or in special cases, "optimum-like" - tuning techniques are adopted. Part of them, for example [II-3] – [II-7], are developed in frequency domain and known as Modulus Optimum Methods (MO-m). Two of these methods are representative:

- The basic MO-m, shortly presented in paragraph 2.2, using an approach based on papers [II-3], [II-25] and later [II-26]-[II-30], [II-34].
- The Symmetrical Optimum Method (SO-m) [II-24], [II-25], [II-36], [II-3], [II-9], [II-30], [II-34] shortly presented in par.2.3.

They are based on conditions imposed upon the magnitude-frequency characteristics of the closed loop.

2.1. Control structure and basic relations. Optimization techniques

2.1.1. Control structure and basic relations

To define the control problem the CS presented in Fig. 2.1-1 is used. The notations are: C – controller, with t.f. $H_c(s)$, P – plant, with t.f. $H_p(s)$, F-r – reference-filter, $F_r(s)$, r_0 – main reference, r – pre-filtered reference, e – error, u – control signal, y – measured output signal, d_1 – disturbance acting on the plant output, d_2 – disturbance acting on the plant input (load disturbance), n – measurement noise in feedback (its present will be not considered).

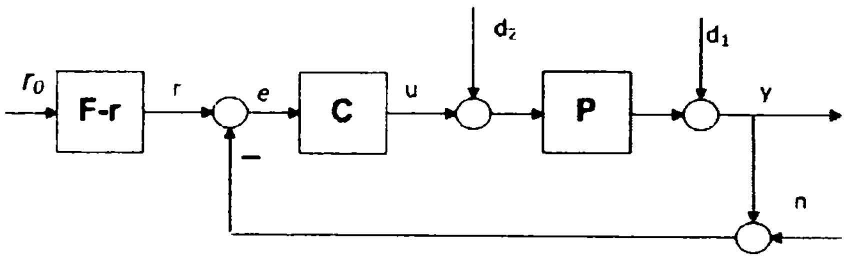


Fig. 2.1-1. Basic control structure

The basic relations between the inputs {r, d_1, d_2 } and the considered outputs {y, u, e} are (continuous time description t.f.s):

- Set-point response (r → y),
- Disturbance rejection (d_1 → y, d_2 → y),
- Robustness to model uncertainties.

$$y(s) = H_c(s)H_p(s)S(s)r(s) + S(s)d_1(s) + H_p(s)S(s)d_2(s) \quad (2.1\text{-}1)$$

$$u(s) = H_c(s)S(s)r(s) - H_c(s)S(s)d_1(s) - H_c(s)H_p(s)S(s)d_2(s) \quad (2.1\text{-}2)$$

$$e(s) = S(s)r(s) - S(s)d_1(s) - H_p(s)S(s)d_2(s) \quad (2.1\text{-}3)$$

$$r(s) = F_r(s)r_0(s) \quad (2.1\text{-}4)$$

The sensitivity function S(s) and the complementary sensitivity function T(s) are

$$S(s) = \frac{1}{1 + H_c(s)H_p(s)} \quad (2.1\text{-}5)$$

$$T(s) = \frac{H_c(s)H_p(s)}{1 + H_c(s)H_p(s)} = H_r(s) \quad (2.1\text{-}6)$$

$$S(s) + T(s) = 1 \quad \text{or} \quad T(s) = 1 - S(s) \quad (2.1\text{-}7)$$

$$L(s) = H_0(s) = H_c(s)H_p(s) \quad \text{- the open loop t.f.} \quad (2.1\text{-}8)$$

From equations (2.1-1)-(2.1-3) four transfer functions (t.f.) can be synthesized

$$H_r(s) = \frac{H_c(s)H_p(s)}{1 + H_c(s)H_p(s)} \quad (2.1\text{-}9)$$

$$H_{d2}(s) = \frac{H_p(s)}{1 + H_c(s)H_p(s)},\ H_{d1}(s) = \frac{1}{1 + H_c(s)H_p(s)},\ H_u(s) = \frac{H_c(s)}{1 + H_c(s)H_p(s)} \quad (2.1\text{-}10)$$

2.1.2. Optimization in frequency domain

Frequency domain optimization is treated differently in different applications. A first approach is presented in papers [II-24], [II-25] (due to C. Kessler) and [II-3], [II-9], [II-36], where the requirements for an "optimal" named behaviour are formulated through (2.1-11) and (2.1-12), respectively:

$$M_r(\omega) = |H_r(j\omega)| \approx 1, \text{ for values of } \omega \geq 0 \text{ as large as possible,} \quad (a)(2.1\text{-}11)$$

$$M_{d1,d2}(j\omega) = |H_{d1,d2}(j\omega)| \approx 0 \text{ , for values of } \omega \geq 0 \text{ as large as possible} \quad (b)$$

The so named magnitude optimality conditions were formulated by Whiteley. Based on these, "optimality (named) relations" can be developed between the coefficients of the characteristic equation. Not respecting all the conditions leads to sub-optimal design relations. This aspect is treated further in this part of the thesis.

A second approach is specific to robust control system design [II-66], where the controller is developed ensuring desired maximum value in frequency domain for the sensitivity function S(s) or for the complementary sensitivity function T(s):

$$M_s = \max|S(j\omega)| \quad , \quad M_p = \max|T(j\omega)| \qquad \text{for } \omega \geq 0 \qquad (2.1\text{-}12)$$

and their typical values are within the intervals [II-3], [II-8]:

$$1.2 \leq M_s \leq 2 \quad , \quad 1.0 \leq M_p \leq 1.5$$

For good set-point tracking it is necessary that M_p should be as close to 1 as possible, and for very good disturbance rejection it is necessary that M_s should be as small as possible. $|S(j\omega)|$ and/or $|T(j\omega)|$ are usually used to express conditions of robust performance [III-34] [II-8]. The magnitude function $|S(j\omega)|$ and/or $|T(j\omega)|$ can be obtained as function of ω:

$$|S(j\omega)| = f_1(\omega) \quad , \quad |T(j\omega)| = f_2(\omega) \text{ , } f_1(\omega), f_2(\omega) : [0 \to \infty) \to R$$

Solving the optimization problem in (2.1-12) means the maximization of the function $f_1(\omega)$ or $f_2(\omega)$ with respect to ω. This approach is applied in part III of the thesis. Such an approach can be also applied to the frequency function of the open loop $L(j\omega)=H_0(j\omega)$.

2.2. The Modulus Optimum method

2.2.1. Basics of Modulus Optimum method (MO-m)

The MO-m is based on frequency domain requirements related to relation (2.2-1):

$$\begin{aligned} &\text{t.f. } H_r(s): && |H_r(j\omega)| = M_r(\omega) = 1 && (a) \\ &\text{t.f. } H_{d1}(s): && |H_{d1}(j\omega)| = M_{d1}(\omega) = 0 && (b) \\ &\text{t.f. } H_{d2}(s): && |H_{d2}(j\omega)| = M_{d2}(\omega) = 0 && (c) \end{aligned} \qquad (2.2\text{-}1)$$

for values of ω as large as possible. By decomposing the expressions of $M_r(\omega)$, $M_{d1}(\omega)$, $M_{d2}(\omega)$ into Mc-Laurin series, the design conditions could be established if the following requirements were fulfilled ([II-9], [II-25], [II-36]):

$$M_r(0) = 1 \quad (1) \text{ and } \quad \left.\frac{d^v|M_r(\omega)|}{d\omega^v}\right|_{\omega=0} = 0 \quad \text{for } \quad v = \overline{1,n} \quad (2) \qquad (a)(2.2\text{-}2)$$

$$M_{d1,d2}(0) = 0 \quad (1) \quad \text{and} \quad \left.\frac{d^{v}|M_{d1,d2}(\omega)|}{d\omega^{v}}\right|_{\omega=0} = 0 \quad \text{for} \quad v = \overline{1,n} \quad (2) \qquad (b)$$

(for details see [II-36], in Romanian).

Condition (1) from expression (2.2-2) can be ensured by poles of L(s) placed in the origin of the s plane. The tuning method based on MO-m tries to fulfill "as good as possible" these requirements [II-25]. The MO-m can be applied in two variants:

- The first variant is based on determining domains of variation of controller parameters that satisfy the imposed requirements, finally determining "the best solution". The method requires huge amount of calculations [II-37].
- The second variant is based on direct tuning relations. Applying this variant is closer to engineering practice [II-3], [II-25].

For many practical situations regarding low order plants, different variants and extensions for applying the MO-m related to Kessler's method are presented in literature [II-3], [II-9], [II-34]-[II-36], [II-48], [II-78]. In practical applications the MO-m is considered as basic in controller design for electrical driving systems and it will be used as basis for comparison.

2.2.2. The MO-m variant given by Kessler for low order plants and PI (PID) controllers

For many practical situations, for the plant description low order models (benchmark type models) can be used, Table 2.2-1, [II-3], [II-15], [II-9]. T_Σ is obtained as a small time constant or the sum of small time constants, (marked in (2.2-3) by τ_i), resulting from the "theorem of small time constants". The method can also be applied for plants with small dead-time T_m that fulfill statements from expressions:

$$T_\Sigma = \Sigma T_i + T_m \qquad (2.2\text{-}3)$$

The table includes also alternative design methods, regarded to the same t.fs.

A. Tuning relations

Accepting that the plant's parameters are (relatively) well known, for all cases marked with MO- from Table 2.2-1, it can be written that:

$$L(s) = H_c(s)H_p(s) = \frac{k_c k_p}{s(1 + sT_\Sigma)} \qquad (2.2\text{-}4)$$

$$H_r(s) = \frac{k_c k_p}{s(1 + sT_\Sigma) + k_c k_p} = \frac{a_0}{a_2 s^2 + a_1 s + a_0}, \quad |H_r(j\omega)| = \left[\frac{a_0^2}{a_0^2 - (2a_0 a_2 - a_1^2)\omega^2 + a_2^2\omega^4}\right]^{1/2} \qquad (2.2\text{-}5)$$

In literature the "optimality condition" [II-3] is given in form of (2.2-6) and makes possible the calculation of controller parameter k_c:

$$2a_0 a_2 = a_1^2 \qquad (2.2\text{-}6)$$

$$k_c = \frac{1}{2k_p T_\Sigma} \qquad (2.2\text{-}7)$$

and the calculation of optimized t.fs., marked with lower index "0":

$$L_0(s) = \frac{1}{2T_\Sigma(1 + sT_\Sigma)} \qquad (2.2\text{-}8)$$

$$H_{r0}(s)=\frac{\omega_0^2}{\omega_0^2+2\zeta\omega_0 s+s^2};\quad H_{r0}(s)=\frac{1}{1+2T_\Sigma s+2T_\Sigma s^2} \tag{2.2-9}$$

A short overview of achievable performances regarding to the CS inputs are given.

Table 2.2-1. Practical situtions for optimization based on Kessler's variant of MO-m

Variant	$H_p(s)$	$H_c(s)$	Remarks
0	1	2	3
1.	$\frac{k_p}{1+sT_\Sigma}$	$\frac{k_c}{s}$	MO-1.1
2.	$\frac{k_p}{(1+sT_\Sigma)(1+sT_1)}$	$\frac{k_c}{s}(1+sT_c)$, $T_c=T_1$	MO-2.1 and **2p-SO-m**
3.	$\frac{k_p}{(1+sT_\Sigma)(1+sT_1)(1+sT_2)}$ $T_1>T_2>T_\Sigma$	$\frac{k_c}{s}(1+sT_c)(1+sT_c')$ $T_c=T_1;\quad T_c'=T_2$	MO-3.1 and **2p-SO-m**
4.	$\frac{k_p}{s(1+sT_\Sigma)}$	k_c	MO-1.2
		$\frac{k_c}{s}(1+sT_c)$	SO-1 (ESO-m)
5.	$\frac{k_p}{s(1+sT_\Sigma)(1+sT_1)}$ $T_\Sigma/T_1<0.2$	$\frac{k_c(1+sT_d)}{1+sT_f}$ $T_d=T_1;\quad T_d/T_f\approx 10$	MO-2.2
		$\frac{k_c}{s}(1+sT_c)\frac{(1+sT_c')}{(1+sT_f)}$ $T_c'=T_1;\quad T_c'/T_f\approx 10$	SO-2 (ESO-m)
6.	$\frac{k_p}{s(1+sT_\Sigma)(1+sT_1)(1+sT_2)}$ $T_1>T_2>T_\Sigma,\quad T_\Sigma/T_1<0.2$	$\frac{k_c(1+sT_{d1})(1+sT_{d2})}{(1+sT_{f1})(1+sT_{f2})}$ $T_{d1}=T_1;\quad T_{d1}/T_{f1}\approx 10$ $T_{d2}=T_2;\quad T_{d2}/T_{f2}\approx 10$	MO-3.2
		$\frac{k_c}{s}(1+sT_c)\frac{(1+sT_c')(1+sT_d)}{(1+sT_f')(1+sT_f)}$ $T_c'=T_1;\quad T_c'/T_f'\approx 10$ $T_d=T_2;\quad T_d/T_f\approx 10$	SO-3 (ESO-m)

Remark: Cases 4, 5 and 6 were analysed in detail in [II-72].

B. Control system performances

- In time domain: regarding to a step reference:
  - the overshoot, $\sigma_1=4.3\%$;
  - the settling time, $t_s=8.4T_\Sigma$;
  - the first settling time, $t_1=4.7T_\Sigma$. (2.2-10)

- the static coefficient $y_n = y_\infty / d_{1,2\ \infty | r=0\ (r=const)}$ depends on the placement of the integral component and type of constant disturbance [II-25].

- In frequency domain:
  - phase margin (reserve) $\varphi_r = 0^0$; crossover frequency, $\omega_c = 1/2T_\Sigma$;
  - maximum magnitude of the frequency response: $M_{pmax} = \max|T(j\omega)| = 1$ for $\omega \to 0$; (2.2-11)
  - the maximum value of the loop sensitivity function $M_{smas} = 1.272$ (or its inverse $M_{S0}^{-1} = 0.786$) is in the typically recommended range of $(1.2 < M_s < 2$, [II-3]).

The method is adequate for developing controllers for systems with constant reference; in case of variable reference the transients are slow.

***C.* External constant type disturbance rejection**

The property must be treated separately for disturbances acting on the input (d_2) and on the output (d_1) of the plant. For d_2 type disturbances called also as load disturbances the behaviour of the system is satisfactory only for case MO-1.1. For cases MO-2.1 and MO-3.1, the transients lead to a slow rejection of it:

$$H_{d2o}^{(2.1)}(s) = H_{d2o}^{(1.1)}(s)\frac{1}{(1+sT_1)} \qquad \text{(MO-2.1)} \qquad (2.2\text{-}12)$$

$$H_{d2o}^{(3.1)}(s) = H_{d2o}^{(1.1)}(s)\frac{1}{(1+sT_1)(1+sT_2)} \qquad \text{(MO-2.2)} \qquad (2.2\text{-}13)$$

where $H_{d2o}^{(1.1)}(s) = \dfrac{2k_pT_\Sigma s}{1+2T_\Sigma+2T_\Sigma^2s^2}$, in the case of disturbance acting on the output.

The presence of factors $(1+sT_1)^{-1}$ and $[(1+sT_1)(1+sT_2)]^{-1}$ results in a worsening of the response time $t_{s(d2)}$ [II-9], [II-34]. The lower the T_Σ / T_1 ratio is the bigger the worsening is [II-72]. In figure 2.2-1 only the cases focused on MO-2.1 case are exemplified, simulated for $\{k_p = 1, T_\Sigma = 1\}$ and different values of $m = T_\Sigma / T_1$, m={0.05, 0.1, 0.15, 0.2, 0.25, ... , 0.5)}.

***D.* Solutions for improving load disturbance rejection**

The solutions are approached in different ways:

- Applying different alternative optimization criteria for calculating the parameters ([II-2]-[II-7], [II-30], [II-37], [II-40] a.o.). Papers [II-3], [II-4], [II-5], [II-30] are survey papers (based on references from the end of the nineties and beginning of 2000) and that offer a relatively comprehensive view of the latest research. Even these methods accept – at least, partially – pole-zero compensation.
- Efficient tuning methods as simple as possible, easily applicable in practice: these are the methods based on the modification of conditions for applying the MO criterion (for example Vrančič et.al. in papers [II-26]-[II-29]).
- New tuning methods,
- Application of combined CS that use both facilities and elegance of PID control and facilities offered by complex control structures (CCS, IMC, Smith Predictor techniques etc.).

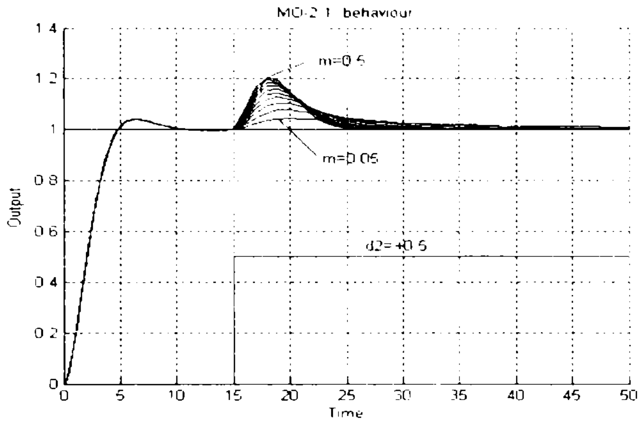


Fig.2.2-1. System response to a step reference input followed by a load disturbance, m={0.05 ... 0.5)}

- Different design methods associated to different robustness conditions imposed to the system [II-41]-[II-44], [II-67], [II-80], [II-81].
- Giving up to classical control structures based on PI (PID) controllers. Practitioners sometimes deny this solution [II-2], [II-3].

The first two directions are based on the idea that the strict application of pole-zero cancellation of big time constants has disadvantageous effect on load disturbance rejection.

The approach suggested by Vrancič [II-26] - [II-29] belongs to the first category. It is based on the fact that the deficiency can be suppressed by an appropriate modification of the optimization criterion associated to the magnitude optimum. Gorez and Klàn give a similar approach in [II-45].

Other contributions can also be mentioned into the second category: - contributions oriented to plants with first order proportional model with dead time (FOPD) [II-6], [II-7] - iterative techniques to solve transcendental equations, graphical tuning based on the parametric D-stability partitioning [II-40] - [II-42], frequency technique based methods Kessler's Symmetrical Optimum [II-3], [II-9], [II-35], [II-36] etc.

2.3 The Symmetrical Optimum method

2.3.1. Basic variant of Symmetrical Optimum method (SO-m)

The basic variant was given by C. Kessler [II-31], as a particular method of the MO-m. The method is presented also in literature under different forms adapted to benchmark type models [II-9], [II-24], [II-34], [II-36], [II-38], [II-39], [II-46], [II-47]. In [II-30], [II-36] and [II-38] it is highlighted the fact that SO-m handles well also nonlinearities and time varying parameters.

In its practical form the target consist in realization of a second order pole in the origin for L(s) that ensures zero steady state error ($e_{\infty t}=0$) for ramp changes of the reference. In [II-25], [II-31], [II-36] the plant is given as:

$$H_p(s) = \frac{k_p}{(1 + sT_\Sigma)\prod_1^n (1 + sT_k)} \quad \text{where } T_\Sigma << \text{oricare}T_k \qquad (2.3\text{-}1)$$

T_Σ is obtained as a small time constant or the sum of small time constants, τ_i. The use of a generalised (PID-m) controller with an accepted t.f. of form:

$$H_c(s) \approx \frac{k_c \prod_1^m (1 + s\tau_k)}{s} \qquad (2.3\text{-}2)$$

is recommended. Fullfiling condition $T_\Sigma << \text{oricare}T_k$, the approximation (2.3-3) is considered (expressed in frequency domain):

$$1 + j\omega T_k \approx j\omega T_k \qquad (2.3\text{-}3)$$

This approximation leads finally to the approximate open loop t.f.:

$$L(s) = \frac{k_c k_p}{s} \frac{\prod_1^m (1 + s\tau_k)}{(1 + sT_\Sigma)\prod_1^n (1 + sT_k)} \approx \frac{k_c k_p}{s} \frac{\prod_1^m (1 + s\tau_k)}{(1 + sT_\Sigma)\prod_1^n (sT_k)} \qquad (2.3\text{-}4)$$

The deduction of basic equations is presented in [II-9], [II-31], [II-36]. Finally, the optimized expression of the open-loop t.f. results in the reduced in form of:

$$L_0(s) = \frac{1 + 4T_\Sigma s}{8T_\Sigma^2 s^2 (1 + sT_\Sigma)} \quad , \quad k_0 = k_c k_p = \frac{1}{8T_\Sigma^2} \qquad (2.3\text{-}5)$$

Consequently, the optimized closed-loop t.f. is given by:

$$H_{r0}(s) = \frac{L_0(s)}{1 + L_0(s)} = \frac{1 + 4T_\Sigma s}{1 + 4T_\Sigma s + 8T_\Sigma^2 s^2 + 8T_\Sigma^3 s^3} = \frac{1 + 4T_\Sigma s}{(1 + 2T_\Sigma s)(1 + 2T_\Sigma s + 4T_\Sigma^2 s^2)} \qquad (2.3\text{-}6)$$

The controller parameters can be calculated based on supplementary conditions. Regarding the open loop t.f. $L_0(s)$, relation (2.3-5), it can be stated that:

- The crossover frequency ω_c of the compensated system should be placed at $\omega_c = 1/(2T_\Sigma)$ and the slope of the Bode diagram at the crossover frequency is -20dB/dec.;
- The PI (PID) controller is chosen such that it preserves the slope of -20dB/dec for one octave to the right and m-octaves to the left of the crossover frequency.

2.3.2. Version of SO-m given by Voda& Landau (the KVL-relations)

The method - named Kessler's SO tuning rules modified by Voda and Landau - is not based on cancellation of the poles. In [II-38] the starting point is equation (2.3-7), which is a particular form of the t.f. (2.3-1) for n=1 and 2:

$$H_{p0}(s) = \frac{k_p}{(1 + sT_1)(1 + sT_2)(1 + s\tau_1)\cdots(1 + s\tau_v)} e^{-sT_m} \qquad (2.3\text{-}7)$$

and T_m has the order of magnitude of small time constants. Applying (2.3-2), the relation (2.3-7) can be approximated with a benchmark model of form:

$$H_p(s) = \frac{k_p}{(1 + sT_1)(1 + sT_2)(1 + sT_\Sigma)} \quad \text{where, } T_\Sigma = \Sigma T_i + T_m \qquad (2.3\text{-}8)$$

with the conditions definied by Voda and Landau:

$T_1 \gg 4T_\Sigma$ and $T_2=0$ (a) or $T_1>T_2>4T_\Sigma$ (b) .

Particularizing in (2.2-4) for n=1 or 2, the parameter calculus relation for the PI (a) or PID (b) controller obtains the forms given in [II-38]:

- For n=1: the plant t.f. results of form (2.3-9) (a) and a PI controller is used;

$$H_p(s) = \frac{k_p}{(1+sT_1)(1+sT_\Sigma)} \quad (a) \quad H_c(s) = k_C(1+\frac{1}{sT_i}) = \frac{k_c}{s}(1+sT_c) \quad (b) \quad (2.3\text{-}9)$$

$$k_C = \frac{1}{8k_pT_\Sigma^2}\ , \quad T_i = 4T_\Sigma \quad (1) \qquad k_c = \frac{1}{8k_pT_\Sigma^2}\ , \quad T_c = 4T_\Sigma \quad (2)$$

The closed loop t.f. results as:

$$H_{r0}(s) = \frac{1+4T_\Sigma s}{1+4T_\Sigma s+8T_\Sigma^2 s^2+8T_\Sigma^3 s^3} = \frac{1+4T_\Sigma s}{(1+2T_\Sigma s)(1+2T_\Sigma s+4T_\Sigma^2 s^2)} \quad (2.3\text{-}10)$$

- For n=2, the plant t.f. is (2.3-11) (a). Accepting that $T_1 > T_2 > 4T_\Sigma$, a PID controller (b) is used, and the tuning relations of its parameters are as follows:

$$H_p(s) = \frac{k_p}{(1+sT_1)(1+sT_2)(1+sT_\Sigma)} \ (a)\ , \ H_c(s) = k_{Cp}(1+\frac{1}{sT_i}+sT_d) = \frac{k_c}{s}(1+sT_c)(1+sT_c') \ (b)$$

$$k_C = \frac{k_c}{T_i} = \frac{T_1(T_2+4T_\Sigma)}{8k_pT_\Sigma^2}\ , \quad T_i = T_2+4T_\Sigma\ , \quad T_d = \frac{2T_2T_\Sigma}{T_2+T_\Sigma} \quad (1) \quad (2.3\text{-}11)$$

$$k_c = \frac{T_1T_2}{2k_pT_\Sigma}\frac{1}{T_cT_c'} = \frac{T_1T_2}{2k_p 64T_\Sigma^3}\ , \quad T_c = T_c' = 8T_\Sigma \quad (2)$$

The closed loop results in a fourth order t.f.:

$$H_{r0}(s) = \frac{(1+8T_\Sigma s)^2}{(1+11{,}2T_\Sigma s)(1+2.4T_\Sigma s)(1+2.3T_\Sigma s+4.7T_\Sigma^2 s^2)} \quad (2.3\text{-}12)$$

Remark: Relations (2.3-10) and (2.3-12) are valid only if approximations (2.3-3), (2.3-4) are taken into account. Otherwise the double parameterization proposed in paragraph 2.4 must be used. In [II-39] also an "auto-calibration" method for PI and PID controllers is presented, based on frequency technique and the possibility to use the SO-m techniques is presented.

2.3.3. Version of the SO-m for low order benchmark-type plants

A modified version of SO-m was given in [II-3], [II-9], [II-34], [II-48] restricted to benchmark-type plants with integral component (a pole in the origin), having a t.f. $H_p(s)$ of form (2.3-13) (in Table 2.2-1, the cases marked with SO-1 - SO-3):

$$H_p(s) = \frac{k_p}{s(1+sT_\Sigma)(1+sT_1)(1+sT_2)} \quad (2.3\text{-}13)$$

Using adequate controllers, for al cases L(s) result with a double pole in origin:

$$L(s) = H_c(s)H_p(s) = \frac{k_ck_p(1+sT_c)}{s^2(1+sT_\Sigma)} \qquad (T_c>T_\Sigma) \qquad (2.3\text{-}14)$$

Due to this, for $H_r(s)$ a t.f. in form of (2.3-15) results:

$$H_r(s)=\frac{k_ck_pT_cs+k_ck_p}{s^3T_\Sigma+s^2+k_ck_pT_cs+k_ck_p}=\frac{b_1s+b_0}{a_3s^3+a_2s^2+a_1s+a_0}\quad \text{with}\quad \begin{matrix} b_0=a_0\\ b_1=a_1\end{matrix} \qquad (2.3\text{-}15)$$

$$a_0=b_0=k_ck_p,\quad a_1=b_1=k_ck_pT_c,\quad a_2=1,\quad a_3=T_\Sigma\ . \qquad (2.3\text{-}16)$$

Based on (2.3-14) the "modulus optimum" conditions results [II-3], [II-9], [II-34]:

$$2a_0a_2=a_1^2\quad ,\qquad 2a_1a_3=a_2^2 \qquad (2.3\text{-}17)$$

which ensure for $H_r(s)$ an optimal "in modulus" form, $H_{r0}(s)$:

$$|H_{r0}(j\omega)|=[\frac{1+(a_1/a_0)^2\omega^2}{1+(a_3/a_0)^2\omega^6}]^{1/2}=[\frac{1+(4T_\Sigma)^2\omega^2}{1+(8T_\Sigma^3)^2\omega^6}]^{1/2} \qquad (2.3\text{-}18)$$

A. Tuning relations

Applying conditions (2.3-17) the controller parameters result in a simple form (in [II-72] presented for all cases), easy to be used in practice.

- SO-1 case: $k_C=\frac{1}{2k_pT_\Sigma}$, $T_i=4T_\Sigma$ (1) $k_c=\frac{k_C}{T_i}=\frac{1}{8k_pT_\Sigma^2}$, $T_c=4T_\Sigma$ (2)(a)

(2.3-19)

- SO-2 case: $k_C=\frac{T_i}{8k_pT_\Sigma{}^2}$, $T_i=T_1+4T_\Sigma$, $T_d=\frac{4T_1T_\Sigma}{T_1+4T_\Sigma}$ (1) (b)

$$k_c=\frac{1}{8k_pT_\Sigma^2},\quad T_c=4T_\Sigma,\quad T_c'=T_1 \qquad (2)$$

Then the "optimal" t.f.s $L_0(s)$, $H_{r0}(s)$ and $S_0(s)$ can be easy found:

$$L_0(s)=\frac{(1+4T_\Sigma s)}{8T_\Sigma^2s^2(1+sT_\Sigma)}\quad ,\quad k_0=k_ck_p=\frac{1}{8T_\Sigma^2} \qquad (2.3\text{-}20)$$

$$H_{r0}(s)=\frac{L_0(s)}{1+L_0(s)}=\frac{1+4T_\Sigma s}{(1+2T_\Sigma s)(1+2T_\Sigma s+4T_\Sigma^2s^2)} \qquad (a)$$

$$S_0(s)=\frac{1}{1+L_0(s)}=\frac{8T_\Sigma{}^2s^2(1+sT_\Sigma)}{1+4T_\Sigma s+8T_\Sigma^2s^2+8T_\Sigma^3s^3} \qquad (b) \qquad (2.3\text{-}21)$$

***B.* Control system performances**

In Fig.2.3-1 some features of the system performance are presented (for k_p=1 and T_Σ =1).

- In time-domain. Simulation of the control system (for SO-2 case):
  - the overshoot $\sigma_{1,r}\approx 43{,}0\,\%$ (very high)
  - the settling time $t_{s,r}=16.5T_\Sigma$;
  - the first settling time $t_{1,r}=3.1\,T_\Sigma$;
  - the steady state error for a step and for a ramp changes of the reference are zero (e_∞=0 , $e_{r\infty}$=0);
- In frequency domain:
  - the phase margin $\varphi_{rm}=36^0$, (maximum value) at the crossover frequency $\omega_c=1/(2T_\Sigma)$;
  - Magnitude plot of the complementary sensitivity function with $M_{p\max}(\omega)=1{,}682$ for $\omega=0.414/T_\Sigma$;

- Nyquist diagram $h_+\{L_0(j\omega)\}$, fig.2.3-1;
- the maximum value of sensitivity function: $M_{S\max} = 1.682$ for $\omega = 0{,}6$ is big. Due to these, the robustness towards plant nonlinearities and time-varying characteristics is reduced.

The main characteristics of SO-m tuning technique can be synthesized as mentioned for example in [III-72].

C. External constant type disturbance rejection

Regardless of the disturbance type, d_1 or d_2, the static coefficient is equal to zero, $\gamma_n = 0$. Based on the resulting t.f.s, the following results are available:

- d_1 type disturbances: for all situations of SO-m from Table 2.2-1 good behavior is ensured, see Fig. 2.3-1 (d) .

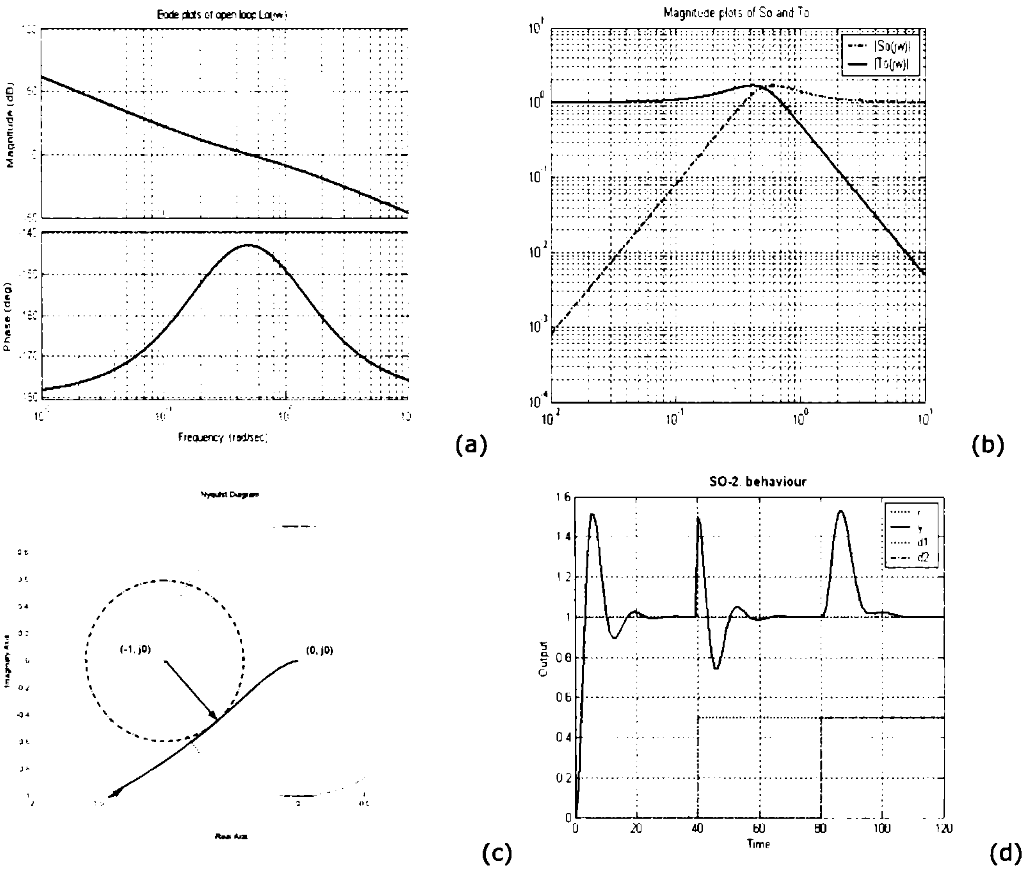


Fig.2.3-1. Significant diagrams for the SO optimized system

- d_2 type (load) disturbances, Fig. 2.3-2: - for case SO-1 the behaviour of the system is satisfactory; - for cases SO-2 and SO-3 the presence of factors $(1+sT_1)^{-1}$ and $(1+sT_1)^{-1}(1+sT_2)^{-1}$ in the corresponding t.f. results in a slower control; the settling time $t_{s(d2)}$ is increased.

2.3.4. Parameterization of SO-m: the Extended Symmetrical Optimum method

The Extended Symmetrical Optimum Method (abbreviated the ESO-method) was introduced in papers [II-46], [II-47] and [II-48], for low order plants with integral component (see Table 2.2.1). The adequate types for plants and controllers are combined as in Table 2.2-1. For the cases marked with 5 and 6 a parcial pole-zero cancellation is applied.

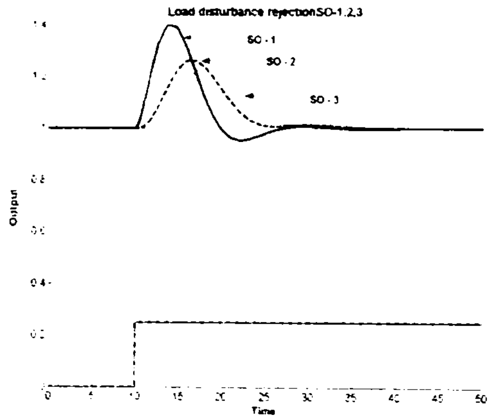


Fig. 2.3-2. Unit step load disturbance for case SO-1, SO-2, SO-2

The method consists in a parameterization of the modulus optimum conditions under the following form:

$$\beta^{1/2}a_0a_2 = a_1^2 \quad , \quad \beta^{1/2}a_1a_3 = a_2^2 \qquad (2.3\text{-}22)$$

where β is a design parameter which modifies the frequency characteristics from modulus-optimum form (2.3-18). Through this parameterization the method leads to an improvement of CS performances (for β=4 all the specific SO-m variants are obtained). Applying the conditions (2.3-22) leads in $|H_r(j\omega)|$ to the "optimal" form:

$$|H_{r0}(j\omega)| = \left[\frac{a_0^2 + a_1^2\omega^2}{a_0^2 - (\beta^{1/2}a_0a_2 - a_1^2)\omega^2 - (\beta^{1/2}a_1a_3 - a_2^2)\omega^4 + a_3^2\omega^6}\right]^{1/2} = \left[\frac{1 + (a_1^2 / a_0^2)\omega^2}{1 + (a_3^2 / a_0^2)\omega^6}\right]^{1/2}$$

(2.3-23)

A. Tuning relations

- Case SO-1: $k_c = \frac{k_C}{T_i} = \frac{1}{\beta^{3/2}k_pT_\Sigma^2}, T_c = \beta T_\Sigma = T_i$ (1) $k_C = \frac{1}{\beta^{1/2}k_pT_\Sigma}, \quad T_i = \beta T_\Sigma$ (2) (2.3-24)

- Case SO-2:

$$k_c = \frac{1}{\beta^{3/2}k_pT_\Sigma^2} \quad , \quad T_c = \beta T_\Sigma \quad , \quad T_c' = T_1 \qquad (1)$$

$$k_C = \frac{T_1}{\beta^{3/2}k_pT_\Sigma^{\,2}} \quad , \quad T_i = T_1 + \beta T_\Sigma \qquad T_d = \frac{\beta T_1 T_\Sigma}{T_1 + \beta T_\Sigma} \qquad (2)$$

(2.3-25)

Remark: Case SO-3 is detailed in [II-72].

Applying (2.3-22), for all mentioned cases the t.f.-s $L_0(s)$ and $H_{r0}(s)$ obtain the same form:

$$L_0(s) = H_c(s)H_p(s) = \frac{(1+\beta T_\Sigma s)}{\beta^{3/2}T_\Sigma^2 s^2(1+sT_\Sigma)} \quad \text{with} \quad \frac{1}{\beta^{3/2}T_\Sigma^2} = k_0 = k_c k_p \qquad (2.3\text{-}26)$$

$$H_0(s) = \frac{1+\beta T_\Sigma s}{1+\beta T_\Sigma s+\beta^{3/2}T_\Sigma^2 s^2+\beta^{3/2}T_\Sigma^3 s^3} = \frac{1+\beta T_\Sigma s}{(1+\beta^{1/2}T_\Sigma s)(1+(\beta-\beta^{1/2})T_\Sigma s+\beta T_\Sigma^2 s^2)} \qquad (2.3\text{-}27)$$

B. Control system performances

In figure 2.3-3 the CS performance indices versus β are depictured in form of diagrams [II-47] and [II-53]. These diagrams are very useful for controller design by allowing to fixing the value of β according to the desired performance.

- In time domain: the performance indices, $\{\sigma_{1,r}, \hat{t}_{s,r} = t_{s,r}/T_\Sigma, \hat{t}_{1,r} = t_{1,r}/T_\Sigma\}$;
  Mainly, by increasing the value of β, the overshoot decreases and the oscillations of the error signal are diminished.
- In Frequency domain
  - The phase margin φ_m versus β is given in figure 2.3-3 in a graphical form.
  - Bode Diagrams and phase margins. In [II-47] the Bode diagrams are ploted for different β values. The improvement of the phase-margin obtained by the increase of β is favourable.

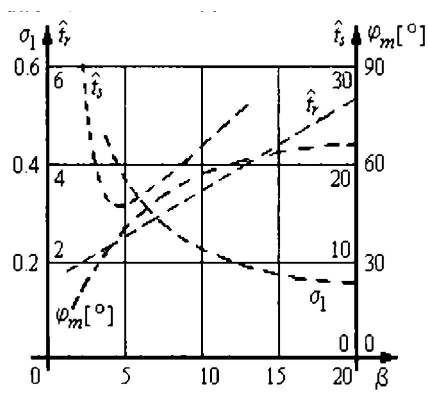


Fig. 2.3-3. Control system performance indices versus β

  - Sensitivity function analysis. Based on the relation of $S_0(s)$ in figure 2.3-4 (a), (b), (c) the Nyquist diagrams are presented for β= 4, 9, 16; the $M_{S0}=f(\beta)$ circles and the values of M_{S0}^{-1} are also marked. The curves point out the increase of robustness when the value of β is increased.
  - Magnitude plot of the complementary sensitivity function $M_p(\omega,\beta) = |H_0(j\omega,\beta)|$; the dependencies depictured in Fig.2.3-5 (a) - (c) and $M_{pmax(\beta=4)} \approx 1.6823$, $M_{pmax(\beta=9)} \approx 1.2990$, $M_{pmax(\beta=16)} \approx 1.1978$.

C. External constant type disturbance rejection

The static γ_n coefficient is always zero. Simulation results for k_p =1, T_Σ =1, T_1 =10 and T_2 =4 illustrate the situation, fig.2.3-6 [III-72].

- d_1 - type disturbances: for all cases quick transients are ensured.
- d_2 - type disturbances, fig.2.3-6 (a), (b), (c) the step responses). Only for case SO-1 the behaviour of the system is satisfactory. The increase of overshoot is observed by increasing the value of β. For cases SO-2 and SO-3 the CS-s presents the factors $(1+sT_1)^{-1}$ and $(1+sT_1)^{-1}(1+sT_2)^{-1}$ which

leads to a lengthening of the response time $t_{s,d2}$; this bigger inertia of the plant reduces the overshoot.

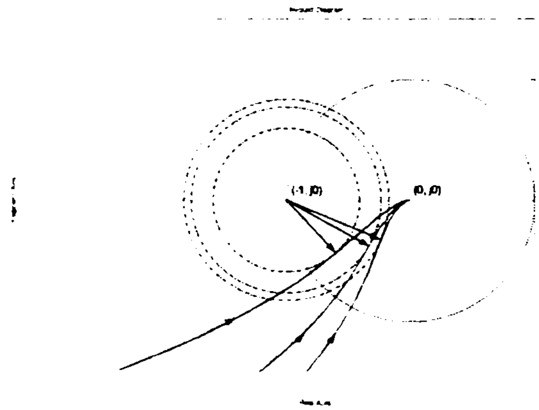


Fig.2.3-4. Nyquist curves and M_{S0}^{-1} circles for β = 4, 9, 16 and the $M_{S0}^{-1}=f(\beta)$ circles

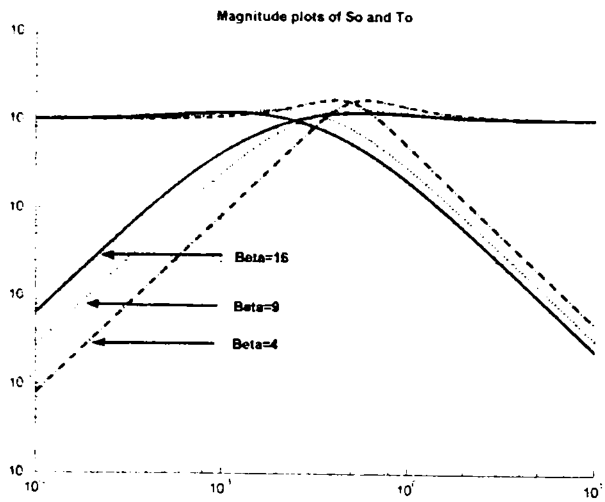
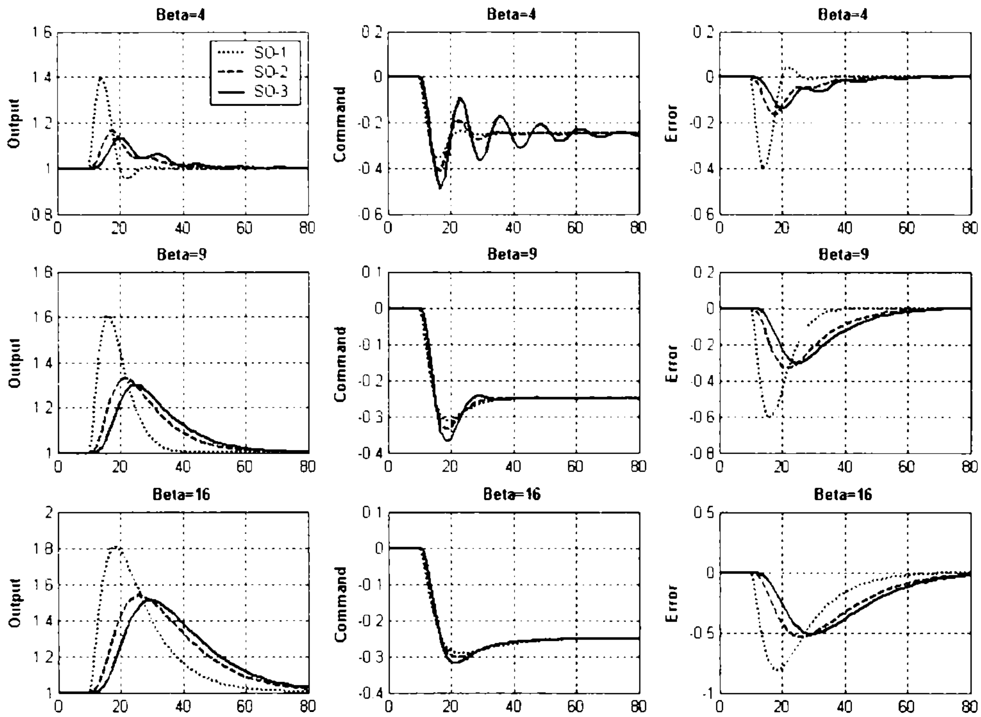


Fig.2.3-5. Magnitude plot of the $M_S(\omega,\beta)$ and $M_P(\omega,\beta)$ for β=4, 9, 16

The increase of the phase margin (accompanied with the decreasing of ω_c) leads also to increasing of the settling time. These effects are highlighted in fig.2.3-6 (b), (c). The reference behaviours can be corrected using adequate reference filter F-r.



(a) (b) (c)
Fig.2.3-6. Output response (shifted with steady state value 1 (a), control signal (b) and error (c) for cases SO-1, -2, -3 and different values of β

3. Reference tracking, load disturbance rejection and robustness enhancement through double parameterization of the Symmetrical Optimum method: the 2p-SO-method

3.1. The method

Based on positive results given in [I-5], [I-9] and [II-47], the proposed double parameterization of the Symmetrical Optimum method (2p-SO-m) was introduced by the author in papers [II-21], [II-23] and is oriented to fulfill:

- good tracking performances,
- efficient disturbance-rejection for a special case of applications for plants without integrating components characterised by $T_1 > T_2 >> T_\Sigma$

For these requirements the MO-m does not give satisfaction. The double parameterization ensures the satisfaction of both requirements. Detailed research results were presented also in [II-97], part C, for which I was task responsible.

The t.f. of the plant corresponds to the initial approach of SO-m, (2.3-1), for plants with t.f.

$$\hat{H}_p(s) = H_p(s)e^{sT_m} \quad (3.1\text{-}1)$$

$$H_p(s) = \frac{k_P}{(1+sT_1)(1+sT_2)(1+s\tau_1)(1+s\tau_2)\ldots(1+s\tau_k)} \quad (3.1\text{-}2)$$

If the time delay T_m is small enough it will be included in the small time constant. Applying the theorem of small time constants, (3.1-2) can be rewritten as:

$$H_p(s) = \frac{k_p}{(1+sT_\Sigma)(1+sT_1)(1+sT_2)} \ , \ T_1 > T_2 >> T_\Sigma \ , \ T_\Sigma = \sum_1^k \tau_v + T_m \quad (3.1\text{-}3)$$

If $T_2 << T_1$ and $T_\Sigma = \sum_1^k \tau_v + T_2 + T_m$, then

$$H_p(s) = \frac{k_p}{(1+sT_\Sigma)(1+sT_1)} \ , \ T_1 >> T_\Sigma \quad (3.1\text{-}4)$$

For this class of plants with dominant time-constant(s) a double parameterization (marked 2p-) is proposed. The resulted tuning technology was applied later in detail in [II-16], [II-18], [II-20], [II-49], [II-53] and verified through simulation.

The double parameterisation is based on the followings:

(1) First, with the condition that $T_\Sigma / T_1 << 1$, the parameter m is defined

$$m = T_\Sigma / T_1$$

So, the approximations given by (2.3-3) and (2.3-4) are more completely treated and become possible to analyse the situations when - regarding load disturbance rejection - this approach is more advantageous compared to MO-m.

(2) Second, the use of the optimisation relations (2.3-22) specific for the case of plants with integral component:

(2.3-22): $$\beta^{1/2} a_0 a_2 = a_1^2 \quad , \quad \beta^{1/2} a_1 a_3 = a_2^2 \quad (3.1\text{-}5)$$

Through this an improvement of the phase margin can be reached:

The method was called Extension through a double parameterization of the Symmetrical (Optimum) method and is marked with 2p-SO-m. Since the tuning relations do not satisfy the "optimum" conditions (2.3-17) the term "Optimum" could be omitted. The controller can ensure:

- Use of pre-calculated (crisp) tuning relations, based on the model of the plant;
- The possibility of improving the CS's phase margin, reducing its sensitivity and increasing its robustness;
- The possibility of using both types of controllers: - with homogenous structure regarded to the inputs; - with non-homogenous structure regarded to the inputs;
- The possibility of improving good reference signal tracking by using reference filters with parameters that can be easily fixed.
- The possibility of improving load disturbance rejection for some specific cases.

It must be mentioned that only a minority of the tuning methods presented in the literature deals with load disturbance rejection (for example [II-10], [II-40], [II-50], [II-51], [II-52]), even if - in most cases - the CS operate with constant reference and is subject to disturbances. Because of this, an efficient rejection of the effect of load disturbance becomes often dominant.

3.1.1. Basic relations

Three basic situations taken into account are synthesized in Table 3.1.1. For each case the corresponding controllers are of PI, PID or PID^2 type, used in their ideal forms.

A. Tuning relations of controller parameters

Accepting the controller-plant combination given in Table 3.1-1, and applying the indicated pole-zero cancellation, one gets:

$$L(s) = H_c(s)H_p(s) = \frac{k_c k_p (1 + sT_c)}{s(1 + sT_1)(1 + sT_\Sigma)}, H_r(s) = \frac{L(s)}{1 + L(s)} \tag{3.1-6}$$

$$H_r(s) = \frac{k_c k_p + s k_c k_p T_c}{k_c k_p + s(1 + k_c k_p T_c) + s^2(T_1 + T_\Sigma) + s^3 T_1 T_\Sigma} \tag{3.1-7}$$

Table 3.1-1. The basic situations (see also Table 2.2-1)

Case	$H_p(s)$	$H_c(s)$	Remarks
0	1	2	3
1.	$\frac{k_p}{(1+sT_\Sigma)(1+sT_1)}$	$\frac{k_c}{s}(1 + sT_c)$, $T_c = T_1$	**2p-SO-m-1** and MO-2.1
2.	$\frac{k_p}{(1 + sT_\Sigma)(1 + sT_1)(1 + sT_2)}$ $T_1 > T_2 > T_\Sigma$	$\frac{k_c}{s}(1 + sT_c)\frac{(1 + sT_c')}{(1 + sT_f)}$ $\frac{k_c}{s}(1 + sT_c)(1 + sT_c')$ $T_c' = T_2$; $(T_c' / T_f' \approx 10)$	**2p-SO-m-2** and MO-3.1
3.	$\frac{k_p}{(1+sT_\Sigma)(1+sT_1)(1+sT_2)(1+sT_3)}$ $T_1 > T_2 > T_3 > T_\Sigma$, $T_\Sigma / T_1 < 0.2$	$\frac{k_c}{s}(1+sT_c)\frac{(1+sT_c')(1+sT_d)}{(1+sT_f')(1+sT_f)}$ $T_c' = T_2$; $(T_c' / T_f' \approx 10)$ $T_d = T_3$; $(T_d / T_f \approx 10)$	**2p-SO-m-3** (this case is not detailed here, [II-72])

Relatively to the general Proportional-Derivative with Lags (PDL^3) form (2.3-15), the coefficients a_v, b_μ are:

$$\begin{aligned} &a_0 = k_c k_p \quad , \quad a_1 = 1 + k_c k_p T_c \quad , \quad a_2 = T_1 + T_\Sigma \quad , \quad a_3 = T_1 T_\Sigma \\ &b_0 = k_c k_p \quad , \quad b_1 = k_c k_p T_c \end{aligned} \tag{3.1-8}$$

Upon these coefficients the conditions (3.1-5) are imposed.

In order to discuss the approximations (2.3-3) and (2.3-4) the m parameter is introduced:

$$m = T_\Sigma / T_1 \tag{3.1-9}$$

In accordance with the conditions imposed by Kessler, the situations for interest are characterised by values of m<(<<)1. Replacing (3.1-8) into the second parameterization, (3.1-5), results:

$$\beta^{1/2} k_c k_p (T_1 + T_\Sigma) = (1 + k_c k_p T_c)^2 \text{ (a)} \tag{3.1-10}$$

$$\beta^{1/2}(1 + k_c k_p T_c) T_1 T_\Sigma = (T_1 + T_\Sigma)^2 \text{ (b)} \tag{3.1-11}$$

The tuning relation of k_c given in a double-parameterised form, is:

$$k_c = \frac{(1+m)^2}{\beta^{3/2} k_p T_1 \frac{T_\Sigma}{T_1} m}(1+m) \quad \text{or, with} \quad T'_\Sigma = \frac{T_\Sigma}{1+m} \qquad (3.1\text{-}12)$$

$$k_c = \frac{(1+m)^2}{m}\frac{1}{\beta^{3/2} k_p T'_\Sigma} = \frac{(1+m)^3}{m}\frac{1}{\beta^{3/2} k_p T_\Sigma}. \qquad (3.1\text{-}13)$$

T_c can be determined by replacing k_c into (3.1-11):

$$T_c = \beta T_\Sigma \frac{[1+(2-\beta^{1/2})m+m^2]}{(1+m)^3} \quad \text{or} \quad T_c = \beta T_{\Sigma m} \quad \text{with} \qquad 3.1\text{-}14)$$

$$\Delta_m(m) = [1+(2-\beta^{1/2})m+m^2] \quad \text{and} \quad T_{\Sigma m} = T'_\Sigma \frac{\Delta_m(m)}{(1+m)^2} = T_\Sigma \frac{\Delta_m(m)}{(1+m)^3} \qquad (3.1\text{-}15)$$

For the particular values β=4, 9, 16 (that give integer square roots) the controller parameters $\{k_c, T_c\}$ get more compact forms, Table 3.1-2.

Table 3.1-2. The controller parameters $\{k_c, T_c\}$ for particular values β=4, 9, 16

β	$k_c = \frac{(1+m)^2}{\beta^{3/2} k_p T'_\Sigma m}$	$\Delta_m(m)$	$T_c = \beta T'_\Sigma \frac{[1+(2-\beta^{1/2})m+m^2]}{(1+m)^2}$
4	$k_c = \frac{(1+m)^2}{8 k_p T'_\Sigma m}$	$(1+m)^2$	$T_c = 4T'_\Sigma \frac{(1+m^2)}{(1+m)^2}$
9	$k_c = \frac{(1+m)^2}{27 k_p T'_\Sigma m}$	$(1-m+m^2)$	$T_c = 9T'_\Sigma \frac{(1-m+m^2)}{(1+m)^2}$
16	$k_c = \frac{(1+m)^2}{64 k_p T'_\Sigma m}$	$(1-m)^2$	$T_c = 16T'_\Sigma \frac{(1-m)^2}{(1+m)^2}$

B. Optimized forms for the main t.f.s

The "optimized" expressions for $L_0(s)$, $H_{r0}(s)$, $S_0(s)$, $H_{d20}(s)$ were determined as follows:

$$L_0(s) = \frac{1+\beta T_{\Sigma m} s}{\beta^{3/2} T'_\Sigma \frac{m}{(1+m)^2} s(1+sT_1)(1+sT_\Sigma)} \qquad (3.1\text{-}16)$$

$$H_{r0}(s) = \frac{(1+\beta T_{\Sigma m} s)}{\beta^{3/2} T'^3_\Sigma s^3 + \beta^{3/2} T'^2_\Sigma s^2 + \beta T'_\Sigma s + 1}\ (a),$$
$$H_{r0}(s) = \frac{(1+\beta T_{\Sigma m} s)}{(1+\beta^{1/2} T'_\Sigma s)[1+(\beta\ \ \beta^{1/2}) T'_\Sigma s + \beta T'^2_\Sigma s^2]}\ (b) \qquad (3.1\text{-}17)$$

$$S_0(s) = \frac{\beta^{3/2} T'_\Sigma \frac{m}{(1+m)^2} s(1+sT_1)(1+sT_\Sigma)}{\beta^{3/2} T'^3_\Sigma s^3 + \beta^{3/2} T'^2_\Sigma s^2 + \beta T'_\Sigma s + 1} \qquad (3.1\text{-}18)$$

The t.f. regarding load disturbance $H_{d20}(s)$ depends essentially on the t.f. of the plant:

- Case 1: $H_p(s) = \frac{k_p}{(1+sT_\Sigma)(1+sT_1)}$ and using a PI controller:

$$H_{d20(1)}(s)=\frac{H_p(s)}{1+L_0(s)}=\frac{\beta^{3/2}k_pT'_{\Sigma}\frac{m}{(1+m)^2}s}{\beta^{3/2}T_{\Sigma}'^{3}s^3+\beta^{3/2}T_{\Sigma}'^{2}s^2+\beta T'_{\Sigma}s+1} \quad (3.1\text{-}19)$$

- Case 2: $H_p(s)=\frac{k_p}{(1+sT_{\Sigma})(1+sT_2)(1+sT_1)}$ and using a PID controller

$$H_{d20(2)}(s)=H_{d20(1)}(s)\frac{1}{(1+sT_2)} \quad (3.1\text{-}20)$$

Discussions upon the particular case $T_1>>T_{\Sigma}$. For $T_1>>T_{\Sigma}$, by accepting Kessler's simplifying conditions, it can be written:

$$\frac{k_p}{1+sT_1}\approx\frac{k_p}{sT_1}=\frac{k_p'}{s} \text{ and } H_p(s)=\frac{k_p}{s(1+sT_{\Sigma})\ldots} \quad (\text{where } k_p=k_p') \quad (3.1\text{-}21)$$

C. Significant particular cases

For particular values β=4 , 9 , 16 the expressions of $L_0(s)$, $H_{r0}(s)$, $S_0(s)$, $H_{d20}(s)$ result as shown in Table 3.1-3. For:

$$m=\frac{T_{\Sigma}}{T_1}\to 0,\ T'_{\Sigma}=\frac{T_{\Sigma}}{(1+m)}=T_{\Sigma},\ T_{\Sigma m}=T'_{\Sigma}\frac{\Delta_m(m)}{(1+m)^2}=T_{\Sigma}\frac{\Delta_m(m)}{(1+m)^3} \quad (3.1\text{-}22)$$

and consequently, the tuning relations for PI or PID parameters are similarly with that are given for the (E)SO-m which can be considered as a particular case.

D. Analysis of changes in controller parameter values

Particular values m€[0.05, 0.20 (0.25)] and 4≤β≤16 are considered. The normalised values of k_c and T_c , denoted k_{kc}^* = $f_1(m)$ and k_{Tc}^* = $f_2(m)$, are defined based on relations (3.1-14) and (3.1-15). Using the notations:

$$k_{c0}=\frac{1}{\beta^{3/2}k_pT_{\Sigma}^2},\ T_{c0}=\beta T_{\Sigma} \text{ or: } k_c=\frac{(1+m)^3}{m}T_{\Sigma}k_{c0},\ k_{kc}^*=\frac{k_c}{k_{c0}}=(1+m)^3T_1 \quad (3.1\text{-}23)$$

$$T_c=T_{c0}\frac{\Delta_m(m)}{(1+m)^3} \quad \text{and} \quad k_{Tc}^*=\frac{T_c}{T_{c0}}=\frac{\Delta_m(m)}{(1+m)^3} \quad (3.1\text{-}24)$$

For m having values between m $\in$[0.05, 0.5] the expressions of k_{kc}^* and k_{Tc}^* can be calculated. Relations (3.1-23) and (3.1-24) highlight that for values of the ratio $T_{\Sigma}/T_1\geq 0.05$ the parameter changes cannot be neglected and the approximations (2.3-3) must be revised.

Based on these results, the comparison of solutions with those given by MO-m tuning is theoretically founded and the use of 2p-SO-m can be advantageous. For this purpose only the case (1), 2p-SO-1 is treated; the case (2), 2p-SO-2 can be treated in a similar way:

$$H_p(s)=\frac{k_p}{(1+sT_{\Sigma})(1+sT_1)} \ (1),\quad H_p(s)=\frac{k_p}{(1+sT_{\Sigma})(1+sT_2)(1+sT_1)} \ (2)$$

Table 3.1-3. The equations of $L_0(s)$, $H_{r0}(s)$, $S_0(s)$, $H_{dz0}(s)$ in its double parameterised form for particular β values: β=4 , 9 , 16

β	$L_0(s)=\dfrac{1+\beta T_{\Sigma m}s}{\Delta_L(s)}$	$H_{r0}(s)=\dfrac{(1+\beta T_{\Sigma m}s)}{\Delta(s)}$	$S_0(s)=\dfrac{\beta^{3/2}T_\Sigma \frac{m}{(1+m)^2}s(1+sT_1)(1+sT_\Sigma)}{\Delta(s)}$	$H_{dz0}(s)=S_0(s)P(s)=\dfrac{\beta^{3/4}k_p T_\Sigma \frac{m}{(1+m)^2}s}{\Delta(s)}$
4	$\dfrac{1+9\frac{T_1T_\Sigma}{(T_1+T_\Sigma)^3}(T_1^2-T_1T_\Sigma+T_\Sigma^2)s}{\Delta_L(s)\vert_{\beta=9}}$	$\dfrac{1+9\frac{T_1T_\Sigma}{(T_1+T_\Sigma)^3}(T_1^2+T_\Sigma^2)s}{\Delta(s)\vert_{\beta=4}}$	$\dfrac{8T_\Sigma' \frac{m}{(1+m)^2}s(1+sT_1)(1+sT_\Sigma)}{\Delta(s)\vert_{\beta=4}}$	$\dfrac{8T_\Sigma' \frac{m}{(1+m)^2}s}{\Delta(s)\vert_{\beta=4}}$
9	$\dfrac{1+9\frac{T_1T_\Sigma}{(T_1+T_\Sigma)^3}(T_1^2-T_1T_\Sigma+T_\Sigma^2)s}{\Delta_L(s)\vert_{\beta=9}}$	$\dfrac{1+9\frac{T_1T_\Sigma}{(T_1+T_\Sigma)^3}(T_1^2-T_1T_\Sigma+T_\Sigma^2)s}{\Delta(s)\vert_{\beta=9}}$	$\dfrac{27T_\Sigma' \frac{m}{(1+m)^2}s(1+sT_1)(1+sT_\Sigma)}{\Delta(s)\vert_{\beta=9}}$	$\dfrac{27T_\Sigma' \frac{m}{(1+m)^2}s}{\Delta(s)\vert_{\beta=9}}$
16	$\dfrac{1+16\frac{T_1T_\Sigma}{(T_1+T_\Sigma)^3}(T_1-T_\Sigma)^2s}{\Delta_L(s)\vert_{\beta=16}}$	$\dfrac{1+16\frac{T_1T_\Sigma}{(T_1+T_\Sigma)^3}(T_1-T_\Sigma)^2s}{\Delta(s)\vert_{\beta=16}}$	$\dfrac{64T_\Sigma' \frac{m}{(1+m)^2}s(1+sT_1)(1+sT_\Sigma)}{\Delta(s)\vert_{\beta=16}}$	$\dfrac{64T_\Sigma' \frac{m}{(1+m)^2}s}{\Delta(s)\vert_{\beta=16}}$

β		
β	$\Delta_L(s)\vert_\beta=\beta^{3/2}T_\Sigma'\frac{m}{(1+m)^2}s(1+sT_1)(1+sT_\Sigma)$	$\Delta(s)\vert_\beta=\beta^{3/2}T_\Sigma'^3s^3+\beta^{3/2}T_\Sigma'^2s^2+\beta T_\Sigma's+1=(1+\beta T_\Sigma's)[1+(\beta-\sqrt{\beta})T_\Sigma's+\beta T_\Sigma'^2s^2]$
4	$\Delta_L(s)\vert_{\beta=4}=8\frac{T_\Sigma^2T_1^2}{(T_1+T_\Sigma)^3}s(1+sT_1)(1+sT_\Sigma)$	$\Delta(s)\vert_{\beta=4}=8T_\Sigma'^3s^3+8T_\Sigma'^2s^2+4T_\Sigma's+1=(1+2T_\Sigma's)(1+2T_\Sigma's+4T_\Sigma'^2s^2)$
9	$\Delta_L(s)\vert_{\beta=9}=27\frac{T_\Sigma^2T_1^2}{(T_1+T_\Sigma)^3}s(1+sT_1)(1+sT_\Sigma)$	$\Delta(s)\vert_{\beta=9}=27T_\Sigma'^3s^3+27T_\Sigma'^2s^2+9T_\Sigma's+1=(1+3T_\Sigma's)[1+6T_\Sigma's+9T_\Sigma'^2s^2]$
16	$\Delta_L(s)\vert_{\beta=16}=64\frac{T_\Sigma^2T_1^2}{(T_1+T_\Sigma)^3}s(1+sT_1)(1+sT_\Sigma)$	$\Delta(s)\vert_{\beta=16}=64T_\Sigma'^3s^3+64T_\Sigma'^2s^2+16T_\Sigma's+1=(1+4T_\Sigma's)[1+12T_\Sigma's+16T_\Sigma'^2s^2]$

Additional relations: $\Delta_m(m)=[1+(2-\beta^{1/2})m+m^2]$, $T_{\Sigma m}=T_\Sigma'\dfrac{\Delta_m(m)}{(1+m)^2}$, $T_\Sigma'=\dfrac{T_\Sigma}{(1+m)}$ (3.1-15)

The plant parameters' are presented in Tables 3.1-4, $k_p=1$, $T_\Sigma=1$, T_1. The controller is calculated using the following expressions:

- for MO-m case: (1): k_c given by rel.(2.2-7) and $T_c = T_1$ (2) $T_c' = T_2$
- for 2p-SO-m case (1): k_c given by (3.1-14) and T_c given by (3.1-15)
- for 2p-S0-m case (2): only rel. $T_c' = T_2$ is added.

For the analysis is useful to table the parameters T_Σ'/T_Σ and $T_{\Sigma m}/T_\Sigma$, Table 3.1-5.

Table 3.1-4. Values of controller parameters, case (1)

			$m = T_\Sigma / T_1$ $(T_\Sigma=1)$					
			0.05	0.1	0.15	0.2	0.25	0.5
Plant Parameters		T_1	20	10	6.66	5	4	2
$k_p=1$, $T_\Sigma=1$		T_2	-	-	-	-	-	-
Controller Param.								
MO-m		k_c	0.5	0.5	0.5	0.5	0.5	0.5
		T_c	20	10	6.66	5	4	2
2p-SO-m	β=4	k_c	3.4	1.25	0.836	0.625	0.5	0.25
		T_c	3.46	3.03	2.69	2.41	2.18	1.48
	β=9	k_c	0.75	0.37	0.25	0.19	0.15	0.09
		T_c	7.46	6.16	5.17	4.47	3.75	2.00
	β=16	k_c	0.3.12	0.116	0.104	0.078	0.064	0.035
		T_c	12.48	9.74	7.6	5.92	4.61	1.19

Table 3.1-5. Values of coefficients T_Σ'/T_Σ and $T_{\Sigma m}/T_\Sigma = f_2(m, \beta)$

β		$m = T_\Sigma / T_1$					
	$\frac{T_\Sigma'}{T_\Sigma} = \frac{1}{(1+m)}$	0.05	0.1	0.15	0.2	0.25	0.5
		0.952	0.91	0.87	0.83	0.8	0.66
4	$\frac{T_{\Sigma m}}{T_\Sigma} = \frac{(1+m^2)}{(1+m)^3}$	0.866	0.759	0.672	0.602	0.544	0.370
9	$\frac{T_{\Sigma m}}{T_\Sigma} = \frac{(1-m+m^2)}{(1+m)^3}$	0.823	0.684	0.574	0.486	0.416	0.222
16	$\frac{T_{\Sigma m}}{T_\Sigma} = \frac{(1-m)^2}{(1+m)^2}$	0.780	0.609	0.475	0.370	0.288	0.074

3.1.2. Control system performance

A. **Performances in the time-domain.**

- **System performance regarding the reference input.** Figures 3.1-1 (a),(b),(c) and (d) highlights the unit step reference response, y(t). The values for the performance indices are synthesized in Table 3.1-6. Some conclusions are available:
  - For small values of m (0.05, 0.1) the first settling time $t_{1,r}$ proves to be convenient even if the overshoot and the settling time is bigger. By increasing the value of β and m, this advantage disappears;
  - For the same m value, by increasing the value of β, the overshoot σ_1 decreases and the oscillations are diminished; for values of β between 4 ... 6 the settling time decreases, then it increases by further increasing β.

- By increasing the value of m, the overshoot decreases.

Table 3.1-6. The values for performance indices regarding the step reference input

		2p-SO-m					
		β					
m		4	5	6	7	8	9
0.05	$\sigma_{1,r}$	37.5	30.3	25.1	21.3	17.8	15.2
	$\hat{t}_{1,r}$	3.2	3.7	4.3	5.0	5.7	6.4
	$\hat{t}_{s,r}$	32.0	26.7	24.2	26.7	30.5	35.3
0.10	$\sigma_{1,r}$	29.6	22.0	17.0	13.9	11,5	9.7
	$\hat{t}_{1,r}$	3.9	4.7	5.2	6.3	7.4	8.5
	$\hat{t}_{s,r}$	29.0	24.2	23.7	25.3	28.7	33.0
0.15	$\sigma_{1,r}$	21.7	15.6	11.7	9.0	7.0	5.6
	$\hat{t}_{1,r}$	4.5	5.3	6.7	8.0	9.7	11.3
	$\hat{t}_{s,r}$	26.8	22.4	21.2	23.7	27.0	30.8
0.20	$\sigma_{1,r}$	18.2	13.4	8.0	4.7	2.0	<2.0
	$\hat{t}_{1,r}$	5.1	6.3	8.0	9.4	12.0	14.5
	$\hat{t}_{s,r}$	24.6	20.7	19.0	21.0	24.7	28.4

- In case of a ramp input the static error is non-zero, having the value:

$$e_{r\infty} = \lim_{s \to 0}\{s\,S_0(s)\frac{1}{s^2}\} = \lim\{s\,\frac{\beta^{3/2}T'_\Sigma \frac{m}{(1+m)^2}\,s(1+sT_1)(1+sT_\Sigma)}{\beta^{3/2}T'^3_\Sigma s^3 + \beta^{3/2}T'^2_\Sigma s^2 + \beta T'_\Sigma s + 1}\,\frac{1}{s^2}\} = \beta^{3/2}T'_\Sigma\frac{m}{(1+m)^2}$$

or, with $T'_\Sigma = T_\Sigma/(1+m)$ results:

$$e_{r\infty} = \beta^{3/2}T_\Sigma\frac{m}{(1+m)^3} \quad (3.1\text{-}25)$$

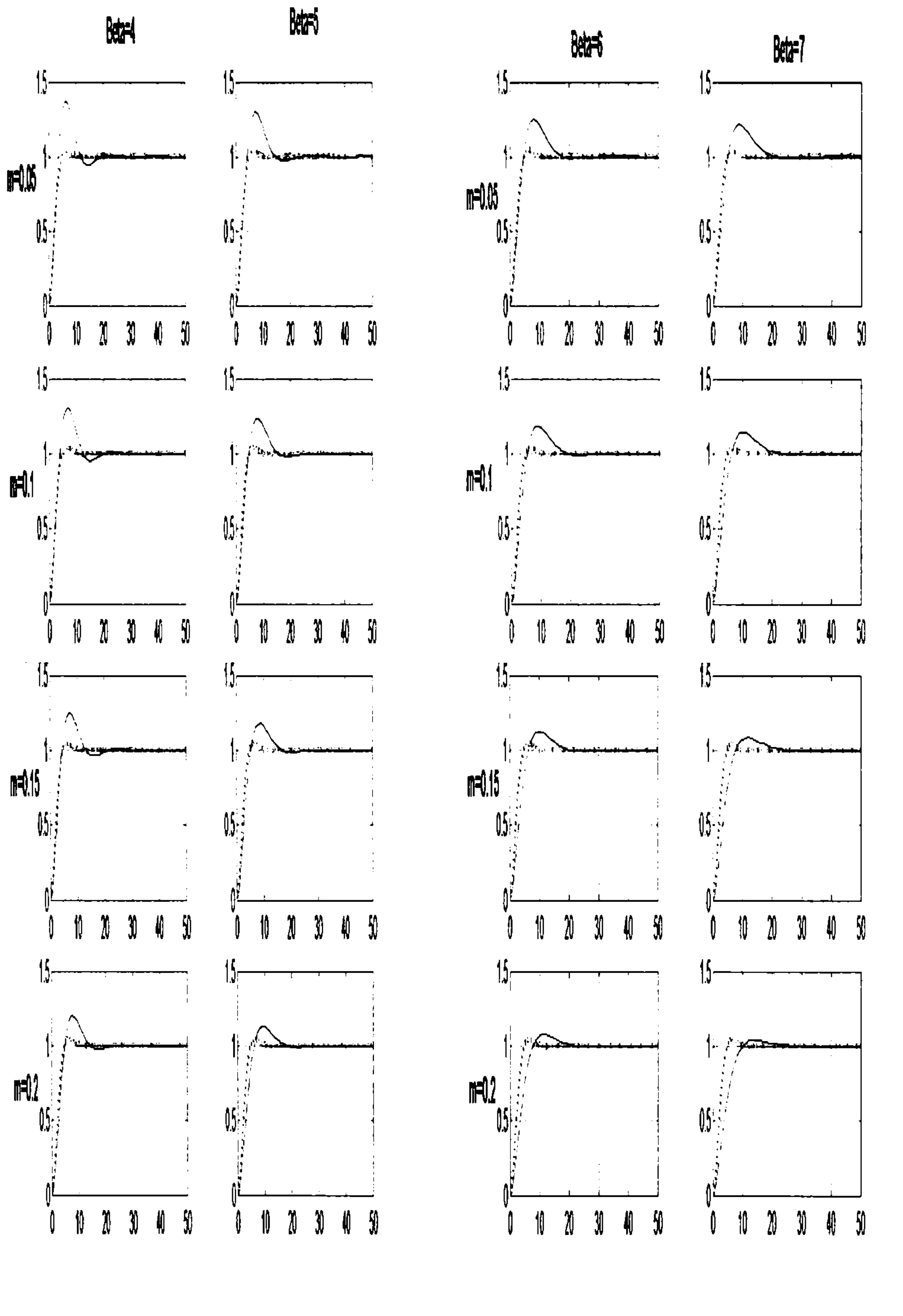
For values of β bigger than 9, the transients are slower if m is increased, the trend being to become a-periodic. The performance indices depicted in Fig.3.1-2 are based on simulation results and interpolation; the curves have m as parameter. Because of pole-zero cancellation the values for the performance indices for MO-m are independent from m.

***B.* Improving the reference performances.**

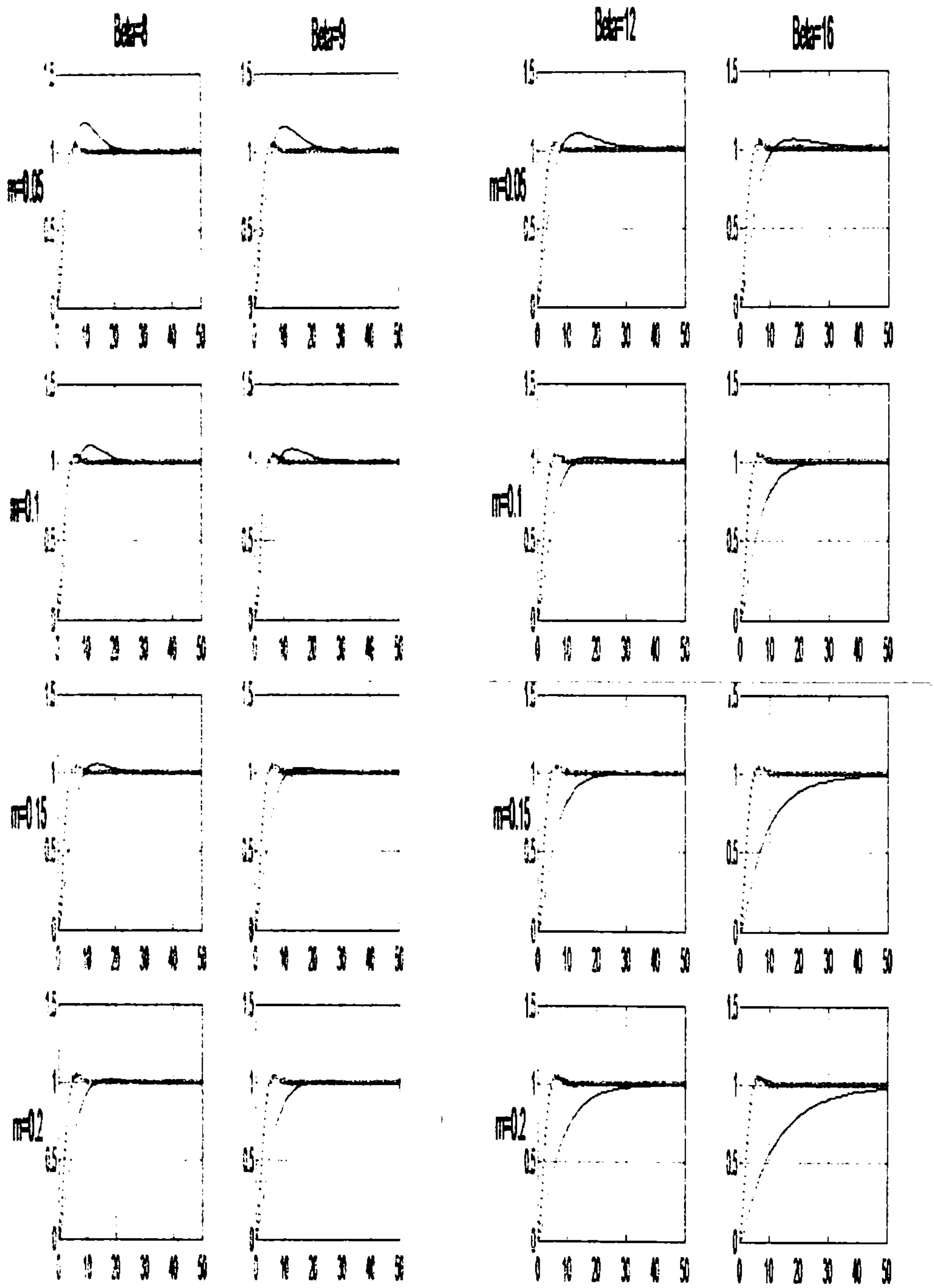
Regarding the reference performances, two ways for correcting the CS performances are applicable.

- o Using a proper reference filter: referring to rel. (3.1-17), two types of proper reference filters $F_{r0}(s)$ are available:
- One which suppresses the effect of the zero in the close-loop t.f.:

$$F_{r0}(s) = \frac{1}{1+\beta T_{\Sigma m}s} \quad \text{and} \quad \tilde{H}(s) = \frac{1}{(1+\beta^{1/2}T'_\Sigma s)[1+(\beta-\beta^{1/2})T'_\Sigma s+\beta T'^2_\Sigma s^3]} \quad (3.1\text{-}26)$$


Beta=4
Beta=5
Beta=6
Beta=7
m=0.05
m=0.1
m=0.15
m=0.2
1.5
1
0.5
0
0 10 20 30 40 50

(a) (b)



(c) (d)

Fig.3.1-1. Unit step reference response with β and m- parameters

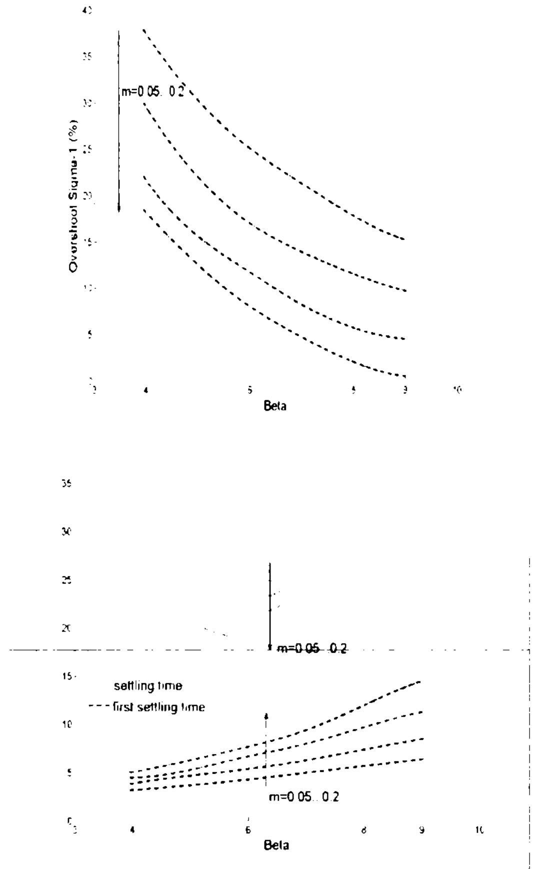


Fig.3.1-2. System performances regarding the reference input $\sigma_{1,r}, t_{s,r}, t_{1,r} = f(\beta)$, m parameter

- A second one that suppresses the effect of the zero and the pair of poles in square parentheses:

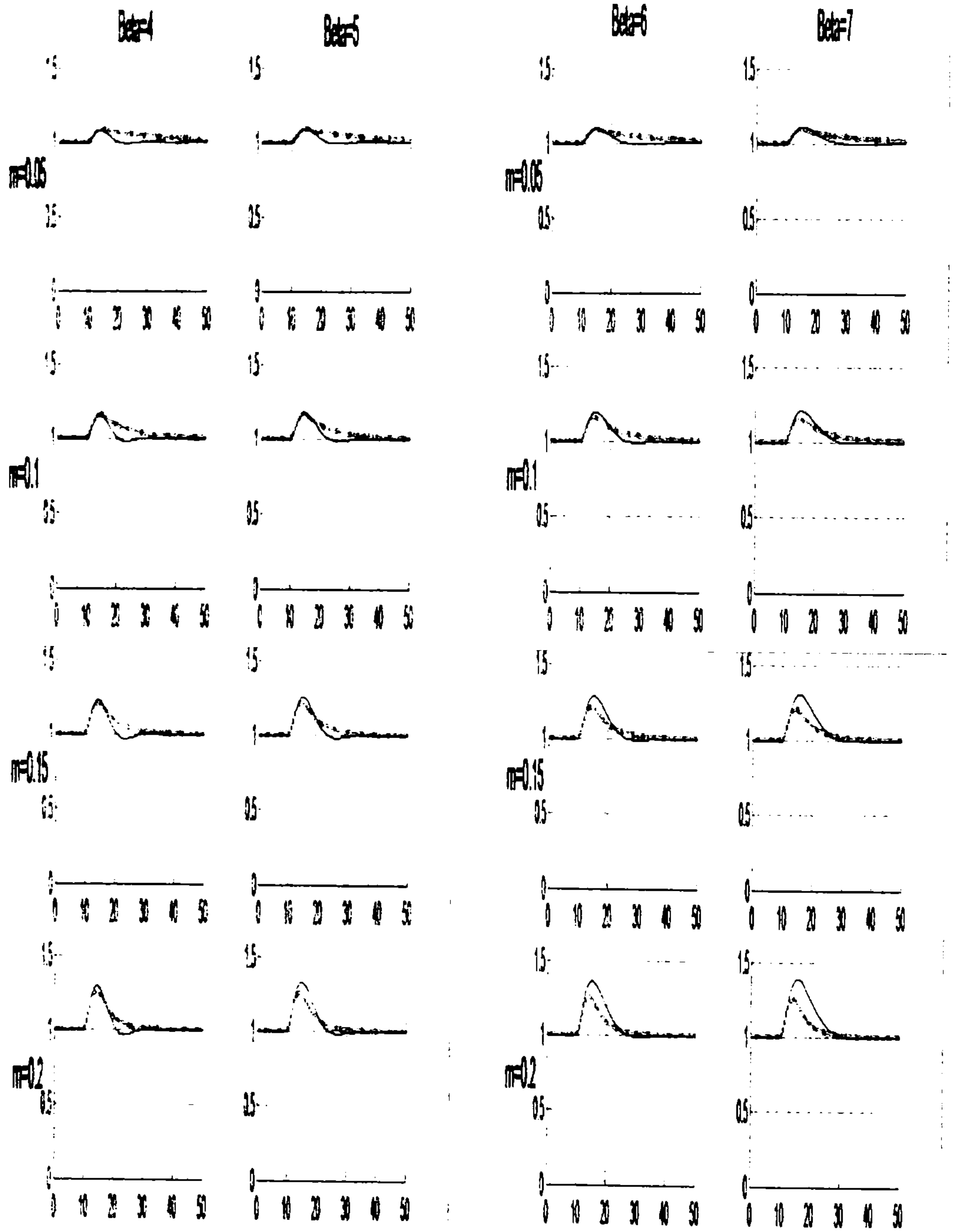
$$F_{r0}(s) = \frac{1 + (\beta - \beta^{1/2})T_{\Sigma}'s + \beta T_{\Sigma}'^{2}s^{2}}{(1 + \beta T_{\Sigma m}s)(1 + sT_{f}s)} \quad \text{and} \quad \tilde{H}_{r}(s) = \frac{1}{(1 + \beta T_{\Sigma}'s)(1 + sT_{f})} \qquad (3.1\text{-}27)$$

- o Using controllers with non-homogenous structure, which ensure different behaviour regarding the reference tracking and disturbance rejection. If PI (PID) controllers are used, the derivative component is shifted only into the feedback, chapter1, Fig.1.1-1 (b). The forms were used in applications described in [II-18] - [II-49]. This approach leads to a 2-DOF controller [II-70].

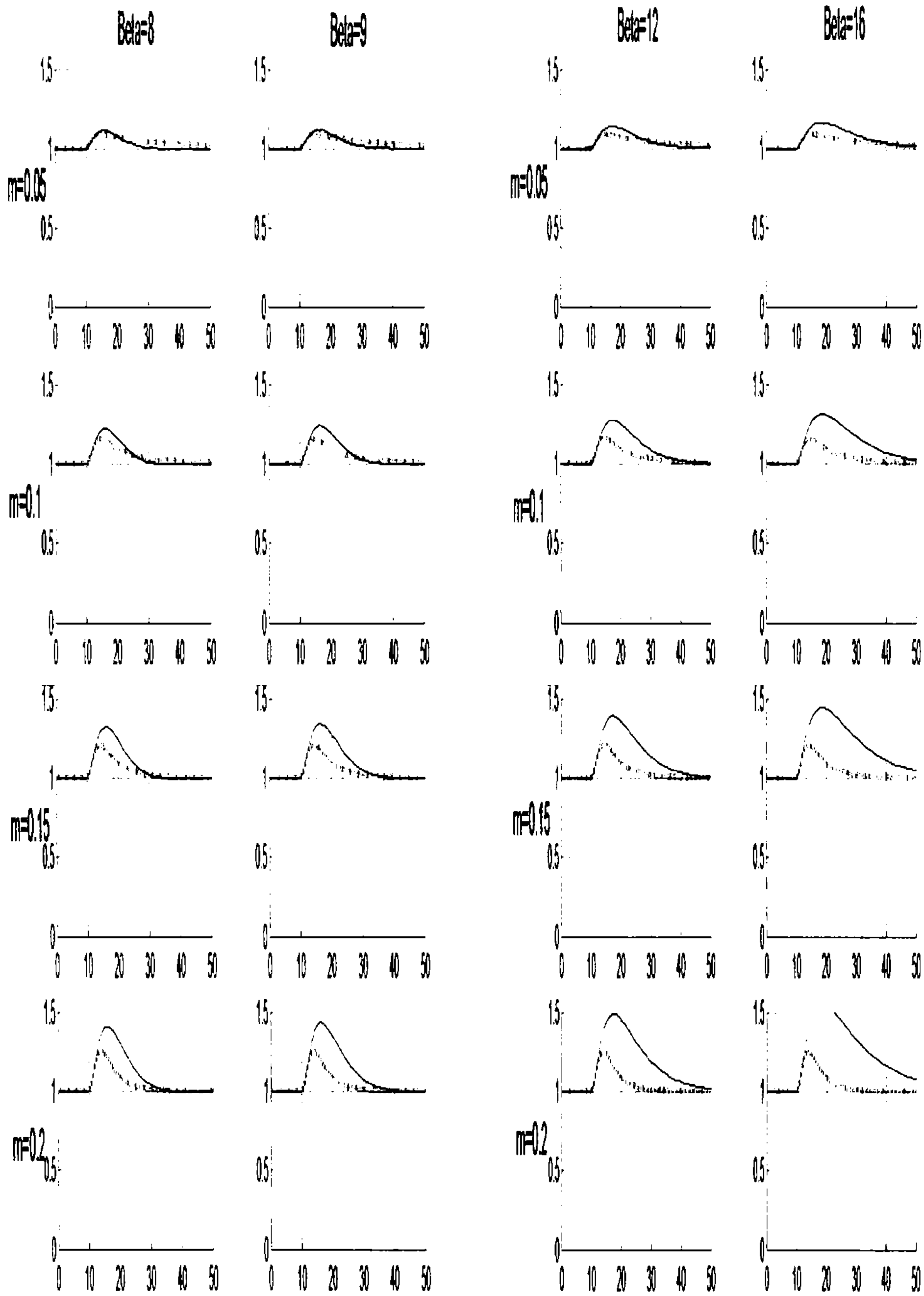
***C.* System performance regarding the load disturbance**

The duration of the load disturbance response and the maximum deviation of the output are of main interest [II-51], [II-52]. Figures 3.1-3 (a),(b),(c) and (d) highlights the step load disturbance response, y(t). From Figures 3.1-3 the following conclusions can be drawn:

- For the same value of $m=T_{\Sigma}/T_{1}$, by increasing the value of β, the overshoot increases;
- By increasing the value of m the overshot increase.



(a) (b)



(c) (d)

Fig.3.1-3. 2p-SO-m Unit step load disturbance response, with β and m – parameters

- Compared with MO-m performances, for m< 0.15 (also depending on β, recommended domain 4 < β ≤ 9 (12)) the effect of load disturbance is faster rejected. The smaller m is ($T_1 >> T_\Sigma$) the more favourable this property is.

It can be observed that from this point of view at, an increase of β over the value of 9 the use 2p-SO-m does not offer benefits. Also for MO-m differences in transients in disturbance rejection occur (highlighted with bold). Table 3.1-7 synthesizes the values of performance indices regarding the load disturbance obtained through simulation and interpolation. The domains of {m, β} for which the use of 2p-SO-m is advantageous/partially advantageous regarding the load disturbance are highlighted with bold.

Table 3.1-7. The values of performance indices regarding the load disturbance

		MO-m	2p-SO-m Value of β					
m			4	5	6	7	8	9
0.05	$\hat{t}_{s,d2}$	**45,5**	**9.2**	**11.1**	**13.0**	**14.9**	**17.5**	**19.8**
	$\sigma_{1,d2}$	**9.3**	**7.7**	**8.7**	**9.7**	10.5	11.4	12.3
0.10	$\hat{t}_{s,d2}$	**28.7**	**10.6**	**12.6**	**14.5**	**17.1**	**19.6**	**23.4**
	$\sigma_{1,d2}$	**15.7**	**15.3**	17.4	19.1	20.8	22.1	2352
0.15	$\hat{t}_{s,d2}$	**19.7**	**15.2***	**17.9***	**13.9**	**16.7**	**19.7**	22.7
	$\sigma_{1,d2}$	**21.3**	22.9	25.4	28.3	30.1	32.1	34.0
0.20	$\hat{t}_{s,d2}$	**17.6**	**17.6***	**13.1**	**16.1**	19.5	22.4	26.8
	$\sigma_{1,d2}$	**25.9**	29.7	32.8	36.1	38.1	40.8	42.7

Remark: $t_{s,d2} = \hat{t}_{s,d2} T_\Sigma$. For the cases marked with * the system is quite oscillatory, consequently the steady state value (± 2 %) is reached later. The dashed values are in the recommended domain or strictly near of.

Fig.3.1-4 presents the performance indices regarding the load disturbance with m in horizontal axis and β parameter.

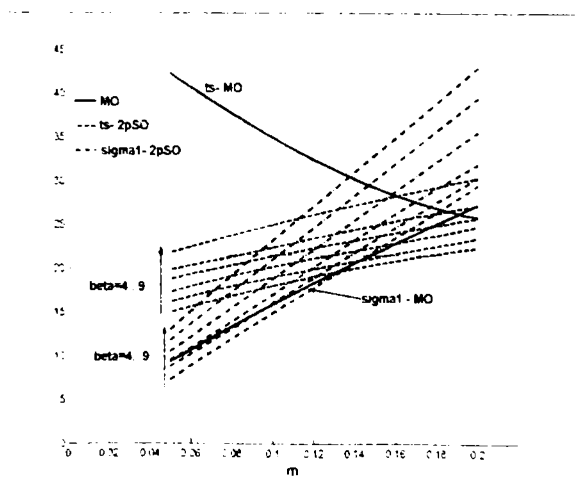


Fig.3.1-4. System performances regarding the load disturbance $\sigma_{1,d2}$, $t_{s,d2} = f(\beta)$, m -parameter

***D.* Analysis in frequency domain.**

- **Sensitivity function analysis.** To characterise the sensitivity of CS, for β= **4,** 5, **6,** 7, 8, **9,** 12, **16** and m={0.05, 0.10, 0.15, 0.20} the modulus of S(jω) and the maximum sensitivity value M_{s0} and its inverse M_{S0}^{-1} were calculated, Table 3.1-9. The values of ω_c and φ_r are calculated and represented in Table 3.1-9 for the cases 2p-SO-1,

Table 3.1-8. The values for M_{S0} and M_{S0}^{-1}

m ╲ β		4	5	6	7	8	9	12	16
		M_{s0} / M^{-1}_{s0}							
0.05	M_{s0}	1.602	1.45	1.36	1.303	1.263	1.235	1.180	1.14
	M^{-1}_{s0}	**0.624**	**0.690**	**0.735**	**0.767**	**0.792**	**0.810**	**0.847**	**0.876**
0.10	M_{s0}	1.529	1.385	1.302	1.248	1.212	1.185	1.136	1.103
	M^{-1}_{s0}	**0.654**	**0.722**	**0.768**	**0.801**	**0.825**	**0.844**	**0.880**	**0.907**
0.15	M_{s0}	1.464	1.330	1.255	1.206	1.172	1.149	1.106	1.076
	M^{-1}_{s0}	**0.683**	**0.752**	**0.797**	**0.829**	**0.853**	**0.870**	**0.904**	**0.929**
0.20	M_{s0}	1.406	1.285	1.217	1.172	1.143	1.122	1.083	1.058
	M^{-1}_{s0}	0.711	0.778	0.822	0.853	0.875	0.891	0.923	0.945

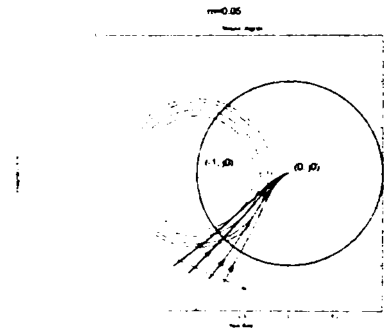
Remark: for MO-m $M_{s\max} = \max\{|S_0(j\omega)|\} = 1.272$ and M_{S0}^{-1}=0.786; the dashed values are in the recommended domain ($1.2 < M_{s\max} < 2$) or strictly near of.

Table 3.1-9. The crossover frequency and the value of the phase reserve; the maximum magnitude of the frequency response: M_p=max|T(jω)|=1 (ω→0)

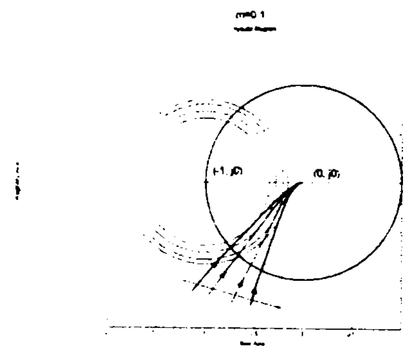
m		4	5	6	7	8	9	12
		β						
0.05	ω_c	0.461	0.406	0.365	0.334	0.308	0.287	0.241
	φ_r	**39.4**	**45.0**	**49.4**	**53.0**	**56.1**	**58.7**	**64.9**
0.10	ω_c	0.428	0.371	0.328	0.295	0.268	0.246	0.196
	φ_r	**42,4**	**48,7**	**53.8**	**58.0**	**61.7**	**64.9**	**72.7**
0.15	ω_c	0.400	0.340	0.295	0261	0.232	0.208	0.155
	φ_r	**45.8**	**52.8**	**58.5**	**63.3**	**67.5**	**71.1**	**79.7**
0.20	ω_c	0.374	0.312	0.265	0.228	0.199	0.174	0.122
	φ_r	**49.6**	**57.2**	**63.4**	**68.5**	**72.7**	**76.4**	**84.1**

Remark: For MO-m $\varphi_r=60^0$ at $\omega_c = 1/2T_\Sigma$ (for $T_\Sigma = 1$ results $\omega_c = 0.5$); the dashed values are in the recommended domain or strictly near of.

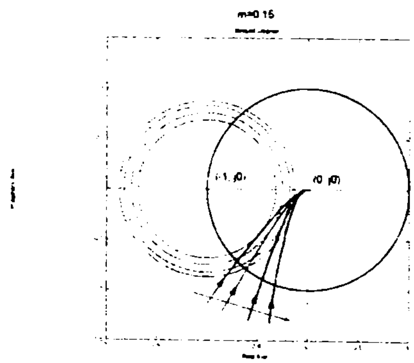
Figure 3.1-5 (a),(b),(c),(d) present the calculated Nyquist plots and the M_{S0}^{-1} circles are marked for the values with bold. The curves point out for each m the increase of robustness when the value of β is increased.



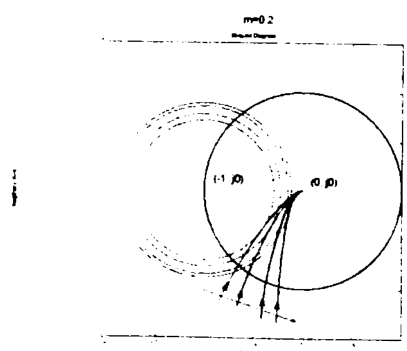
(a)



(b)



(c)



(d)

The sens for β - increase

Fig.3.1-5. Nyquist curves and M_{S0}^{-1} circles for different m and β and the $M_{S0}^{-1}=f(\beta)$ circles

BUPT

Based on (3.1-16) the expression of the phase margin φ_r can be expressed as:

$$\varphi_r = \arg\{H_0(j\omega_c)\} + \pi = \operatorname{arctg}(\beta T_{\Sigma m}\omega_c) - \operatorname{arctg}(T_1\omega_c) - \operatorname{arctg}(T_\Sigma\omega_c) + \pi/2 \quad (3.1\text{-}28)$$

the calculated diagram of curves φ_r as function of {β, m} are depicted in Fig. 3.1-6.

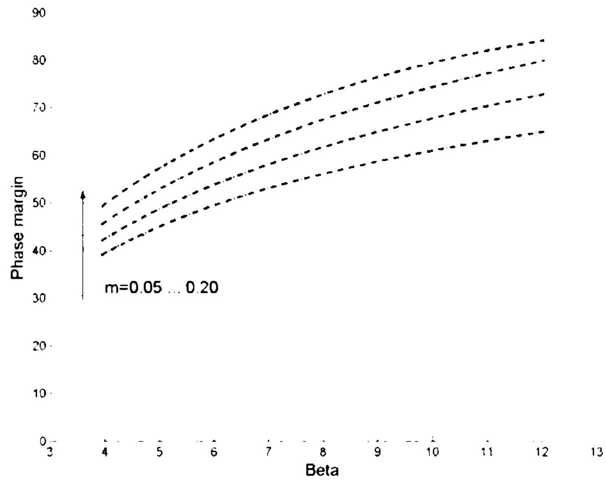


Fig.3.1-6. The phase margin (φ_r) versus β curves, m-parameter

- **Bode Diagrams and phase margins.** Based on relation (3.1-18) for the same m and β parameter the Bode diagrams were calculated and represented in figure 3.1-7 (a) ...(d). Some main remarks can be mentioned:
  - Even for small values of m the approximation of a lag element with an integrating one (see relations (2.3-3), (2.3-4)) is questionable.
  - The increase of the phase margin through m and β sustain a low sensitivity design of systems with time varying parameters.
  - For every value of β, as far as the phase margin is concerned, the 2p-SO tuning method ensures the cut-off frequency in the immediate vicinity of the maximal phase value. This situation is obvious for the cases $m \le 0.05$ (at the limit $m < 0.10$) .
  - The phase reserve diagram allows a purpose oriented supplementary gain adjustment.
    - For the case of variable k_p, solving equation (3.1-28) for $\varphi_r = \varphi_{r,min}$ (an imposed value), the 2p-SO-m method can guarantee a frequency margin for which the phase margin is larger than a minimum value.
    - The minimal value for β which ensures positive phase margin can be determined based on relations (3.1-19) and (3.1-28), respectively of the condition:

$$\beta T_{\Sigma m} = \beta T_\Sigma \frac{\Delta_m(m)}{(1+m)^3} > T_\Sigma \quad (3.1\text{-}29)$$

- **Magnitude plot of the complementary sensitivity function.** For m and β - parameters, the graphics of $M_p(\omega) = |H_{r0}(j\omega)|$ are calculated and depictured in Figure 3.1-8; its maximal value, M_{pmax} is synthesized in Table 3.1-10. By increasing the value of β the value of $M_{p\ max}$ decreases, the system becomes less and less oscillatory.

Table 3.1-10 The maximal value $M_{p\,max}$

	β						
m	4	5	6	7	8	9	12
0.05	**1.573**	**1.415**	**1.321**	**1.257**	**1.211**	**1.176**	**1.104**
0.10	**1.456**	**1.303**	**1.210**	**1.147**	**1.102**	**1.067**	**1.008**
0.15	**1.343**	**1.199**	**1.114**	**1.058**	**1.023**	**1.004**	0.998
0.20	**1.241**	**1.113**	**1.042**	**1.006**	0.999	0.998	0.997

Remark: the dasched values are in the recommended domain or strictly near of.

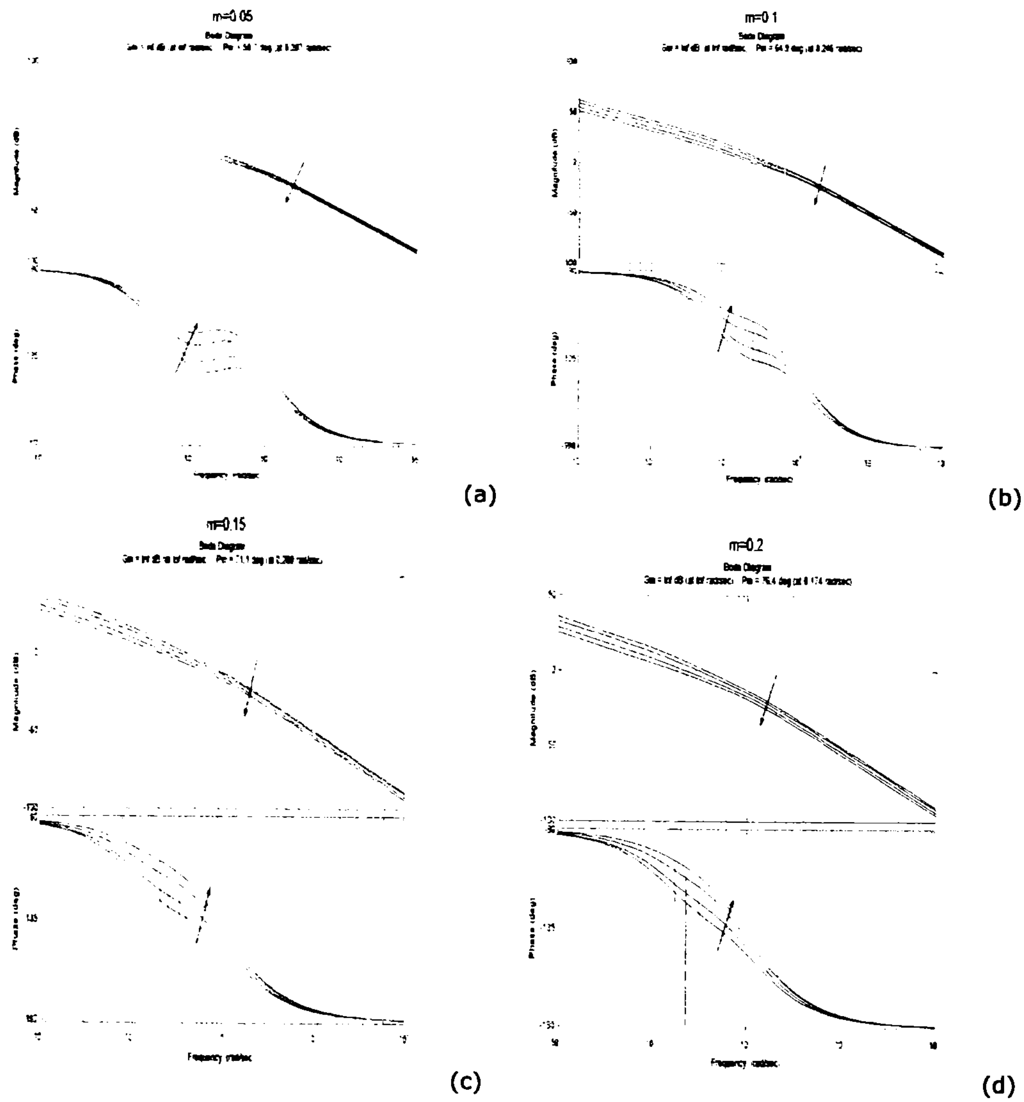
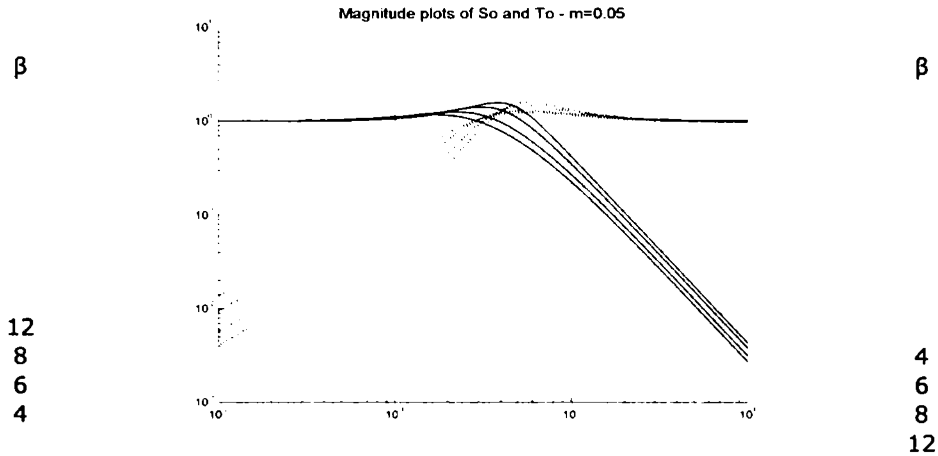
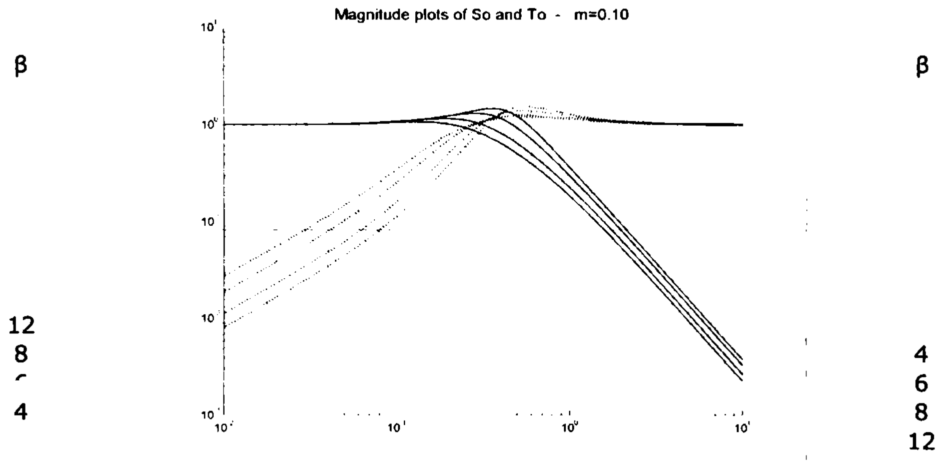


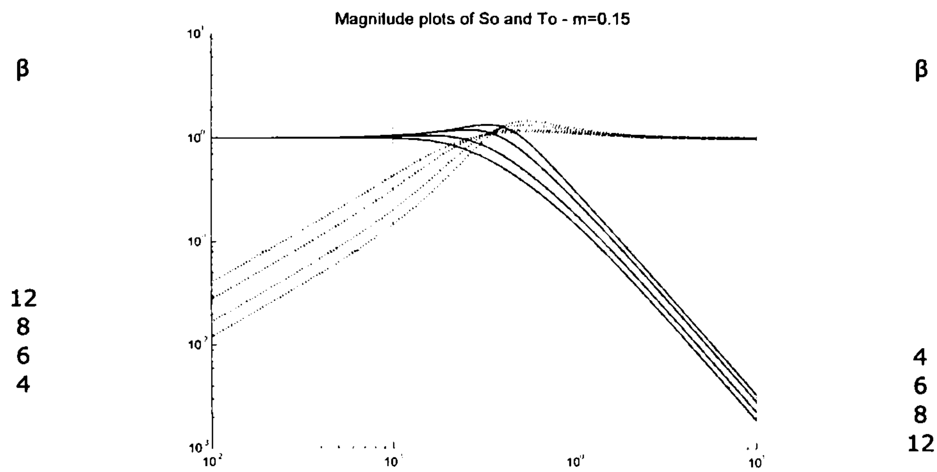
Fig.3.1-7. Bode diagrams for m and β parameters


Magnitude plots of So and To - m=0.05
β
β
12
8
6
4
4
6
8
12

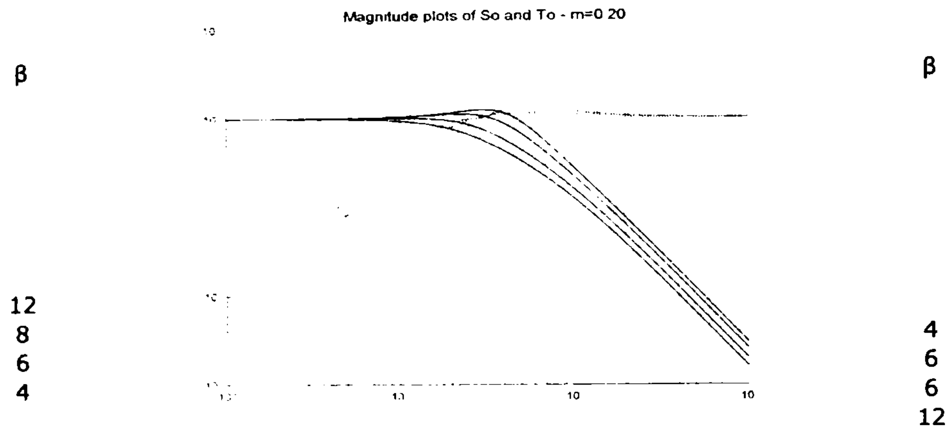
(a)


Magnitude plots of So and To - m=0.10
β
β
12
8
4
4
6
8
12

(b)


Magnitude plots of So and To - m=0.15
β
β
12
8
6
4
4
6
8
12

(c)



(d)

Fig.3.1-8. Magnitude plot of the sensitivity and complementary sensitivity function, {m, β} parameters

3.2. Choose of controller design method. Design methodology and steps

3.2.1. Choose of controller design method

For many electric driving applications the t.f. of the plant can be accepted to have benchmark type form, see tables 2.2-1 and 2.3-1. Imposing certain requirements (see Chapter 1) regarding variable reference tracking, load disturbance rejection and robustness to parameter changes reduces the area of usable design methods. The 2p-SO-m is an alternative to the MO-m for the cases defined in paragraph 3.1. The comparison of the two methods is interesting only for the value (driving systems with big inertia) values for which the method is destined. The parameterization with β: $4 < \beta < 9\ (16)$ extends the design options.

The design is performed in continuous time while the implementation can be also in discrete time [II-95], [II-89]. The main aspects which must be taken into account when designing a controller for driving system with big inertia are:

- Reduced control error and good behaviour regarding a slowly changing reference;
- Good behaviour regarding load disturbances;
- The possibility to control the phase margin through an adeguately chosen value for β ;
- Reduced sensitivity regarding plants parameter changes through an adequately chosen value for β ;
- The CS performances can improved by use reference filters;

Tables A to D give information for helping decide between the MO- and 2p-SO-methods to methods:

- Table A highlights the t.f. for the cases when 2p-SO-m is recommended to be used;

- Table B shows that from the point of view of reduce values for the first settling time $\hat{t}_{1,r}$, 2p-SO-m is advantageous for small values for m and values of β in interval $\beta <$ (6) 9;
- Table C shows that from the point of view of variable input the 2p-SO-m is much better than MO-m;
- Table D. The performance indices regarding the load disturbance reflect a better behaviour of 2p-SO-m compared with MO-m;

Tables which sinthesize the 2p-SO-m compared with MO-m.

Table A. Recommended situations for design with 2p-SO-m compared with MO-m

Case	The plant t.f. $H_p(s)$	Recom. method
1.	$\frac{k_p}{1+sT_\Sigma}$	MO-1.1
2.	$\frac{k_p}{(1+sT_\Sigma)(1+sT_1)}$ $T_1 >> T_\Sigma$	MO-2.1
		2p-SO-m
3.	$\frac{k_p}{(1+sT_\Sigma)(1+sT_1)(1+sT_2)}$ $T_1 > T_2 >> T_\Sigma$	MO-3.1
		2p-SO-m

Table B. Performances regarding the step reference input: 2p-SO-m compared with MO-m

m \ β		4	5	6	7	8	9	12
0.05	$\sigma_{1,r}$							
	$\hat{t}_{1,r}$							
	$\hat{t}_{s,r}$							
0.10	$\sigma_{1,r}$							
	$\hat{t}_{1,r}$							
	$\hat{t}_{s,r}$							
0.15	$\sigma_{1,r}$							
	$\hat{t}_{1,r}$							
	$\hat{t}_{s,r}$							
0.20	$\sigma_{1,r}$							
	$\hat{t}_{1,r}$							
	$\hat{t}_{s,r}$							

Table C. Performances regarding the ramp reference input: 2p-SO-m compared with MO-m

β / m	4	5	6	7	8	9	10	12	14	16
0.05										
0.10										
0.15										
0.20										
0.25										

Table D. The performance indices regarding the load disturbance: 2p-SO-m compared with MO-m

m	β	4	5	6	7	8	9
0.05	$\hat{t}_{s,d2}$						
	$\sigma_{1,d2}$						
0.10	$\hat{t}_{s,d2}$						
	$\sigma_{1,d2}$						
0.15	$\hat{t}_{s,d2}$						
	$\sigma_{1,d2}$						
0.20	$\hat{t}_{s,d2}$						
	$\sigma_{1,d2}$						

Remarks:

- much better performance of the 2p-SO-m compared with MO-m — RECOMMENDED
- similar performance of the 2p-SO-m compared with MO-m — Recommended
- not so good performance of the 2p-SO-m compared with MO-m — Acceptable, and other points of view can be taken into account

The simultaneous fulfillment of all requirements necessitates some compromises, subsequently the 2p-SO-m recommended in the following cases:

- The value of parameter m lies within $0.05<m<0.20$ (0.25) values. If in such cases the approximation $m \approx 0$ was made and ESO-m was chosen then a correction in controller parameters would be required;
- Through the possibility of choosing β in the recommended domain $4<\beta<9\,(12)$ the design compromises can be fulfilled;
- For values of $m< 0.05$ the approximation $m \approx 0$ is justified.

3.2.2. Design methodology and steps

Considering the choice of the 2p-SO-m design method justified, the design steps are the following:

- Starting form the approximated form of the plant t.f. $H_p(s)$, one of the variants of the 2p-SO-m is chosen and value for m is determined (m_0);

- The possible modifications of m are estimated, due to the modification of the dominant time constant;
- The value of β is determined which satisfies the imposed performances for the nominal value m_0 and the parameter changes are evaluated for changes of a m;
- The controller $H_c(s)$ is chosen and its parameter calculated; if necessary, a reference filter $F_r(s)$ is added;
- The control structure is extended with supplementary functions:
    - Limitations of the control signal, AWR measure,
    - Limitations of the actuator output,
    - Feed-forward filters;
- Taking into account the neglected aspects of the plant (or design), further corrections and fine-tuning can be expected;
- Step by step the solutions are verified.

3.3. Conclusions. The main advantages of the 2p-SO-m

The 2p-SO-m can be considered as a generalized form of the classical SO-m and of the ESO-m. The main advantages of the 2p-SO-m are synthesized in the followings:

- The method is "optimal in the magnitude" regarding the (slow) variable reference only for small values of m ($m \leq 0.05$) and $\beta=4$ by guaranteeing a value close to the maximum value of the phase margin for constant plant parameters;
- The increase of β ensures the reduction of system sensitivity and can ensure the increase of robustness (see Figure 3.1-5). Anyhow, values of $\beta>9$ are not justified.
- For $0.05<m\leq 0.20$ (0.25) the 2p-SO-m design method ensures good behavior regarding load disturbances.
- For the case of variable k_p, solving equation (3.1-30) for an imposed value $\varphi_r = \varphi_{r,min}$, the 2p-SO-m method can guarantee a frequency margin for which the phase margin is larger than a minimum value.

3.4. Youla parameterization for the MO-m, ESO-m and 2p-SO-m

The Youla parameterization (called also Q-parameterization) is a design method applied for both stable and unstable plants [II-54]-[II-59], [II-62]. The Youla parameterization requires polynomials relative to the system's properties. The disadvantage of the method consists in the fact that in case of high order, non-minimum phase or unstable plants the controller results as a non conventional one. In this chapter the controller design method based on MO-m, ESO-m and 2p-SO-m [II-60], [II-61] is transposed in a Youla-parameterization form.

The use of Youla-parameterization in case of ESO-m and 2p-SO-m is justified by the possibility of imposing favorable forms of the expressions of S(s) and T(s) that ensures the desired system performances.

3.4.1. Preliminary aspects

If $G(s)$ is a bounded rational form $|G(j\omega)|<\infty$, with real coefficients there exists a co-prime factorization over the set of all bounded rational forms with real coefficients [II-61]:

$$G(s)=\frac{N(s)}{M(s)} \quad \text{with} \quad G(s)\in\varphi \tag{3.4-1}$$

$$N(s)X(s)+M(s)Y(s)=1 \quad \text{(Bezout's Identity)} \tag{3.4-2}$$

$N(s), X(s), M(s), Y(s)\in\varphi$, where φ is the set of all bounded rational forms with real coefficients.

The set of all stabilizing controllers for $G(s)$ (3.4-1) is defined as:

$$C(s)=H_c(s)=\frac{X(s)+M(s)Q(s)}{Y(s)-N(s)Q(s)} \quad \text{where } Q(s)\in\varphi \tag{3.4-3}$$

$Q(s)$ - represent the so-called Youla parameterization polynomial.

If the plant with t.f. $H_p(s)$ is stable, the co-prime factorization (3.4-1) and (3.4-2) can be particularized as:

$$N(s)=H_p(s) \quad , \quad M(s)=1 \quad , \quad X(s)=0 \quad , \quad Y(s)=1 \tag{3.4-4}$$

Accordingly, the controller (3.4-3), noted with $H_c(s)$, is determined as:

$$H_c(s)=\frac{Q(s)}{1-H_p(s)Q(s)} \tag{3.4-5}$$

Figures 3.4-1 present the basic block diagram (a) and the restructured block diagram (b) regarding the Youla parameterization of controller design. The following notations have been used: H_c – the controller, with t.f. $H_c(s)$, H_p – the plant, with t.f. $H_p(s)$, Q – the Youla parameterization polynomial; r_0 – reference, r – pre-filtered reference, e – error signal, u – control signal, y – measured output signal, d_1 – output acting plant disturbance, d_2 – input acting plant disturbance.

In all the cases treated here the plants are stable, characterized by bounded $H_p(s)$ t.f.-s $|H_p(j\omega)|<\infty$. The equations which characterize the structure in fig. 3.4-1 (a) (MISO - case) are (2.1-1) to (2.1-10).

For stable plants the Youla parameterization controller design consists in establishing Q(s) so that well stated requirements are fulfilled for S(s). From (2.1-1) and (2.1-9) it results:

$$S(s)=1-H_p(s)Q(s) \tag{3.4-6}$$

From relations (3.4-5) and (3.4-6) it results that S(s) and $H_c(s)$ depend only on Q(s) and $H_p(s)$.

Consequently, in controller design the following steps are made [II-60]:

Step (1): For the given stable plant with t.f. $H_p(s)$, calculation of $H_c(s)$ is performed with Q(s) as parameter. Calculations of S(s) and T(s) follow.

Step (2): Establish of a Q(s) through which the imposed performances for S(s) or T(s) are ensured.

Step (3): Establish the controller $H_c(s)$ that fulfills the imposed requirements.

Step (4): Verification of desired performances, sensitivity analysis of the system.

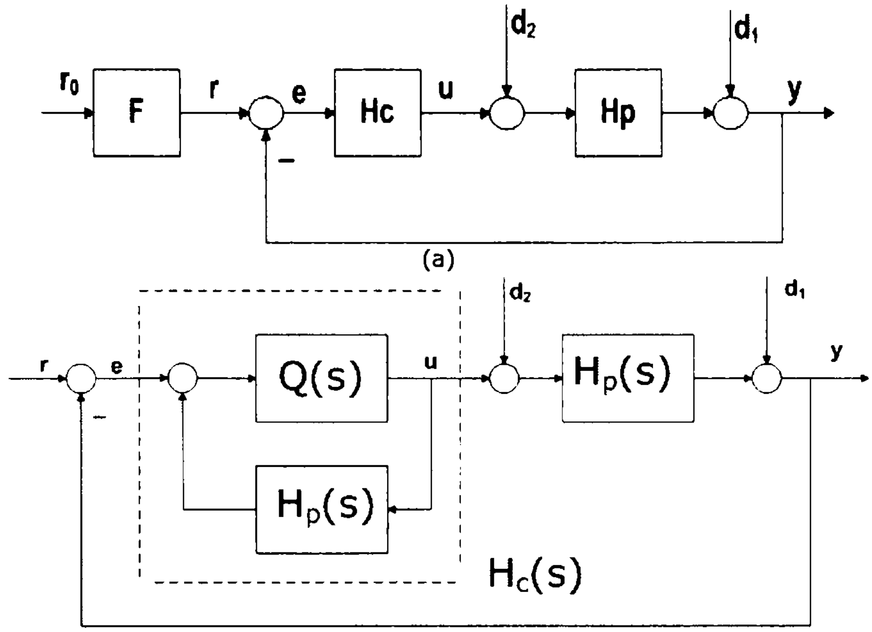


Fig. 3.4-1. Basic control structures

3.4.2. Youla parameterization for the MO-m

The plant t.f. is either (3.4-7) (1) or (3.4-7) (2):

$$H_P(s) = \frac{k_p}{(1+sT_\Sigma)(1+sT_1)} \ (1) \text{ or } H_P(s) = \frac{k_p}{(1+sT_\Sigma)(1+sT_1)(1+sT_2)} \ (2) \quad (3.4\text{-}7)$$

The design is exemplified only for t.f. (3.4-7) (1), the case (2) being solved similarly.

(1) Calculation of $H_c(s)$ with Q(s) parameter. Replacing (3.4-7) into (3.4-5) results:

$$H_c(s) = Q(s)\frac{(1+sT_\Sigma)(1+sT_1)}{(1+sT_\Sigma)(1+sT_1) - k_pQ(s)} \quad (3.4\text{-}8)$$

with the realizable expression for S(s) and T(s):

$$S(s) = \frac{(1+sT_\Sigma)(1+sT_1) - k_pQ(s)}{(1+sT_\Sigma)(1+sT_1)} \quad (3.4\text{-}9)$$

$$T(s) = \frac{k_pQ(s)}{(1+sT_\Sigma)(1+sT_1)} \quad (3.4\text{-}10)$$

(2) Establish the stable Q(s). Using (3.4-9) or (3.4-10) results:

$$Q(s) = \frac{1}{k_p}T(s)(1+sT_\Sigma)(1+sT_1) \quad \text{or}$$

$$Q(s)=\frac{1}{k_p}[1-S(s)](1+sT_\Sigma)(1+sT_1) \tag{3.4-11}$$

Using T(s) with its specific MO-m form and replacing into (3.4-10) results Q(s):

$$T(s)=\frac{1}{1+2T_\Sigma s+2T_\Sigma^2 s^2} \tag{3.4-12}$$

$$Q(s)=\frac{1}{k_p}\frac{(1+sT_\Sigma)(1+sT_1)}{1+2T_\Sigma s+2T_\Sigma^2 s^2} \tag{3.4-13}$$

(3) Establish a controller $H_C(s)$. finaly, the controller results as a PI controller:

$$H_c(s)=\frac{1}{k_p}\frac{(1+sT_\Sigma)(1+sT_1)}{1+2T_\Sigma s+2T_\Sigma^2 s^2}\frac{(1+sT_\Sigma)(1+sT_1)(1+2T_\Sigma s+2T_\Sigma s^2)}{(1+sT_\Sigma)(1+sT_1)[1+2T_\Sigma s+2T_\Sigma^2 s^2-1]}=\frac{1}{2k_pT_\Sigma s}(1+T_1 s)$$

The controller result as a:

$$k_c=\frac{1}{2k_pT_\Sigma}\ ,\quad T_c=T_1 \quad (3.4\text{-}14) \quad \text{and} \quad T_c'=T_2 \quad (3.4\text{-}15)$$

3.4.3. Youla parameterization for the ESO-m

The t.f. of plant is (3.4-16) (a) or (b):

$$H_P(s)=\frac{k_p}{s(1+sT_\Sigma)}\ \text{(a)} \quad \text{or} \quad H_P(s)=\frac{k_p}{s(1+sT_\Sigma)(1+sT_1)}\ \text{(b)} \tag{3.4-16}$$

The design is exemplified only for t.f. (3.4-16) (a); the step.

(1) Calculation of $H_c(s)$ with Q(s) parameter. Replacing results $H_c(s)$, S(s) and T(s)

$$H\ (s)=Q(s)\frac{s(1+sT_\Sigma)}{s(1+sT_\Sigma)-k_pQ(s)} \tag{3.4-17}$$

$$S(s)=\frac{s(1+sT_\Sigma)-k_pQ(s)}{s(1+sT_\Sigma)} \quad (3.4\text{-}18) \qquad T(s)=\frac{k_pQ(s)}{s(1+sT_\Sigma)} \quad (3.4\text{-}19)$$

(2) Establish a stable Q(s):

$$Q(s)=\frac{1}{k_p}T(s)s(1+sT_\Sigma) \quad \text{or} \quad Q(s)=\frac{1}{k_p}(1-S(s))s(1+sT_\Sigma) \tag{3.4-20}$$

The system performances are imposed through T(s), which is specific for ESO-m:

$$T(s)=\frac{1+\beta T_\Sigma}{1+\beta T_\Sigma s+\beta^{3/2}T_\Sigma^{\ 2}s^2+\beta^{3/2}T_\Sigma^{\ 3}s^3} \tag{3.4-21}$$

(3) Establish the controller's t.f. $H_c(s)$:

$$H_c(s)=\frac{1}{k_p}\frac{s(1+sT_\Sigma)}{\beta^{3/2}T_\Sigma^{\ 2}s^2(1+sT_\Sigma)}(1+\beta T_\Sigma s)=\frac{1}{\beta^{3/2}k_pT_\Sigma^{\ 2}s}(1+\beta T_\Sigma s) \tag{3.4-22}$$

The resulting controller is a PI type controller with parameters:

$$k_c=\frac{1}{\beta^{3/2}k_pT_\Sigma^{\ 2}} \quad T_c=\beta T_\Sigma \quad (3.4\text{-}23) \quad \text{and} \quad T_c'=T_2 \quad (3.4\text{-}24)$$

3.4.4. Youla parameterization for the 2p-SO-m

The plant t.f. is (3.4-7) (a) or (b). The design is exemplified only for t.f. (3.4-7) (a).

(1) Calculation of $H_c(s)$ with Q(s) parameter. Replacing results $H_c(s)$, S(s) and T(s)

$$H_c(s) = Q(s)\frac{(1+sT_\Sigma)(1+sT_1)}{(1+sT_\Sigma)(1+sT_1)-k_pQ(s)} \quad (3.4\text{-}25)$$

$$S(s) = \frac{(1+sT_1)(1+sT_\Sigma)-k_pQ(s)}{(1+sT_1)(1+sT_\Sigma)} \quad (3.4\text{-}26)$$

$$T(s) = \frac{k_pQ(s)}{(1+sT_1)(1+sT_\Sigma)} \quad (3.4\text{-}27)$$

(2) Establish a stable Q(s):

$$Q(s) = \frac{1}{k_p}T(s)(1+sT_1)(1+sT_\Sigma) \quad \text{or} \quad Q(s) = \frac{1}{k_p}(1-S(s))(1+sT_1)(1+sT_\Sigma) \quad (3.4\text{-}28)$$

The system performances are imposed through an adequate choice of T(s). For this case a proportional-derivative-with 3rd order lag model is adequate (PDL3). Accepting the use of a PI controller, one gets:

$$T(s) = \frac{1+sT_c}{a_3s^3+a_2s^2+a_1s+1} \quad (3.4\text{-}29)$$

where $a_3 = \frac{T_\Sigma T_1}{k_pk_c}$, $a_2 = \frac{T_\Sigma + T_1}{k_pk_c}$, $a_1 = \frac{1+k_pk_cT_c}{k_pk_c}$, $a_0 = 1$ (3.4-30)

(3) Establish the controller $H_c(s)$.

$$H_c(s) = \frac{1}{k_p}\frac{(1+sT_1)(1+sT_\Sigma)(1+sT_c)}{(a_1-T_c)s[\frac{a_3}{a_1-T_c}s^2+\frac{a_2}{a_1-T_c}s+1]} \quad (3.4\text{-}31)$$

If the general conditions (3.4-5) and (2.4-5) are imposed and the notation (2.4-9) is used, it results:

$$k_c = \frac{(1+m)^3}{m}\frac{1}{\beta^{3/2}k_pT_\Sigma} \quad (3.4\text{-}32) \quad \text{and:} \quad T_c = \beta T_{\Sigma m} = \beta T_\Sigma'\frac{[1+(2-\beta^{1/2})+m^2]}{(1+m)^2} \quad (3.4\text{-}33)$$

Using successive replacements, one gets

$$a_3 = \frac{T_\Sigma^3}{(1+m)^3}\beta^{3/2} = T_\Sigma'^3\beta^{3/2}, \quad a_2 = \frac{T_\Sigma^2}{(1+m)^2}\beta^{3/2} = T_\Sigma'^2\beta^{3/2}, \quad a_1 = T_\Sigma'\beta, \quad a_0 = 1 \quad (3.4\text{-}34)$$

$$Q(s) = \frac{1}{k_p}\frac{(1+sT_1)(1+sT_\Sigma)(1+\beta T_{\Sigma m}s)}{\beta^{3/2}T_\Sigma'^3s^3+\beta^{3/2}T_\Sigma'^2s^2+\beta T_\Sigma's+1} \quad (3.4\text{-}35)$$

For the second case the supplementary time constant is $T_c' = T_2$. In this case, for Youla parameterization design the following choice can be made:

- Place the zero $z_3 = -\frac{1}{\beta T_{\Sigma m}}$ as a function of the values of $\{\beta, T_\Sigma, m = T_\Sigma / T_1\}$;
- The poles of the system are function of $\{\beta, T_\Sigma, m\}$:

$$\Delta(s) = \beta^{3/2}T_\Sigma^{'3}s^3 + \beta^{3/2}T_\Sigma^{'2}s^2 + \beta T_\Sigma^{'}s + 1 \qquad (3.4\text{-}36)$$

$$p_1 = -\frac{1}{\beta^{1/2}T_\Sigma}\ , \quad p_{2,3} = \frac{-(\beta - \beta^{1/2})T_\Sigma^{'} \pm \left[(\beta - \beta^{1/2})^2 T_\Sigma^{'2} - 4\beta T_\Sigma^{'2}\right]^{1/2}}{2\beta T_\Sigma^{'2}} \qquad (3.4\text{-}37)$$

The controller parameters result from relations (3.4-32) and (3.4-33). In this context the poles' placement given by (3.4-37) leads to a Q(s) of form (3.4-28), and finally the controller is:

$$H_C(s) = \frac{1}{\beta^{3/2}k_p T_\Sigma^{2} s}(1 + \beta T_{\Sigma m}s) \qquad (3.4\text{-}38)$$

3.4.5. Conclusions

In this paragraph an interpretation of the Youla-parameterization design is presented for controller design for cases of benchmark type stable models, $H_p(s)$, which can characterize electrical driving systems with large mechanical time constants.

The study presents in detail the way of transposing the positive results gained from classical design methods based on modulus conditions (MO-m, SO-m) or conditions derived from these (ESO-m and 2p-SO-m) into a Youla-parameterization formulation. If the imposed conditions are adequately chosen, the controller is easy to implement. If the conditions are inadequate then the controller structure results as more difficult to comprehend and the solutions are less accepted in the practice.

4. Cascade Control Solution for an elctric traction system

4.1. Mathematical Modelling of the plant

Low power traction motors in electrical drive vehicles are frequently oriented on DC-machines (DC-m) or brushless DC motors (BLDC-m) [II-84] - [II-88] (but other solutions are also used, see also part I, [I-16]). From the point of view of mathematical modelling, the two solutions differ only insignificantly, mainly on parameter calculus relations.

The functional block diagram of an electrical driving system as part of an electric vehicle was presented in part I chapter 2 and detailed in figure 2.2-1 as an electric vehicle. In electrical traction, the operating range of a DC-m is divided into four quadrants: forward motoring, forward breaking, reverse motoring and reverse braking [II-86],[II-88],[II-87], [II-90].

4.1.1. Modeling the DC-m and the vehicle dynamics

The hypotheses accepted at modelling imply that in normal regimes the DC-m works in the linear domain where the flux (current) is constant in value (valid especially for BLDC-m). An eventually change in the excitation regime will modify the basic model, but a linearization in the new working point results in the basic situation. As a result, the block diagram of the DC-m is depicted in figure 2.2-3, part I.The basic equations that characterize the functionality of the system are given in (2.2-1) - (2.2-10), pat I: The modelling of the other functional blocks of the electric vehicle is not subject of this study, but they are described in [II-83], [II-100].

4.1.2. Numerical values of the plant

The numerical data for the considered application are defined in [II-85], [III-99].

❑ **Numerical values of the DC-m in the nominal functioning point**

The values are presented in Table 4.1-1

Table 4.1-1. Numerical values for the DC-m

Torque	Rotation	Useful power	Voltage	Current	Absorbed Power	Efficiency	Electrical time const
[Nm]	[rot/min]	[kw]	[V]	[A]	[kw]	[%]	[sec]
50,16	1605	8,43	77,6	126	9,78	86,18	0.1

Other electrical data:

- $R_a = 0.1\ \Omega$, - estimated value from the car builder,
- Gain and time constant of actuator: k_A=30 V/V, T_A=0.02 sec;
- Gains for current and speed sensors: k_{Mi}=0.0238 V/A; $k_{M\omega}$=0.0178 V/(rad/sec).

❑ **Numerical values regarded to the vehicle** ([II-85])

Nominal numerical values for the plant are:

- Total mass of vehicle, including an 80kg heavy driver: m_{tot}=1860 kg;
- Frontal area of vehicle: A_d=2.4 m^2;
- Air drag coefficient: C_d=0.4;
- Air density: ρ=1.225 kg/m^3;
- Rolling resistance coefficient: C_r=0.015;
- Wheel radius: w_r=0.3 m;
- Final drive ratio: f_r=4.875 Nm/(rad/sec).

Starting from the energy conservation principle, relations (4.1-1)-(A.1-3) were deduced; the numerical value is:

$$J_{veh} = 1860\ \frac{0.3^2}{4.875^2} = 7.04\ kg m^2 \quad (4.1\text{-}1)$$

Considering the moment of inertia of the wheels and electric motor having the value:

$$J_W = 1.56\ kg m^2 \quad (4.1\text{-}2)$$

The total inertia results as:

$$J_{tot} = J_{veh} + J_W = 7.04 + 1.56 = 8.6\ kgm^2 \quad (4.1\text{-}3)$$

which results in a mechanical time constant of T_m = 5.43 sec.

The two time constants are T_m=5.43 sec and T_a=0.1 sec resulting in a value of m: m<0.02.

4.2. Control structures. Controller design, simulation results

4.2.1. Control aim and imposed performances

In spite of their simplicity, the CCS can substantially improve the speed and the linearity of the closed-loop system. The aims of the control structures applied to the DC-m are grouped as follows:

- To ensure good reference signal tracking (speed) with small settling time and small overshoot (good transients and zero-steady-state error at ω_∞=const.); the steady state variable reference (ramp) error should be as small as possible; the relatively slowly changing ramp can be considered decomposed in a sequence of step variation.
- To ensure load disturbance rejection due to modifications in the driving conditions.
- To show reduced sensitivity [II-84] to changes in the total inertia of the system.

4.2.2. Control solutions

Due to previous experiences in domain [II-83], [II-85] there were adopted two classical solutions, both having two control loops in a CCS:

- One internal control loop of the current, consisting in a PI controller extended with an AWR measure [II-90], [II-91].
- One external control loop of speed ω [rad/sec] (motor speed) with a PI controller.

The two controllers are designed separately.

- The first control solution is a clssical one, figure 4.2-1.
- The second control structure, figure 4.2-2, differs from the first through the outer loop, in which a forcing block was added to correct the current reference for the inner loop. This can decrease the response time of the system.

❑ **The inner loop: the current control loop.** The inner loop is identical in both cases, and it consists of a PI controller with AWR measure [II-90], [II-91]. The parameters of the current controller were based on the MO-m having the design relation (2.2-7) and Table 2.2-1:

$$H_{Cc}(s) = \frac{k_{ci}}{s}(1 + sT_{ci}), \quad k_{ci} = \frac{1}{2k_{pi}T_{\Sigma i}}, \quad T_{ci} = T_a \qquad (4.2\text{-}1)$$

where k_{pi} – the gain of the inner part of the plant, containing the actuator, electric circuit and current sensor), T_a – electric time constant, $T_{\Sigma i}$ – equivalent of small (parasitic) time constants (resulting from the power electronics time constant, the sensor and filter time constants), with $T_a > T_{\Sigma i}$. For the given application the plants' t.f. results

$$H_{PCI}(s) \rightarrow \widetilde{H}_{PCI}(s) = \frac{k_{PCI}}{(1 + sT_{\Sigma i})(1 + sT_a)} \qquad (4.2\text{-}2)$$

)

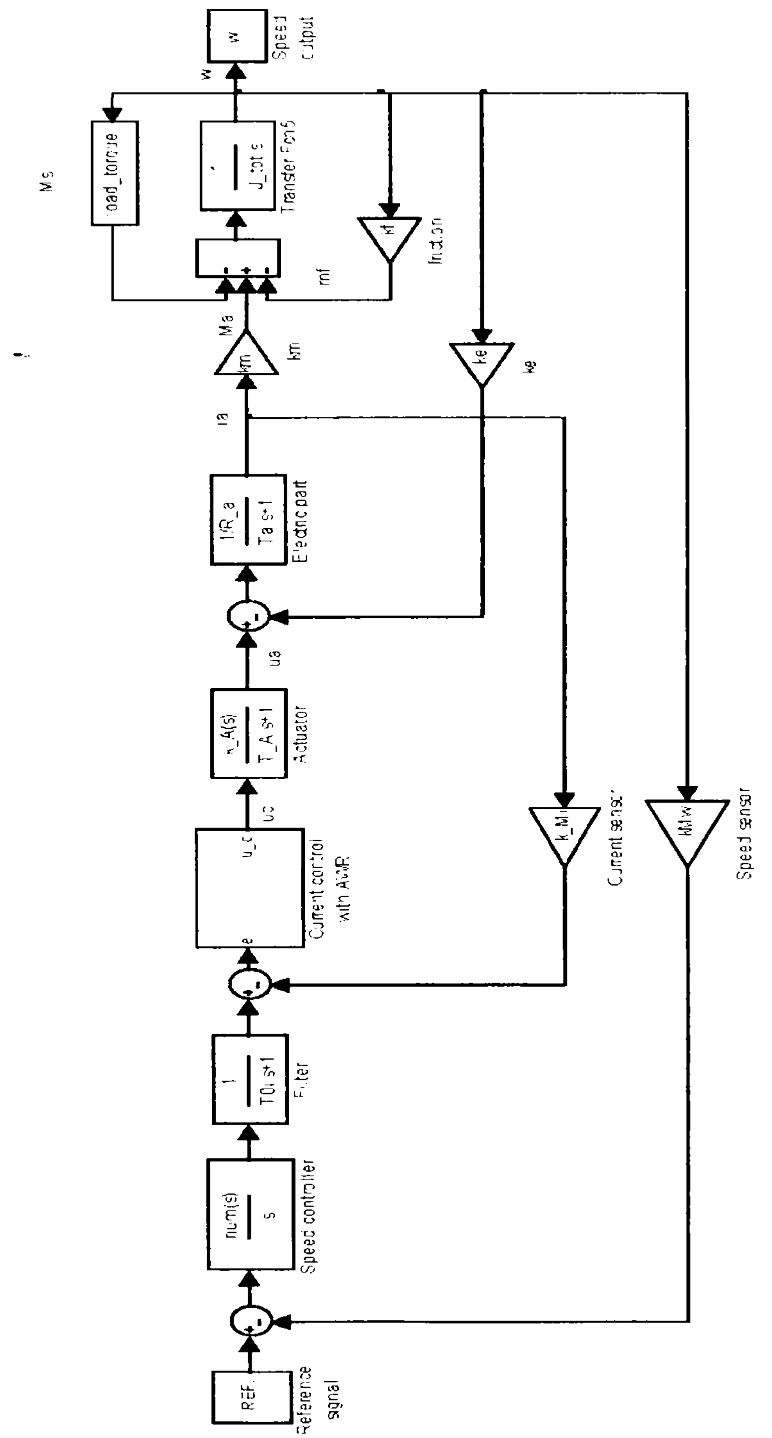


Fig.4.2-1. First cascade control structure for the DC-m (Simulink scheme)

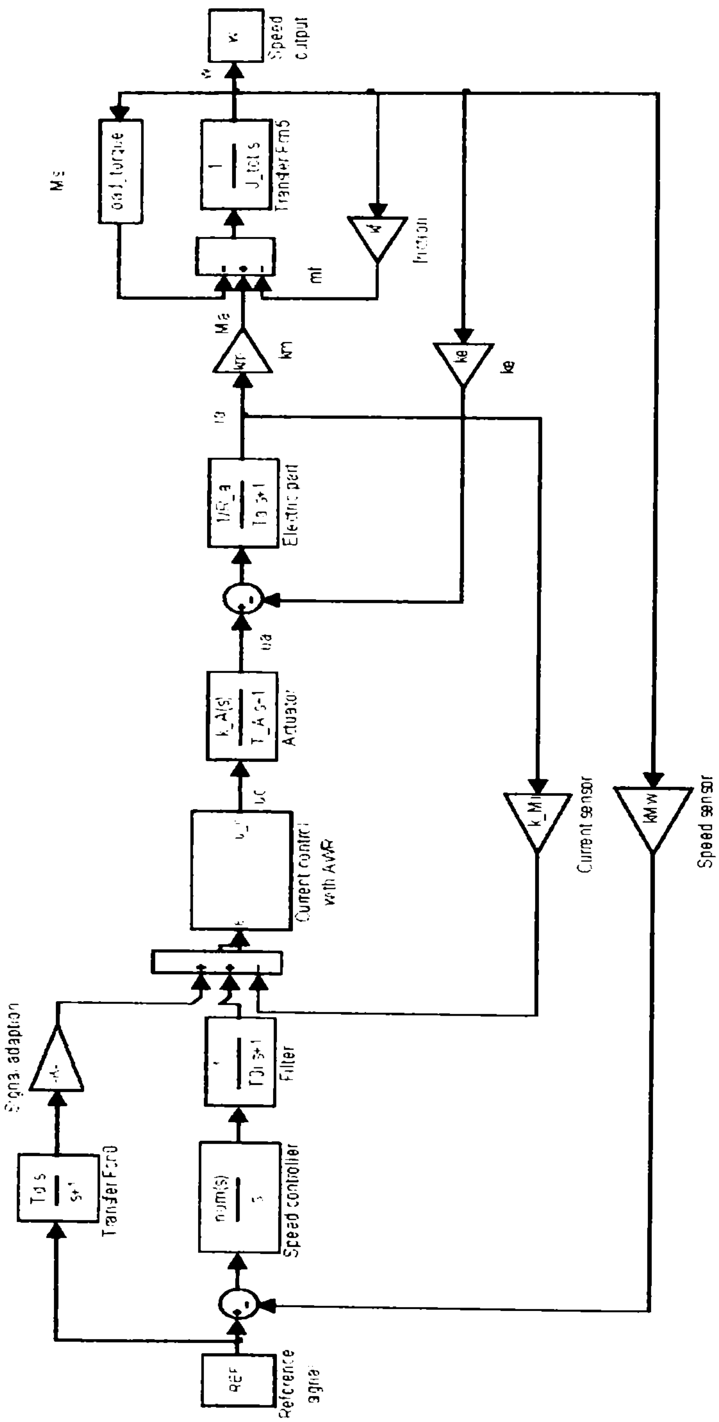


Fig.4.2-2. Second cascade control structure for the DC-m: speed control with reference forcing block (Simulink scheme)

where $k_{PCI} = k_{pi} = k_A \frac{1}{R_a} k_{Mi} k_{fi} = 1.75$ and $T_{\Sigma i} = T_A + T_a + T_{fi} = 0.04$ sec. and the controller parameters results as $k_{ci} = 7.0$, $T_{ci} = 0.1$ sec.

The AWR measure was introduced to attenuate the effects of going into limitation of the controller and realized according to [II-90], [II-91] having the value of T_t=0.005.
Other methods for handling constraints of the control signal are also used, for example a solution where the controller itself is by a dynamic feedback of a static saturation element [II-91]. The inner optimized control loop can be approximated as follows:

$$H_{ri}(s) = \frac{1}{k_{M1}} \frac{1}{1 + 2T_{\Sigma i} s} \quad , \ k_{Mi} = 0.0238 \text{ V/A} , \quad 2T_{\Sigma i} = 0.08 \text{sec.}$$

❑ **The external loop: the speed control loop**. As the outer loop, it consists of a PI controller in two variants for implementation:

(a) one homogenous variant,
(b) One case when a forcing filter for the reference value was introduced.

Neglecting the friction coefficient, k_f, for the simplified design method ($m \approx 0.02$) the plant t.f. used for speed controller design can be considered:

$$H_p(s) = \frac{k_p}{s(1 + sT_{\Sigma\omega})} \qquad (4.2\text{-}3)$$

where $T_{\Sigma\omega} = 2T_{\Sigma i} + T_{io}$ stands for the current loop and parasitic time constants ($T_{io} = 0.05$), k_p characterizes the dynamics of the mechanical part of the driving system (J_{tot}), the inverse of the current sensor k_{Mi}^{-1} and the speed sensor $k_{M\omega}$.
The speed controller is PI type-controller having the t.f.

$$H_{C\omega}(s) = k_{C\omega}(1 + \frac{1}{sT_{c\omega}}) = \frac{k_{c\omega}}{s}(1 + sT_{c\omega}) \qquad (4.2\text{-}4)$$

Based on desired control performances the parameter β is chosen $\beta \approx 16$. A larger value of β ensures less oscillating transients and a bigger phase margin. Due to the fact that in this case m was very small, no corrections were applied at this stage. Finally the parameters of the speed controller results as: $k_{c\omega} \approx 35.0$ and $T_{c\omega}$=1.75 .
For the second case the controller is the same and the feed forward correction term is a Derivative with first order Lag type filter (DL1) with t.f.:

$$H_{ff}(s) = \frac{56.0 \quad s}{1 + s} \qquad (4.2\text{-}5)$$

4.2.3. Simulation results

The simulation scenarios are the following: the first control structure is simulated, followed by the second cascade structure simulations (comparison of the currents' and dynamics), ended by simulations for the first case regarding sensitivity aspects for a change in the mass of the plant.

The reference signal is the same for all three cases, consisting in an acceleration part, a part with constant velocity and a part of deceleration until a stop is reached. The load of the system is taken into account as in [II-84], [II-93]. The registered variables are: the velocity (speed), the current and the electromagnetic torque M_a vs. disturbance torque M_s

- **Case of simple cascade structure**

The simulation results are depictured in figures 4.2-3, 4.2-4 and 4.2-5.

- **Case of cascade structure with correction of the current reference**

In this case the differences in the current behaviour are depicted, together with the active power (dashed line - simple cascade structure, solid line - structure with current correction). The differences in the speed dynamics are not significant, the active power differences are proportional with the current, figures 4.2-6 and 4.2-7.

- **Simple cascade structure with modified load**

The simulations are presented in figures 4.2-8 and 4.2-9 (for the first cascade structure): the mass of the vehicle is changed with +25% of it (solid line - original load, dashed line - increased load): $m_{veh}=m_{veh0}+\Delta m=1860+0.25*1860=2332kg$.

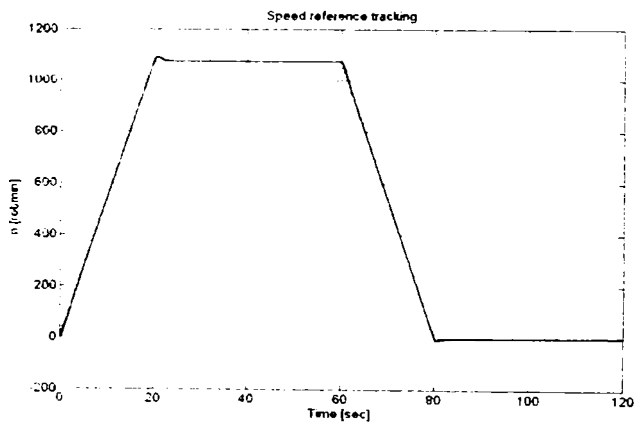


Fig.4.2-3. Speed reference tracking

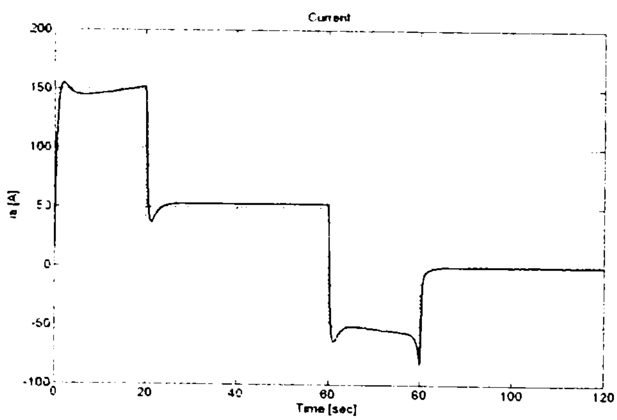


Fig.4.2-4. Behaviour of the current

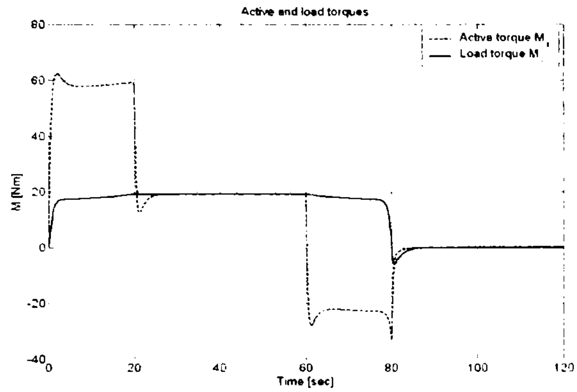


Fig.4.2-5. Active torque M_a vs. disturbance torque M_s

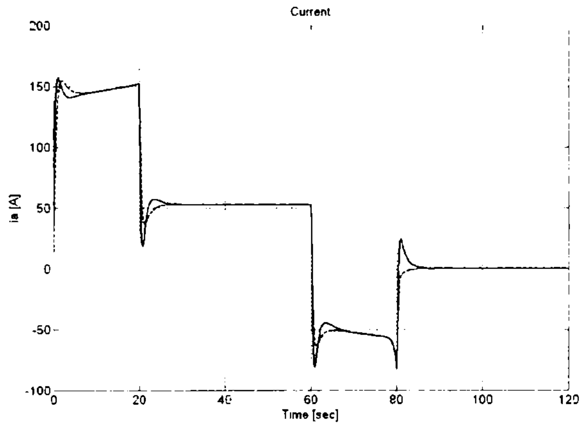


Fig.4.2-6. Comparison of the currents

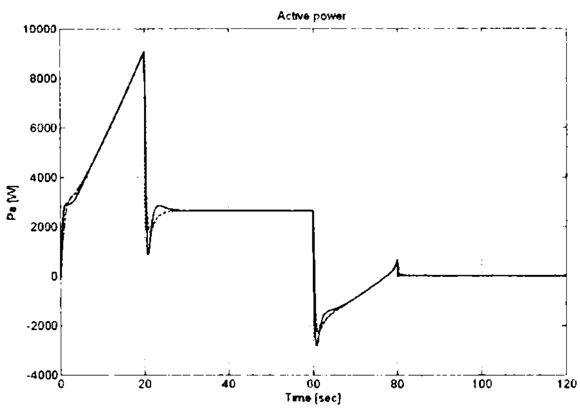


Fig.4.2-7. Comparison of the active powers

The speed was not presented since almost the same behaviour resulted. But in order to achieve this performance, the current is higher (since it needs more power to carry the increased weight). Still the current does not reach its maximal admissible value (4 times the nominal value of 126 A).

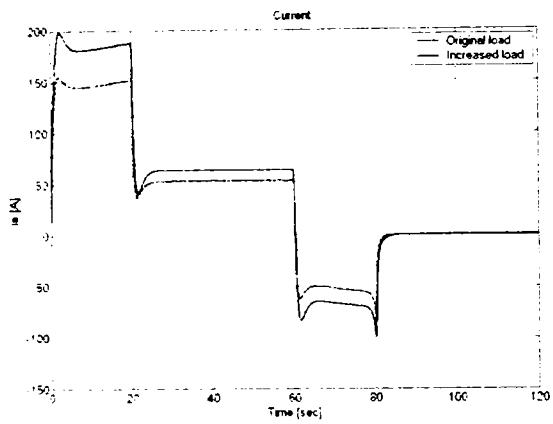


Fig.4.2-8. Behaviour of the current

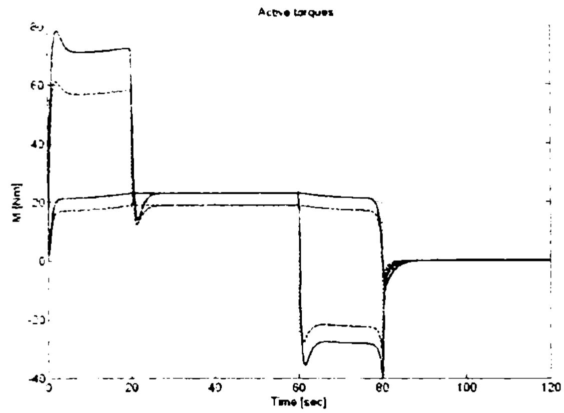


Fig.4.2-9. Active torque M_a vs. disturbance torque M_s

The active power is higher (12 kW compared to 9kW at starting), but without exceeding the maximal power of 15 kW of the machine. Both the active torque and the load torque are higher, as expected.
In the simulations non-linear phenomena induced by the limitations did not occur.

4.3. Conclusions

The chapter presents an efficient CCS for electrical drives used for electrical traction vehicle in two variants – without and with a forcing feed forward term for the current reference. The numerical data's regarded to the application are based on a real application of a hybrid solar vehicle.
In order to ensure superior performances, for controller design different variants of the modulus optimum tuning method: the MO-method for the inner loop, and due to the fact that the ratio $m = T_\Sigma / T_1$ is very small, accepting Kessler's approximation, the ESO-m were used; due to the very small value of $m \approx 0.2$ the correction of the results based on the tuning method 2p-ESO-m can be neglected.

Simulations were performed using the Matlab/Simulink environment, for a reference drive cycle, derived from the NED Cycle [II-86]. The simulated cases reflect a very good behaviour of the system both regarding reference tracking and also sensitivity to parameter changes.

5. Part conclusions and contributions

This part of the thesis is a detailed study based on 94 papers, including papers to which I have contributions. Out of these at 8 papers I am first or single author; also the three PhD reports sustained at U.P.Timişoara are mentioned. The main contributions of part II of the thesis are:

1. In Chapter 2 an overview on optimal design methods based on Modulus Optimum criteria, detailing the MO method, SO method and ESO method.
2. In Chapter 3 a novel controller design method based on a double parameterization of the optimality conditions specific for the SO method is introduced: the 2p-SO-method. The method (paragraph 3.1) refers to plants with very time constants and for which the application of the ESO-m requires approximations. The double parameterization is based on the followings:
   - First, with the condition that $T_\Sigma / T_1 << 1$, the parameter m is defined
   $$m = T_\Sigma / T_1$$
   - Second, the use of the optimisation relations:
   $$\beta^{1/2} a_0 a_2 = a_1^2 \quad , \qquad \beta^{1/2} a_1 a_3 = a_2^2$$

   Through this an improvement of the phase margin can be reached. Design cases are presented and the specific tuning relations deduced. The achievable performances, the method efficiency are compared with the MO-method (much preferred design method for PID controllers for driving systems. Specific particular cases are presented and analyzed, accompanied by performance diagrams and methods for improving these. Simulation data allow a good specification for the cases when the method proves to be very efficient (paragraph 3.1.2, 3.1.3).
3. Taking into account that the robust design based on the Youla parameterization proves to be very efficient in many cases a Youla parameterization approach of the MO-m, ESO-m and 2p-SO-method is given.
4. For the electric drive of a vehicle with electric traction (based on real data [II-85]) Chapter 4 presents a detailed design for a cascade control system. Two variants of the cascade structure are presented, controller with homogenous and non-homogenous structure, using AWR measure. The simulation results reflect the expected behaviour.
5. Connected to this part of the thesis, appendix I present a 2-DOF approach for PI and PID controllers and a design method which can easy applied in practice.

Finally to underline the relevance of the research report with comparable results from September 2007, [II-66] must be mentioned.

Part III. New Design Solutions in Speed Control for Hydrogenerators

"Predictive control is usually considered when a better performance than that achievable by non-predictive control is required"
(Camacho, E.F. and Bordons, C. [III-31])

1. Introduction. The structure of part

Part III of the thesys present control solutions dedicated to the speed control for hydro-generators. In control of complex plants it is often needed to use advanced CS-s. To choose an adequate control solution, the mathematical modelling of the plant is necessary. The widely used mathematical models for each subsystem of a HG and the speed control control structures were presented in part I chapter 3, figure 3.1. The load - frequency control is active during normal operating conditions. The input to this CS is the speed deviation.

Chapter 2 synthesises design solutions in speed control for hydrogenerator.

In Chapter 3 a cascade control structure and a mixed design technique of hydro-turbine application (speed control) is presented [III-26], [III-27]; the aim of this structure is based on fact that the plant can be separated into two decoupled parts and the proposed solution is based on a cascade control structure with an internal minimax controller, to reject internally located deterministic disturbances and a main Generalized Predictive Controller (GPC) loop, to reject external (stochastic) disturbances induced by the power system. The external loop consists of a GPC structure [III-31] used in its polynomial representation, which has been transformed into an Internal Model Control (IMC) structure. The disturbance acting on this part of the plant is the active power induced by the power system and considered as a stochastic one. Under these conditions the combined cascade control structure can be an attractive solution to ensure the control system performance enhancement. Finaly, the design steps are presented. The solution was tested through simulation for a numerical case study regarding to a real application.

Chapter 4 propose a fuzzy control (FC) solution based on a new Takagi-Sugeno Fuzzy Controller (TS-FC) dedicated to the speed control of hydro turbine generators [III-15]; the controlled plant is characterized by models presented in part I chapter 3. The servosystem is accepted to be stabilised using pole asgnment technique and represented finaly as a first order with lag model. In the first phase conventional PI controllers are developed ensuring desired maximum values both for the sensitivity function and for the complementary sensitivity function in frequency domain. Further, a design method for a four inputs-two outputs TS-FC is presented. The FC system guarantees maximum imposed sensitivity functions. The proposed solution was tested and validated through simulation of a case study. The controller is compared with two PI controllers designed separately with respect to the reference input and with respect to one of the disturbance inputs.

Chapter 5 synthesises the contributions presented in this part of the thesis.

2. Design solutions in speed control for hydrogenerators

The review on classical design solutions in speed control of HG highlights the fact that most of the practical solutions are based on a local stabilization of the servosystem and the use of a PI (PID) controller in the main loop. The main PI (PID) controllers are tuned according to a large variety of methods. In the case of the classical design methods the speed controller is designed for a nominal or a specific regime based on the robustness of the controller; the solutions are verified for different values of the parameters as well, accepting the fact that in a real PSat normal functioning regime these modifications are not significant.

2.1. Cascade control structures. Trends

The relevance of cascade control structures is still very actual, a multitude of papers regarding these are continuously appearing (see for example the papers of the 17th IFAC World Congress in Prague, 2005). Cascade control is a wide-spread control technique for industrial applications. Its use is motivated by its advantages and also by the necessity to use in some cases such multiple loop structures due to the physical features of the plant. The objective of cascade control is to split a control process into two (or even more) parts, where the outer controller generates on its output the reference signal for the secondary, inner controller.

State feedback can be considered as a generalization of cascade control, where not only some, but all state variables are fed back. Also inner variables can be kept limited using limitation elements inserted in the cascade loop. If the inner loop is unstable, a local stabilization can be perfomed, and thus the primary controller can handle the stabilized part.

The main advantages of using cascade structures can be briefly summarised in the following:

- Improved disturbance rejection which acts distributedly on the plants, in the form of allowing a faster rejection by the secondary, inner controller. In this case the choice of the secondary controller is very important, so that it can allow a fast recovery from the effect of perturbations.
- A better control in the primary loop and improved dynamical performance through optimizing the inner loop.

There are also disadvantages of applying cascade controllers, which usually come from the nature of the plant (see [III-66].

The main interest in using cascade control structures are:

- Development of new design methods based on combining different structures and design methods, by which the particularities of the given plant can be taken into account;
- Spreding of the application areas to new domains of automatic control.

In this context the following papers are mentioned [III-67] – [III-73] in a form of a survey.

In [I-63] an excitation controller for a single generator based on modern multi-loop design methodology is presented in this paper. The proposed controller consists of two-loops: a stabilizing (damping injection) loop and a voltage regulating loop. The task of the stabilizing loop is to add damping in the face of voltage oscillations.

In [III-68] the design and experimental testing of a prototype cascade feedback control system is treated. The overall controller architecture is constituted by three control sub-tasks. In [III-69] the objective of the application is a cascade model-based predictive control technology for increase of a boiler efficiency.

In [III-70] two new two-degree-of-freedom control structures are proposed for cascade control systems, both of which are identical in the controller design procedure.The primary controller used for setpoint tracking is derived in terms of H_2 optimal performance objective. The secondary controller is responsible for rejecting load disturbances that act into the intermediate process and therefore is called a load disturbance estimator, which is inversely figured out by proposing the desired complementary sensitivity function of the inner loop for disturbance rejection.

In [III-71] based on experience and engineering insight a systematic procedure for cascade control structure design for complete chemical plants (plantwide control) is presented. In [III-72] the aim of improving control performances using cascade control and Smith predictor is treated. The combination of different control strucures and methods is highlighted to be advantageous. .

In [III-73] a new simultaneous online automatic tuning method for cascade control using a relay feedback approach is presented. Departing from the traditional approach towards tuning of cascade control systems where the secondary and primary loops are tuned in strict sequence, the proposed idea is to carry out the entire tuning process in one experiment. For ease of practical applications, the entire procedure of controller design may be automated and carried out online. A direct controller tuning approach to tune the controllers is proposed here.

In [III-74] constrained cascade controls for parallel processes are treated. One method employs the traditional cascade controllers, applied to serial transfer functions. The second uses cascade controllers applied to parallel transfer function processes. The latter method shows sensitivity to disturbance and tuning of inner loops. A third innovative method, called a pseudo-cascade controller, is introduced for parallel transfer functions.

In [I-75] a cascaded Nonlinear Receding-Horizon Control of Induction Motors is presented with application to induction motors in cascade structure. The controller is based on a finite horizon continuous time minimization of nonlinear tracking errors. Both the rotor flux and load torque are estimated by Kalman filter. Computer simulations show the flux-speed tracking performances and the disturbance rejection capabilities of the proposed controller in the nominal and mismatched parameter case.

In [III-37] a Predictive Control solution of Fast Unstable and Nonminimum-phase Nonlinear Systems is treated. With standard predictive control, the time required for optimization is typically larger than the sampling interval that is needed for stabilization of the fast dynamics. On the other hand, due to the nonminimum-phase behavior, control based on input-output feedback linearization leads to unstable internal dynamics. In this paper, a cascade structure is proposed, with control based on input-output feedback linearisat ion forming the inner loop and predictive control the outer loop. Assuming high-gain feedback for the inner loop, a stability analysis of the global scheme is provided based on singular perturbation theory. The approach is illustrated via the simulation.

In [III-77]a method for Tuning the cascade control systems by means of magnitude optimum method is presented. The proposed tuning approach is based on the so called MOMI tuning method which requires only one simple steady-state change of the process. The efficiency of the proposed approach has been tested on three process models and compared to one existing tuning method by means of simulation.

Other intresting aspects regarded to CCSs are presented in papers [III-79] (a nuclear reactor application), [III-79] (dual RST-control of an Inverted Pendulum with Simulink S-functions Implementation), [III-37] (Robust GPC-QFT approach), [III-81] (AWR applications in CCS) (a.o., see for example the Proceedings of the European Control Conference 2007 Kos, Greece, July 2-5, 2007).

2.2. An overview of hydrogenerators speed control development

Based on significant papers present in the literature, different trends in HG speed control design were analysed. The main conclusions drawn from this are synthesized in the following.

(1) Derivation of PI (PID) controller parameters in the frequency domain by using the open loop Bode diagrams is frequently used in applications (see references [III-5], [III-6], [III-9], [III-10], [III-12], [III-64], [III-65].

(2) Optimal tuning of controller parameters based on integral cost functions has the well-known advantageous(see part II, references [II-5], [II-9], [II-10], [III-44], [III-48]):
- Use of integral cost functions can ensure good performances regarding to input and disturbace;
- The most used integral cost functions are quadratic integral cost functions (ISE) ,due to the fact that the evaluation and minimising the expression of ISE cost functions or combined quadratic forms with different components (poly-optimisation) is relatively easy. For example, in [III-5] a lot of such integral indices are analysed and pertinent conclusions concerning the use of integral indices in tuning PI(D) controllers for hydro-generator applications are given.

(3) In its different form, the MBC structures [III-66], [III-41], [III-42] represent an attractive solution that can ensure good performances and studies for its implementation are of a permanent actuality. The IMC solutions having a generic form of the structure presented in figure 2.2-1 [III-66] are for special interest.
The figure highlights four blocks:
- The real plant, with the controlled output, (measured values),
- The plant model, which computes the predicted values for the plant evolution,
- The disturbance estimation or the model parameter adaptation, the block makes adjustments to the disturbance estimation or the model parameters so that the predicted values are brought closer to the controlled output

The controller computes the control action (manipulated inputs on the plant/model) needed to achieve the target (setpoint, reference).
The basic structures can be combined and extended in various forms resulting in combined structures which can ensure better control performances. For example in [III-83], [III-84], [III-85], a new hybrid control system design technique in delta domain for IMC structure based on

the delta model of the plant is presented and applied for benchmark-type test models (including non-minimum phase models). Also a generalised form for delta domain DB control algorithm is given, a hybrid implementation of the controller (Z-domain combined with delta domain) is presented and its architecture is compared with the pure delta-domain implementation.

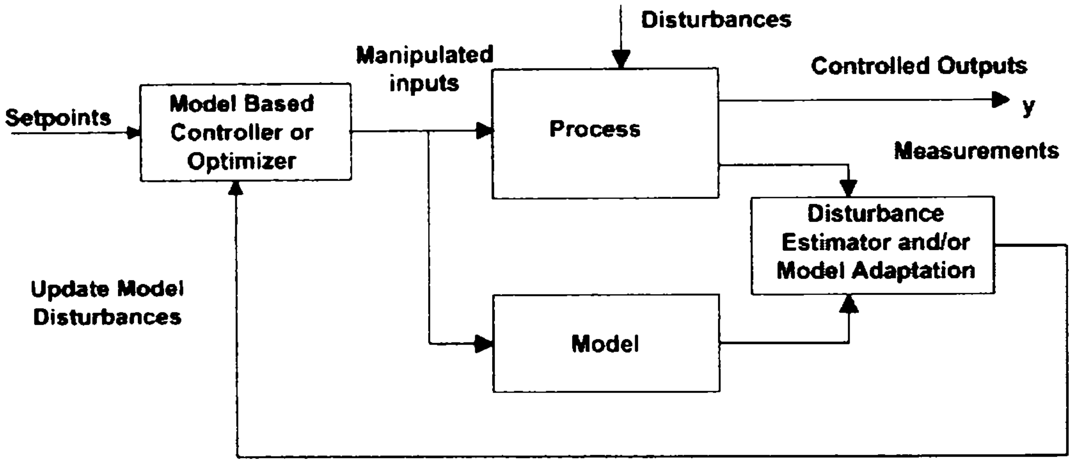


Fig.2.2-1. The generic form of MBC control structure

(4) MPC in its basic form and advanced forms and GPC has gained much attention and developed considerably in the past years, both in research and in industry. Good overview of predictive control is given in [III-31], [III-32], [III-86] applied in continuous time, in discrete domain, follow by an approach in the delta domain [III-19]. In [III-30] is remarked the fact that the treatement of the external disturbances is obviously necessary. In [III-46] a predictive feedforward control of a hydroelectric plant is presented.

(5) Structures based on FC techniques are successfully applied in various industrial domains. In [III-5] various control solutions using FCs with dynamics are presented and developing solutions are given:
- standard quasi-PI FC solutions with integrating component introduced on the input or on the output of the basic fuzzy block;
- quasi-PI controllers with modified rule base;
- quasi-PI controllers with modified rule base (RB) and modified distribution of membershipfunctons (MF) for control signal linguistic terms;
- variable structure quasi-PI FC-s;

There are presented also new development strategies for fuzzy tuning of parameters controllers meant for nonminimum-phase control system that are easily implementable. Other later research results in FC solutions for nonminimum-phased control system were presented in [III-45], [III-50], [III-63], [III-89].

As a result of the trends presented in the previous paragraph, of mixture of two control solutions in a CCS, in Part III of the thesis cascade control solutions are used in two applications:

- First a new CCS design solution was introduced and presented (papers [III-26], [III-27]);
- Second a new design of a Fuzzy Control (FC) solution for HG-s with imposed maximum sensitivity functions (mainly in paper [III-15]).

3. Applied GPC cascade control solution for hydrogenerators

3.1. Introduction

Based on [III-26], [III-27] and [III-41], [III-42], the chapter presents a new two-stage control solution which deals separately with the disturbances acting on the plant. In paragraph 3.5 the particular structure is applied speed control of a HG (the main application), but, it can be applied also in other practical applications, where the plant can be separated into two parts.

The advantages of classical CCS are well known (see chapter 2). In the proposed CCS, the internal loop solves the rejection of the deterministic part of the disturbance using minimax control for worst case of the inner disturbance. Further, a GPC is used in its polynomial representation, which has been transformed into an IMC structure. The disturbances acting on this part of the plant are considered stochastic ones.

3.2. Proposed cascade control structure

Many practical applications are dealing with plants that can be decomposed into sub-processes P_1, P_2, figure 3.2-1.

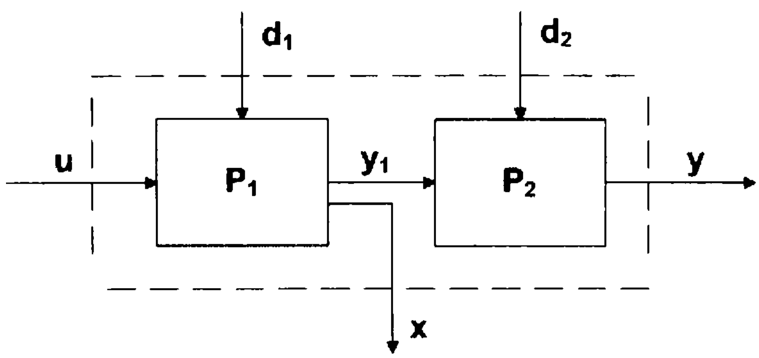


Fig.3.2-1. Decomposition of plant

The independently acting disturbances d_1 and d_2 can be unmeasurable but estimable or, in some cases, measurable. It is assumed that $d_1(t), d_2(t) \in L_2$, i.e.

$$\int_0^\infty d_1^T(t)d_1(t)dt < \infty \quad , \quad \int_0^\infty d_2^T(t)d_2(t)dt < \infty \tag{3.2-1}$$

The nature of these disturbances is considered to be different and the simultaneous rejection of both disturbances by a single controller can become a difficult task. In the proposed CS their rejection becomes an independent task for two controllers connected in cascade loop.

The efficient rejection of internal disturbances d_1 upon the output can be ensured through a local loop, which is optimized both as far as the control signal u and the disturbance is concerned. This way the structure will have a favourable behaviour regarding both transfer chanels, u→y and $d_1 \rightarrow y_1$. Figure 3.2-2 presents the cascade control structure.

3.3. Optimal controller design for disturbance rejection based on minimax criterion

As introduced in the previous section, the design of the "inner" control loop, which is affected by d_1 disturbances, can represent a linear quadratic minimax problem, where the controller minimizes a given cost function when the disturbances maximize this cost function [III-28], [III-29], [III-34]. A state feedback controller is used (Fig.3.2-2) for which a minimax gain is calculated, which is able to reject disturbances up to the value of the worst case disturbance (also determined in the algorithm) [III-28], [III-29].

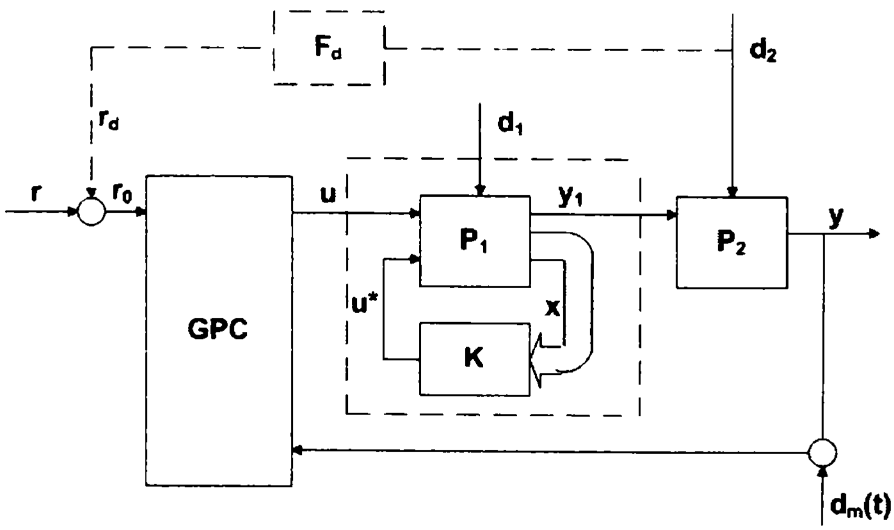


Fig.3.2-2. Cascade GPC control structure

Consider the P1 plant (figure 3.2-2) as a linear system, and its dynamics described as follows:

$$\begin{aligned} \dot{x}(t) &= Ax(t) + Bu(t) + Ld_1(t) \\ y_1(t) &= Cx(t) \end{aligned} \tag{3.3-1}$$

$d_1(t)$ representing the inner disturbance. For linear systems a quadratic cost function is defined in order to obtain the optimal control signal, and the effect of the disturbance is included explicitly:

$$J(u,d_1)=\frac{1}{2}\int_0^{\infty}\left[y_1^T(t)y_1(t)+\rho^2u^T(t)u(t)\quad \gamma^2 d_1^T(t)d_1(t)\right]dt \tag{3.3-2}$$

ρ is a design parameter and γ is a free parameter which has to be chosen so that it is in line with the Riccati equation to be solved. The disturbance term is with negative sign due to the solution possibilities of the Control Algebraic Riccati Equation (for more details see for example [III-28], [III-29] (Athans and Bokor).

The target is to minimize the cost function and obtain the worst case disturbance rejection state-feedback (minimax-game), under which conditions disturbances up to the worst case disturbance are rejected. Reformulating the task, a differential-game problem results:

$$\min_{u(t)}\{\max_{d_1(t)}[J(u,d_1)]\} \tag{3.3-3}$$

Supplementing the cost function with a co-state vector and with its help solving the Euler-Lagrange equation and deriving then the Hamiltonian system, finally the optimal control signal u^* and the worst case disturbance d_1^* are obtained:

$$u^*(t) = -\frac{1}{\rho^2}B^TPx^*(t) = -K_ux^* \quad , \quad d_1^*(t)=\frac{1}{\gamma^2}L^TPx^*(t)=K_dx^* \tag{3.3-4}$$

Here x^* stands for the optimal state resulting from the Modified Control Algebraic Equation (MCARE) (3.3-5), while P represents a positive definite and symmetric ($P=P^T>0$) solution to the MCARE:

$$PA + A^TP + C^TC - P\left(\frac{1}{\rho^2}BB^T - \frac{1}{\gamma^2}LL^T\right)P = 0 \tag{3.3-5}$$

Finally the feedback gain matrix K is given by its two components K_u and K_d, where K_u is the gain component related to the optimal control and K_d is the gain component related to the maximal disturbance. This is emphasized in figure 3.3-1.

In this figure d^* represents the worst case disturbance (in the sense of H_∞ norm) which can still be successfully rejected by the control system. d^* is computed only for simulation purposes, to emphasize the actual value of the worst case disturbance. In reality this d_1 is the existing disturbance which must be rejected, so for real cases d^* does not exist, but disturbance d_1.

For the optimized system which is physically implemented the closed loop transfer functions (for the SISO case) regarding the disturbance – the sensitivity function $S_1(s)$) - and regarding the reference - the complementary sensitivity function $T_1(s)$ - is:

$$S_1(s) = C(sI - A + BK_u)^{-1}L \quad , \quad T_1(s) = P_{S1}(s) = C(sI - A + BK_u)^{-1}B \tag{3.3-6}$$

This also simplifies the calculus of the external control loop. With the help of a Matlab program the optimal solution is obtained in an efficient way [III-62].

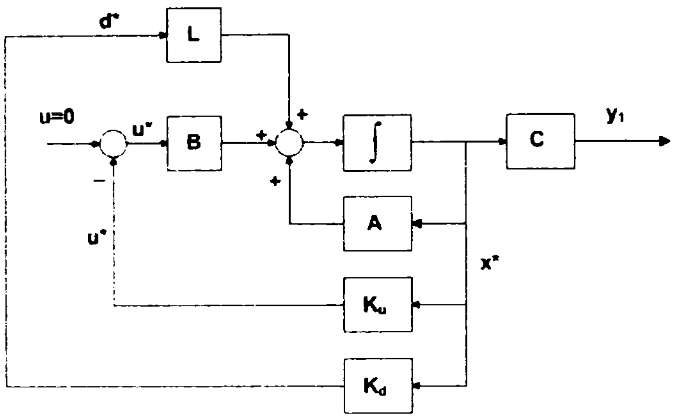


Fig.3.3-1. Minimax control loop for worst case disturbance for theoretical representation

3.4. GPC controller design in its IMC representation

The main controller is designed on basis of GPC algorithm. There exists a polynomial form of GPC in case if there are no constraints. The plant model (linear or linearized) can be described by the following CARIMA model:

$$A(q^{-1})y(t) = q^{-D_T}B(q^{-1})u(t-1) + \frac{C(q^{-1})}{\Delta}d_2(t) \qquad (3.4\text{-}1)$$

u(t) is the control sequence, y(t) the output sequence, $d_2(t)$ is a zero mean white noise, D_T is the physical dead time. Polynomials A, B and C are described in the backward shift operator q^{-1}, $\Delta = 1 - q^{-1}$. The $C(q^{-1})$ polynomial is chosen for 1, for simplicity. The minimized cost function, in order to obtain the control, sequence is:

$$J = \sum_{j=N_1}^{N_2} \delta(j)[\hat{y}(t+j\,|\,t) - r(t+j)]^2 + \sum_{j=1}^{N_u} \lambda_u(j)[\Delta u(t+j-1)]^2 \qquad (3.4\text{-}2)$$

where N_1 and N_2 are the limits of the prediction horizon, N_u is the control horizon, $\hat{y}(t+j\,|\,t)$ is the j-step ahead prediction of the output, r(t+j) is the future reference trajectory and δ(j) and λ(j) are weighting sequences.

After the minimization the control law is obtained:

$$\Delta u(t) = W(r - f) = \sum_{i=N_1}^{N_2} k_i [r(t+i) - f(t+i)] \qquad (3.4\text{-}3)$$

W is the first row of the matrix $(G^T G + \lambda I)^{-1} G^T$, f is the free response, r is the reference signal and G is a matrix containing elements of the plant's unit step response. If there are no constraints, the GPC algorithm can be transformed into a two-degree-of-freedom (2DOF) polynomial structure, (see figure 3.4-1):

$$R(q^{-1})\Delta u(t) = T(q^{-1})r(t) - S(q^{-1})y(t), \qquad (a)\ (3.4\text{-}4)$$

where R, S, T are polynomials in the backward shift operator. The $R(q^{-1})$ and $S(q^{-1})$ polynomials can be calculated and the result will be (see Appendix 2):

$$R(q^{-1}) = \frac{T(q^{-1}) + q^{-1}\sum_{i=N_1}^{N_2} k_i I_i}{\sum_{i=N_1}^{N_2} k_i}, \quad S(q^{-1}) = \frac{\sum_{i=N_1}^{N_2} k_i F_i}{\sum_{i=N_1}^{N_2} k_i} \qquad (b)\ (3.4\text{-}4)$$

where the polynomial $T(q^{-1})$ is a free parameter (often chosen to be 1, see Appendix 2).

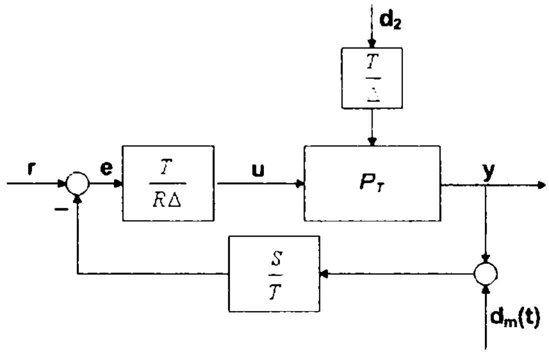


Fig.3.4-1. RST structure of GPC

The main idea for IMC structures is to include the model of the process in the control loop, and the serially connected controller is chosen to the best realizable inverse of it, [III-26], [III-33]. In figure 3.4-2 the IMC structure is presented, enriched with two filters F_r and F_y and handling constraints. Of course, this is valid only with stable plants.

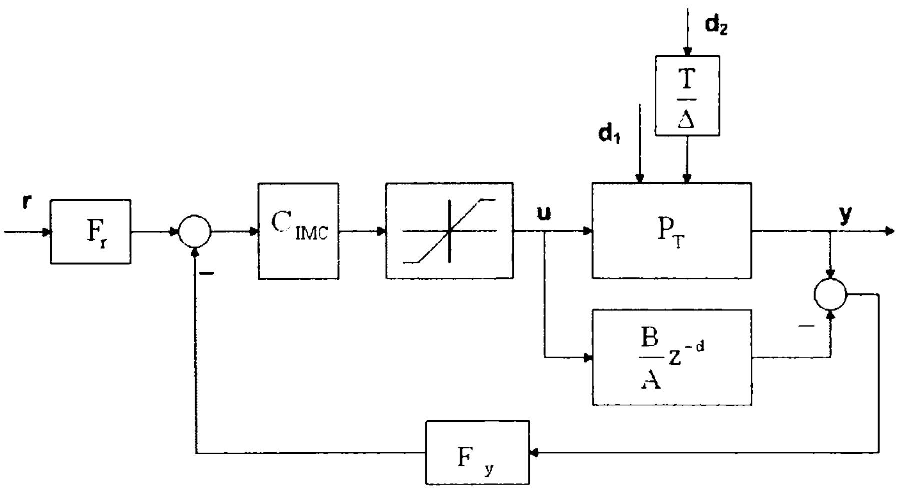


Fig.3.4-2. IMC structure with limitation

Making the two structures equivalent (the RST and the IMC), it results that:

$$F_r = T \quad , \quad F_y = S \quad , \quad C_{IMC} = \frac{A}{R\Delta A + BSz^{-d}} \tag{3.4-5}$$

Based on [III-26], [III-33] and [III-41] in Appendix 2 a detailed analysis of the IMC structure is presented:

- constraint handling in case of RST and IMC structures,
- influence of predictive parameters on closed loop poles.

3.5. An applied GPC cascade control solution for hydrogenerators

Based on [III-26] and [III-41], the paragraph presents a new concept in speed control of HGs (the plant was presented in Part I, chapter 3, figure 3.1-1), a Cascade GPC with Minimax Optimal Inner Control Loop, combining the advatages of LQ design technique for the inner loop and the GPC algoritm for the main loop. The plant can be separated into two decoupled parts. The MMs for the subsystems were presented in Part I, chapter 3, table 3.2-1. The proposed solution is based on a cascade control structure and design steps are presented.

The internal loop consists in a double integrating servo-system, which must be stabilized; the loop solves the rejection of inner deterministic disturbance. The solution is based on a minimax control for worst case of the inner disturbance.

The external loop consists in a GPC structure used in its polynomial representation, which has been transformed into an IMC structure. The disturbance acting on this part of the plant is the active power induced by the power system and considered as a stochastic one. Under these conditions the combined cascade control structure can be an attractive solution to ensure the control system performance enhancement. The solution has been tested through simulation for a numerical case study regarding to a real application [III-5].

3.5.1. The plant and its mathematical model

The plant is decomposed into sub-processes figure 3.2-1, (P1, P2), represents the structure of the plant: u(t) is the control signal, y(t) the controlled output. Different types of disturbances, $d_1(t)$ and $d_2(t)$, act on the plant, which must be particularized as function of the involved application. Local loops can be used for achieving good performances. The t.f. regarding the $u \rightarrow y$ is of form

$$H_p(s) = P(s) = P_1(s)P_2(s) \qquad (3.5\text{-}1)$$

In the considered application, the inner part P1 of the plant is the unstabilised hydraulic servo-system, which represents the actuator (the positioning system), Fig.3.5-1:

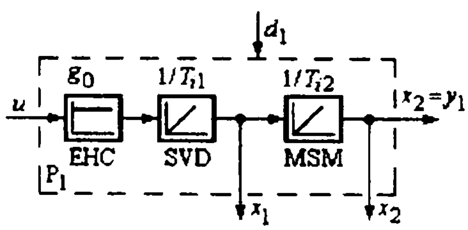


Fig.3.5-1. Simplified structure of unstabilised servo-system

where, EHC – the electro-hydraulic converter; SVD – the slide-valve distributor; MSM – the main servo-motor; u - the control signal, y_1 – the gate position of the turbine; x_1 and x_2 – the state variables associated with the SVD and MSM, respectively.

The local stabilization of the servosystem is solved practically based on a feedback loop [III-14]; the stabilising algorithms can be different. For example, the classical Mannesmann-Rexroth solution [III-15] is based on a pole-allocation technique, resulting a second order with lag mathematical model, which can be approximated with a first order system. In [III-91] a FC solution of an electrohydraulic servosystem under nonlinearity constraints is presented.

The external part corresponds to the hydraulic subsystem and the synchronous generator coupled to the power system (part I chapter 3), with modeling details given there. It is accepted that the servo-system can be stabilized in the form of $P_1(s)$:

$$P_1(s) = \frac{k_s}{1 + 2\varsigma T_s s + T_s^2 s^2} \qquad (3.5\text{-}2)$$

The load-type disturbance $d_1(t)$ acts at this level: the whole water column (having the height of the dam) [III-3], [III-6] is weighing on the system and modelled as a deterministic disturbance. The worst case is at 10% open blades of the turbine, the best case is when the blades are completely open. Suppose that the modifications of $d_1(t)=x_3(t)$ are given by:

$$T_e \dot{d}_1(t) + d_1(t) = d_1^0(t) \qquad (3.5\text{-}3)$$

$d_1^0(t)$ - constant.

So, the state-space mathematical model of P1 can be expressed as:

$$\begin{aligned} \dot{\underline{x}}(t) &= \underline{A}\,\underline{x}(t) + \underline{b}\,u(t) + \underline{L}\,d_1^0(t) \\ y_1(t) &= \underline{c}^T \underline{x}(t) \end{aligned} \qquad (3.5\text{-}4)$$

in the case of an additive to $x_1(t)$ - type disturbance $d_1(t)$:

$$\underline{A} = \begin{bmatrix} 0 & 0 & 0 \\ 1/T_{i2} & 0 & 1/T_{i2} \\ 0 & 0 & -1/T_e \end{bmatrix}, \underline{b} = \begin{bmatrix} g_0/T_{i1} \\ 0 \\ 0 \end{bmatrix}, \underline{c}^T = \begin{bmatrix} 0 & 1 & 0 \end{bmatrix}, \underline{L} = \begin{bmatrix} 0 \\ 0 \\ 1/T_e \end{bmatrix} \qquad (3.5\text{-}5)$$

The P2 part of the plant corresponds to the HT (T in figure), the penstock system (PsS) and the SG connected to the PS. For normal operating regimes the simplified linearized models were synthesized in part I chapter 3 table 3.1. It is assumed that the turbine is ideal, without losses and at full load, and P2 is given by [III-3] – [III-10]:

$$P_2^*(s) = k_{P2}^* \frac{1 - sT_w}{1 + sT_w/2} \qquad (3.5\text{-}6)$$

where $k_{P2}^* = k_w$ is the plant gain, T_w represents the water starting time. The HG (including the turbine) connected to the PS can be approximated with the t.f. $P_2^{**}(s)$:

$$P_2^{**}(s) = \frac{k_{P2}^{**}}{(a_m + sT_m)} \qquad (3.5\text{-}7)$$

where T_m stands for the mechanical time constant of the HG-SG depending on the rotating part's inertia. The parameter a_m has the values between $0.3 \le a_m \le 1.3$ with its extreme values $a_m = 0$ (before synchronization of the SG with the PS); the largest value for a_m is for SGs operating connected to the PS with infinite load.

In the design phase of the controller, values of $a_m=1$ are considered. In many situations the extreme case of $a_m = 0$ needs an alternative controller.

Connecting these subsystems the resulting t.f. for the plant results in $P_2(s)$:

$$P_2(s) = \frac{k_{P2}(1 - sT_w)}{(1 + sT_w / 2)(a_m + sT_m)} \tag{3.5-8}$$

Load disturbances $d_2(t)$ are induced by the power sysem and characterise the active power request. The t.f. of the plant controlled by the GPC is

$$H_p(s) = \frac{k_s}{(1 + 2\varsigma T_s s + T_s^2 s^2)} \frac{k_{P2}(1 - sT_w)}{(1 + sT_w / 2)(a_m + sT_m)} \tag{3.5-9}$$

In accordance with IEEE Committee reports, [III-7], [III-8], the resulted model (3.5-9) is widely used in studies regarded to modelling speed control systems of hydrogenerators, and current approaches with this respect include suboptimal control problems solved by various control techniques (see also paragraph 2.2).

The simultaneous rejection of the independently acting disturbances $d_1(t)$ and $d_2(t)$ by a single controller can become a difficult task. This is the reason why in the presented control structure their rejection becomes an independent task for two cascade controllers.

3.5.2. Disturbance rejection in cascade control structure

Under these conditions the combined cascade control structure can be an attractive solution to ensure the control system performance enhancement. The proposed cascade control structure is presented in Fig.3.2-2. The structure contains two controllers:

- the inner, a state-feedback controller that realizes the stabilized electro-hydraulic system (EHS) and insures adequate transients and behaviour regarding disturbance $d_1(t)$,
- the main controller, a GPC which deals with reference tracking and rejection of disturbance $d_2(t)$.

The stabilized EHS must ensure an aperiodic response (in (3.5-9) $\zeta \geq 1$, with distinct time constants T_{s1} , T_{s2}) or slightly oscillating, $0 << \zeta < 1$). The dynamics is imposed by the hydraulic part of the system. In steady-state regime $d_1(t)$ can be considered almost constant or slowly variable.

In case of disadvantageous behaviour of the system (see for example [III-10], [III-11], [III-14], [III-56]) the random disturbance on the loop, $d_2(t)$, induced by the electrical PS can be characterized as an oscillating disturbance with a relatively small value for ζ, which can be considered as the output of a second order lag filter with t.f. [III-30]:

$$F_{dist}(s) = \frac{\omega_0^2}{\omega_0^2 + 2\varsigma\omega_0 s + s^2} \quad \text{and} \quad d_2(s) = F_{dist}(s)v(s) \qquad \text{(a)} \quad (3.5\text{-}10)$$

$v(s)$ - step or impulse. ω_0 characterizes the frequency, specific for the power system in different functioning regime. Generally the value of ω_0 is not constant, but in a given point of the power system its variation range is not too large. In discrete form, they can be approximated also as random variations, described by discrete relation given in [III-30]:

$$d_2(t) = \frac{C(q^{-1})}{D(q^{-1})} \xi(t) \qquad \text{(b)} \ (3.5\text{-}10)$$

where $\xi(t)$ is a zero mean white noise.

Using spectral analysis techniques it is possible to detect the frequency around which the main disturbance is located. In the literature it is mentioned that such a disturbance filter can be tuned according to [III-30]:

$$\frac{C(q^{-1})}{D(q^{-1})} = \frac{1}{(1 - q^{-1})(1 - ae^{j\alpha} q^{-1})(1 + ae^{j\alpha} q^{-1})} \qquad (3.5\text{-}11)$$

where $\alpha=2\pi f_0 h$ (h the sampling period, f_0 – a characteristic frequency of power system [III-10]) and $0 << a < 1$.

3.5.3. The validation of the proposed control solution. Simulation results

The case study is dedicated to the speed control of a real-like application. The parameters of the plant in (3.5-8) and (3.5-9) according to [III-5], [III-15] are: $g_0 = 0.0625$, $T_{i1} = 0.001872$, $T_{i2} = 0.0756$, $k_\omega = 1$, $T_w = 2.2$ sec, $\alpha_m = 1$ (the SG is accepted to operate connected to the power system with infinite load), $T_m = 6.8$ sec.

The gain components for minimax control are calculated as presented in section 3.3. For this calculus a Matlab program was written. As a first step, the value of the parameter γ was established using interval halving (the parameter ρ is chosen to be $\rho=0.1$) applied to the CARE, and the final minimal value is $\gamma=\gamma_{min}=0.028$.

Solution to the CARE leads finally to the minimax game:

$$K_u = [43.706 \quad 99.311 \quad 40.879], \quad K_d = [149.58 \quad 345.35 \quad 140.65] \qquad (3.5\text{-}12)$$

Frequency domain analysis can be performed in order to check the performance of the control loop. Here only the closed-loop t.f. $P_1(s)$ is presented which is further needed for the cascade control structure. It is the following t.f. regarding to the reference u(t):

$$P_{CI}(s) = \frac{441.6}{s^2 + 1459s + 4.86 \bullet 10^4} \qquad (3.5\text{-}13)$$

In its simplified form the stabilized EHS can be modelled according to (3.5-1). Accepting the first-order Pade approximation for the dead-time element, the sub-plant P2 can be characterized by a t.f.:

$$P_2(s) = \frac{1 + 2.2s}{(1 + 1.1s)(1 + 6.8s)} e^{-4.4\,s} \qquad (3.5\text{-}14)$$

Since the stabilized EHS has a very small (neglectable) time constant compared to the rest of the process, and it disappears at sampling, at the design phase of the GPC outer loop only the steady-state value of $P_1(s)$ is taken into consideration, so the composed process t.f. is

$$P_T(s) = 0.0101 \cdot P_2(s)\ .$$

The second step is to apply GPC to the process, having the following GPC parameters:

$$N_1 = 1, \quad N_2 = 70, \quad N_u = 1, \quad \delta = 1, \quad \lambda_u = 0.1 \qquad (3.5\text{-}15)$$

Filter $T(q^{-1})$ was chosen for simplicity to be T=1. A sampling time of h=1.45 sec was chosen (so the discrete dead time results as $T_D=3$). The R and S polynomials are computed also with a Matlab program, and the results in this case are:

$$R(z^{-1}) = 0.183 + 0.0093z^{-1} + 0.0094z^{-2} + 0.0095z^{-3} + 0.0041z^{-4}$$
$$S(z^{-1}) = 11.1385 - 23.3389z^{-1} + 7.2004z^{-2} \quad (3.5\text{-}16)$$

The IMC structure parameters result:

$$F_r(z^{-1}) = T(z^{-1}) \quad , \quad F_W(z^{-1}) = T(z^{-1})$$
$$C(z^{-1}) = \frac{5.465 - 10.05z^{-1} + 5.84z^{-2} - 1.09z^{-3}}{1 - 2.788z^{-1} + 2.815z^{-2} - 1.214z^{-3} + 0.1896z^{-4}} \quad (3.5\text{-}17)$$

The testing through simulation of the proposed cascade control structure was performed in two different ways.

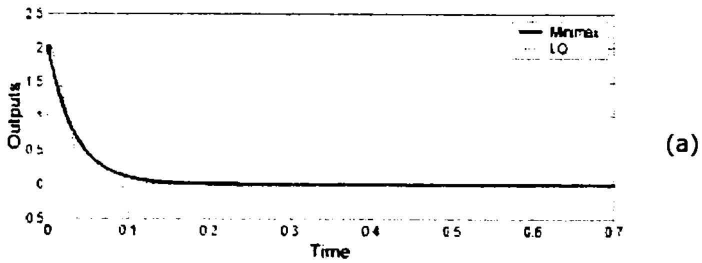
- First the inner loop was simulated as a continuous subsystem (this simulation is sustained by the real implementation solutions – analogue electronics, local microcontroller). A comparative test was made:
  - the control structure using minimax controller,
  - an LQ controller having tuning parameters Q=1 and R=0.01.

  Zero reference signal was supposed, and the initial conditions for the state variables were chosen for:

  $$x_1(0) = 1 \ , \quad x_2(0) = 2 \ , \quad x_3(0) = 1 \ , \quad (3.5\text{-}18)$$

  The figure 3.5-2 presents the output signal's and the states' evolution without disturbance.

  An exponentially decreasing disturbance of type d_1 is applied to both systems, acting as in (3.5-10); its time-evolution is depicted in figure 3.5-3. Applying this disturbance, the states and outputs are presented for both minimax and LQ control (Fig.3.5-4). It can be noticed that rejection of the disturbance is better in the case of the minimax control.
- Secondly, the GPC structure was tested through simulation. The following scenario was chosen: a step reference followed by two non-simultaneous disturbances, d_1 – type step disturbance (acting at time moment 100 sec) having the amplitude of -1, then d_2 – type disturbance (amplitude -0.05), according to the specific application and modelled with a second-order filter having the t.f.:



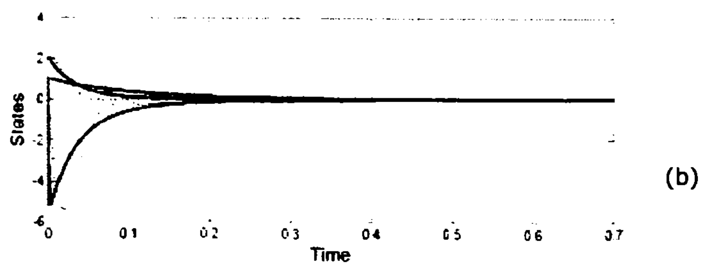


Fig.3.5-2. Outputs and states of minimax and LQ control

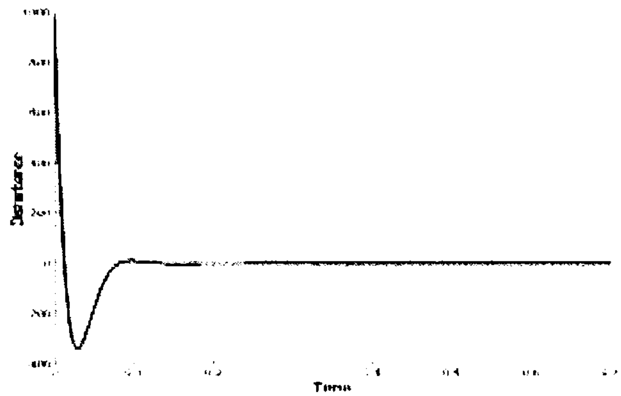


Fig.3.5-3. Generated disturbance d1

$$F_{dist}(s) = \frac{1}{1 + 0.5s + 0.4s^2} \tag{3.5-19}$$

The amplitude of the step is -0.05, acting at time moment 170sec. (this is similar to the one presented for example in [III-9], [III-10], Fig.3.5-5 emphasizes the simulation results.

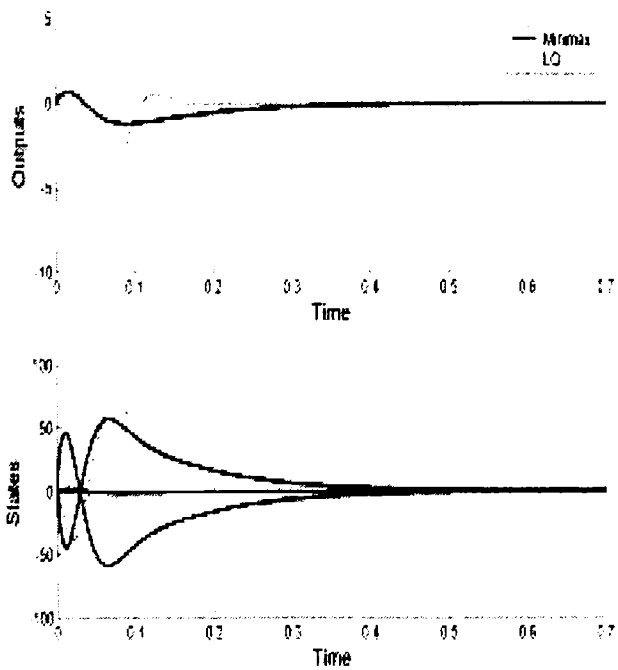


Fig.3.5-4. Disturbance rejection in minimax and LQ loops

From the analysis of the simulation it results a good behavior of the system both regarding reference tracking and disturbance rejection.

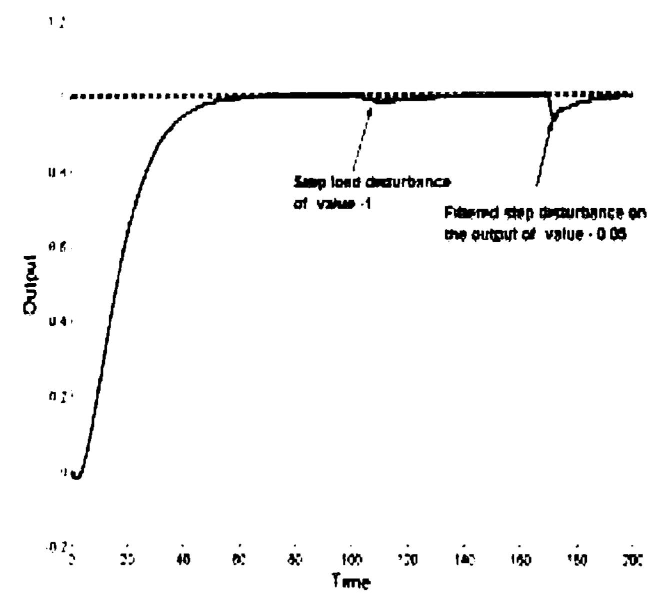


Fig.3.5-5. Simulation of cascade control system

3.6. Chapter conclusions

The chapter presents a two-stage CCS with an internal minimax state controller dedicated for rejecting internally located deterministic disturbances and a main GPC loop. In case of many applications the maximal value of the inner disturbance can be estimated or calculated. Since the design effort in case of the minimax controller is increased, a computer aided design is used both in this case and for GPC controller (and in IMC representation) as well. The use of the GPC controller under IMC representation based on the GPC's polynomial RST structure has the advantage of easy implementation.

Simulations give good results and sustains the efficiency both of the inner loop and of the GPC controller, regarding disturbances that are specific for the aimed applications.

Then, the control solution is applied to the speed control of hydro turbine generators. The solution involves a cascade control structure with an internal minimax state-feedback controller to reject internally located deterministic disturbances and a main GPC loop. The proposed approach is justified because in case of many applications (e.g. the presented one) the maximal value of the inner disturbance can be estimated or calculated.

Since the design effort in case of the minimax controller is increased, a computer-aided design is used both in this case and for GPC in IMC structure representation. Digital simulation results of the case study show that the control system ensures good performances. The results validate this control structure and its design method.

4. Fuzzy Control Solution for Hydrogenerators with Imposed Maximum Sensitivity Functions

The chapter is based on published researches results [III-15] and presents a new FC solution based on a TS-FC dedicated to the speed control of HTG; the controlled plant is characterized by models given in part I. In the first phase two conventional PI controllers are developed ensuring desired maximum values both for the sensitivity function and for the complementary sensitivity function in frequency domain.

Further by accepting the approximate equivalence between FCs and linear ones in certain conditions [III-54] (see also part IV of thesis), an attractive design method for four inputs-two outputs TS-FC is presented. The FC system guarantees maximum imposed sensitivity functions and therefore good responses with respect to modifications of reference and disturbance inputs, and robustness with respect to model uncertainties.

The proposed solution was tested through simulation of a case study regarding a modeled HG. The controller is compared with two PI controllers designed separately with respect to the reference input and with respect to one of the disturbance inputs.

4.1. Introduction

The controlled plant used in the speed and active power control of hydro generators was introduced in part I, chapter 3 and presented in details in paragraph 3.5. The simplified linearized models are synthesized in part I, chapter 3 table 3.1. The values of the variable positive parameters $\{T_w, k_\omega, a_\omega, T_m\}$ depend on the operating point.

The considered plant belongs to the class of second order non-minimum phase systems (NFSs) with two negative real poles and one positive zero called second-order "right half plane zero" systems [III-5]. The CS structure is presented in figure 4.1-1 (for abbreviations see part I chapter 3).

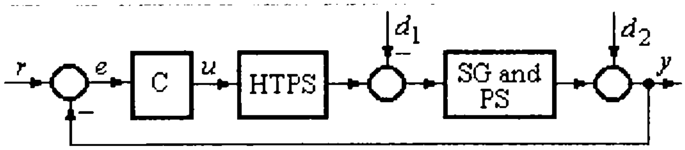


Fig.4.1-1. Control system structure

For NFS-s is often needed the use advanced, complex CS structures to ensure good control system performance [III-5]. As synthesized in chapter 2, current approaches to speed control of HTG include gain scheduling [III-45], robust control [III-9], optimal control [III-48] or predictive control [III-46] and chapter 3.

Due to the main advantages of FC in its basic version without dynamics this approach is also used in speed control of HTGs [III-5], [III-7], [III-63]. But, it is well considered that nowadays more than 90 % of control loops use conventional PI / PID controllers due to the very good control system performance they can offer

[III-96], including the applications in speed control of hydrogeneratorss [III-3], [III-7], [III-8] [III-9]. The introduction of dynamics in the structure of FCs leads to PD-, PI- or PID-FCs [III-76]. Under certain well-stated conditions the approximate equivalence between FCs and linear ones [III-54] is generally acknowledged. This is one of the reasons why there are widely accepted development methods for FCs by employing the merge between the knowledge on conventional linear PI controllers and the experience of experts in controlling the plant.

For the CS structure in Fig. 4.1-1 there are defined the sensitivity function $S(s)$ and the complementary sensitivity function $T(s)$ (see part II):

$$S(s) = \frac{1}{1 + H_C(s)H_P(s)} \quad , \quad T(s) = \frac{H_C(s)H_P(s)}{1 + H_C(s)H_P(s)} = 1 - S(s) \qquad (4.1\text{-}1)$$

where $H_C(s)$ is the t.f. of the controller and $H_P(s)$ is the t.f. of the plant:

$$H_P(s) = P_1(s)P_2(s) = \frac{k_\omega(1 - T_w s)}{(1 + (T_w/2)s)(\alpha_m + T_m s)} \qquad (4.1\text{-}2)$$

This can be done also in the time domain [III-57]. The maximum sensitivity M_s and the maximum complementary sensitivity M_p are defined in the frequency domain:

$$M_s = \max_{\omega \in \Re} |S(j\omega)| \quad , \quad M_p = \max_{\omega \in \Re} |T(j\omega)| \quad , \qquad (4.1\text{-}3)$$

and their typical values are within the domains [III-58]:

$$1.2 \le M_s \le 2.0 \quad , \quad 1 \le M_p \le 1.5 \qquad (4.1\text{-}4)$$

The CS specifications are expressed in terms of set point response, load disturbance response and robustness with respect to model uncertainties.
For very good set point tracking it is necessary that M_p should be as close to 1, and for very good disturbance rejection it is necessary that M_s should be as small as possible. It is very difficult to fulfill all the three specifications and a compromise is necessary to be made. With this respect, M_s and M_p can be used as design parameters.

Based on the accepted approximate equivalence between fuzzy controllers and linear ones, first linear PI controllers are developed to ensure the desired values of M_s and M_P. Then, the TS-FC is developed by accepting the well established property of Takagi-Sugeno fuzzy systems to be bumpless interpolators between linear controllers [III-59]. In this case, the TS-FC performs the interpolation between the separately developed linear PI controllers.

The FC solution is made up of a TS-FC structure and of its development method. Another advantage of the solution is that it guarantees (in terms of the accepted equivalence) the maximum sensitivities M_s and M_p of the developed fuzzy control system. In addition, since $|S(j\omega)|$ and $|T(j\omega)|$ are usually used to express conditions of robust performance [III-34], accompanied by the proper definitions of the weights, the proposed solution ensures quasi-robust fuzzy control systems of low cost.

4.2. Development of PI Controllers with Imposed Maximum Sensitivity Functions

In the first phase of the FC development, to control the nonminimum-phase plant (4.1-2) a PI controller is considered (Figure 4.1-1) with the t.f. $H_C(s)$:

$$H_C(s) = \frac{k_C}{sT_i}(1 + sT_i) \quad (4.2\text{-}1)$$

where k_C – the controller gain and T_i – the integral time constant. Based on recommendations given in [III-6] T_i is tuned to compensate the large time constant of the controlled plant fulfilling the supplementary recommendation:

$$T_i = T_m / a_m, \ a_m > 0 . \quad (4.2\text{-}2)$$

Using the connection (4.2-2), the design reduces to only one parameter to be tuned, k_C, as function of the imposed maximum sensitivities M_s and M_p as design parameters, presented in this paragraph.
Remark: The tuning equation (4.2-2) can be avoided, but that leads to both two controller parameters to be tuned; however, this approach will not be used here because it is relatively complex for the control systems designer.

4.2.1. Design using parameter M_s

The sensitivity function $S(s)$ can be expressed by using the t.f.s $H_C(s)$ and $H_P(s)$:

$$S(s) = \frac{T_m s(1 + (T_w / 2)s)}{(T_m T_w / 2)s^2 + (T_m - k_\omega T_w k_C)s + k_\omega k_C} . \quad (4.2\text{-}3)$$

Then, the magnitud function $|S(j\omega)|$ can be obtained immediately:

$$| S(j\omega) | = \frac{T_m \omega \sqrt{4 + T_w^2 \omega^2}}{\sqrt{T_m^2 T_w^2 \omega^4 + 4(T_m^2 - 3k_\omega T_m T_w k_C)\omega^2 + 4k_\omega^2 k_C^2}} = f(\omega), \quad f:[0, \infty) \to R \quad (4.2\text{-}4)$$

Solving the first optimization problem in (4.1-3) means the maximization of the function $f(\omega)$ in (4.2-4) with respect to ω. The maximum of $f(\omega)$ is achieved for $\omega = \omega_s$ given by (4.2-5):

$$\omega_s = \frac{1}{T_w} \sqrt{\frac{k_\omega T_w k_C + \sqrt{3k_\omega T_w k_C (4T_m - k_\omega T_w k_C)}}{3T_m \quad k_\omega T_w k_C}} \quad (4.2\text{-}5)$$

The substitution of $\omega = \omega_s$ in (4.2-4) leads to the maximum value of the modulus $|S(j\omega)|$, which is imposed to be equal with M_s. After some computations, the first condition in (4.1-3) becomes equivalent to the following:

$$\begin{aligned} &T_m^2(M_s^2 - 1)\sqrt{3k_\omega T_w k_C (4T_m - k_\omega T_w k_C)} \\ &- M_s^2[2(k_\omega T_w k_C)^3 - 12T_m(k_\omega T_w k_C)^2 + 19T_m^2 k_\omega T_w k_C - 6T_m^3] - T_m^2(6T_m - k_\omega T_w k_C) = 0. \end{aligned} \quad (4.2\text{-}6)$$

Equation (4.2-6) must be solved with respect to the controller gain k_C. Solving (4.2-6) for the desired values of the design parameter M_s requires the use of numerical techniques [III-62]. However, it can be guaranteed that (4.2-6) has a single root within the interval (4.2-7) which is equivalent to the Hurwitz-type stability constraint regarding the closed-loop control system:

$$0 < k_C < T_m / (k_\omega T_w) . \quad (4.2\text{-}7)$$

4.2.2. Design using parameter M_p

By using the transfer functions $H_C(s)$ and $H_P(s)$, the complementary sensitivity function T(s) can be expressed as:

$$T(s) = \frac{k_\omega k_C(1 - T_w s)}{(T_m T_w / 2)s^2 + (T_m - k_\omega T_w k_C)s + k_\omega k_C} \quad , \qquad (4.2\text{-}8)$$

and its magnitude function $|T(j\omega)|$ can be expressed in terms of (4.2-9):

$$| T(j\omega) |= \frac{2k_\omega k_C\sqrt{1 + T_w^2\omega^2}}{\sqrt{T_m^2 T_w^2 \omega^4 + 4(T_m^2 - 3k_\omega T_m T_w k_C)\omega^2 + 4k_\omega^2 k_C^2}} = g(\omega)\,, g:[0, \infty)\rightarrow R \qquad (4.2\text{-}9)$$

To solve the second optimization problem in (4.1-3) it is necessary to maximize the function $g(\omega)$ in (4.2-9) with respect to ω [III-62]; the maximum of $g(\omega)$ is obtained for $\omega = \omega_p$:

$$\omega_s = \frac{1}{T_w} \sqrt{-1 + \sqrt{3(4k_\omega T_w k_C / T_m - 1)}} \qquad (4.2\text{-}10)$$

The substitution of $\omega = \omega_p$ in (4.2-9) results in the maximum value of $|T(j\omega)|$, which is imposed to be equal with M_p. Computations yield the second condition in (4.1-3):

$$M_p^2 T_m \sqrt{3T_m(4k_\omega T_w k_C - T_m)} + M_p^2[2(k_\omega T_w k_C)^2 - 6T_m k_\omega T_w k_C + T_m^2] - 2(k_\omega T_w k_C)^2 = 0. \qquad (4.2\text{-}11)$$

Equation (4.2-11) must be solved with respect to the controller gain k_C for the desired values of the design parameter M_p, which can be done by numerical techniques. Equation (4.2-11) has at least two real roots within the interval (4.2-12):

$$T_m /(4k_\omega T_w) < k_C < T_m /(k_\omega T_w)\,, \qquad (4.2\text{-}12)$$

which fulfills the Hurwitz-type stability constraint (4.2-12) regarding the closed-loop control system.

Observing that T(s) is the closed-loop t.f. with respect to the reference input r(t), it is justified that the linear PI controllers tuned by imposing (4.2-2) and solving (4.2-11) ensure the optimization in the case of the dynamic regimes characterized by the modifications of r. These controller types are referred to as PI-C-r and have the tuning parameters denominated k_C^r (the solution to (4.2-11) as function of M_p) and T_i , tuned according (4.2-2).

Since S(s) represents the closed-loop transfer function with respect to $d_2(t)$, it can be accepted that the linear PI controllers tuned by imposing (4.2-2) and solving (4.2-6) ensure the optimization in the case of the dynamic regimes characterized by the modifications of $d_2(t)$. These controller types are referred to as PI-C-d and have the tuning parameter k_C^d, the solution to (4.2-6) as function of M_s) and T_i tuned with (4.2-2).

Remark: A similar design technology for $H_C(s)$ can be developed based on minimization of the phase reserve or the modulus reserve, definied in L(s), the open loop t.f. of the system:

$$L(s) = H_C(s)H_P(s) = \frac{k_C}{sT_i}(1 + sT_i)\frac{k_\omega(1 - T_w s)}{(1 + (T_w / 2)s)(a_m + T_m s)}$$

where the supplementary recommendation:

$$T_i = T_m / a_m, \ a_m > 0 . \qquad (4.2\text{-}2)$$

can be taken into account. The tuning equation (4.2-2) can be avoided, but that leads to both two controller parameters to be tuned.

4.3. Takagi-Sugeno Fuzzy Controller Structure and Development Method

To improve the CS performance the controller will be realized as a TS-FC. The development starts with the two continuous-time PI controller types developed in the previous paragraph, PI-C-r and PI-C-d. These linear PI controllers are discretized resulting in two incremental quasi-continuous digital PI controllers with the discrete-time Equations (4.3-1) and (4.3-2):

- for the incremental digital PI controller obtained from PI-C-r:

$$\Delta u_k = \Delta u_k^r = K_P^r \Delta e_k + K_I^r e_k , \qquad (4.3\text{-}1)$$

- for the incremental digital PI controller obtained from PI-C-d:

$$\Delta u_k = \Delta u_k^d = K_P^d \Delta e_k + K_I^d e_k , \qquad (4.3\text{-}2)$$

where $\Delta e_k=e_k-e_{k-1}$ and $\Delta u_k=u_k-u_{k-1}$ stand for the increments of control error and of control signal, respectively. The parameters of the two incremental digital PI controllers can be calculated in terms of (4.3-3) and (4.3-4) in the case of Tustin's discretization method (based on [III-76], see the short overview presented in part IV):

- for the parameters $\{K_P^r, K_I^r\}$ of the incremental digital PI controller obtained from PI-C-r:

$$K_P^r = k_C^r(1 - h/(2T_i)), \ K_I^r = k_C^r h / T_i \qquad (4.3\text{-}3)$$

- for the parameters $\{K_P^d, K_I^d\}$ of the incremental digital PI controller obtained from PI-C-d:

$$K_P^d = k_C^d(1 - h/(2T_i)), \ K_I^d = k_C^d h / T_i , \qquad (4.3\text{-}4)$$

where h is the sampling period.

The proposed TS-FC structure is presented in Fig. 4.3-1, and it consists of the basic four inputs-two outputs fuzzy controller (B-FC) and the linear blocks with dynamics.

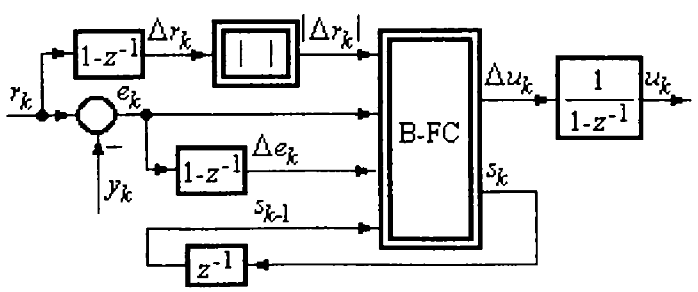


Fig.4.3-1. Takagi-Sugeno fuzzy controller structure

The fuzzy block B-FC represents a Takagi-Sugeno fuzzy system, it uses the max and min operators in the inference engine and employs the weighted average method for defuzzification ([III-49], [III-76]). The fuzzification is done by the membership functions illustrated in Fig. 4.3-2 ($\Delta r_k = r_k - r_{k-1}$ – increment of reference input), which points out the strictly speaking positive parameters of the TS-FC to be determined by the development method: $\{S_e, S_{\Delta e}, S_{\Delta r}, S_s\}$.

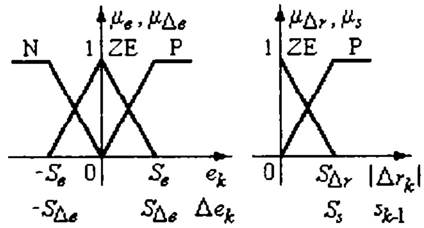


Fig. 4.3-2. Input membership functions

The B-FC block has two roles:

- to elaborate the control signal Δu_k,
- the role of observing the current dynamic regime by computing the variable s_k, with the linguistic terms "ZE" and "P" corresponding to the dynamic regimes caused by the modification of d_2 or r, respectively.

The inference engine of B-FC operates on the basis of the rule base presented in Table 4.3-1 and Table 4.3-2 as decision tables. The rule base for Δu_k can be reduced to only four rules and the rule base for s_k can be reduced to only two rules.

Table 4.3-1. Decision table to compute Δu_k.

				$\|\Delta r_k\|$					
				ZE			P		
				e_k			e_k		
				N	ZE	P	N	ZE	P
Δu_k	P	Δe_k	P	a_1	a_1	a_2	a_1	a_1	a_2
			ZE	a_1	a_3	a_1	a_1	a_1	a_1
			N	a_2	a_1	a_1	a_2	a_1	a_1
	ZE	Δe_k	P	a_3	a_3	a_4	a_1	a_1	a_2
			ZE	a_3	a_3	a_3	a_1	a_1	a_1
			N	a_4	a_3	a_3	a_2	a_1	a_1

Table 4.3-2. Decision table to compute s_k.

				$\|\Delta r_k\|$					
				ZE			P		
				e_k			e_k		
				N	ZE	P	N	ZE	P
s_{k-1}	P	Δe_k	P	S_s	S_s	S_s	S_s	S_s	S_s
			ZE	S_s	0	S_s	S_s	S_s	S_s
			N	S_s	S_s	S_s	S_s	S_s	S_s
	ZE	Δe_k	P	0	0	0	S_s	S_s	S_s
			ZE	0	0	0	S_s	S_s	S_s
			N	0	0	0	S_s	S_s	S_s

The following notations were used in Table 4.3-1: $a_1=\Delta u_k^r$ for a large value of M_p, $a_2=\Delta u_k^r$ for a small value of M_p, $a_3=\Delta u_k^d$ for a large value of M_s, $a_4=\Delta u_k^d$ for a small value of M_s. The presence of a_2 and of a_4 is necessary to alleviate the overshoot and the downshoot (specific to this class of NFSs) when e_k and Δe_k have the same signs.

Finaly, the proposed development method for the TS-FC consists of the following steps:

- Express the simplified mathematical model of the controlled plant in terms of (4.1-2);
- Choose the values of the design parameters M_p (two values) and M_s (two values) for the two continuous-time PI controllers, PI-C-r and PI-C-d, respectively, as function of the desired / imposed control system performance (general recommendations are given in [III-46];
- Tune the parameters of the linear PI controller types PI-C-r and PI-C-d: T_i by (4.2-2), k_C^r (two values) by solving (4.2-6) and k_C^d (two values) by solving (4.2-9);
- Choose an adequate value of the sampling period h, accepted by quasi-continuous digital control and take into account the presence of a zero-order hold (ZOH);
- Discretize the continuous-time PI controllers and compute in terms of (4.3-3) and (4.3-4) the parameters of the incremental quasi-continuous digital PI controllers $\{K_P^r, K_I^r\}$ (two sets of parameters) and $\{K_P^d, K_I^d\}$ (also two sets or parameters);
- Choose the value of the parameter S_e of the TS-FC in accordance with the experience of the control systems designer and apply Equation (4.3-5) corresponding to the modal equivalence principle to obtain the value of $S_{\Delta e}$:

$$S_{\Delta e} = [\max\{K_I^r / K_P^r, K_I^d / K_P^d\}] S_e \qquad (4.3\text{-}5)$$

  where the maximum is calculated for all four linear PI controller types;
- Choose the values of the other two TS-FC parameters, $S_{\Delta r}$ and S_s, by using (4.3-6) representing an adaptation of the recommendation given in [III-76] based on the fact that $S_{\Delta r}$ must be sufficiently small to point out clearly the constant values of r_k (exemplified here for an accepted unit step modification of r and a 2 % settling time) and that S_s must create a clear difference between the dynamic regimes concerning the modifications of r and of d_2:

$$S_{\Delta r} = 0.02, \ S_s = 1 \qquad (4.3\text{-}6)$$

Equations (4.3-5) and (4.3-6) will ensure the approximate equivalence between the proposed TS-FC and the linear PI controllers developed in paragraph 4.2.

4.4. Case Study. Digital Simulation Results

To validate the proposed fuzzy controller and its development method a case study dedicated to the speed control of a SG (like in power station Iron Gate 1, in Romania). For the considered case study the parameters of the plant in (4.1-2) are [III-5], [III-45]: $k_\omega = 1$, $T_w = 2.2$ s, $a_m = 1$ (the SG is accepted to operate connected to the PS with infinite load, known as the single machine infinite bus system situation [III-36]), $T_m = 6.8$ s.

As it was mentioned, first the linear PI controllers must be obtained. Applying (4.2-2) leads to T_i = 6.8 s. Solving equations (4.2-4) and (4.2-11) for the values of M_s and M_p in the intervals (5) results in the diagrams with k_C as function of M_s and M_p illustrated in Fig. 4.4-1.

For the sake of comparison, firstly two linear PI controllers are developed, one PI-C-r corresponding to α_2 in Table 4.3-1 and one PI-C-d corresponding to α_4 in Table 4.3-1. Imposing M_p= 1 and M_s= 1.2, the two controller gains will be k_C^r= 1.0306 and k_C^d= 0.4079.

To develop the TS-FC the design steps from paragraph 4.3 are applied. The main parameters involved in the TS-FC development are: k_C^r= 1.5665 (for α_3 in Table 4.3-1 and M_p= 1.5) and k_C^d= 1.3715 (for α_4 in Table 4.3-1 and M_s= 2), h=0.05 s, S_e= 0.3, $S_{\Delta e}$= 0.0022.

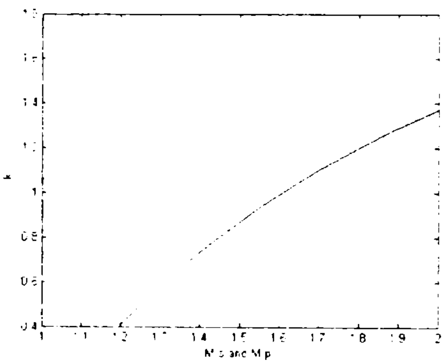


Fig. 4.4-1. k_C versus M_s (solid) and k_C versus M_p (dotted)

The developed control systems were tested through simulation, for the following scenario exemplified for both linear PI controllers and the TS-FC: a unit step reference followed by two non-simultaneous disturbances, d_1 – step disturbance acting at time moment 100 s, then d_2 – step disturbance acting at time moment 150 s.

The amplitude of both disturbance steps is 0.25, acceptable for the considered application. Part of the digital simulation results are presented in Fig. 4.4-2, Fig. 4.4-3 and Fig. 4.4-4.

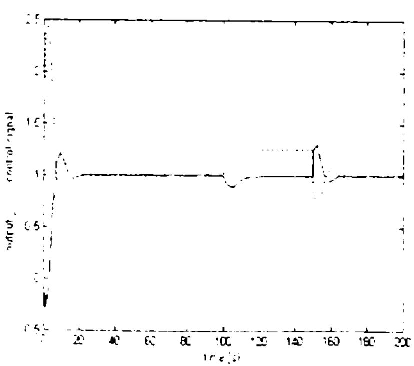


Fig. 4.4-2. Simulation results for CS with PI-C-r

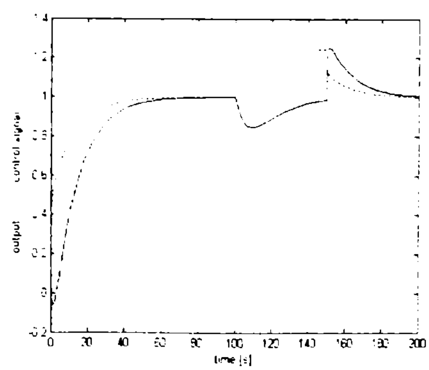


Fig. 4.4-3. Simulation results for CS with PI-C-d

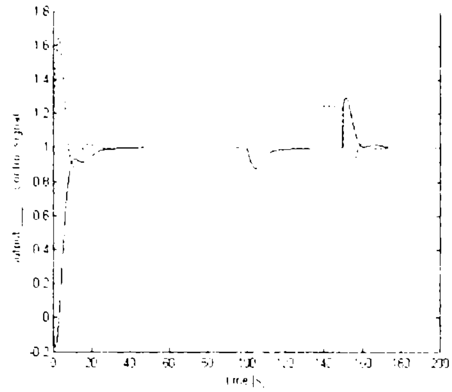


Fig. 4.4-4. Simulation results for control system with TS-FC

It can be noticed that the CS performances are very good:

- The PI controller tuned regarding the reference (PI-C-r) ensures a small overshoot but the behaviour regarding the disturbances is less good;
- The PI controller tuned regarding the load disturbances (PI-C-d) ensures much smaller overshoot but the behaviour regarding the reference is less good;
- The TS-DC with parameters tuned according to reference and load disturbance ensures regarding both imputs good behaviour, selcting the controllers which must be used.

4.5. Chapter conclusions

The chapter presents a new control development solution dedicated to the speed control of HTGs. The solution involves a four inputs-two outputs TS-FC, developed by starting with the design of two sets of conventional PI controllers ensuring the desired maximum values of the sensitivity function and of the complementary sensitivity function in the frequency domain.

Due to the accepted approximate equivalence between FCs and linear ones, it is justified to consider that the proposed approach guarantees the maximum imposed sensitivity functions for the developed fuzzy control system.

Simulation results for a case study prove the possible enhancement of control performances with respect to the modifications of the reference input and of the two disturbance inputs ensured by the proposed solution, in comparison with two solutions comprising conventional PI controllers. These good results can validate the new fuzzy controller and its development method. The presented FC solution can be implemented as low cost automation solution.

5. Part conclusions and contributions

This part of the thesis is dedicated to new solutions and development methods for speed control of a HTG conected to a PS.

Based on the synthesis presented in part I, chapter 3, the symplified MMs of the plant were presented.

Then, a new two-stage CCS with an internal minimax state controller dedicated for rejecting internally located deterministic disturbances and a main GPC loop is introduced. The use of the GPC controller under IMC representation based on the GPC's polynomial RST structure has the advantage of easy implementation. The control solution is applied to the speed control of HTGs. The solution involves the CCS with an internal minimax state-feedback controller to reject internally located deterministic disturbances and a main GPC loop. The proposed approach is justified because in case of many applications (e.g. the presented one) the maximal value of the inner disturbance can be estimated or calculated. Since the design effort in case of the minimax controller is increased, a CAD is used both in this case and for GPC in IMC structure representation. Digital simulation results of the case study show that the CS ensures good performances. The results validate this CS and its design method.

Chapter 4 introduce a new control development solution dedicated to the speed control of HTG. Due to the accepted approximate equivalence between FCs and linear ones, the contribution involves a four inputs-two outputs TS-FC, developed by starting with the design of two sets of conventional PI controllers ensuring the desired maximum values of the sensitivity function and of the complementary sensitivity function in the frequency domain.

Simulation results for a case study prove the possible enhancement of CS performances with respect to the modifications of the reference input and of the two disturbance inputs ensured by the proposed solution.

Both proposed solutions can satisfy good performance needs, and are viable alternative for HTG speed control. Their introduction on real applications depends on the acceptance of the control system designers, for whom tradition and safety are of high priority.

Part IV. Development of Fuzzy Controllers in delta domain

"My crystal ball is fuzzy" (Lotfi Zadeh)

1. Introduction. The structure of part

Part IV of the thesis presents the delta approach in design of control solutions; the theoretical part and the applications are based on the aim of delta approach defined by Middleton, R.H. and Goodwin, G.C. [IV-1], [IV-2] and synthesized in a unitary system-theory approach [IV-3]. The results of algorithmic design were verified by digital simulation.

To sustain the advantages of delta domain design instead a discrete one Chapter 2 presents CS design techniques in delta domain based on the delta model of the plant aiming to compare them with similar methods specific to continuous and discrete time design. Design of PI, PID and DB algorithms are presented and exemplified for second order benchmark type plants [IV-5], [IV-6].

Then a design technique in delta domain for IMC structure is presented and a generalised form for delta domain DB control algorithm is given [IV-7], [IV-8]. A hybrid implementation of the controller - Z-domain representation combined with delta domain representation - is presented and its architecture is compared with the pure delta-domain implementation. The effect of placing the limitations in the IMC structure is also studied. Paragraph 2.3.4 presents the IMC-based Smith predictor controller CS for plants with dead time.

The hybrid control structure is compared with a similar DB controller in Z-domain for second and third order plants (benchmarks). Illustrative sensitivity analysis between the hybrid control system and the Z-domain system was performed to underline the advantages of system design in delta domain.

Based on positive conclusions regarding the delta transform and its application in control system design (chapter 1), in Chapter 3 a design method in delta domain - referred to as delta domain design of low cost fuzzy control systems - meant for a class of servo-systems widely used in industry (energetic, see part III the servo-actuator, mechatronic systems) will be introduced, [IV-21], [IV-22].

The controlled plants are characterized by second-order dynamics of integral type, controlled by two-degree-of-freedom PI-fuzzy controllers. The method consists of three design steps based on continuous-time linear case design results expressed in terms of the ESO-method or 2p-SO-method, applied in delta domain, followed by the transfer of these results to the fuzzy case. The new design method and Mamdani PI-fuzzy controllers are validated by real-time experiments in controlling a servo-system with nonlinearities. The novelty consists in approaching the design from a delta point of view and highlighting the advantages of this approach:

- better control performances,
- a more reduced sensitivity when implementing the controller using a finite wordlength.

Finally, Chapter 4 synthesise the main conclusions of the part and the contributions.

2. Control Structure Design using Delta-Operator

2.1 The Delta transformation

The sampled models are traditionally represented using shift-operator parameterizations. In order to avoid loss of information, one should sample as fast as possible, but unfortunately the traditional shift-operator models fail to provide meaningful descriptions of the sampled system at fast sampling rates [IV-1]-[IV-4]. Sampling in digital control systems may cause unwanted behaviour in some cases; one disadvantage is the reduction of the stability region in discrete case to the unit circle, which is a much smaller area than the left half plane. Therefore the poles and zeros need a much more precise description (4 - 6 digits) at least.

To overcome these difficulties the delta parameterization or delta transformation based on signal differencing was introduced in 1990, [IV-1] by means of:

$$\delta = \frac{q-1}{h} \qquad \text{or} \qquad \gamma = \frac{z-1}{h} \tag{2.1-1}$$

where, q^{-1} and z^{-1} are the shift operator, γ (or δ) is the variable associated to the delta operator. Equivalently it results:

$$q = \delta h + 1 \qquad \text{or} \qquad q = \gamma h + 1 \tag{2.1-2}$$

Exists a close relationship between the discrete-time delta operator and the continuous-time differential operator at fast sampling rates, ($h \to 0$). The delta operator offers the same flexibility as the shift operator does in the description of discrete systems, yet has several advantages over it [IV-5], [IV-6].

One important property of the delta transformation is very useful in controller design. In the case of the delta transformation the stability domain expands as the sampling period gets smaller. The link between the three regions (s-continuous / Laplace, z- discrete and δ-delta domain) is shown in figure 2.1-1 [IV-2]. From (2.1-1) and (2.1-2) it is obvious that the stability region of the delta-operator is enclosed by a circle of radius 1/h and centered in (-1/h, 0).

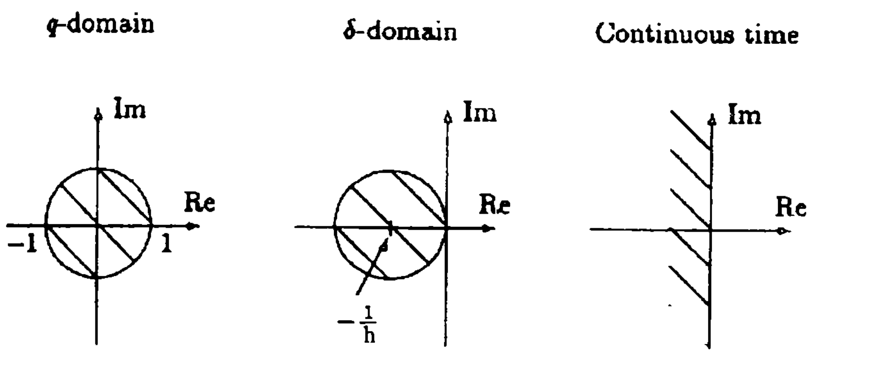


Fig.2.1-1. Stability regions [IV-2]

For $h \to 0$ the δ stability region converges to the open left-half plane which coincides with the stability region for the continuous-time differential equation. In [IV-5] is presented how do the dominant complex-conjugate poles of a second order continuous system with (ξ=0.707) behave in Z domain and in Delta domain. When using the Z transformation the well-known „heart-shaped" territory limited inside the unit circle is obtained (see Fig.2.1-2).

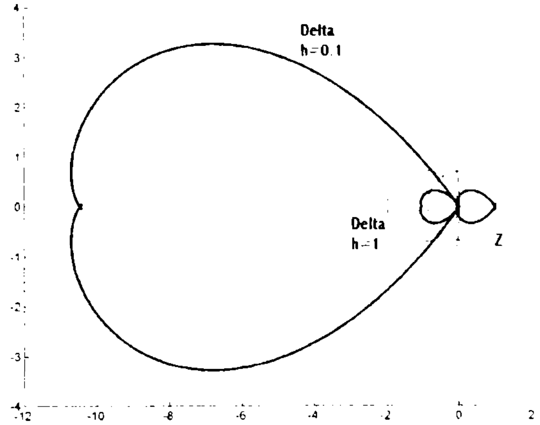


Fig.2.1-2. Dominant complex-conjugate poles locus in continuous, delta and Z domain

For the delta representation for h=1 it is shifted in the circle centered in (-1,0), and for h= 0.1 it becomes 10 times larger.
By reducing the sampling period even more the poles of the system in delta domain get closer to the continuous poles, especially the dominant ones. This remark sustains the idea of controller design techniques in delta domain being close to the design methods from continuous time.

2.2 Modelling in delta domain. Short overview

The aim of the modeling consists in finding a shift form model for the continuous system and then converting it to the delta model. In [IV-2] and [IV-3] many methods are presented; some of the aspects were synthesized and analyzed in [IV-6]:

- using the delta transformation formula;
- based on the state-space representation (in different variant)
- using the transfer function representation. In [IV-2] it is proved that the delta-t.f., when sampling with a ZOH, has the following expression:

$$H(\gamma) = \frac{\gamma}{1+h\gamma} T\{L^{-1}\{\frac{1}{s}H(s)\}\} \qquad (2.2\text{-}1)$$

where T{ } represents the generalized transform of a function:

$$F(\gamma) = T\{f(t)\} = \int_0^{\infty} f(\tau)E(\gamma-\tau)d\tau \qquad (2.2\text{-}2)$$

The use of these formulas to get the delta-t.f. has the following advantages:

- it is obtained directly from the continuous t.f. without needing to calculate the Z t.f. as well, intermediately;
- it already includes the Zero-Order-Hold in its expression;

- it highlights the similarities rather than the differences to the continuous t.f. (in opposite to the Z discrete form);
- the poles and zeros are much closer to the continuous form, in fact the closer the sampling time tends to zero the closer the poles and zeros would be to the continuous ones (see example in par.2.1).

Comparison studies regarded the continuous, the discrete and the delta representation and identification are made and presented in [IV-6] using the Delta-Toolbox. The use of the Delta Toolbox helps to calculate the different representational forms (transfer function, state-space), and allows us to make similar analyses like the Control System Toolbox [IV-11]. Using the delta domain representation it is relatively easy to check if the delta model is approximately the correct answer for a given discretization of a continuous system, but it is extremely difficult, if not impossible, for shift models.

In order to be able to simulate in SIMULINK just like in the continuous and shift case, in [IV-6] some S-Functions were written for t.f.s for both process and controllers.

For plants some examples has taken: Proportional with first order Lag (PL1) type, Proportional with second order Lag (PL2) and Proportional with third order Lag (PL3) type MMs. For controller a PI, a PID and a Proportional Derivative with first order Lag (PDL1) controller have been studied.

Such S-functions were used in different studies regarded to control structures development in delta domain, published at several international conferences [IV-5], [IV-7] - [IV-10]. Models are deduced and presented in [IV-5] and [IV-6] and will be appealed in this part of the thesis.

2.3. Controller design techniques in delta domain. Analysis and case studies

In [IV-3] there are presented different ways of designing delta-controllers:

- The controller design in continuous time according to the continuous model of the plant followed by the replacement of the variable s with the corresponding delta-domain variable, γ;
- The direct controller design in delta domain, based on the delta model of the process.

One advantage of this direct design is that the delta domain model of the plant already contains the zero-order-hold (ZOH). Based on the continuous t.f. the delta t.f. $H(\gamma)$ is obtained using relation (2.2-1).

Regarded to the results presented in papers [IV-5], [IV-6], [IV-7], [IV-8] this paragraph synthesizes the following controller design techniques based on the delta model of the plant:

- Optimization in delta domain using conventional (PID) controllers based on the extension of the MO-method, the ESO-method and the 2p-SO-method;
- Pole cancellation method and design in frequency domain,
- DB control.

The results obtained using the design in delta domain, were compared with results from the Z-discrete domain. Benchmarks for the second and third order plants (proportional with lag) are used in controller design. A new implementation technique is presented; simulation results exemplify the results of the study and sustain the main conclusions.

Other design techniques were developed and are presented in chapter 4.

2.3.1. PI(D) controller design in the delta domain based on MO-m and 2p-SO-m

Based on the first control application described in part I (chapter 2) the case of a second order continuous plant with real-negative poles and without zeros with a t.f. (2.3-1) was treated first [IV-6]:

$$H_P(s) = \frac{A_S}{(1 + T_{1S}s)(1 + T_{2S}s)} \tag{2.3-1}$$

Taking h as sampling period, the corresponding delta t.f. is calculated with relation:

$$H_P(\gamma) = A\frac{\tau\gamma + 1}{(T_1\gamma + 1)(T_2\gamma + 1)} \tag{2.3-2}$$

with $T_1 > T_2 >> \tau > 0$. τ result as the delta-transformation zero.

Two different conventional controller design techniques based on pole cancellation are presented:

- Based on optimization using (1) the MO-m; for the case $T_1 >(>) T_2 >> \tau > 0$ and (2) in case of disturbance acting on plant input, the 2p-SO-method, (see part II of the thesis).
- Using pole cancellation, design in the frequency domain imposing a phase margin of $\varphi_m = 60°$.

A. Design based on MO-method

Using a delta PI controller with a delta t.f. in form of [IV-6]:

$$H_{C\ PI}(\gamma) = k_C\frac{T_I\gamma + 1}{\gamma} \quad \text{with} \quad T_I = T \tag{2.3-3}$$

and applying the pole cancellation $T_I = T_1$, he corresponding closed-loop delta t.f. $H_r(\gamma)$ results in form of

$$H_r(\gamma) = \frac{1 + \tau\gamma}{[T_2/(k_C A)]\gamma^2 + [(1 + k_C A\tau)/(k_C A)]\gamma + 1} \quad (b_0 = a_0 = 1) \tag{2.3-4}$$

The parameter of the controller can be calculated by extension to the delta domain based on the optimization relations specific to the MO-method:

$$2a_0a_2 = a_1^2 \tag{2.3-5}$$

By applying this relation to the control system we obtain optimal behaviour; the parameter k_C is calculated using the relation (2.3-4) and (2.3-5) which takes into account the presence of the zero $z_1 = -1/\tau$. Parameter k_C will be calculated on basis of a second order equation for which the solution (+) can generate an unstable system; we take the second solution of:

$$k_C = \frac{2A(T_2 - \tau) \pm 2A\sqrt{T_2^2 - 2\tau T_2}}{2A^2\tau^2} \tag{2.3-6}$$

In continuous case the system performances are considered "optimal" (part II). In delta domain the system's performances can be considered as "sub-optimal", situated in the near vicinity of those from the continuous case. The difference is due to the design technique (these performances can be calculated on analytical way depending on the proportion of τ/T_2, see numerical example 1).

B. Design based on 2p-SO-method.

The delta t.f.s are calculated in form of relation (2.3-2). Supposing $T_1 >> T_2$ and using a delta PI controller with a delta t.f. in form of (2.3-3) the corresponding closed-loop delta t.f. $H_r(\gamma)$ results in form of:

$$H_r(\gamma) = \frac{T_i\tau\gamma^2 + (T_i + \tau)\gamma + 1}{\frac{T_1T_2}{k_CA}\gamma^3 + \frac{(T_1 + T_2) + k_CAT_i\tau}{k_CA}\gamma^2 + \frac{k_CA(T_i + \tau) + 1}{k_CA}\gamma + 1} \quad (2.3\text{-}7)$$

The design conditions are (see part II):

$$\beta^2 a_0 a_2 = a_1^2 \qquad \beta^2 a_1 a_3 = a_2^2 \quad (2.3\text{-}8)$$

β - design parameter with recommended values between 4,0 - 20. Through the second parameterization $m = T_2 / T_1$, the design relations can be easily obtained.

A simple simulation result illustrates the design. Let the continuous and the delta t.f. of the plant be:

$$H_P(s) = \frac{1}{(0.67s + 1)(0.33s + 1)} \quad \text{(a) and}$$

$$H_P(\gamma) = \frac{0.0538\gamma + 1}{(0.7212\gamma + 1)(0.3825\gamma + 1)} \quad \text{(b)} \quad (2.3\text{-}9)$$

$T_2 = 0.33$, $T_1 = 0.67$ with $T_2 < T_1$.

The model correspond to an electrical driving system (see Part I, paragraph 2.2.2, relation (2.2-5)). The MO-m method is applied in the mentioned two versions: (i) delta domain design, using a sampling time h=0.1; (ii) continuous design and discrete (Z) implementation.

The PI controllers - based on pole cancellation ($T_i = T_1$) and MO-m optimization has the following t.f-s:

$$H_{C\ PI}(\gamma) = 1.5315\,\frac{0.7212\gamma + 1}{\gamma} \quad \text{(delta);}$$

$$H_{C\ PI}(s) = 1.5152\,\frac{0.67s + 1}{s} \quad \text{(continuous)} \quad (2.3\text{-}10)$$

Discretizing the PI delta and the PI continuously working controllers using the trapezoidal rule and a sampling time of h=0.1, the discrete controllers result as:

$$H^{(\delta)}_{C\ PI}(z) = 1.5315\,\frac{0.7212z - 0.6212}{z - 1}\,, \quad H^{(c)}_{C\ PI}(z) = \frac{1.0910z - 0.9394}{z - 1} \quad (2.3\text{-}11)$$

Figure 2.3-1 presents the simulation results regarding the two systems. It can be noticed that the design in delta domain in this case leads to better results than the results in discrete Z-domain (obtained as described above). Compared with the continuous time design, the delta domain design results in a damping coefficient near to $\zeta \cong 0.7$, overshoot $\sigma_1 \cong 8$ %, settling time t_s and phase margin φ_m approximately, $t_s \cong 10 \cdot T_2$ and $\varphi_m \cong 60°$.

Other representative examples are given in [IV-5]. In all the presented comparative simulation results the differences are quite small but it can be concluded that the design in delta domain gives a slightly better performance than the design in Z domain.

The next main task was compare the "sensitivity" of the two control structures, the values of the poles and zeros were given with only two digits. In all analyzed cases, the control system using the discrete (Z) controller shows a bigger sensitivity [IV-5], [IV-9].

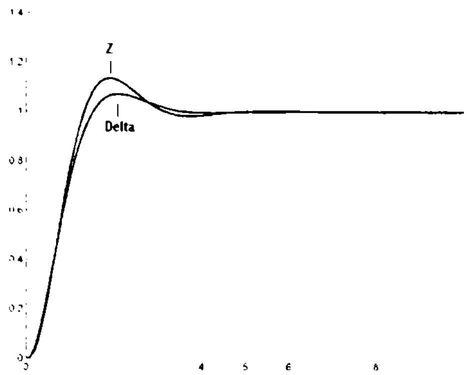


Fig.2.3-1. Simulation results for a PI controller design in delta domain and a discrete PI controller (MO-method)

2.3.2. Dead-Beat controller design in delta domain

Dead-beat (DB) control is an alternative, efficient control solution for plants with real-negative poles and without zeros [IV-12], [IV-21] and piecewise constant reference. In the analyzed case, [IV-5], [IV-6], the plant given by t.f. (2.3-1) (PL2 without zeros) and a delta model in form (2.3-2). The DB controller settling time is necessary of two sampling periods (for other supplementary constrains the number of sampling periods must be greater). In delta design, under condition of not allowing the plant zero ($z=-1/\tau$) cancellation, the DB condition for the closed loop t.f. would be:

$$H_r(\gamma) = G(\gamma) = \frac{\tau\gamma + 1}{(h\gamma + 1)^2} \qquad (2.3\text{-}12)$$

Based on [IV-12], [IV-21], the DB controller can be designed using (2.3-13):

$$H_{C\ DB}(\gamma) = \frac{G(\gamma)}{H_P(\gamma)[1 - G(\gamma)]} \qquad (2.3\text{-}13)$$

This delta DB controller ensures a control time in two sampling periods and no inter-sampling oscillations (see the numerical example). Limitations in the control signal have not been taken into account; a such an example [IV-9] is not presented here.

The design method is exemplified considering the same continuous time plant, the same sampling time h=0.1 and the delta process model (2.3-9). Based on (2.3-13) the corresponding DB controller results with t.f. (2.3-14):

$$H_{C\ DB}(\gamma) = \frac{(1 + 0.7134\gamma)(1 + 0.3904\gamma)}{0.1462\gamma(1 + 0.0684\gamma)} \qquad (2.3\text{-}14)$$

For comparison, let us consider the following discrete plant model (2.3-15)

$$H_P(z) = \frac{0.0195z + 0.0168}{z^2 - 1.5999z + 0.6362} \qquad (2.3\text{-}15)$$

and a DB discrete controller designed in Z-domain with a settling time of two sampling periods [IV-12]:

$$H_{C\text{-}DB}(z) = \frac{B^{-}(z)z^{-2}}{H_P(z)[1 - B^{-}(z)z^{-2}]}$$

$$H_{C\text{-}DB}(z) = \frac{27.5482z^2 - 44.0744z + 17.5262}{z^2 - 0.5372z - 0.4628} \qquad (2.3\text{-}16)$$

Comparing the step response simulation results of these two systems, see Figure 2.3-2, it can be noticed that the behaviour of the two systems is mainly coinciding.

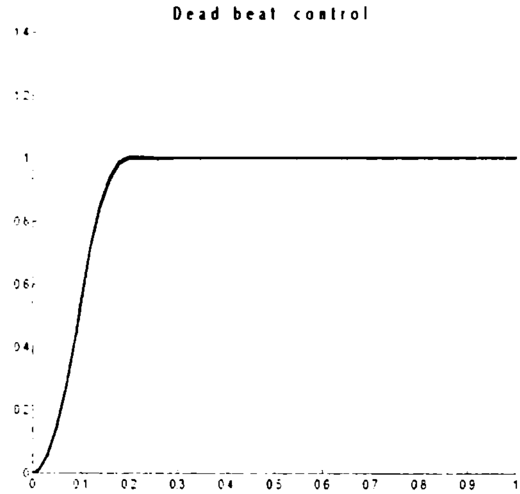


Fig.2.3-2. Simulation results for DB control systems in delta and in Z domain

Other detailed case studies regarding to third order (benchmark-type) plants are presented in [IV-5]. All case studies provide that the main advantage of use of delta controller instead of discretized PI(D) controller consists in better control performances.

These conclusions justified the delta approach design of FCs.

2.3.3. Hybrid IMC Dead-Beat Controller Design in Delta Domain. Case studies

Based on the results presented mainly in papers [IV-7]-[IV-11], the paragraph presents a new hybrid control system design technique in delta domain for IMC structure based on the delta model of the plant extended with the dead time represented directly in discrete domain (Z).

Also a generalised form for delta domain DB control algorithm is given, a hybrid implementation of the controller (Z-domain combined with delta domain) is presented and its architecture is compared with the pure delta-domain implementation. In this case the effect of placing the limitations in the IMC structure is also studied.

The proposed control structure was compared with a similar DB controller in Z-domain for second and third order benchmark-type plants. An illustrative sensitivity analysis between the hybrid control system and the Z-domain system was performed; the sensitivity analysis points out some useful conclusions which sustain the viability of the proposed method. Benchmark type plant model was used for the controller design. Simulation results exemplify the good results of the study.

A. Generalized Dead-beat control using IMC structure with hybrid delta domain and Z-domain implementation

Considering the t.f. of a continuous plant and its discrete t.f. (h the. sampling time) in form of:

$$H_P(s) = \frac{B_C(s)}{A_C(s)} e^{-sT_m}, \quad H_P(z) = \frac{B_D(z)}{A_D(z)} z^{-d} \qquad (2.3\text{-}17)$$

where $d = T_m / h$ - integer

The corresponding delta t.f., $H_P(\gamma)$, results in form of :

$$H_P(\gamma) = \frac{B(\gamma)}{A(\gamma)} \frac{1}{(h\gamma + 1)^d} \qquad (2.3\text{-}18)$$

The zeros of the plant can be decomposed into cancelling zeros B^+ and non-cancelling zeros B^- (zeros inside and outside respectively of the "heart-shape", Fig.2.1-2, including inverse unstable zeros as well), it results:

$$B(\gamma) = B^+(\gamma) B^-(\gamma) \qquad (2.3\text{-}19)$$

Taking into account the requirement of zero steady-state error, the following condition is imposed:

$$B^-(0) = 1 \qquad (2.3\text{-}20)$$

If the design of a DB-controller in delta domain is wanted, then for the closed loop t.f. $H_r(\gamma)$, must be imposed the form:

$$H_r(\gamma) = \frac{H_C(\gamma) H_P(\gamma)}{1 + H_C(\gamma) H_P(\gamma)} = B^-(\gamma) \frac{1}{(h\gamma + 1)^d} \frac{1}{(h\gamma + 1)^n} \qquad (2.3\text{-}21)$$

where n is defined as $n = 1 + \deg(B^-)$. The imposed closed loop t.f. (2.3-21) results from the considerations that first of all the minimum number of sample times in which DB behaviour can be achieved is n. Secondly, also the dead time must be taken into account when imposing certain closed loop behaviour. Last but not least the non-cancelling zero has to be excluded from the controller's expression in order to avoid inter-sampling oscillations and also zero steady-state error.

The general expression of the controller can be calculated from equation (2.3-21):

$$H_C(\gamma) = \frac{A(h\gamma + 1)^d}{B^+[(h\gamma + 1)^{d+n} - B^-]} \qquad (2.3\text{-}22)$$

The main disadvantage is that when using small sampling period and the dead time is high, the expression of the controller becomes complicated, the powers of the dead time in the denominator result in very small numbers, thus the use of the delta transformation looses its advantage.

One alternative solution proposed in [IV-7], [IV-8], [IV-10] which eliminates this disadvantage and combines both delta models and Z-domain representation using an IMC structure, is presented in Fig.2.3-3. If the plant is open-loop stable (see [IV-13]-[IV-15]), then the IMC structures are very attractive for some control applications.

The design is relatively simple; in [IV-16] is shown that the classical IMC structure is actually equivalent to the Youla-Kucera parameterization of all stabilizing controllers [IV-17],[IV-18].

In case of delta domain design the use of the proposed hybrid structure (Fig.2.3-2) could be advantageous: the structure combines both Z-domain representation and delta-domain representation and so it combines the advantages of the delta parameterization and the possibility of dealing easily with the dead time by introducing it in Z-domain.

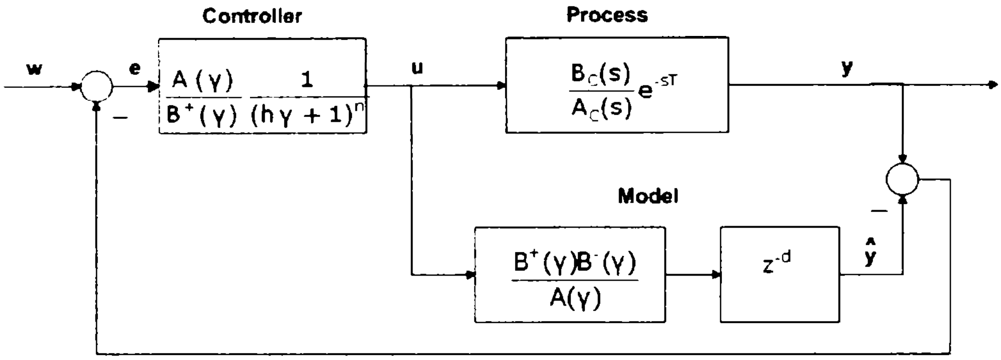


Fig.2.3-3. Hybrid IMC structure

Following this way, the controller can be adapted to different values of the dead time of the plant T_m by changing the value of d in the model (where $d=T_m/h$). Thus this controller is very sensitive to mismatch in the dead time (as well as pure Z - domain controllers of the kind).

The controller is obtained by imposing the closed loop transfer function the same condition as in (2.3-26). Accordingly, the controller t.f., $H_C(\gamma)$, obtains the expression:

$$H_C(\gamma) = \frac{A(\gamma)}{B^+(\gamma)} \frac{1}{(h\gamma+1)^n} \qquad (2.3\text{-}23)$$

and the internal model of the plant is hybrid; composed of a part in delta model and another one in the Z - domain (expression of the dead time) (figure 2.3-4).

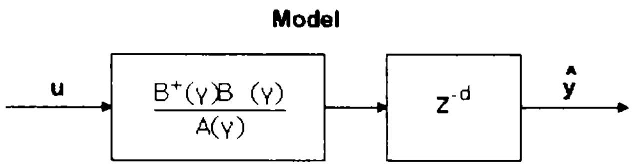


Fig.2.3-4. Internal hybrid model of the plant

A great advantage of the proposed hybrid architecture, figure 2.3-3, is that when modifying the value of the dead time T_m, the controller does not need to be redesigned, only a simple parameter change in the plant model in Z-domain solves in a favourable way the problem, the solution is very simple to implement.

The IMC delta controller resulting is also (2.3-23), and the plant model would be changed only in part of z^{-d}. Simulation results presented in [IV-8] and [IV-10] point out that there are no changes is the behavior of the system, only the effect of the increased dead time appears.

B. Effect of limitations in the hybrid IMC structure

A significant binding of control with practice is represented by the introduction of limitations in the controller's structure. The convenient placement of the limiting element represents a very important practical problem. In this context, the effect of the place of the control signal limitation within IMC structure has been also analyzed. Two significant cases are considered:

(1) When the limitation is inside the IMC structure (Fig.2.3-5),
(2) When the limitation is outside the IMC structure (Fig.2.3-6).

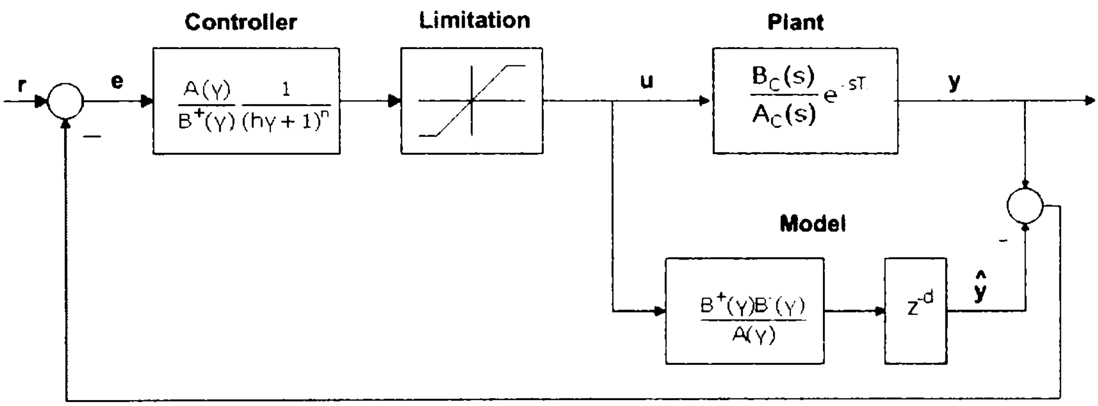


Fig.2.3-5. Hybrid IMC structure with limitation inside the model, case (1)

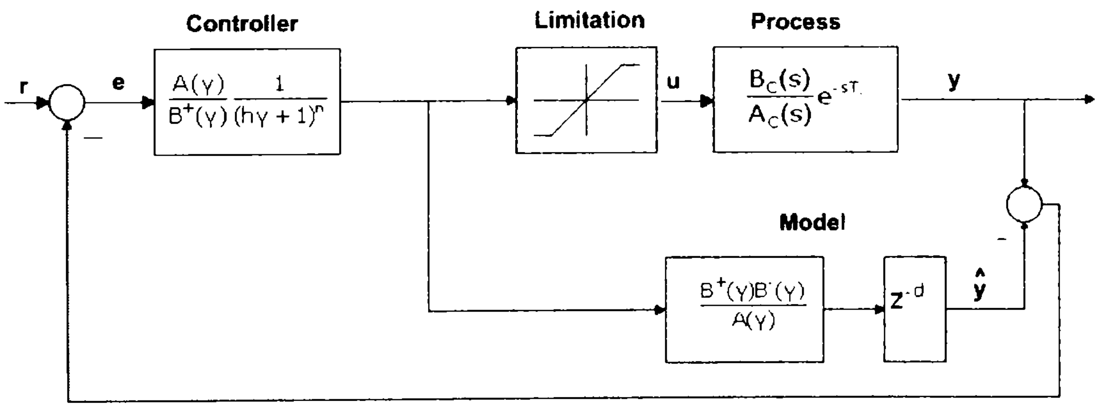


Fig.2.3-6. Hybrid IMC structure with limitation outside the model, case (2)

Taking a PL2 plant with dead time (the numerical example, was treated in details in [IV-7], [IV-10]) with a sampling period of h=0.05, it leads to a discrete dead time d=2.

$$H_P(s) = \frac{1}{(0.67s+1)(0.33s+1)} e^{-0.1s} \qquad (2.3\text{-}24)$$

(the model correspond to an electrical driving system (see Part I, paragraph 2.2.2, relation (2.2-5), where the power electronic is taken into account).

Based on (2.2-1), the delta t.f. of the continuous plant results in form of:

$$H_P(\gamma) = \frac{0.0259\gamma + 1}{(0.6953\gamma + 1)(0.3556\gamma + 1)} \frac{1}{(0.05\gamma + 1)^2} \qquad (2.3\text{-}25)$$

Implementing the control structure according to the hybrid architecture developed, the controller results in form of:

$$H_C(\gamma) = \frac{(0.6953\gamma + 1)(0.3556\gamma + 1)}{(0.05\gamma + 1)^2} \qquad (2.3\text{-}26)$$

For a comparison of the solution let the Z-domain DB controller obtained on the same basis be analyzed. For the considered plant, the Z t.f. result as:

$$H_P(z) = \frac{0.0052z + 0.0049}{z^2 - 1.7875z + 0.7976}\frac{1}{z^2} \qquad (2.3\text{-}27)$$

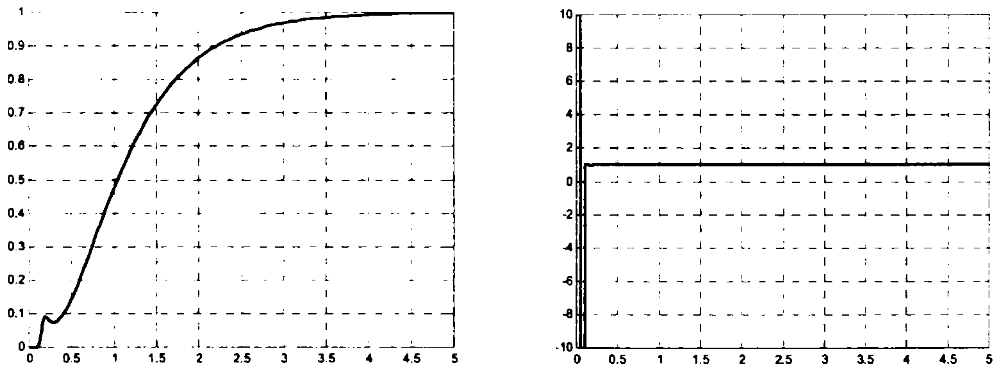
The developed DB controller in Z - domain, $H_C(z)$,is:

$$H_c(z) = \frac{z^2 - 1.7875z + 0.7976}{0.0101z^2} \qquad (2.3\text{-}28)$$

It has to be mentioned that the same IMC structure was used as in delta-domain, but using a controller only in Z-domain. There are no significant differences in the system responses but the control signal is quite high [IV-10].

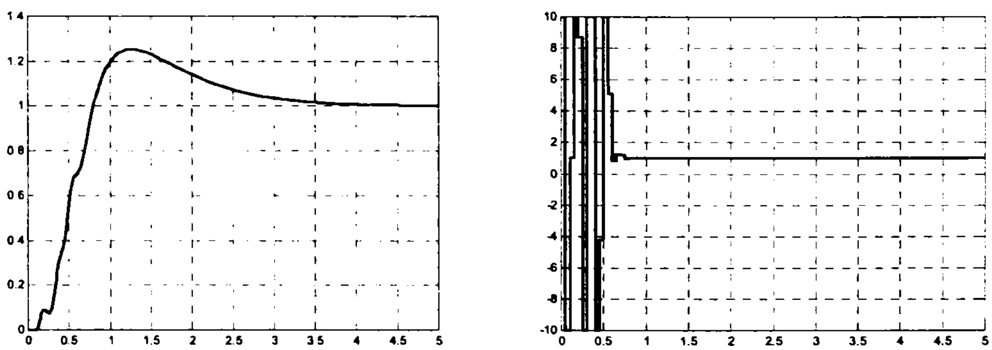
Imposing a limitation of ± 10 units (correlated with a concrete application these values can be changed adequately) at the output of the controller we obtain:

- for the first case a system response without overshoot ($\sigma_1=0$), (Fig.2.3-7),
- in the second case a response with overshoot (Fig.2.3-8).



(a) (b)

Fig.2.3-7. Step response and control signal in case of limitation inside the IMC structure



(a) (b)

Fig.2.3-8. Step response and control signal in case of limitation outside the IMC structure

Focused on comparing only the system outputs for both cases, it can be seen that when the limitation is incorporated inside the IMC structure not only there is no overshoot but the control signal does not oscillate so much between the extreme limits.

Similar investigations related to the location of the saturation have been considered in [IV-19]. It has to be also mentioned that the results are almost the same for hybrid IMC as well as for pure Z-domain IMC structure but the design in delta seems to be more user friendly.

C. Illustrative sensitivity analysis in case of a second order plant

For comparison suppose there are mismatches between the identified model and the actual plant; for example, the model's coefficients are established with only three digits accuracy. Also suppose the Z-domain DB controller is given by (2.3-28).

The delta model of the plant is (having three digits coefficients):

$$H_P(\gamma) = \frac{0.025\gamma + 1}{0.247\gamma^2 + 1.050\gamma + 1} \frac{1}{(0.050\gamma + 1)^2} \qquad (2.3\text{-}29)$$

and Z-domain model is (the same three digits coefficients convention has been followed):

$$H_P(z) = \frac{0.005z + 0.004}{z^2 - 1.787z + 0.797} \frac{1}{z^2} \qquad (2.3\text{-}30)$$

According to the IMC structure the internal model would be different from the plant, and also the actual controller part would be designed accordingly. This way the delta controller results in form of:

$$H_C(\gamma) = \frac{0.247\gamma^2 + 1.050\gamma + 1}{(0.050\gamma + 1)^2} = \frac{(0.694\gamma + 1)(0.355\gamma + 1)}{(0.050\gamma + 1)^2} \qquad (2.3\text{-}31)$$

and in Z-domain the controller would be:

$$H_C(z) = \frac{z^2 - 1.787z + 0.797}{0.010\, z^2} \qquad (2.3\text{-}32)$$

In the ideal case there is no mismatch and there are no disturbances, and so the control is open loop.

Comparing the step responses and the control signals of the two systems in closed loop Fig.2.3-9, it results that the pure Z-domain structure (dot-line) is more sensitive to parameter mismatch than the hybrid structure (line) but the static error is zero, due to the integrating effect of the controller.

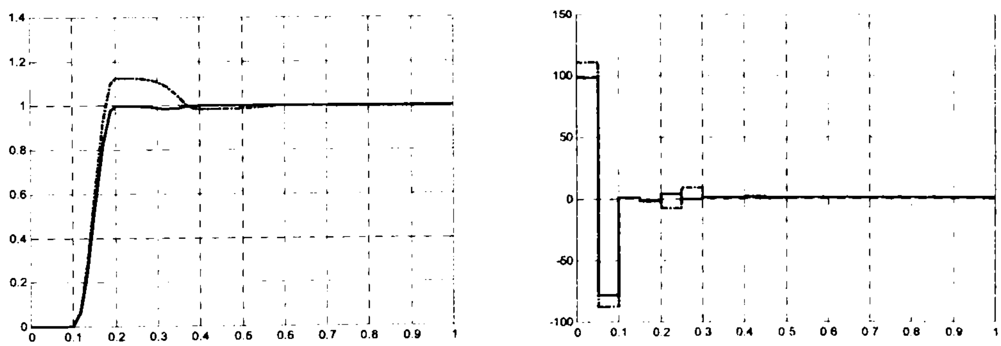


Fig.2.3-9. Sensitivity analysis between hybrid IMC structure and pure Z-domain structure

2.3.4. IMC-based Smith predictor for plants with dead time

In this case first the controller C is designed and in accordance to this, the Smith-predictor is calculated, followed by the verification of the system behavior [II-35], [IV-12], [IV-15]. The basic idea in designing a Smith-predictor is reflected in Figure 2.3-9 (a) and (b), where the dead time is virtually excluded from the closed loop, and the controller that is designed accordingly will be finally converted to the Smith predictor (design exemplified in z-domain):

$$C_{SM}(z) = \frac{H_{C\text{-}PID}(z)}{1 + H_{C\text{-}PID}(z)H_P(z)(1 - z^{-d})} \qquad (2.3\text{-}33)$$

Transposed to the hybrid z-delta domain IMC structure representation, the difference to the DB control scheme is depicted in Fig.2.3-10. In this case the controller t.f. $C_{SM}(\gamma)$ has the expression:

$$C_{SM}(\gamma) = \frac{H_{C\text{-}PID}(\gamma)}{1 + H_{C\text{-}PID}(\gamma)\dfrac{B(\gamma)}{A(\gamma)}} \qquad (2.3\text{-}34)$$

where, $H_{C\text{-}PID}(\gamma)$ denoting the t.f. of PID controller designed according to the plant without dead-time. The dead-time $d = T_m/h$ (d- integer) is represented in Z-domain and it can be easily adapted to the dead-time of the plant (T_m).

To exemplify the procedure, consider the same plant with t.f.-s (2.3-9) (a) and (b). The PID controller is designed to the plant model without dead-time, using pole cancellation and imposing a phase margin of $\varphi_m \approx 60°$. The delta t.f. of a PID controller is:

$$H_{C\text{-}PID}(\gamma) = \frac{K_C}{T_I\gamma}\,\frac{(T_I\gamma+1)(T_D\gamma+1)}{T_f\gamma+1} \quad \text{with} \quad k_C = \frac{K_C}{T_I} \qquad (2.3\text{-}35)$$

where, if applying the pole cancellation technique then $T_I = T_1$ and $T_D = T_2$ and k_C can be calculated imposing the desired phase margin. T_1 and T_2 are the biggest two time constants of the plant. T_f will be chosen in order to make the system quicker, but keeping in view the limitations in the control signal (in this case $T_f = T_2/4$ was chosen).

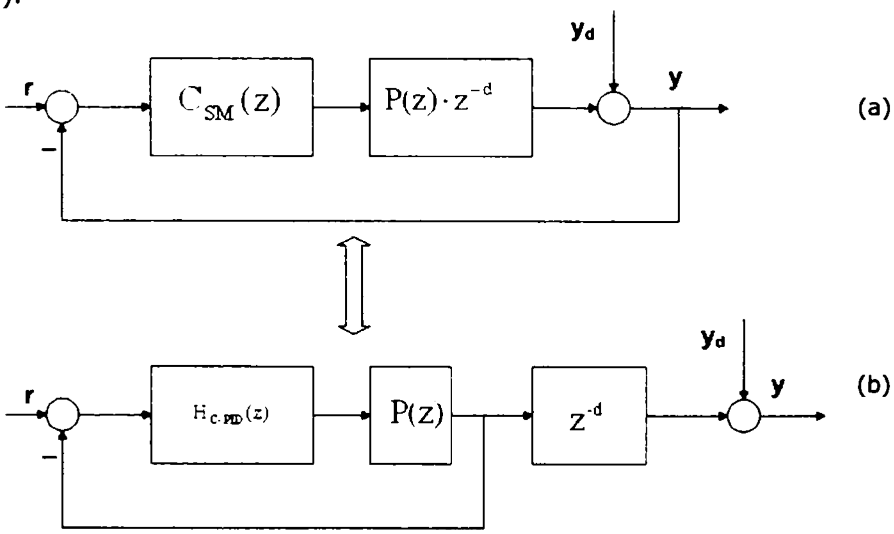


Fig.2.3-10. Theoretical transformation of dead-time compensation scheme

The numerical values of the PID controller are:

$$H_{C\text{-}PID}(\gamma) = \frac{1.6237\gamma^2 + 6.9007\gamma + 6.5662}{\gamma(0.0889\gamma + 1)} \qquad (2.3\text{-}36)$$

The controller from the IMC structure (see figure 2.3-11) will be in this case:

$$C(\gamma) = \frac{18.2623\,\gamma^2 + 77.6169\,\gamma + 73.8546}{\gamma^2 + 13.1635\,\gamma + 73.8546} \quad (2.3\text{-}37)$$

In case of a perfect match between plant and the model, the control is actually open loop, its response to unit step input is shown in Fig.2.3-12. In this case the overshoot is quite small, and the control signal does not have such a high value as at DB control (of course, settling time is much bigger). In paper [IV-10] the case of limitations in the control signal is also treated.
Remark: For this application two other adequate design techniques were used: once the use in delta domain of the well-known MO-method and other, the use in delta domain of the 2p-SO-m (see part II of the tesis).

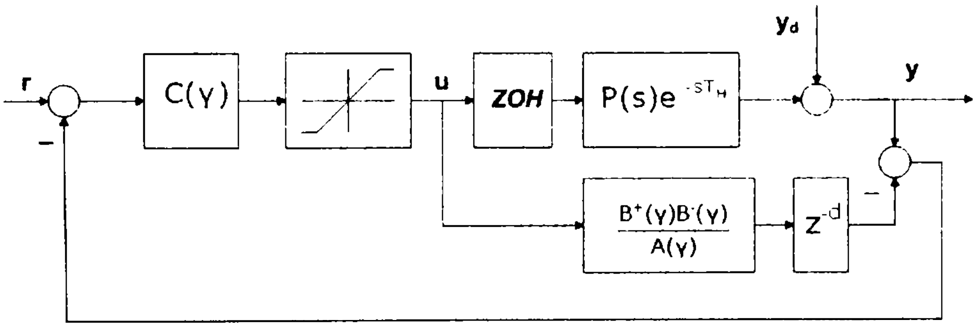


Fig.2.3-11. Hybrid z-delta domain Smith-predictor using IMC structure

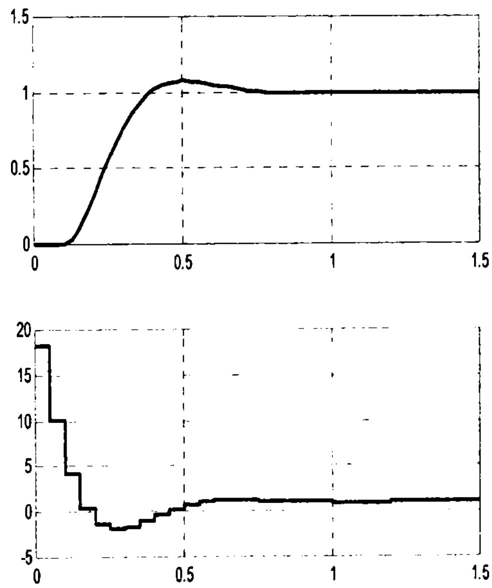


Fig.2.3-12. Dead-time compensation using PID controller

In case of Smith predictor algorithm also a detailed sensitivity analysis can be performed [IV-10]. Simulating the behaviour of the two systems it can be noticed that the z-domain representation (dot-line) is worse than the hybrid z-delta domain (solid line), Fig.2.3-13 (for a numerical example, given in [IV-10]).

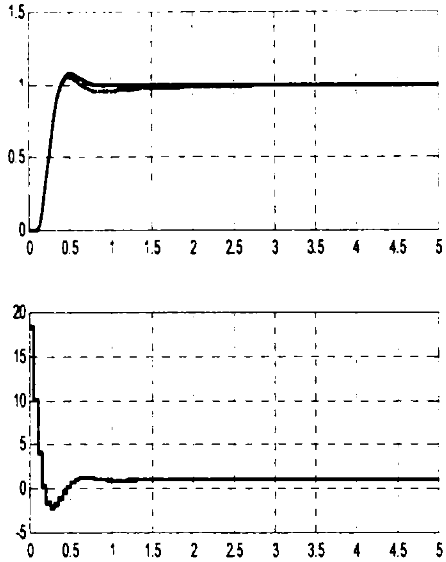


Fig.2.3-13. Closed loop behaviour in case of plant-model mismatch

2.4. Chapter conclusions

The chapter is based mainly on papers [IV-5], [IV-6], [IV-7], [IV-8] and [IV-10] (at all of this papers the author of the thesis is first author) and presents different controller design techniques based on the delta model of the plant:

- Optimization in delta domain using conventional (PID) controllers based on MO-method and in a similarly manner for 2p-SO-m; tuning relations in delta domain are deduced;
- Pole cancellation method, design in the frequency domain,
- DB control in different variants,
- A new approach to control system design based on the IMC in delta domain, with a mixed representation of the plant model within the IMC controller, [IV-7]. The method is based on the dual representation in delta and Z discrete domain of the plant, which has the advantage of the delta parameterization and the possibility of dealing easily with the dead time by introducing it in Z-domain.

As comparison criteria of the efficiency of these methods the followings were chosen:

- discrete controllers (PI, PID) designed under similar conditions;
- discrete DB controllers designed under similar conditions.

The comparison criteria were the difference in the step responses under same conditions obtained through simulation. The main conclusions that can be drawn are that the delta domain representation can stay for an attractive alternative in controller design.

The theoretical results have been verified through simulation on low order plants – benchmark types – frequently applied for verification of control solutions.

3. Delta Domain Design of Low Cost Fuzzy Controlled Servo systems

3.1. Introduction. The structure of the chapter

Within this chapter aspects regarding Mamdani-type fuzzy controller design are dealt with ([IV-23], [IV-24], [IV-25]), based on the approximate equivalence between linear controllers and fuzzy controllers. The aim is generally acknowledged, accepted and used [IV-26], [IV-27]. General aspects regarding the case of 1-DOF fuzzy controllers based mainly on research of Prof. R.-E. Precup (PhD supervisor) and his collective, and at which I have participated in the analysis and development of applications using PID and fuzzy quasi-PID controllers (for example papers [IV-23], [IV-29], [IV-31] (a case of sensitivity analysis)) are presented I paragraph 3.2.

Further, a design method in delta domain - referred to as delta domain design - of low cost fuzzy control systems meant for a class of servosystems widely used in industry (energetic, see part III the servo-actuator, mechatronic systems) is introduced, paragraph 3.3 two recent research results are synthesized:

As complementary research results (applicable also in delta domain) Appendix IV-1 presents the structure and the design of 2-DOF fuzzy controllers (2-DOF FC) (see for example papers [IV-39], [IV-22], [IV-33]), where I have elaborated the structure and design aspects of 2-DOF FCs based on linear design described in [IV-35]. The results were presented also in [IV-37].

3.2. PI and PID (1-DOF) fuzzy controllers with dynamics: a short overview

The paragraph is based mainly on references [IV-22], [IV-23], [IV-24], [IV-25]. The use of PI (PID) control strategies (in its different forms) can ensure well-known good control features:

- zero steady-state control error and constant (load) disturbance effect rejection, required by the majority of applications,
- enhancement of CS dynamics regarding to input (reference and (or) load disturbance),
- the positive practical experience gained in implementing the linear PI controller.

The fuzzy control is useful due to its capability to use a linguistic characterization of the controlled plant, but also to combine the two design approaches [IV-23], [IV-24], [IV-25], [IV-40], [IV-41], [IV-62].

The basic FC is without dynamics, having a specific nonlinear behaviour. The dynamics can be introduced using anticipative, derivative, integral and – more generally – predictive effects and adaptation to the actual operating conditions. The derivative (D) or the integral (I) components can be accomplished efficiently in its digital version, which creates a quasi-continuous (Q-C) equivalent of the analogue D and I components, respectively.

Two versions of quasi-PI (PID) fuzzy controllers (generically PI-FCs), are widely used:

- the position type Q-C PI-FC;
- the velocity type Q-C PI-FC .

The implementation form of the FCs is based on the discretized quasi-continuously working (Q-C) equivalent. They can be characterized using a pseudo-transfer function (p-t.f.). The Q-C PI FC-s can be systematically developed, by starting from the features of a basic linear PI(D) controller. For the standard PI-FC the integral effect can be introduced:

- on the output of the FC, the result being the standard version of the quasi-PI fuzzy controller with output integration (PI-FC-OI);
- on the input of the FC, the result being the standard version of the quasi-PI fuzzy controller with input integration (PI-FC-II).

❑ The block diagram of the standard with output integration (PI-FC-OI) controller is presented in figure 3.2-1.

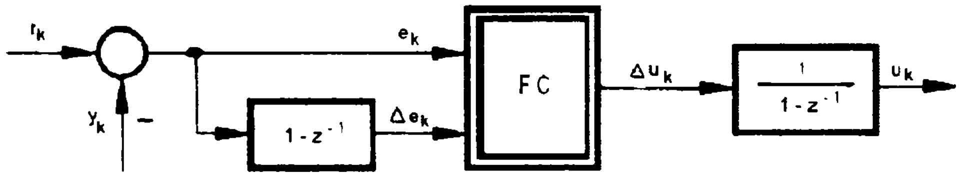


Fig.3.2-1. Block diagram of standard PI-FC-OI (see [IV-23])

In the development phase three parameters are available $\{B_e, B_{\Delta e}, B_{\Delta u}\}$; these - strictly positive parameters - are in correlation with the shapes of the m.f.s of the LTs corresponding to the input and output linguistic variables (LVs). The complete rule base is expressed as decision table (see for example Table 3.2-1).

Table 3.3-1 Decision table of the Fuzzy block B-FC

$\Delta^2 r_k \backslash \Delta r_k$	**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

The main steps for tuning the parameters $\{B_e, B_{\Delta e}, B_{\Delta u}\}$ are presented mainly in [IV-23], [IV-24], [IV-25].

The parameters of the basic linear continuous PI controller, $H_C(s)$,$\{k_C$ and $T_i,\}$ are obtained using a classical development method and they are taken into consideration in $\{B_e, B_{\Delta e}, B_{\Delta u}\}$ by applying this method for tuning the FC parameters.

The choice of the inference and defuzzification method represent the user's option (for example MAX-MIN composition rules and Centre of Gravity defuzzification method).

The incremental form Δu_k can be used in the CS in two basic ways:

- directly, if the actuator is of integral type
- by computing the effective value of control signal according to rel. (3.2-1):

$$u_k = u_{k-1} + \Delta u_k \,. \qquad (3.2\text{-}1)$$

- The block-diagram of the standard PI-FC with integration on the input of the controller (PI-FC-II). is presented in figure 3.2-2; it is based on a parallel version of the continuously working PI controller:

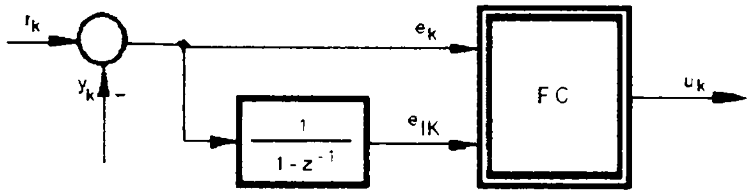


Fig.3.2-2. Block diagram of standard PI-FC-II (see [IV-23])

The standard PI-FC-II can be developed based on the analogy with the standard PI-FC-OI.
The structures of standard PI-FCs were used with specific adaptations in control of synchronous generators, electrical driving systems and mechatronic systems (mainly verified based on simulation results, [IV-29], [IV-30], [IV-33], [IV-34], [IV-62] with satisfactory results.
The particular features of the controlled plant could require modifications in the CS. These modifications can deal with:

- modifications in the rule base, imposed by the effective behaviour of the controlled plant caused by some special features (for example, the case of the non-minimum phase character);
- modifications in the fuzzification and defuzzification modules.

The introduction of dynamic components in the FC structure can create some difficulties mainly concerning:

- the modification of the introduced dynamics in different working points (regimes) of the controller;
- an increase in the number of the degrees of freedom in the development and implementation of the controller.

Based on the presented approach to PI-FC development, special fuzzy controllers can be mentioned (see [IV-29] [IV-30], [IV-34], [IV-62]):

- variable structure quasi-PI FCs and controller structures with fuzzy adaptation strategy of the parameters of the standard PI-FC;
- conventional controllers with fuzzy adaptation of parameters successfully applied to several applications (for example [IV-34]);
- PID fuzzy predictive FCs [IV-34].

These developments can contribute to fuzzy control system (FC-S) performance enhancement.

3.3. Delta domain design of fuzzy controllers

The extension of continuous design methods in delta domain allows the design of a new generation of low-cost fuzzy controllers [IV-52] - [IV-54]. One advantage of the direct design is that the delta domain model of the process already contains the ZOH. Based on the continuous t.f. the delta t.f. H(γ) is obtained using relation (2.2-1). Based on accumulated experience (references in PhD Report 3, [IV-6]) such continuous design methods are in particular the ESO method - for servosystems with integral component - or 2p-SO-method for second order plants with different time constants ($T_1 >> T_\Sigma$) and load disturbance.

3.3.1. Controller design

The use of a linear PI controller $H_C(s)$ is considered. The controller can be fuzzified as PI-FC with integration on the input of the controller (PI-FC-II) or on the output of the controller (PI-FC-OI), figure 3.3-1.

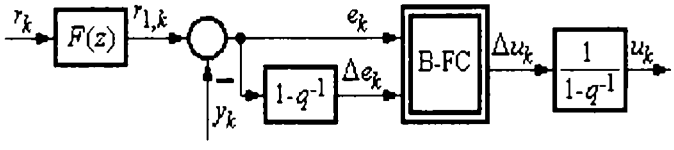


Fig.3.3-1. PI-FC OI structure with reference filter

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. Fuzzifying the discrete-time linear set-point filter F(z) may lead to other versions of PI-FCs.

The two-input-single-output nonlinear fuzzy-block B-FC 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 (MF) (see Appendix 3, chapter 2, figure A-2-3). The inference engine in B-FC employs Mamdani's MAX-MIN compositional rule of inference assisted by the rule base presented in Table 3.2-1, and the centre of gravity method is used for defuzzification. The PI-FC has only three tuning parameters, $B_e, B_{\Delta e}$ and $B_{\Delta u}$, to be obtained in terms of delta domain design.

Summing up all aspects presented in Sections 3.1 and 3.2, the delta domain design method dedicated to the benchmark type plants using FCs of the following design steps. The first two steps are dedicated to linear controller design, mainly in delta-domain and the third to B-FC design:

(1) Express the simplified plant model under the form of t.f. $H_P(s)$ (or P(s)) (low order one) with reduced number of parameters (for example, two). Then, apply an adequate design method in delta domain, calculate the delta domain form of the controller t.f. $H_C(\gamma)$ and choose an adequate developed set-point filter.

(2) Set the sampling period h, in accordance with the requirements of quasi-continuous digital control taking into account the presence of ZOH, and express the discrete-time equation of the set-point filter F(z). Apply the delta transform (2.3-1) to obtain the discrete-time equation of digital controller $H_C(z)$ in incremental version;

(3) Apply the modal equivalence principle [IV-26], calculate the equivalent fuzzy parameters $B_e, B_{\Delta e}$ and $B_{\Delta u}$, where the parameter B_e represents designer's option. The stability analysis of fuzzy CS or its sensitivity analysis with respect to parametric variations of controlled plant can be performed in order to get information on setting B_e.

3.3.2. Extension to 2-DOF controller design

For low order plants the 2-DOF design can be performed also for the continuous case [IV-35]. In this sense, based on the facts presented regarding the design of 1-DOF controllers in delta domain using reference filters, the design of 2-DOF FC controllers in delta domain is easy to deduce.

3.3.3. Application of ESO-m in delta domain: integrating with first order lag (IL1) type of plant and PI-FC-OI with reference filter design

The procedure was presented in [IV-21]. The plant is characterized by the continuous t.f. P(s) (see part II of the thesis):

$$H_P(s) = \frac{k_P}{s(1 + T_\Sigma s)} \qquad (3.3\text{-}1)$$

with k_P – controlled plant gain and T_Σ – small time constant or sum of parasitic time constants. The dynamics of actuator and measuring element are included in $H_P(s)$. The control solution based on ESO-method, [IV-49], ensures good performances with respect to both inputs. The control performances can be improved by using the reference filter (see part II). The PI controller $H_C(s)$ with reference filter F(s) is characterized by continuous t.f.:

$$H_C(s) = \frac{k_c}{s}(1 + sT_i) = \frac{k_C}{sT_i}(1 + sT_i) \quad \text{with} \quad k_c = k_C / T_i \qquad (3.3\text{-}2)$$

k_C – controller gain, T_i – integral time constant and

$$F(s) = \frac{1}{1 + T_i s}. \qquad (3.3\text{-}3)$$

Applying (2.2-1) to the controlled plant $H_P(s)$ the resulting delta domain t.f. is $H_P(\gamma)$ (τ result as the delta-transformation zero):

$$H_P(\gamma) = \frac{k_P[(1 + \tau\gamma]}{\gamma(1 + T_T\gamma)} \qquad \tau = (T_T - T_\Sigma), \qquad (3.3\text{-}4)$$

where the integral character is kept and the time constant of the delta t.f. is T_T is:

$$T_T = T_\Sigma \frac{\exp(h / T_\Sigma)}{\exp(h / T_\Sigma) - 1} > 0. \qquad (3.3\text{-}5)$$

Remarks: For $h \to 0$ (h - the sampling time) $T_T \to T_\Sigma$ and the parameters in the delta t.f. $P(\gamma)$ converge to the parameters in the continuous-time t.f. P(s). The introduced zero $z = -\frac{1}{(T_T - T_\Sigma)}$ is positioned left-side of the pole $p = -\frac{1}{T_T}$.

The design steps are those mentioned previously.

(1) Express the simplified plant model under the form of t.f. P(s) in (3.3-1) with two parameters, k_P and T_Σ. ;

(2) Apply the ESO method in delta domain by choosing the design parameter β as function of desired/imposed CS performance indices, use the PI tuning conditions (part II):

$$k_c = \frac{1}{\beta^{3/2} T_\Sigma^2 k_P} \quad , \quad T_i = \beta T_\Sigma . \tag{3.3-6}$$

to calculate the delta domain form of PI controller t.f. :

$$H_{C\text{-}PI}(\gamma) = k_c \frac{T_c \gamma + 1}{\gamma} = \frac{k_C}{T_i}(1 + \gamma T_i) \quad \text{with } T_c = T_i \quad k_c = k_C / T_i \tag{3.3-7}$$

and obtain the set-point filter parameters in (3.3-3).
Set the sampling period h, $h << T_\Sigma$ in accordance with the requirements of quasi-continuous digital control taking into account the presence of ZOH, and express the discrete-time equation of the-point filter F(z).

(3) Following this, apply the delta transform (2.2-1), (2.2-2) to obtain the discrete-time equation of digital PI controller $H_C(q^{-1})$ in incremental version [IV-25]:

$$\Delta u_k = K_P \Delta e_k + K_I \;\rightarrow\; e_k = K_P(\Delta e_k + \alpha e_k) \tag{3.3-8}$$

where:

$$K_P = k_c(T_i - h) > 0, K_I = k_c h > 0, \alpha = K_I / K_P = h/(T_i - h) \tag{3.3-9}$$

Apply the equivalence principle particularized as [IV-24], [IV-25]:

$$B_{\Delta e} = \alpha B_e, B_{\Delta u} = K_I B_e \tag{3.3-10}$$

where, the parameter B_e represents designer's option.

The fuzzification is solved in terms of 5 regularly distributed input and 7 output membership functions (singletons) shown in Fig. 3.3-2.

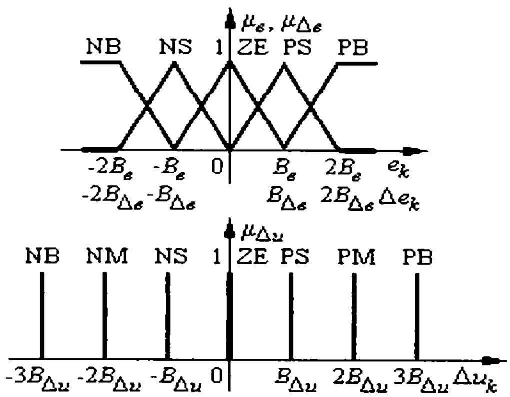


Fig.3.3-2. Membership functions associated to B-FC

The stability analysis of fuzzy CS or its sensitivity analysis with respect to parametric variations of controlled plant can be also performed [IV-25].

3.3.4. Application of MO-m in delta domain

The plant is a second order proportional with lag (PL2) type plant and the controller will be a PI-FC-OI (the plant model corresponds to an electrical driving system (see Part I, paragraph 2.2.2, relation (2.2-5). Let the t.f. of the continuous plant with real-negative poles and without zeros:

$$H_P(s) = \frac{k_p}{(1+T_{1S}s)(1+T_{2S}s)} \quad , \quad T_{1S} > (>>) T_{2S} \tag{3.3-11}$$

The corresponding delta t.f. is calculated with (2.2-1) results in form of:

$$H_P(\gamma) = \frac{k_p(\tau\gamma+1)}{(T_1\gamma+1)(T_2\gamma+1)} \tag{3.3-12}$$

with $T_1 > T_2 >> \tau > 0$, h - as sampling period and τ result as the delta-transformation zero.

Two different design techniques are presented (given from part II of thesis):

- The use of Modulus Optimum method (MO-m) based on pole cancellation;
- The use of 2p-SO-method, for the case $T_1 >> T_2 >(>) \tau > 0$ and disturbance acting on plant input (load disturbance).

(1) Design based on the MO-method. Suppose $T_1 > T_2$. Use a delta PI controller with a delta t.f. in form of:

$$H_{C\ PI}(\gamma) = k_c \frac{T_c\gamma+1}{\gamma} \quad \text{with} \quad T_i = T_c \quad k_c = k_C / T_i \tag{3.3-13}$$

and apply the pole cancellation $T_i = T_1$, the corresponding open-loop respectively closed-loop delta t.f.s result as presented in paragraph 2.3-1 (a).

(2) Design based on the 2p-SO-method. The 2p-SO-m was developed by the author and presented in part II. The delta t.f.s is considered in form of relation (2.2-1). Suppose $T_1 >> T_2$. and load disturbance. Use a delta PI controller with a delta t.f. in form of (3.3-7), the corresponding open-loop respectively closed-loop delta t.f.'s , $H_0(\gamma)$ and $H_r(\gamma)$ results in form presented in paragraph 2.3-1 (b).

The design steps are those mentioned previously and must be applied regarded to ESO-m or MO-method. Following this, apply the delta transform to obtain the discrete-time equation of digital PI controller $H_C(q^{-1})$ in incremental version (relations (3.3-8) – (3.3-10)); the parameter B_e represents designer's option.

3.4. Case Study and Real-Time Experiments

To validate the new design method of Mamdani PI FCs, it is considered a case study with the controlled plant consisting in a DC-m drive+load (DC-generator operating on a resistive load) which can be characterized through a linearized simplified MM of a proportional, second order with lag (PL2) type t.f. [IV-29].

The experimental setup can be applied in mechatronic systems (for example, see [IV-57]), and it consists of speed control of a laboratory DC-m driving system (AMIRA DR300) [IV-58], existing in research laboratory B-028-B of "Politehnica" University of Timisoara, Dept. of Automation and Applied Informatics). The complete picture of experimental setup (Fig. 3.4-1) illustrates the elastic coupling reflected in servosystem nonlinearity.

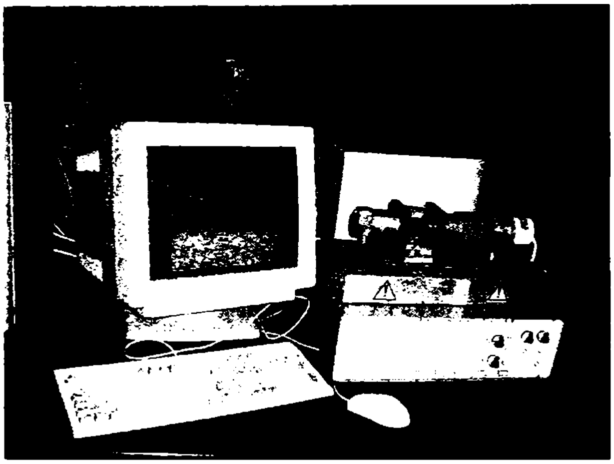


Fig. 3.4-1. Complete picture of experimental setup (DC driving system)

The DC-m 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 free drive-shaft.

The mathematical model of controlled plant can be well approximated by the transfer function P(s) in (3.3-1), decomposed as in figure 3.4-2 in two subsystems:

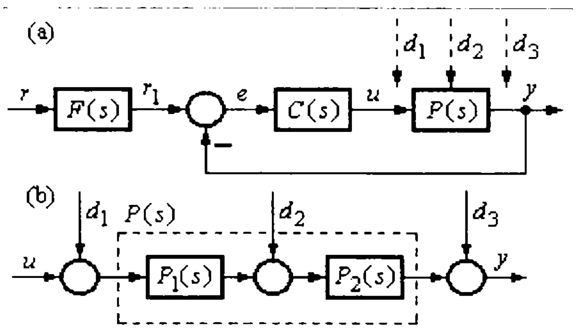


Fig. 3.4-2. Control system structure (a) and definition of load disturbance input scenarios (b)

$$H_{P1}(s)=\frac{k_{P1}}{1+T_{\Sigma}s} \quad \text{and} \quad H_{P2}(s)=\frac{k_{P2}}{s},\ k_P=k_{P1}k_{P2}, \qquad (3.4\text{-}1)$$

with $k_P = 4900, k_{P1} = 1, k_{P2} = 4900, T_{\Sigma} = 0.035$ sec. (the numerical data were determinate in research laboratory B-028-B in previous experiments and used also in diploma works [IV-61], [IV-62]). Applying (2.2-1) to the controlled plant $H_P(s)$ the resulting delta domain t.f. is $H_P(\gamma)$ (τ - the delta-transformation zero):

$$H_P(\gamma)=\frac{k_P[(1+\tau\gamma)]}{\gamma(1+T_T\gamma)} \qquad \tau=(T_T-T_{\Sigma}), \qquad (3.4\text{-}2)$$

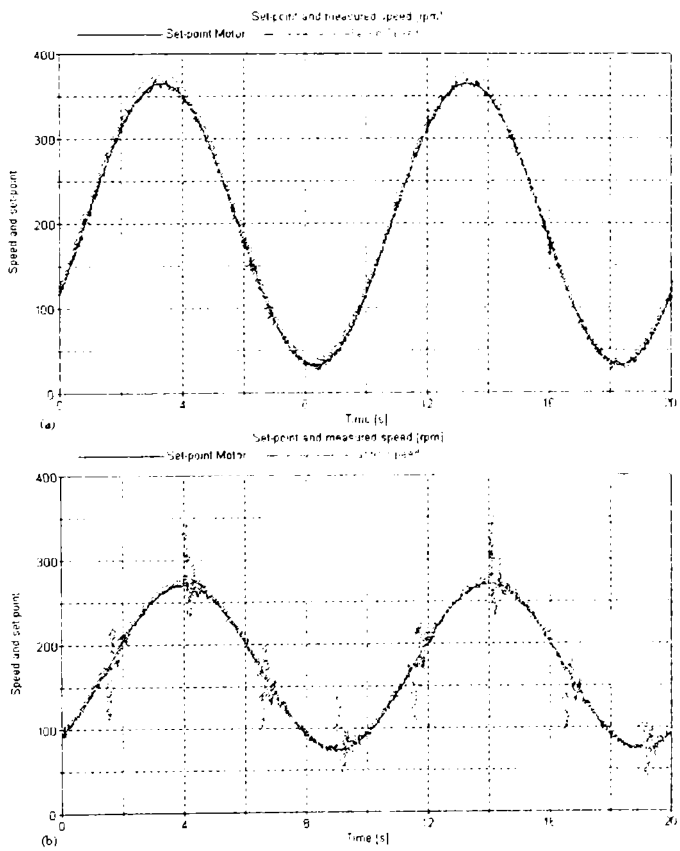
$k_P = 4900, k_{P1} = 1, k_{P2} = 4900$, h=0.01 sec; T_T calculated with relation (3.3-5) results $T_T = 0.037$ sec.

The delta design method presented in the previous paragraph continues with choosing the design parameter, $\beta = 6$ and the continuous-time PI controller tuning parameters obtain the values $k_C = 0.0113$ (or $k_C = 0.0024$) and $T_i = 0.21$ sec.

Next, the sampling period is set to the relatively small value, h=0.01 sec and the digital PI controller parameters (see rel.(3.3-8) result as $K_P = 0.0023$, $K_I = 1.13\ 10^{-4}$ and $a=0.05$.

Finally, the parameter B_e is set taking into consideration the range of set-point values, $B_e = 100$, and (3.3-9) leads to the other two PI-FC tuning parameters, $B_{\Delta e} = 5$, $B_{\Delta u} = 0.011$.

The fuzzification is solved in terms of 5 regularly distributed input and 7 output membership functions (singletons) shown in Fig. 3.3-2.



(a)

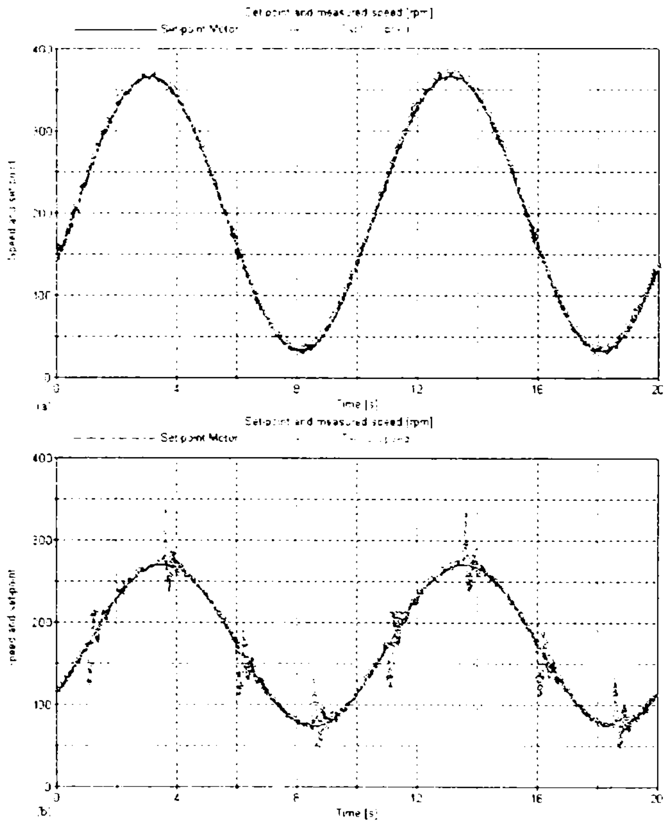
(b)

Fig. 3.4-2. Speed response of CS with PI controller

Part of the results of real-time experiments is presented in figure 3.4-3 and figure 3.4-4. The results are illustrated as variations of reference (the set-point speed with sinus variation with relatively large time period T_{sin}=10 sec.) and y (the controlled speed) versus time, in Fig. 3.4-3 for the linear control structure with PI controller and in Fig. 3.4-4 for the FC structure with Mamdani PI-FC developed in delta domain.

Two experimental scenarios were used: without load in Fig. 3.4-3 (a) and Fig. 3.4-4 (a), and with a T_{dis}=5 sec period of 10 % rated d_2-type load disturbance in Fig. 3.4-3 (b) and Fig. 3.4-4 (b). Note that these experiments have been done at low speeds.

Due to the facts that (1) the plant has only small nonlinearities and (2) at the fuzzification level were used 5 and 7 membership functions, it must be remarked that in this case, the differences between the linear case and the fuzzy case are not significant.



(a)

(b)

Fig. 3.4-4. Speed response of CS with Mamdani PI-FC

3.5. Chapter conclusions

The chapter propose a new design method dedicated to PI-FCs for benchmark-type plants (in the considered application, a second-order dynamics of integral type which models a class of servosystems). The low cost feature of the design method and suggested Mamdani PI-FCs results from the simplicity and transparency of the design method and FC structure as well, having in view relatively simple implementations, shown in Section 3.3.

This attractive design method consists of three design steps. The method was derived by transferring the linear design results, applying the ESO-method in delta domain, to the strictly speaking fuzzy controller design in terms of the equivalence principle.

As results also from chapter 2, the benefits of delta transform can be more interesting, if the suggested sampling period h is smaller than the value used [IV-4]. If the sampling period is set to smaller value the main difference between the Z and delta transform can be bigger. However, the discrete-time implementation is employed behind the delta transform.

Part of the results of real-time experiments conducted with one PI-fuzzy controller designed to control a (small nonlinear) servosystem validate both the delta design method and the fuzzy controllers. The results prove that the control system with Mamdani PI-FC ensures good performance indices; in comparison with the linear control system with PI controller with respect to modifications of set-point and one load disturbance input type, the performances are similarly; this is motivated by the small nonlinearity of the fuzzy controller.

4. Part conclusions

The chapter is structured in two main parts; each of theparts sythesised own conclusions, based on extended research results.

In chapter 2, based mainly on papers [IV-5], [IV-6], [IV-7], [IV-8] and [IV-10] (at all of this papers the author of the thesis is first author) and presents different controller design techniques based on the delta model of the plant. The theoretical results were verified through simulation on low order plants - benchmark types - frequently applied for verification of control solutions.

Chapter 3, based mainly on paper [IV-21] based on delta representation of the systems, a new design method dedicated to a Mamdani type PI-FC-OI for benchmark-type plants model is proposed. The design is based on simplicity and transparency of the method and controller structure as well, having in view relatively simple implementations.

This attractive design method consists of three design steps. The method was derived by transferring the linear design results, applying the MO-methods in delta domain, to the strictly speaking fuzzy controller design in terms of the modal equivalence principle.

In conclusion the benefits of delta transform can be more interesting, if the suggested sampling period is smaller than the value used. In fact, this design ensures a good relation between discrete- and continuous-time models because relation (2.1-1) can be considered as the forward rectangular method.

The delta design technology can be also used with good results also on design of TS-FC, see for example paper [[IV-53], [IV-54].

Part V. Contributions. Possible further research direction

Part V. synthesizes the contributions of the thesis, the main conclusions and possible further research direction.

1. Contributions

A short overview of the contributions was presented in Part I paragraph 1.2. Based on the contributions presented on each Part of the thesis, a review of them is presented.

1.1. Contributions in Part I.

Based on the main task of the thesis, to develop new controllers, control structures and design methods for speed control application this part includes the following contributions:

1. In Chapter 2 there are synthesized as follows: the first application treated in the thesis, a speed control of an electrical driving system: the structure of plant, a simplified mathematical model oriented on control system design and the equivalence of driving systems with DC-m and BLDC-m is highlighted. Based on the NEDC test cycle a simplified test cycle is defined for such a driving system.
2. In Chapter 3 synthesizes basic aspects regarding speed control of a HG. Based on representative papers in the domain, mathematical models of the component subsystems are highlighted. The models are oriented on control system design.

The two syntheses are the basis for the applications from Part II, III and IV of the thesis.

1.2. Contributions in Part II.

The main contributions from Part II of the thesis were published in papers [I-6], [II-21], [II-22], [II-68], [II-95], [I-87] (2^{nd} PhD report) and are summarised as follows:

1. In **Chapter I** a short overview of the recent tendencies of the past 10 years in controller design using PI(D) controllers based on benchmark-type plant models; only those papers are included which are focused on ensuring good reference tracking and load disturbance rejection (chapter 1, paragraph 2.2).
2. **Chapter 2**: A short overview on optimal design methods based on Modulus Optimum criteria, detailing the MO method, SO method and ESO method (chapter 2, paragraph 2.2). Mainly the contributions of Kessler, C. Follinger,

O., Astrom, K.J., Voda and Landau, I-D., St. Preitl and Precup R.-E. are highlighted. For the mentioned methods their particularities and performances are presented. For some of the methods supplementary remarks are also given (chapter 2, paragraph 2.3).

3. In **Chapter 3** a novel controller design method based on a double parameterization of the optimality conditions specific for the SO method is introduced: the 2p-SO-method. The method (paragraph 3.1) refers to plants with very time constants and for which the application of the ESO-m requires approximations. The double parameterization is based on the followings:
   - First, with the condition that $T_\Sigma / T_1 << 1$, the parameter *m* is defined
   $$m = T_\Sigma / T_1$$
   - Second, the use of the optimization relations:
   $$\beta^{1/2} a_0 a_2 = a_1^2 \quad , \qquad \beta^{1/2} a_1 a_3 = a_2^2$$

   Through this an improvement of the phase margin can be reached. Design cases are presented and the specific tuning relations deduced. The achievable performances, the method efficiency are compared with the MO-method (much preferred design method for PID controllers for driving systems. Specific particular cases are presented and analyzed, accompanied by performance diagrams and methods for improving these. Simulation data allow a good specification for the cases when the method proves to be very efficient (paragraph 3.1.2, 3.1.3).
4. Taking into account that the robust design based on the Youla parameterization proves to be very efficient in many cases a Youla parameterization approach of the MO-m, ESO-m and 2p-SO-method is given.
5. **Chapter 4**. For the electric drive of a vehicle with electric traction (based on real data [II-85]) chapter 4 presents a detailed design for a cascade control system. Two variants of the cascade structure are presented, controller with homogenous and non-homogenous structure, using AWR measure. The simulation results reflect the expected behaviour.
6. Connected to this part of the thesis, **Appendix 1** present a 2-DOF approach for PI and PID controllers and a design method for 2-DOF controllers, which can be easy applied in practice.

Finally to underline the relevance of the research, the report with comparable results from September 2007, [II-66] must be mentioned.

1.3. Contributions in Part III

This part of the thesis is dedicated to new solutions and development methods for speed control of a HG conected to a PS.

1. **Chapter 2**: a short survey on Cascade control structures and actual design solutions in speed control for hydrogenerators.
2. **Chapter 3**: a new two-stage CCS with an internal minimax state controller [III-27], dedicated for rejecting internally located deterministic disturbances and a main GPC loop is introduced (papers [III-33], [III-83], the 2nd PhD report [I-87], the 3rd PhD report, [I-88]. The use of the GPC controller under IMC representation based on the GPC's polynomial RST structure has the

advantage of easy implementation. The control solution is applied to the speed control of HTGs. Digital simulation results of the case study show that the CS ensures good performances. The results validate this CS and its design method.

3. **Chapter 4** introduces a new FC development solution dedicated to the speed control of HG (paper [III-15] and [I-88] (the 3rd PhD report)) . Due to the accepted approximate equivalence between FCs and linear ones, the contribution involves a four inputs-two outputs TS-FC, developed by starting with the design of two sets of conventional PI controllers ensuring the desired maximum values of the sensitivity function and of the complementary sensitivity function. Simulation results for a case study prove the possible enhancement of CS performances.
4. **Appendix 2,** regarded to this part: the IMC equivalent of GPC structure has been derived. In [III-33] from an applicative point of view, also the problem of constraint handling of the control signal with different AWR measures with for quick leaving of the saturation zone is also dealt with for the GPC and IMC structures. For the RST controller a 2-DOF IMC structure was deduced, for which saturation handling of the control signal can be used in different ways, depending on the application.

Both proposed solutions can satisfy good performance needs, and are viable alternative for HTG speed control. Their introduction on real applications depends on the acceptance of the control system designers, for whom tradition and safety are of high priority.

1.4. Contributions in Part IV

1. **Chapter 2**: based mainly on papers [IV-5], [IV-6], [IV-7], [IV-8] and [IV-10] (at all of this papers the author of the thesis is first author) and presents different controller design techniques based on the delta model of the plant:
   - Optimization in delta domain using conventional (PID) controllers based on MO-method and 2p-SO-m; tuning relations in delta domain are deduced;
   - Pole cancellation method, design in the frequency domain,
   - DB control in different variants,
   - IMC-based Smith predictor for plants with dead time; A new approach to control system design based on the IMC in delta domain, with a mixed representation of the plant model within the IMC controller, [IV-7] (hybrid architecture). The method is based on the dual representation in delta and Z discrete domain of the plant, which has the advantage of the delta parameterization and the possibility of dealing easily with the dead time by introducing it in Z-domain.

   The theoretical results have been verified through simulation on low order plants – benchmark types – frequently applied for verification of control solutions.
2. **Chapter 3**, based mainly on paper [IV-21], dealing with delta representation of the systems, a new design method dedicated to a Mamdani type PI-fuzzy controller for benchmark-type plants model is proposed. The design is based on simplicity and transparency of the method and controller structure as well, having in view relatively simple

implementations.
This attractive design method consists of three design steps. The method has been derived by transferring the linear design results, applying the MO-methods in delta domain, to the strictly speaking fuzzy controller design in terms of the modal equivalence principle. In fact, this design ensures a good relation between discrete- and continuous-time models.

3. The **Appendix 3** based on papers [IV-28], [IV-22], [IV-35] and [IV-36] (synthesized also in [IV-45] - [IV-47]) and is focused on elaborate a method for the construction and development of 2-DOF fuzzy controllers. The presented development method for of 2-DOF FC's is easy to understand and to implement as CAD development; it is based on starting with the development of the 2 DOF controller followed by the transfer to the fuzzy processing of the components with dynamics. The integral element specific to the 2 DOF controllers is included in the forward channel of the control loop.

2. Possible further research direction

The approach control topics and given solutions can support further research topics. Some of them would be:

- Analytically deduced procedure for tuning 2-DOF PI and PID controllers considering the control-loop robustness based on sensitivity function (see Part II and Part IV) considering dead time;
- Develop new "auto-calibration" method for PI and PID controllers based on ESO-m and 2p-SO-m (see Part II);
- Implement and perform real experiments of CSs and development methods on a electric vechicle (see Part II);
- Developing Combined Control strategies in speed control of HG and try to implement them on real applications (see part III and IV);
- Handle real applications in constraint cases.

The relevance of the presented design methods and according results is sustained by different papers published during the past years [II-66].

Appendices

Appendix 1. **Equivalence between 1-DOF (PID) and 2-DOF controllers**

Appendix 2. **RST Polynomial Structures of the Generalized Predictive Control**

Appendix 3. **Two Degree of Freedom Fuzzy Controllers (2-DOF-FC). Structure and Design**

Appendix 1. Equivalence Between Conventional 1-DOF (PID) and 2-DOF Controllers

The Appendix is based on research results presented mainly in paper [II-70] and the results were used in papers, [IV-22], [IV-28], [IV-43], [II-82].

1. Basic aspects regarding 2-DOF controllers

The 2-DOF structure two distinct controllers [II-3] - [II-11], figure A.1.1-1.

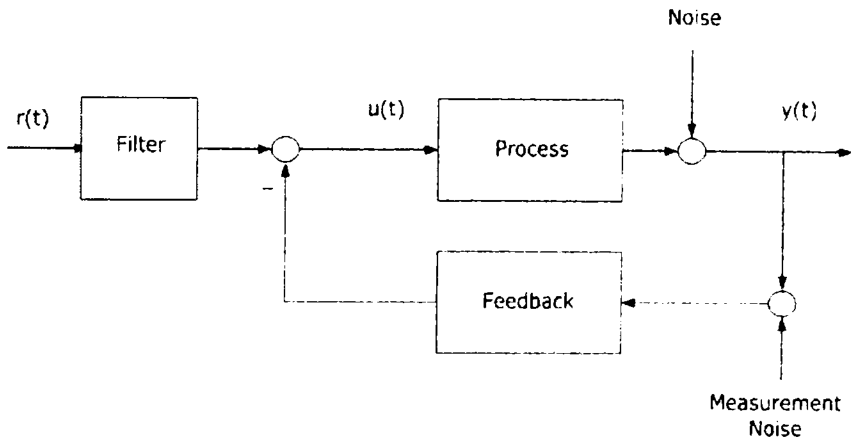


Fig. A.1.1-1. Structure of the 2-DOF controller and control structure

The requirements to be fulfilled by the control loop are:

- the servo performance (zero control error);
- the disturbance rejection;
- robustness.

The traditional 1-DOF structures satisfy these needs equally, because a desired trade-off must be set for each of these properties In case of 2-DOF controllers the attributes enlisted above can be separately adjusted without influencing one another [II-3], [II-10].

For the sake of simplicity it is assumed that the systems is a MISO once but the results can be generalized seamlessly for MIMO systems. Suppose we have the pulse t.f. of the process from the continuous model in the following form:

$$P(z) = (1 - z^{-1})Z\{\frac{P(s)}{s}\} = \frac{B(z)}{A(z)} \qquad \text{(A1-1-1)}$$

Then the block diagram of the system can be seen in Figure A.1.1-2.

The servo performances are given by a reference model $P_m(z) = H_m(z)$ in the form of (A1-1-2) with a condition for zero control error in form of (A1-1-3):

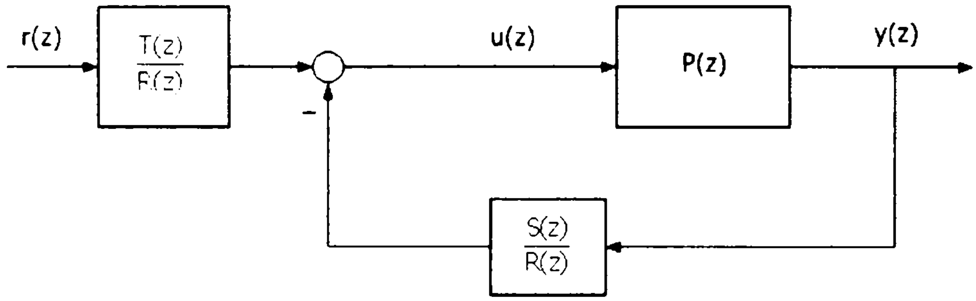


Fig. A.1.1-2. 2-DOF Controller structure (2nd variant). Design level block diagram

$$P_m(z) = H_m(z) = \frac{B_m(z)}{A_m(z)} \quad \text{(A1-1-2)} \qquad \frac{B_m(1)}{A_m(1)} = 1 \quad \text{(A1-1-3)}$$

Based on (A1-1-2) and the pulse t.f. of the closed loop, the main equation to be solved is:

$$\frac{B(z)T(z)}{A(z)R(z)+B(z)S(z)} = \frac{B_m(z)}{A_m(z)} \quad \text{(A1-1-4)}$$

where $T(z)$, $R(z)$, $S(z)$ are unknown polynomials. It is supposed that $B_m(z)$ and $A_m(z)$ do not have common factors. The $A_m(z)$ polynomial determines the poles of the closed loop and a polynomial computation will be performed to set the poles.

2. Design of 2-DOF controllers. Solving the main equation

Equation (A.-1-4) is transcribed in the following form:

$$\frac{B(z)T(z)}{A(z)R(z)+B(z)S(z)} = \frac{B_m(z)}{A_m(z)}\frac{A_o(z)}{A_o(z)}, \quad \text{(A1-2-1)}$$

$A_0(z)$ is an arbitrary polynomial (the observer polynomial) multiplied by the common factors of the model. For causality reasons there are some conditions to be fulfilled:

$$\partial S \le \partial R \qquad \partial T \le \partial R \quad \text{(A1-2-2)}$$

To avoid the inter-sampling oscillations of the continuous system $B(z)$ is decomposed into cancelling zeros ($^+$) and non-cancelling zeros ($^-$)as follows:

$$B(z) = B^+(z)B^-(z) \quad \text{(A1-2-3)} \qquad B_m(z) = B^-(z)B'_m(z) \quad \text{(A1-2-4)}$$

where $B^-(z)$contains the zeros that are zeros of the closed system (and hence zeros of the model) as well. Then it can be simplified the left hand side of (A1-2-1) with $B^+(z)$, because it is not in the numerator of the model, hence $R(z)$ can be factorized:

$$R(z) = B^+(z)R'(z) \quad \text{(A1-2-5)}, \qquad R'(z) = (z-1)^l R_1(z) \quad \text{(A1-2-6)}$$

where $R'(z)$will contain the integrators (l denotes the number of the integrators). After accomplishing the operations elaborated above the equation (A1-2-1) results:

$$\frac{T(z)}{A(z)R'(z)+B\ (z)S(z)}=\frac{B'_m(z)}{A_m(z)}\frac{A_o(z)}{A_o(z)}, \qquad \text{(A1-2-7)}$$

which can be separated into two equations:

$T(z)=B'_m(z)A_o(z)$ (A1-2-8) $A(z)R'(z)+B\ (z)S(z)=A_m(z)A_o(z)$ (A1-2-9)

There is a Diophantine equation over the ring of polynomials, where the $T(z), R'(z), S(z)$ polynomials are unknown. To make the solution easier some degree conditions for the polynomials are considered:

Causality (A1-2-2):	$\partial S \le \partial R$ $\partial T \le \partial R$	(A1-2-10)
From (A1-2-8):	$\partial T=\partial B'_m+\partial A_0$	(A1-2-11)
From (A1-2-9):	$\partial R'=\partial A_m+\partial A_0-\partial A$	(A1-2-12)
Unique solution (A1-2-9) :	$\partial S=\partial A+I$	(A1-2-13)

Finally the steps of the control design are summarized:

(1) Choose the number of integrators (denoted by I)
(2) Specify the degree of $R'(z)$ and then the degree of $A_0(z)$.
(3) Compute degree of $T(z)$ based on (A1-2-11).
(4) Regarding (A1-2-13) we choose $\partial S=\partial A+I-1$, because we need $S(z)$ to set the left hand side coefficients in (A1-2.9) for those required on the right hand side.
(5) Solve the Diophantine equation for found the unknown polynomials.

The CS's block diagram is presented in Figure A.1.2-1, with the integrators within the loop.

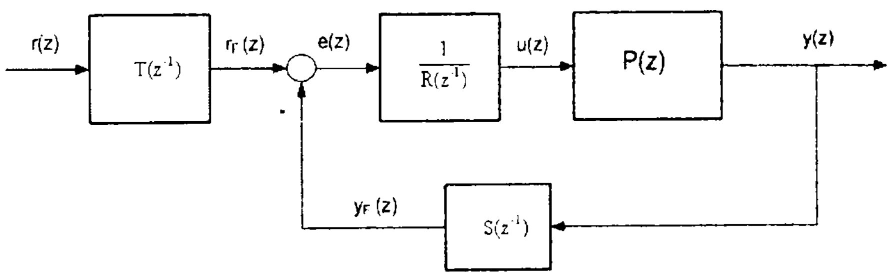


Fig. A.1.2-1 Implementation of the 2-DOF controller

A CAD program in MATLAB is presented as application in [II-71], [IV-35].

3. Equivalence between 1-DOF (PID) and 2-DOF controllers

Let consider the block diagram from fig.A.1.3-1. Replacing the feedback controller $C_S(z)$ on the input channel and the forward loop, the given CS can be transposed into the structure from figure A.1.4-1 [II-70], [II-102]. Corresponding:

$$C(z) = C_S(z) \quad \text{and} \quad F(z) = C_S(z)C_T(z) \qquad \text{(A1-3-1)}$$

For reference signal tracking design relation (A1-1-4) is valid. Consequently 2-DOF controller design can be performed as described in chapter 2, or according to conventional controller design methods.

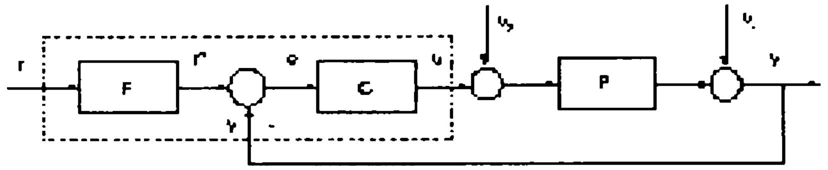


Fig.A.1.3-1 Restructured control system with a conventional controller (C) and a reference filter (F)

Also for v_2 and v_1 (for disturbances, the notation v instead of d is used) type disturbance rejection the two structures have identical behavior (R(z), S(z), T(z) are the 2-DOF polynomials):

$$H_{v2}(z) = \frac{R(z)B(z)}{R(z)A(z) + S(z)B(z)} \quad , \quad H_{v1}(z) = \frac{R(z)A(z)}{R(z)A(z) + S(z)B(z)} \qquad \text{(A1-3-2)}$$

The 2-DOF controller can be restructured in other ways as well, where the presence of a conventional controller (PI or PID) can be highlighted [II-70]. Two types of structure are detailed in figure A.1.3-2. These rearrangements allow:

- To take over design experience from case of PI and PID controllers;
- Introduction of supplementary blocks specific to PI and PID controllers (AWR circuit, bump-less switching a.o.);
- Transformation of PI and PID controllers into 2-DOF structures and vice versa.

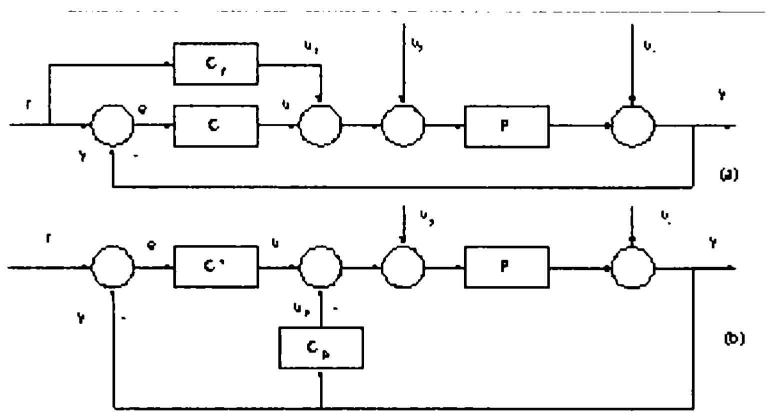


Fig.A.1.3-2. Two alternatives for rearranging a 2DF controller

Following, the controllers from fig.A.1.3-1 will be characterized by continuous t.f. with the "traditional" tuning parameters $\{k_R, T_i, T_d, T_f\}$. Discretizing, the numerical control algorithm is obtained.

Taking the basic controller C of PID type, it can be written (adapted notations for the t.f.s):

- For the structure from fig.A.1.3-1:

$$C(s) = \frac{u(s)}{e(s)} = k_R\left(1 + \frac{1}{sT_i} + \frac{sT_d}{1 + sT_f}\right),$$

$$F(s) = \frac{r^{*}(s)}{r(s)} = \frac{1 + (1 \quad \alpha)\, T_i s + \dfrac{(1 \quad \beta) T_i T_d s^2}{(1 + sT_f)}}{1 + T_i s + \dfrac{T_i T_d s^2}{(1 + sT_f)}} \qquad \text{(A1-3-3)}$$

- For structure (a) from fig.A.1.3-2:

$$C(s) = \frac{u(s)}{e(s)} = k_R\left(1 + \frac{1}{sT_i} + \frac{sT_d}{1 + sT_f}\right),$$

$$C_F(s) = \frac{u_f(s)}{r(s)} = k_R\left(\alpha + \beta \frac{sT_d}{1 + sT_f}\right) \qquad \text{(A1-3-4)}$$

- For structure (b) from fig.A.1.3-2 (with notation $C(s)=C^{*}(s)$):

$$C\ (s) = \frac{u(s)}{e(s)} = k_R\left[(1 \quad \alpha) + \frac{1}{sT_i} + (1 \quad \beta)\frac{sT_d}{1 + sT_f}\right],$$

$$C_P(s) = \frac{u_f(s)}{r(s)} = k_R\left(\alpha + \beta \frac{sT_d}{1 + sT_f}\right) \qquad \text{(A1-3-5)}$$

Depending on the values of α and β parameters, for the presented blocks the behaviors from in Table A.1-3-1 are obtained. The choice of a certain representation of the controller depends on (see for example [II-70], [II-71], [II-102]):

- the structure of the available controller;
- the adopted algorithmic design method and the result of this design.

Within this paragraph these aspects are not analyzed in detail. The comparisons between 2-DOF CS and 1DOF CS which was performed in chapter 3, are based on the structures from fig.A.1.3-1 having $\alpha = 0$ and $\beta = 0$.

Table A.1-3-1 Conections between 2DOF controller and extended 1DOF controller structure

Fig.A.1.4-1		F(s)	-	F(s)C(s)	C(s)	Remarks
Fig.A.1.4-2-a		-	C_F	$C(s)-C_F(s)$	C(s)	-
Fig.A.1.4-2-b		-	C_P	$C^*(s)$	$C^*(s)+C_P(s)$	-
α	β	-	-	(ref. channel)	(feedback)	-
0	0	1	0	PID	PID	1DOF
0	1	PDL2	DL1	PI	PID	1DOF with non-homogenous behavior
1	0	PD2L2	P	PID-L1	PID	
1	1	PL2	PDL2	I	PID	
α	β	PID controller with pre-filtering (2DF controller)				

P – proportional, D – derivative, I – integral, L1(2) – first (second) order lag filter

4. Conclusions and collateral research results

Based on research results detailed in [II-70], [II-71], [IV-35] the Appendix presents an easy to understand methodology for develop 2-DOF controllers. In [II-70], [II-71] the presented methodology was implemented using the facilities given by the MATLAB&SIMULINK and a case study compares the control system performances for the 2-DOF solution with the control system performances for a PID control solution. It was underlined that firstly, the development methodology is very useful, concretized in reduction of development effort, and secondly, the 2-DOF ensure better system performances regarding simultaneously the reference input and to the disturbance. Connections between 2-DOF and conventional controllers are presented: the relations are given in practically useful forms.

The presented design method was extended and then applied to the development of 2 DOF fuzzy controllers. Their design is based on the modal-equivalence principle [IV-22]. The targeted applications were servosystems and algorithms for Trajectory Tracking (mobile robots) [II-22].The presented solutions were verified through simulations.

Appendix 2. Polynomial Structures of Generalized Predictive Control

1. Basic relations.2-DOF (RST) polynomial structure

The GPC algorithm ([III-31]) can be converted in all cases when no constraints appear into a polynomial two-degree-of-freedom structure, the so-called RST structure, figure A.2.1-1. To obtain it, the original GPC algorithm will be considered [III-33], [III-41].

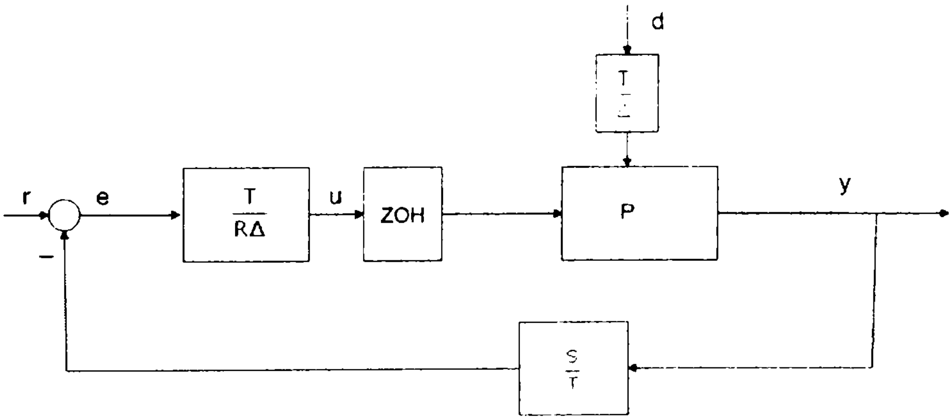


Fig.A.2.1-1. Two-degree-of-freedom (RST) polynomial control structure

The model of the plant is supposed to be of CARIMA type:

$$A(q^{-1})y(t) = z^{-d}B(q^{-1})u(t-1) + C(q^{-1})\frac{e(t)}{\Delta} \qquad \text{(A2-1-1)}$$

where the polynomials are:

$$A(q^{-1}) = 1 + a_1q^{-1} + \ldots + a_{na}q^{-na} \qquad \text{(A2-1-2)}$$

$$B(q^{-1}) = b_0 + b_1q^{-1} + \ldots + b_{nb}q^{-nb} \qquad \text{(A2-1-3)}$$

$$C(q^{-1}) = 1 + c_1q^{-1} + \ldots + c_{nc}q^{-nc} \qquad \text{(A2-1-4)}$$

and

$$\Delta = 1 - q^{-1} \qquad \text{(A2-1-5)}$$

The C(q) polynomial is first chosen for 1, for simplicity [III-31], u(t) is the control sequence and y(t) the output sequence, e(t) is a zero mean white noise, d is the physical dead time. A cost function is defined as follows:

$$J = \sum_{j=N_1}^{N_2} \delta(j)[\hat{y}(t+j \mid t) - r(t+j)]^2 + \sum_{j=1}^{N_u} \lambda_u(j)[\Delta u(t+j-1)]^2 \qquad \text{(A2-1-6)}$$

where N_1 and N_2 are the limits of the prediction horizon, N_u is the control horizon, $\hat{y}(t+j\mid t)$ is the j-step ahead prediction of the output, r(t+j) is the future reference trajectory and $\delta(j)$ and $\lambda(j)$ are weighting sequences.

By minimizing the cost function, an optimal value for the future control sequence is obtained. If only the first element of the control signal sequence is sent to the process (receding horizon strategy), after the minimization the control law is obtained:

$$\Delta u(t) = K(r(t) - f(t)) = \sum_{i=N_1}^{N_2} k_i[r(t+i) - f(t+i)] \tag{A2-1-7}$$

where K is the first row of the matrix $(G^T G+\lambda I)^{-1}G^T$, f is the free response, r is the reference signal, see [III-31].

Further, as mentioned before, if there are no constraints, the GPC algorithm can be transformed into a two-degree-of-freedom (RST) polynomial structure, figure A.2.1-1. Accordingly, the expression of the control signal is:

$$R(q^{-1})\Delta u(t) = T(q^{-1})r(t) - S(q^{-1})y(t) \tag{A2-1-8}$$

where *R, S, T* are polynomials in the backward shift operator. Having a plant model as (A2-1-1), in many cases polynomial $C(q^{-1})$ can not be identified, therefore it is substituted with a T polynomial which can be considered as a pre-filter [III-30].
If the plant model is given by:

$$A(q^{-1})y(t) = q^{-d}B(q^{-1})u(t-1) + T(q^{-1})\frac{e(t)}{\Delta} \tag{A2-1-9}$$

to determine the controller, a Diophantine equation has to be solved [III-93]:

$$T(q^{-1}) = E_j(q^{-1})\Delta A(q^{-1}) + q^{-j}F_j(q^{-1}) \tag{A2-1-10}$$

It plays a role in disturbance rejection, and also it is mentioned [III-31] that it can influence robust stability.

Solving the Diophantine equation and choosing $T(q^{-1})=1$ for simplicity, the final expressions of the R, S and T polynomials will be [III-31]:

$$R(q^{-1}) = \frac{T(q^{-1}) + q^{-1}\sum_{i=N_1}^{N_2} k_i I_i}{\sum_{i=N_1}^{N_2} k_i} \tag{A2-1-11}$$

$$S(q^{-1}) = \frac{\sum_{i=N_1}^{N_2} k_i F_i}{\sum_{i=N_1}^{N_2} k_i}, \qquad T(q^{-1}) = 1 \tag{A2-1-12}$$

where the polynomial $T(q^{-1})$ is a free parameter, chosen according to design requirements, F_i and I_i are elements of the Diophantine equations (I_j are the rows of vector G^T).

The RST polynomial structure can be transformed into a two-degree-of-freedom Internal Model Control (2DOF-IMC) structure, as seen in figure A.2.1-2 (a) or (b) [III-33], [III-42]. Deducing step by step, the following results are obtained:

$$C(q^{-1}) = \frac{S(q^{-1})A(q^{-1})}{F_w(q^{-1})(R(q^{-1})\Delta A(q^{-1}) + S(q^{-1})B(q^{-1})q^{-d})} \tag{A2-1-13}$$

$$\frac{T(q^{-1})}{S(q^{-1})} = \frac{F_r(q^{-1})}{F_w(q^{-1})} \tag{A2-1-14}$$

In equation (A2-1-14) the two members on both sides must be proportional. In the examples presented in [III-42], the two have been taken for equal, where F_r

insures servo performance and F_w disturbance rejection. The IMC structure in form of Fig.A.2.1-2 is valid for stable plants only. In case of unstable plants the Youla parameterization is used [III-92]. The equivalences between the two structures are:

$$F_r = T \qquad \text{(A2-1-15)}$$

$$F_w = S \qquad \text{(A2-1-16)}$$

$$C_{IMC} = \frac{A}{R\,\Delta\,A + B\,S\,z^{-d}} \qquad \text{(A2-1-17)}$$

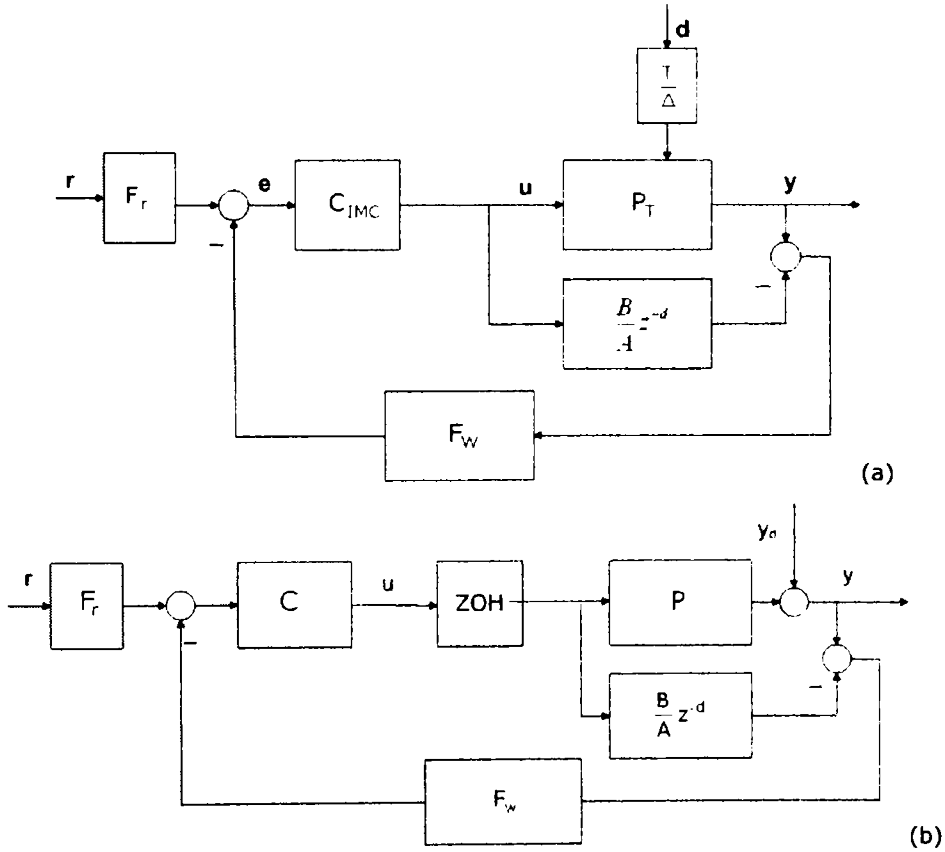


Fig.A.2.1-2. IMC structure of GPC

2. Constraint handling in case of RST and IMC structures

Since the IMC structure is derived from the RST structure, it is valid for unconstrained case. Consequently, some form of constraint handling of the control signal u(t) is required. In an IMC structure constraints of the control signal can be treated in various ways.

Constraints of the control signal can be incorporated into the IMC structure, for example see the structure presented in [III-33], [III-83], figure A.2.2-1. The advantage of this structure is that the same control signal is applied both to the plant and to the internal model; C itself does not contain integrating effect, which is introduced through the IMC feedback.

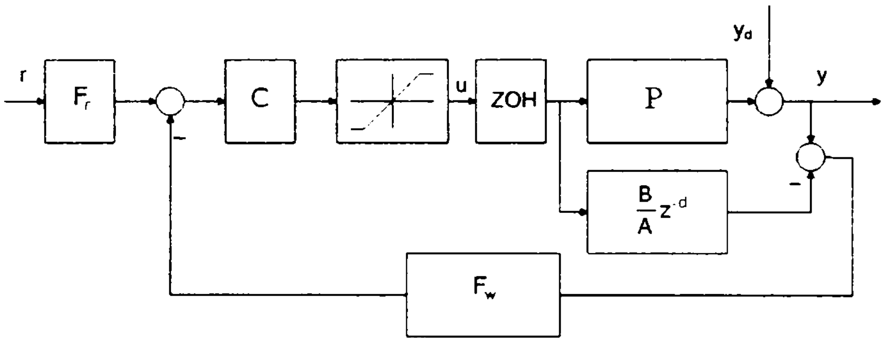


Fig.A.2.2-1. IMC structure with limitation inside the mode

In addition, the anti-windup property of this structure can be improved by realizing the IMC controller in a feedback of the saturation, for example as in figure A.2.2-2 [III-94]). Its advantage is that it takes into account also the dynamics of the controller. The limited input of the plant provides the input for the controller in the feedback.

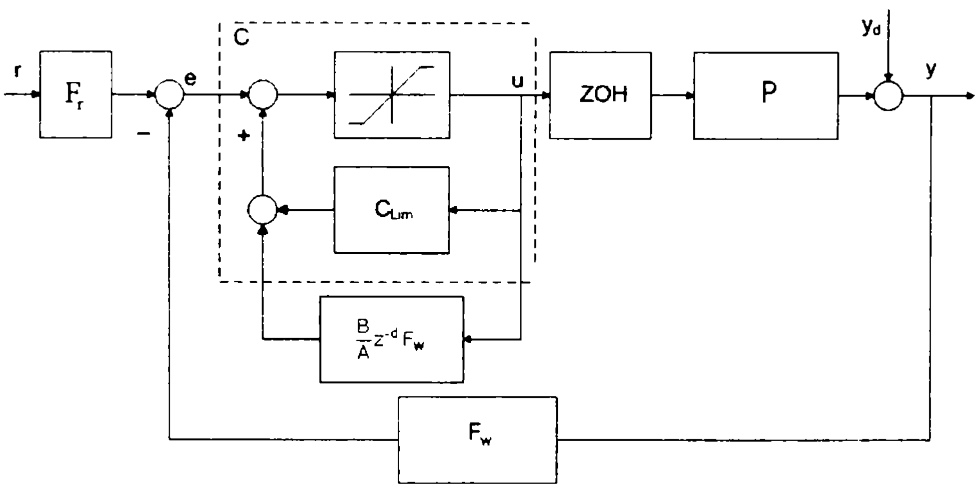


Fig.A.2.2-2. IMC structure with controller in the feedback of the saturation

For a given controller C, the C_{Lim} controller that feeds back the saturating element can be calculated according to the following relation (A2-2-1):

$$C_{Lim}(q^{-1}) = \frac{C(q^{-1}) - 1}{C(q^{-1})} \qquad \text{(A2-2-1)}$$

Special attention must be paid at the implementation of this control structure. An algebraic loop will appear in this case, so measures avoiding it have to be dealt with. One way to handle this is to separate the constant component of the partial fractional representation of C_{Lim} and to incorporate it in the slope of the saturation.

It has been shown [III-33] that handling this way the saturation has advantages over other methods. In [III-40] the effects of saturation are presented according to the schemes from Fig.A.2.2-1 and A.2.2-2, for different benchmark type plants using Simulink simulation. It can be noticed that the output signal in case of feedback saturation proves to have the closest behavior to the unconstrained case. As far as the control signal is concerned, also the same results can be stated.

3. Influence of predictive parameters on the closed loop poles

Predictive parameters in part III paragraph 3.4 relation (3.4-2)

$$J = \sum_{j=N_1}^{N_2} \delta(j)[\hat{y}(t+j\mid t) - r(t+j)]^2 + \sum_{j=1}^{N_u} \lambda_u(j)[\Delta u(t+j-1)]^2 \qquad (3.4\text{-}2)$$

(N_1 , N_2 - the limits of the prediction horizon, N_u - the control horizon, $\delta(j)$ and $\lambda(j)$ - weighting sequences) are chosen according to the desired behaviors; in the literature there exist some guidelines for their choice.

Computing the closed loop t.f. related to RST parameters, it results:

$$H_r(z) = \frac{y(z)}{r(z)} = \frac{T(z)B(z)z^{-d}}{R(z)\Delta A(z) + B(z)S(z)z^{-d}} \qquad \Delta = 1 - z^{-1} \qquad (A2\text{-}3\text{-}1)$$

Also, controller $C_{IMC}(z)$ has the same poles as $H_r(z)$, as it is expected:

$$C_{IMC}(z) = \frac{A(z)}{R(z)\Delta A(z) + B(z)S(z)z^{-d}} \qquad (A2\text{-}3\text{-}2)$$

For different values of the predictive parameters (N_1, N_2, N_u, λ_u) and of the plant model, the poles of $H_r(z)$ are mapped for a second order plant without dead time, with the continuous t.f. and the pulse t.f (h=0.2, ZOH is included):

$$H_P(s) = \frac{1}{(1+0.67s)(1+0.33s)}$$

$$H_P(z) = \frac{0.0674\, z^{-1} + 0.0499\, z^{-2}}{1 - 1.2874\, z^{-1} + 0.4047\, z^{-2}} \qquad (A2\text{-}3\text{-}3)$$

and the T(z) pre-filter was T(z)=1.

In order to explore the possibilities given by simulations, the mappings have been tested through simulation. A study upon the following parameter changes was performed:

- ❑ Changing the tuning parameter λ_u . Let the tuning parameters be:

$$N_1 = 1,\ N_2 = 25,\ N_u = 1,\ d = 0, \quad \lambda_y = 1$$

and let the range of variation of λ_u be:

$$\lambda_u = (0 : 0.01 : 1) \qquad (A2\text{-}3\text{-}4)$$

In this case for this plant the poles of the closed loop system (which are also equal to the poles of the IMC controller C) will vary as shown in Fig.A.2.3-1. The evolution of the poles (if they are changing) is represented by an arrow. It can be observed that one of the poles changing slightly, but not affecting the systems' stability. It must be mentioned that by choosing other tuning parameters the poles are mapping completely differently

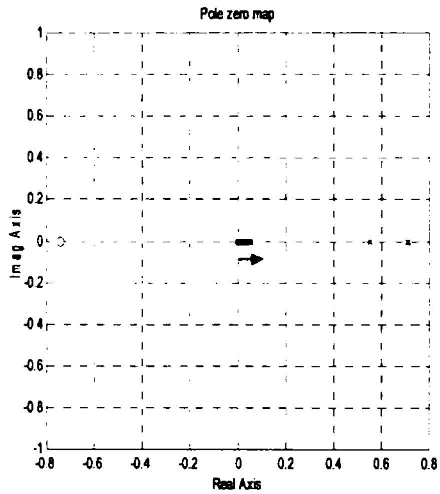


Fig.A.2.3-1. Mapping of closed loop poles for changing λ_u parameter

- Changing the tuning parameter N_1. For this case, let the GPC tuning parameters be:

$$N_2 = 25, N_u = 1, d = 0, \lambda_u = 0.1, \lambda_y = 1$$

and let the range of variation of N_1 be:

$$N_1 = (1:1:25) \qquad \text{(A2-3-5)}$$

The poles of the closed loop system will vary as shown in Fig.A2.3-2 In this case also there is no significant change in the poles' value, not threatening the stability of the system.

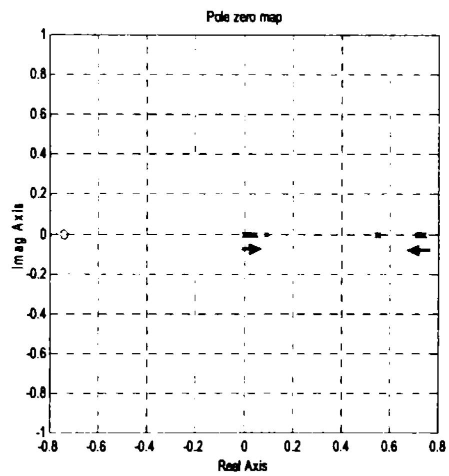


Fig.A.2.3-2. Mapping of closed loop poles for changing N_1 parameter

- Changing the tuning parameter N_2 The tuning parameters are chosen for this example to be:

$$N_1 = 1, N_u = 1, d = 0, \lambda_u = 0.1, \lambda_y = 1$$

and let the range of variation of N_2 be:

$$N_2 = (1:1:25) \qquad \text{(A2-3-6)}$$

In this case for this plant the resulting four poles of the closed loop system (which are also equal to the poles of the IMC controller C) are varying as shown in Fig.A.2.3-2.

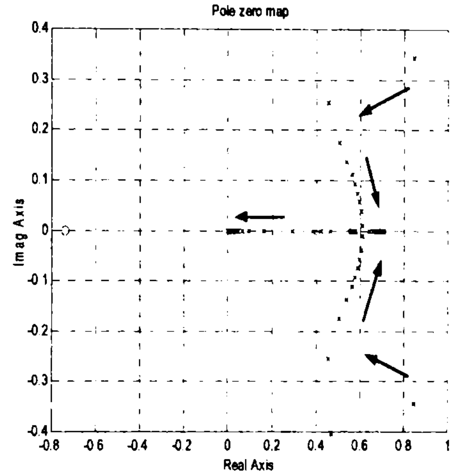


Fig.A.2.3-3. Mapping of closed loop poles for changing N_2 parameter

This case is more interesting, since the choice of parameter N_2 influences much more the control quality. The smaller N_2, the more oscillating the closed loop is, and vice-versa, the larger N_2 is chosen the more damped the system is.

❑Changing tuning parameter N_u . The tuning parameters are chosen for this example to be:

$$N_1 = 1, N_2 = 25, d = 0, \lambda_u = 0.1, \lambda_y = 1$$

and let the range of variation of N_u be:

$$N_u = (1:1:20) \qquad \text{(A2-3-7)}$$

In this case for this plant the poles of the closed loop system are varying as shown in Fig.A.2.3-4.

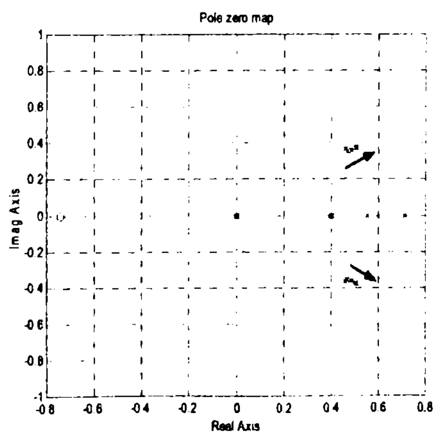


Fig.A.2.3-4. Mapping of closed loop poles for changing N_2 parameter

For this simulation it can be mentioned that for $N_u=1$ all poles are real, but for $N_u>1$ two of them become complex-conjugate.

- Changing the filter polynomial T(z). The T(z) polynomial appears in the noise model that is used for obtaining the predictive controller, it also appears both explicitly and implicitly in the RST structure. The T(z) polynomial (fixed pre-filter of observer polynomial) is considered to improve robustness of the system, if chosen properly [III-31], [III-98]. Detailed results are given also in [III-33], [III-42].
  It can be noticed that in this case the simpler noise filter provides more robust performance.

Further analysis can be performed to show the effect of the noise filter to robustness properties of the control algorithm.

4. Conclusions

In this appendix the IMC equivalent of GPC structure was derived. In [III-33] and [III-42], from an applicative point of view, also the problem of constraint handling of the control signal with different anti-windup measures with for quick leaving of saturation zone is also dealt with for the GPC and IMC structures.

For the RST controller a 2-DOF IMC structure was deduced, for which saturation handling of the control signal can be used in different ways, depending on the application.

In [III-33] and [III-42] the influence of the GPC tuning parameters in case of IMC implementation is presented for second order lag system, exemplified through simulation.

Appendix 3. Two Degree of Freedom Fuzzy Controllers (2-DOF-FC). Structure and Design

The appendix is based on papers [IV-32] and [IV-35] (synthesised also in [IV-45] – [IV-47]).

1. Structure of 2-DOF-FC-s and design

The design methods for classical FCs with dynamics can be successfully extended to the design of 2-DOF-FC-s. In the design phase, the FC can be considered as nonlinear but linearizable near operating points belonging to the control surface [IV-25].

The structure of a CS with 2-DOF controller was presented in part II, Appendix 1. Based on these results there are presented the structure and development principles of 2-DOF FC-s. The developed method is based on the principle of equivalence and design techniques resulting from the linear case [IV-35], [IV-36], [IV-48].

The general structure of a 2-DOF FC is presented in Fig.A.3.1-1, where FC-T is a fuzzy module for the controller T (on the reference channel), and FC-S is a fuzzy module for the controller S (on the feedback channel).

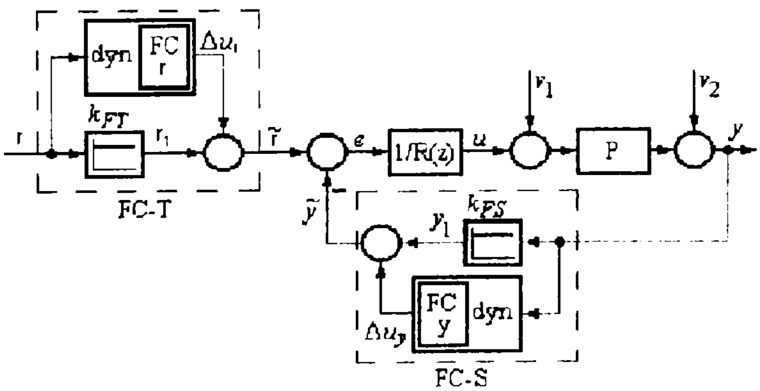


Fig. A.3.1-1. General structure of a 2-DOF fuzzy controller

The integral component brought by the controller is included in the forward channel of the loop and highlighted as follows:

$$R(z) = (z - 1)R'(z) \qquad \text{in the discrete case}$$

$$R(s) = sR'(s) \qquad \text{in the continuous case.} \quad \text{(A3-1-1)}$$

In the case of relatively low order continuous plants, in the construction of the 2-DOF fuzzy controllers it is possible to take over the design experience specific to the 1-DOF controllers and the extensions with filters on the input channels.

The adoption of decentralized feed-forward compensation allows the reduction of the reference tracking error. At the individual joint channel level, the output y(t) must follow the reference signal r(t). The presence of load disturbances due to interactions with other joints is undesirable.

In this context, it becomes possible to use the low order informational modules specific to the fuzzy controllers with dynamics with the detailed computation presented in paragraph 3.2. So, a 2-DOF-FC can be developed on the basis of two versions:

- by starting with the equivalence between a 2-DOF controller and the conventional 1-DOF controllers extended with filters on the input channels;
- by starting with the discrete model of the 2-DOF controller.

In the presence of the integral component, in steady state, the conditions (A3-1-2) are fulfilled:

$$e_\infty = \tilde{r}_\infty - \tilde{y}_\infty = 0, \quad u_\infty = \text{const.}, \qquad \text{(A3-1-2)}$$

and the assurance of a desired value of the output y_∞ depends on obtaining an equilibrium of the two variables:

$$\tilde{r} = k_{FT} r + \Delta u_r, \quad \tilde{y} = k_{FS} y + \Delta u_y \quad , \qquad \text{(A3-1-3)}$$

where k_{FT} and k_{FR} are parameters that adjust the level of the two signals.

The signals Δu_r and Δu_y represent only the dynamic components processed by the FC-s with dynamics FC-r and FC-y. Indeed, from the steady state condition (A3-1-2) it is obtained (A3-1-4):

$$K_{FT\infty} r_\infty = K_{FS\infty} y_\infty \quad , \qquad \text{(A3-1-4)}$$

and it results the necessity for the two components, Δu_r and Δu_y, to have only "transient character":

$$\Delta u_r(t) \to \infty \quad , \quad \Delta u_y(t) \to \infty \qquad \text{for} \qquad t \to \infty \quad . \qquad \text{(A3-1-5)}$$

It is important to outline that generally, the continuous components $k_{FT}\, r_\infty$ (on the reference channel) and $k_{FS}\, y_\infty$ (on the feedback channel) are not allowed to be subject of the fuzzy processing. The modification of k_{FS} would affect the stationary trim-point. If, on the other hand, it were desired that the structure ensures feedback linearization, then k_{FS} could stand for objective of fuzzification.

However, there can be conceivable situations (for example, the case of some reference tracking systems) in which the fuzzy processing can be included in the controller FC-T. For such situations a variable reference input r(t) can be subject to fuzzification, and $\tilde{r}(t)$ will contain corrections as function of the variations of r(t) and of other possible causes.

The two-input-single-output (TISO) nonlinear fuzzy-blocks FC-r an FC-y (without dynamics) includes among its nonlinearities the scaling of inputs and output as part of its fuzzification module.

The structure of the fuzzy modules with dynamic processing, FC-T or FC-S (as part of the T and S controller) can be implemented in the two versions (in Figure A.3.1-1 with respect to the reference signal r): - an analogue version (Fig.A.3.1-2 (a)) and, - a discretized quasi-continuous digital version (Fig.A.3.1-2 (b)).

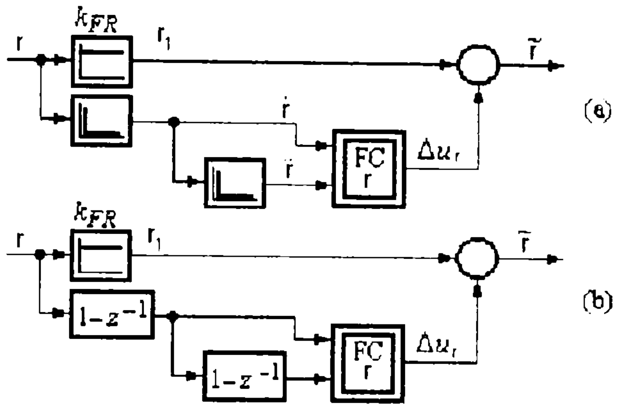


Fig. A.3.1-2. Structure of the module FC-r (T or S): (a) analogue version (b) digital version

The mainly used digital version is based on the computation of the derivatives (as increments) of the variables:

$$\Delta r_k = r_k - r_{k-1}, \quad \Delta^2 r_k = r_k - 2r_{k-1} + r_{k-2} \quad , \qquad \text{(A3-1-6)}$$

the increment of $\Delta u_{r,k}$, depends on Δr_k and $\Delta^2 r_k$:

$$\Delta u_{r,k} = k_1 \Delta r_k + k_2 \Delta^2 r_k = k_1(\Delta r_k + \alpha \Delta^2 r_k). \qquad \text{(A3-1-7)}$$

The parameters k_1, k_2 and α are functions of the parameters of the conventional controller T(s) or S(s) and of the sampling period [IV-25], [IV-26].
Based on equation (A3-1-7) and on representation of the increment $\Delta u_{r,k}$ in the phase plane $<\Delta r_k, \Delta^2 r_k>$, the pseudo-fuzzy features of the algorithm (A3-1-6), (A3-1-7) can be expressed as described [IV-32], [IV-33], fig. A3.1-3 (a):

- there exists a "zero control signal line" $\Delta u_{r,k} = 0$, having the equation:

$$\Delta r_k + \alpha \Delta^2 r_k = 0 \quad ; \qquad \text{(A3-1-8)}$$

- with regard to this line it is obtained that in the upper half-plane: $\Delta u_{r,k} > 0$, and in the lower half-plane: $\Delta u_{r,k} < 0$;
- the distance from any point of the phase plane to the "zero control signal line" corresponds to $|\Delta u_{r,k}|$.

For the fuzzy block FC-r, the fuzzification can be solved as follows:

- for the input variables Δr_k and $\Delta^2 r_k$, an odd number of linguistic terms with regularly distributed triangular type membership functions having an overlap of 1 are chosen;
- for the output variable $\Delta u_{r,k}$, more linguistic terms with regularly distributed singleton type membership functions are chosen, Fig. A3.1-3 (b).

Other shapes and distribution of membership functions can contribute to CS performance enhancement. The strictly positive parameters of the 2-DOF fuzzy controller, $\{B_r, B_{\Delta r}, B_{\Delta u}\}$, are in direct correlation with the shapes of the membership functions of the linguistic terms corresponding to the input and output linguistic variables.

The inference engine of the block FC-r employs the Mamdani's MAX-MIN compositional rule of inference assisted by a complete rule base. The rule base of the block FC-r can be expressed as a symmetrical decision table, illustrated in Table A.3.1-1

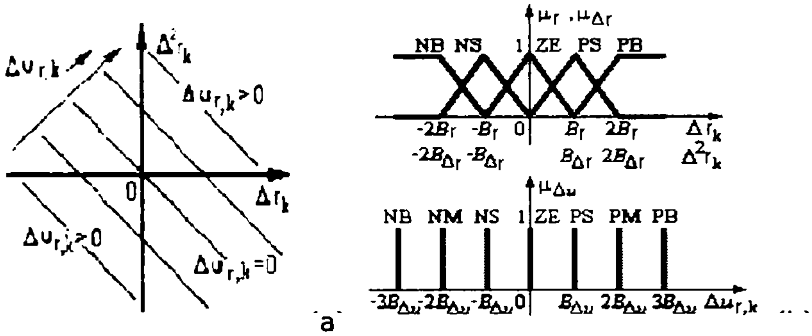


Fig. A.3.1-3. (a) Phase plane representation corresponding to (2.21) and (b) Shapes of the membership functions for the block FC-r

Table A3.1-1 Decision table of the block FC-r

$\Delta^2 r_k \backslash \Delta r_k$	**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

The defuzzification module as part of the block FC-r can be done by the centre of gravity method, but the choice of the inference method and of the defuzzification method represent the user's options. The steps for the 2-DOF FC development are described in [IV-23], [IV-25]. The steps corresponding to the calculus of the 2-DOF controller and to the equivalence with a conventional controller with reference filters have been outlined in Appendix 2.

2. The 2-DOF controller in servo-driving systems

2.1. Basic situations

The role of the servo-driving system is very important for the type motion chosen and its control structure design. The two main benchmark models for the servo-systems (as joint actuators) are:

- a first-order (lag) with integrating component model, with its t.f. (A3-2-1):

$$P(s) = \frac{k_P}{s(1 + sT_1)}; \qquad \text{(A3-2-1)}$$

- a second-order model, with its t.f. (A3-2-2):

$$P(s) = \frac{k_P}{(1 + sT_1)(1 + sT_2)} \qquad \text{(A3-2-2)}$$

For benchmark type models there are control structures and design methods extendable in fuzzy controllers design, with good results ([IV-29], [IV-31], [IV-44] a.o.), applicable inclusively in servo-systems for aircrafts. For some applications the methods were extended in the fuzzy control domain.

Treating the problem in such a manner leads to the idea of controlling the servo-system by means of 2-DOF controllers. The tracking is obtained with an imposed a desired (linearized) reference model $H_{rm}(s)$. In linearized case the resulting reference controller does not change the eigenvalues (regarded to a point) of the compensated plant, which have been previously placed with output feedback design rules.

As mentioned in [IV-37] the mainly used independent joint nonlinear feedback control schemes can be reduced to linear (linearized) design situations. These structures can be themselves brought to the design of some 2-DOF structures. The further fuzzification of the 2-DOF controller (the generation of the 2-DOF FC) is done in terms of the principle presented in this section.

Considering the plant given by the continuous t.f. (A3-2-3)(a):

$$P(s) = \frac{1}{s^2 + 0.5s + 1} \quad \text{with } T_p=1/\omega_p=1,\ \zeta_p=0.25, \qquad \text{(a)} \quad \text{(A3-2-3)}$$

and

$$P(z) = \frac{0.0193\ z + 0.0187}{z^2 - 1.8669\ z + 0.9048}. \qquad \text{(b)} \quad \text{(A3-2-3)}$$

Two developed control structures are considered:

- a 2-DOF controller, which was in the second step fuzzyfied,
- a conventional PID, imposing an imposed phase margin, which was in the second step fuzzyfied.

❑ **The first control structure.** Involving the in [IV-35] mentioned 2-DOF controller which is fuzzified in accordance with the above mentioned aims. For the strictly speaking FC (the FC-r and FC-y blocks in Figures A3.1-1, A3.1-2), the fuzzification was solved in the initial phase as follows:

- for the calculated input variables Δr_k and $\Delta^2 r_k$ (Δy_k and $\Delta^2 y_k$, respectively), five linguistic terms with regularly distributed triangular type membership functions having an overlap of 1 are chosen;
- for the output variables $\Delta u_{r,k}$, $\Delta u_{y,k}$, seven linguistic terms with regularly distributed singleton type membership functions are chosen;
- the inference engine of the FC-r and FC-y blocks employs the Mamdani's MAX-MIN compositional rule of inference assisted by a complete rule base;
- the integrating effect is realized inside the control loop.

❑ **The second control structure**. By accepting the desired phase margin of 60^0, the resulting discrete PID controller has the t.f. (A3-1-12):

$$C_{PID}(z) = \frac{0.9085z^2 - 1.6961z + 0.8220}{z^2 - 1.7647z + 0.7647}. \qquad \text{(A3-2-4)}$$

2.2. Simulation Results

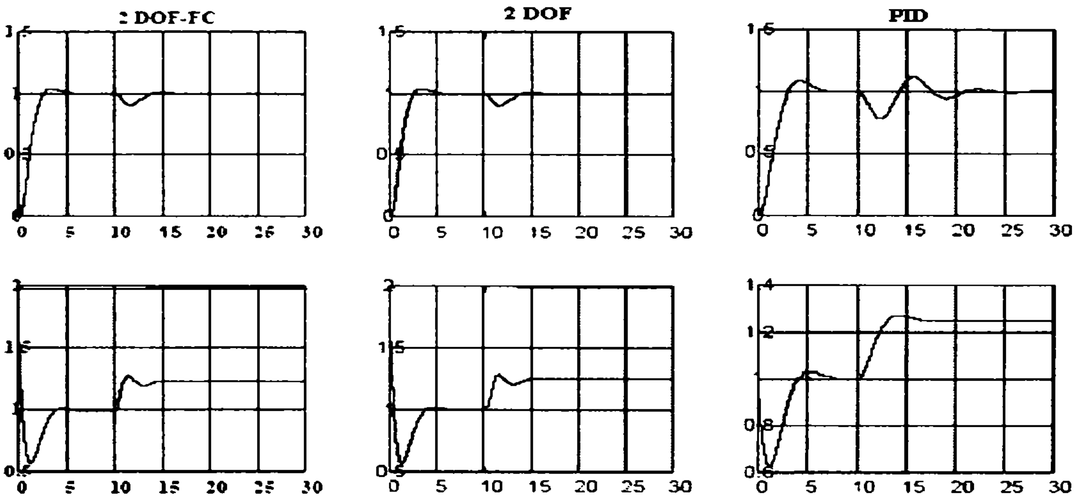
The programs, which implement the designed 2-DOF and PID controllers, are written in Matlab. The controllers are actually calculated using Matlab's facilities, the simulations are run in Simulink. A simultaneous simulation can be performed for a slight comparison between the behaviour of the two structures.

The following tests were made in relation with the simulation results shown in Fig. A.3.2-1 (a), (b) and (c)

- unit step response (reference signal) of the two systems ($0 < t < 10\,\text{sec}.$), followed by

- step disturbance response, t_{ov}=10, 10 < t < 30 sec.

The values for y(t) and u(t) were recorded. It can be observed that due to the linear model of the plant and to the large number of linguistic terms used, the difference between the behaviours of the CS with classical 2-DOF controller and with 2-DOF FC is not remarkable.



(a) (b) (c)

Fig. A3.2-1. Behaviour of CS with: (a) 2-DOF FC; (b) 2-DOF and (c) PID controller: reference unit step response followed by load disturbance step response

Comparing the simulation results for the two systems with 2-DOF-FC and with PID controller, it can be noticed that the 2-DOF CS-s are much more efficient in rejecting the disturbance that appear at the input of the plant.

3. Conclusions

By starting from the requirements concerning control applications for some driving systems, the research was focused on elaborating a method for the construction and development of 2-DOF FC-s.

The presented development method for 2-DOF FC's is easy to understand and to implement as CAD development. It is based on starting with the development of the 2-DOF controller followed by the transfer to the fuzzy processing of the components with dynamics. The integral element is included in the forward channel of the control loop.

The method is applicable relatively simply in the case of plants having not extremely large order. If the system order is increasing, there can appear problems in achievement of the fuzzy processing of the dynamic components.

In the design of the 2-DOF controller alternative approach methods are possible. Applications of 2-DOF fuzzy controllers in the field of robotics (for the control of the servo-systems) can be attractive (see [IV-50], [IV-51]).

References

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