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CONTRIBUTIONS ON IDENTIFICATION IN CLOSED-LOOP AND CONTROL DESIGN (APPLICATION TO DVD PLAYERS)

IDENTIFIKACE V UZAVŘENÉ SMYČCE A ŘÍDICÍ ALGORITMY S APLIKACÍ NA DVD PŘEHRÁVAČE

SHORT VERSION OF PH.D. THESIS

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1 INTRODUCTION

Feedback control is a widely accepted and frequently used technique in order to enforce a dynamical system to behave in a satisfactory manner. The broad concept of a dynamical system can thereby refer to many biological or engineering processes. The application of automatic control can be found in many complex industrial processes or sophisticated mechanical systems to attain a properly operating dynamical system.

Next to feedback control, system identification is used repeatedly to elucidate the dynamical aspects of the system. This procedure enables one to predict the dynamical behavior of an unknown system on the basis of foregoing observations of the system. In this way, knowledge of the dynamical aspects of an industrial process or of a mechanical system is acquired on the basis of experiments and construction of adequate mathematical models. Models are of rapidly increasing importance in engineering and today all control designs are more or less based on them. Models are also extensively used in other, nontechnical areas such a biology, ecology, and economy.

If the physical laws governing the behavior of the system are known we can use them to construct so called *white-box models* of the system. In white-box model, all parameters and variables can be interpreted in terms of physical entities and all constants are known *a priori*. The opposite extreme is known as *black-box modelling*. In this approach the models are constructed from measured input-output data. *A priori* knowledge about the model structure is eliminated and therefore the model and the parameters of it usually have little physical significance.

Linear model structures have been widely used for black-box modelling because they are mathematically attractive. This class of models is rich enough to cover a large number of applications, since systems are often controlled around an operating point and can be considered to behave linearly for small variations of the input.

A third approach is a combination of two extremes, and is called *grey-box modelling*. This approach exploits the *a priori* physical knowledge about the process, but the model structure and the parameters are not assumed to be completely known. The parameters of the model are estimated as for black-box models, by using identification methods. A typical grey box situation is when the model structure is determined by the physical relations of the process, hence the term *physical models*. In this situation, the identification is done with a fixed model structure (only the unknown parameters are estimated).

This thesis focuses on both control design and system identification, with the aim to design a control for a given system on the basis of *white-box model*. This approach is partly illustrated for an industrial Digital Versatile Disk-video player (DVD-video player), a high accuracy opto-electro-mechanical positioning system.

The thesis contains two main parts: The first part deals with theoretical and methodological part on *system identification* in closed-loop of black-box models using state-space representation to define a model structure. The second part of this thesis presents one case-study part on the *modelling and control system design* of a DVD player.

2 STATE-SPACE IDENTIFICATION FOR SISO/MIMO SYSTEMS

2.1 PROBLEM DESCRIPTION

For most physical systems it is easier to construct models with physical insight into continuous-time because most physical laws (Newton's laws of motion, relationships between electrical quantities *etc.*) are expressed in continuous-time. This means that modelling leads to a state-space representation:

$$\dot{x}(t) = F_0(\theta_\mathrm{c})x(t) + G_0(\theta_\mathrm{c})u(t), \tag{2.1}$$

where $F_0(\theta_\mathrm{c})$ and $G_0(\theta_\mathrm{c})$ are the matrices of appropriate dimensions ($n \times n$ and $n \times r$, respectively, for an n-dimensional state and an r-dimensional input) and θ_c is a vector of parameters that typically corresponds to unknown values of physical coefficients in continuous-time model (material constants, *etc.*). Such a selection of the parameter vector θ_c is called the *physical parametrization* of a plant model. The model (2.1) also defines the vector of state variables x which usually have the physical significance (position, velocity, charge, current *etc.*). The measured outputs are known combinations of the states.

The corresponding discrete-time state-space model can be obtained by sampling the input and output signals of the parametrized continuous-time model (2.1):

$$x(kT_\mathrm{s} + T_\mathrm{s}) = A_0(\theta_\mathrm{c})x(kT_\mathrm{s}) + B_0(\theta_\mathrm{c})u(kT_\mathrm{s}), \tag{2.2}$$

where T_s denotes the sampling period and $f_\mathrm{s} = 1/T_\mathrm{s}$ is the sampling frequency. Briefly:

$$x(k+1) = A_0(\theta_\mathrm{c})x(k) + B_0(\theta_\mathrm{c})u(k). \tag{2.3}$$

The discrete-time state-space matrices $A_0(\theta_\mathrm{c})$ and $B_0(\theta_\mathrm{c})$ can be expressed using the following formulae:

$$\begin{aligned} A_0(\theta_\mathrm{c}) &= e^{F_0(\theta_\mathrm{c})T_\mathrm{s}}, \\ B_0(\theta_\mathrm{c}) &= \int_0^{T_\mathrm{s}} e^{F_0(\theta_\mathrm{c})\tau} G_0(\theta_\mathrm{c})\mathrm{d}\tau. \end{aligned} \tag{2.4}$$

Under certain conditions it is possible to replace this complex relationship by a first-order approximation obtained if higher-order terms in the power series expansion of the matrix exponential are neglected:

$$\begin{aligned} A_0(\theta_\mathrm{c}) &\approx I + T_\mathrm{s}F_0(\theta_\mathrm{c}), \\ B_0(\theta_\mathrm{c}) &\approx T_\mathrm{s}G_0(\theta_\mathrm{c}). \end{aligned} \tag{2.5}$$

Considering the formulae (2.4) or (2.5), the discrete-time representation has a disadvantage that the matrices $A_0(\theta_\mathrm{c})$ and $B_0(\theta_\mathrm{c})$ are more complicated functions of the parameter vector θ_c than the matrices $F_0(\theta_\mathrm{c})$ and $G_0(\theta_\mathrm{c})$.

Adding the output equation, the standard discrete-time state-space model is obtained:

$$\begin{aligned} x(k+1) &= A_0(\theta_c)x(k) + B_0(\theta_c)u(k), \\ y(k) &= C_0(\theta_c)x(k) + D_0(\theta_c)u(k). \end{aligned} \tag{2.6}$$

The matrices $A_0(\theta_c)$, $B_0(\theta_c)$, $C_0(\theta_c)$, $D_0(\theta_c)$ of the model (2.6) can depend on the parameter vector θ_c in different ways. A typical case is when certain elements of matrices $A_0(\theta_c)$, $B_0(\theta_c)$, $C_0(\theta_c)$, $D_0(\theta_c)$ may be known or fixed values. The reason may be that the values of the parameters are *a pripori* known physical constants; or, we would like to impose a certain structure on the model. If model (2.6) is a canonical form then the parameter vector θ_c consists of the parameters of the original input-output transfer operator, [Kai80].

Although sampling of input and output signals of a continuous-time model is a natural way to get the discrete-time model (2.6), for certain applications it could be also given directly in discrete-time, with the matrices $A_0(\theta)$, $B_0(\theta)$, $C_0(\theta)$, $D_0(\theta)$ parametrized in terms of the parameter vector θ, rather than via (2.4) or (2.5).

For this assumption, only few specific identification methods in closed-loop are available, [Mel94], [Lju99]. Therefore, new identification algorithms for the discrete-time state-space models have been developed recently, see [Bez01] and [Bez04]. Nevertheless, useful analysis of their properties is missing. This is important for their use and the following improvement.

For these reasons, our study is only focused on the particular case of the identification algorithms where the state-space matrices $A_0(\theta)$, $B_0(\theta)$, $C_0(\theta)$, $D_0(\theta)$ are parametrized in terms of the parameter vector θ. Our contribution in this chapter is an analysis of the state-space identification algorithm properties developed in [Bez01] and [Bez04].

2.2 STATE-SPACE IDENTIFICATION ALGORITHMS FOR SISO SYSTEMS

The aim of the original output-error state-space identification algorithms, which are presented in this section, is to estimate the parameters of the SISO discrete-time state-space model:

$$\boxed{\begin{aligned} x(k+1) &= A_0(\theta)x(k) + B_0(\theta)u(k-d), \\ y(k) &= C_0(\theta)x(k) + p(k), \end{aligned}} \tag{2.7}$$

where $x(k)$ is an n-dimensional state vector, $p(k)$ is an output disturbance noise and θ is a d_0-dimensional parameter vector:

$$\theta = [\theta_1 \ \dots \ \theta_{d_0}]^{\mathrm{T}}. \tag{2.8}$$

One should note that the number of parameters d_0 is not anyhow related to the model order n.

To simplify notation, *backward shift operator* q^{-1} will be omitted in some terms. In the sequel *parameter vector* θ and *estimated parameter vector* $\hat{\theta}$ are also omitted in some terms whenever there is no risk of confusion.

The input-output transfer operator is given by:

$$S_{\mathrm{yu}}(q^{-1}) = q^{-d-1}C_0(I - q^{-1}A_0)^{-1}B_0 = \frac{q^{-d-1}C_0\,\mathrm{adj}(I - q^{-1}A_0)B_0}{\det\,(I - q^{-1}A_0)}\,. \tag{2.9}$$

Denoting the numerator and the denominator:

$$S_{\mathrm{yu}}(q^{-1}) = \frac{q^{-d}B(q^{-1})}{A(q^{-1})}\,, \tag{2.10}$$

$$B(q^{-1}) = b_1q^{-1} + \cdots + b_nq^{-n} = q^{-1}B^*(q^{-1}), \tag{2.11}$$

$$A(q^{-1}) = 1 + a_1q^{-1} + \cdots + a_nq^{-n} = 1 + q^{-1}A^*(q^{-1}), \tag{2.12}$$

one obtains:

$$\begin{aligned} B(q^{-1}) &= q^{-1}C_0\,\mathrm{adj}(I - q^{-1}A_0)B_0, \\ A(q^{-1}) &= \det\,(I - q^{-1}A_0), \end{aligned} \tag{2.13}$$

$$\begin{aligned} B^*(q^{-1}) &= C_0\,\mathrm{adj}(I - q^{-1}A_0)B_0, \\ A^*(q^{-1}) &= q\left(\det\,(I - q^{-1}A_0) - 1\right), \end{aligned} \tag{2.14}$$

and the output of the plant is given by:

$$\begin{aligned} y(k+1) &= S_{\mathrm{yu}}(q^{-1})u(k+1) + p(k+1), \\ &\vdots \\ y(k+1) &= -A^*y(k) + B^*u(k-d) + Ap(k+1). \end{aligned} \tag{2.15}$$

Taking into account that state-space matrices $A_0(\theta)$, $B_0(\theta)$, $C_0(\theta)$ are functions of the parameter vector θ, the equations (2.13) or (2.14) also define the transfer function coefficients $b_1, \ldots, b_n$ and $a_1, \ldots, a_n$ as functions of θ. Therefore using (2.13) or (2.14), one can determine the function $\Gamma(\theta)$, which transforms the parameter vector θ to a vector of the transfer function coefficients:

$$\Gamma^{\mathrm{T}}(\theta) = [a_1(\theta)\ \ldots\ a_n(\theta) \quad b_1(\theta)\ \ldots\ b_n(\theta)]\,. \tag{2.16}$$

This transformational function plays a key role in all the newly proposed algorithms. One can clearly see that $\Gamma(\theta)$ is a $2n$-dimensional vector function of a d_0-dimensional parameter vector θ. It is generally non-linear, the estimated parameters can appear in sums, products, ratios and other aggregate terms. It is assumed differentiable.

Using (2.16), (2.15) can be rewritten to the regressor form:

$$y(k+1) = \Gamma^{\mathrm{T}}(\theta)\varphi(k) + Ap(k+1), \tag{2.17}$$

where $\varphi(k)$ denotes the regressor vector:

$$\varphi(k) = [-y(k)\ \ldots\ -y(k-n+1) \quad u(k-d)\ \ldots\ u(k-n+1-d)]^{\mathrm{T}}, \tag{2.18}$$

and the closed-loop predictor is given by:

$$\hat{y}(k+1) = \Gamma^{\mathrm{T}}(\hat{\theta})\phi(k), \tag{2.19}$$

where $\hat{\theta}$ denotes the estimated parameter vector:

$$\hat{\theta} = \left[\hat{\theta}_1 \ \dots \ \hat{\theta}_{d_0}\right]^{\mathrm{T}}, \tag{2.20}$$

and $\phi(k)$ the predictor regressor vector:

$$\phi(k) = [-\hat{y}(k) \ \dots \ -\hat{y}(k-n+1) \quad \hat{u}(k-d) \ \dots \ \hat{u}(k-n+1-d)]^{\mathrm{T}}. \tag{2.21}$$

Replacing the fixed predictor of the closed-loop (2.19) by an adjustable predictor, one obtains *a priori* predicted output:

$$\hat{y}^{\circ}(k+1) = \hat{y}\left[k+1|\hat{\theta}(k)\right] = \Gamma^{\mathrm{T}}\left[\hat{\theta}(k)\right]\phi(k), \tag{2.22}$$

and *a posteriori* predicted output:

$$\hat{y}(k+1) = \hat{y}\left[k+1|\hat{\theta}(k+1)\right] = \Gamma^{\mathrm{T}}\left[\hat{\theta}(k+1)\right]\phi(k). \tag{2.23}$$

Consequently, the *a priori* prediction error can be defined as:

$$\varepsilon^{\circ}_{\mathrm{CL}}(k+1) = y(k+1) - \hat{y}^{\circ}(k+1), \tag{2.24}$$

and the *a posteriori* prediction error as:

$$\varepsilon_{\mathrm{CL}}(k+1) = y(k+1) - \hat{y}(k+1). \tag{2.25}$$

2.2.1 The SISO GM-2 Algorithm

$$\boxed{\begin{array}{rcl} \hat{\theta}(k+1) & = & \hat{\theta}(k) + \alpha\left(\Gamma'^{\mathrm{T}}_{\theta}\left[\hat{\theta}(k)\right]\phi(k) + \dfrac{\partial\phi(k)}{\partial\hat{\theta}(k)}^{\mathrm{T}}\Gamma\left[\hat{\theta}(k)\right]\right)\varepsilon^{\circ}_{\mathrm{CL}}(k+1) \\ \varepsilon^{\circ}_{\mathrm{CL}}(k+1) & = & y(k+1) - \Gamma^{\mathrm{T}}\left[\hat{\theta}(k)\right]\phi(k) \end{array}} \tag{2.26}$$

One should note that the GM-2 algorithm have the advantage of a closed analytic form.

2.2.2 The SISO IGM-1 and IGM-2 Algorithms

Using the scalar adaptation gains α_1 and α_2, the IGM-1 state-space parameter adaptation algorithm for SISO systems can be summarized:

$$\boxed{\begin{array}{rcl} \hat{\theta}(k+1) & = & \hat{\theta}(k) + \dfrac{\alpha_1\Gamma'^{\mathrm{T}}_{\theta}\left[\hat{\theta}(k)\right]\phi(k)\varepsilon^{\circ}_{\mathrm{CL}}(k+1)}{1+\alpha_2\phi^{\mathrm{T}}(k)\phi(k)} \\ \varepsilon^{\circ}_{\mathrm{CL}}(k+1) & = & y(k+1) - \Gamma^{\mathrm{T}}\left[\hat{\theta}(k)\right]\phi(k) \end{array}} \tag{2.27}$$

Correspondingly, the IGM-2 algorithm is:

$$
\boxed{
\begin{array}{rcl}
\hat{\theta}(k+1) & = & \hat{\theta}(k) + \dfrac{\alpha_1 \Gamma_\theta'^{\mathrm{T}} \left[\hat{\theta}(k)\right] \phi(k) \varepsilon_{\mathrm{CL}}^{\circ}(k+1)}{1 + \alpha_1 \phi^{\mathrm{T}}(k) \Gamma_\theta' \left[\hat{\theta}(k)\right] \Gamma_\theta'^{\mathrm{T}} \left[\hat{\theta}(k)\right] \phi(k)} \\
\varepsilon_{\mathrm{CL}}^{\circ}(k+1) & = & y(k+1) - \Gamma^{\mathrm{T}} \left[\hat{\theta}(k)\right] \phi(k)
\end{array}
}
\tag{2.28}
$$

One should note that the IGM-1 and IGM-2 algorithms have the advantage of a closed analytic form.

2.2.3 The SISO RLS-2 Algorithm

$$
\boxed{
\begin{array}{rcl}
\hat{\theta}(k+1) & = & \hat{\theta}(k) + \Gamma_\theta' \left[\hat{\theta}(k)\right]^{(-1)} F(k+1) \phi(k) \varepsilon_{\mathrm{CL}}^{\circ}(k+1) \\
F(k+1) & = & \dfrac{1}{\lambda_1(k)} \left[F(k) - \dfrac{F(k)\phi(k)\phi^{\mathrm{T}}(k)F(k)}{\frac{\lambda_1(k)}{\lambda_2(k)} + \phi^{\mathrm{T}}(k)F(k)\phi(k)} \right] \\
\varepsilon_{\mathrm{CL}}^{\circ}(k+1) & = & y(k+1) - \Gamma^{\mathrm{T}} \left[\hat{\theta}(k)\right] \phi(k)
\end{array}
}
\tag{2.29}
$$

2.3 STATE-SPACE IDENTIFICATION ALGORITHMS FOR MIMO SYSTEMS

One can consider as a general structure the following PAA (integral type):

$$
\boxed{
\begin{array}{rcl}
\hat{\theta}(k+1) & = & \hat{\theta}(k) + \Gamma_\theta' \left[\hat{\theta}(k)\right]^{(-1)} F(k+1) \phi^{\mathrm{T}}(k) \varepsilon_{\mathrm{CL}}(k+1), \\
F(k+1)^{-1} & = & \Lambda_1(k) F(k)^{-1} + \Lambda_2(k) \phi(k) \phi^{\mathrm{T}}(k), \\
\Lambda_1 & = & \lambda_1 I, \\
\Lambda_2 & = & \lambda_2 I, \\
0 < \lambda_1(k) \leq 1, & & 0 \leq \lambda_2(k) < 2, \\
F(0) > 0, & & F(k)^{-1} > \alpha F(0)^{-1}, \quad 0 < \alpha < \infty, \\
F(k+1) & = & \frac{1}{\lambda_1(k)} \Big\{ F(k) - F(k)\phi^{\mathrm{T}}(k) \Big[\frac{\lambda_1(k)}{\lambda_2(k)} I + \\
& & \phi(k)F(k)\phi^{\mathrm{T}}(k) \Big]^{-1} \phi(k)F(k) \Big\}, \\
\varepsilon_{\mathrm{CL}}(k+1) & = & \varepsilon_{\mathrm{CL}}^{\circ}(k+1) \{ I + \phi^{\mathrm{T}}(k)F(k)\phi(k) \}^{-1},
\end{array}
}
\tag{2.30}
$$

2.4 PROPERTIES OF PARAMETER ADAPTATION ALGORITHMS FOR STATE-SPACE SYSTEMS

2.4.1 Convexity of Criterion Function

$$
J(\hat{\theta}) = \sum_{i=1}^{k} \left(y(i) - \phi(i-1)\Gamma(\hat{\theta}) \right)^{\mathrm{T}} \left(y(i) - \phi(i-1)\Gamma(\hat{\theta}) \right). \tag{2.31}
$$

Assuming a particular parametrization of $\Gamma(\hat{\theta})$ function, $\Gamma(\hat{\theta}) = M\hat{\theta}$, where M is the square matrix given from the parametrization and $\hat{\theta}$ is the estimated parameter vector, then the criterion function $J(\hat{\theta})$, given by (2.31), is a convex function.

2.4.2 Stability of PAA in a Deterministic Environment

Assuming that the closed-loop system (with the controller transfer function matrix K) is stable and the function $\Gamma(\hat{\theta})$ is linear in the parameter vector $\hat{\theta}$, *i.e.* $\Gamma(\hat{\theta}) = M\hat{\theta}$, the recursive parameter estimation algorithm given by (2.30) assures

$$\lim_{k\to\infty} \varepsilon_{\mathrm{CL}}(k+1) = 0,$$
$$\lim_{k\to\infty} \varepsilon^{\circ}_{\mathrm{CL}}(k+1) = 0,$$
$$||\phi(k)|| < C,\ 0 < C < \infty,\ \forall k$$

for all initial conditions $\hat{\theta}(0)$, $\varepsilon^{\circ}_{\mathrm{CL}}(0)$ and $\phi(0)$ if

$$H^{'}_{\mathrm{r}}(z^{-1}) = H_{\mathrm{r}}(z^{-1}) - \frac{\lambda_2}{2}I = \left[I + z^{-1}(A^* + B^*K)\right]^{-1} - \frac{\lambda_2}{2}I \tag{2.32}$$

is strictly positive real matrix, where: $\max_{k}\left(\lambda_2(k)\right) \leq \lambda_2 < 2$.

2.4.3 Parametric Convergence Analysis in a Stochastic Environment

If a particular case of parametrization is assumed $\Gamma(\hat{\theta}) = M\hat{\theta}$, where M is a square matrix, then the *a posteriori* prediction error in the presence of output disturbance noise is given by:

$$\begin{aligned}\varepsilon_{\mathrm{CL}}(k+1) = \left\{I + q^{-1}(A^* + B^*K)\right\}^{-1}\phi(k)M\{\theta - \hat{\theta}(k+1)\}\\ +\left\{I + q^{-1}(A^* + B^*K)\right\}^{-1}Ap(k+1)\,.\end{aligned} \tag{2.33}$$

Consider the recursive parameter estimation algorithm given by (2.30) with $\lambda_1(k) = 1$. Define

$$\phi(k,\,\hat{\theta}) \triangleq \phi(k)|_{\hat{\theta}(k)=\hat{\theta}=\mathrm{const.}}, \tag{2.34}$$
$$\varepsilon_{\mathrm{CL}}(k+1,\hat{\theta}) \triangleq \varepsilon_{\mathrm{CL}}(k+1)|_{\hat{\theta}(k)=\hat{\theta}=\mathrm{const.}}, \tag{2.35}$$
$$D_{\mathrm{s}} \triangleq \{\hat{\theta}:\ \hat{A}(z^{-1})S(z^{-1}) + z^{-d}B(z^{-1})R(z^{-1}) = 0 \Rightarrow |z| < 1\}$$

Assume that $\hat{\theta}(k)$ generated by the algorithm belongs infinitely often to the domain D_{s} for which the stationary process $\phi(k,\,\hat{\theta})$ and $\varepsilon_{\mathrm{CL}}(k+1,\hat{\theta})$ can be defined.

Assume that $p(k)$ is a zero-mean stochastic process with finite moments independent of the reference sequence $r^*(k)$.

If

$$H^{'}_{\mathrm{r}}(z^{-1}) = H_{\mathrm{r}}(z^{-1}) - \frac{\lambda_2}{2}I = \left[I + z^{-1}(A^* + B^*K)\right]^{-1} - \frac{\lambda_2}{2}I. \tag{2.36}$$

is strictly positive real (SPR) transfer function matrix, where: $\max_t (\lambda_2(t)) \leq \lambda_2 < 2$ then

$$\text{Prob}\{\lim_{k\to\infty} \hat{\theta}(k) \in D_\text{c}\} = 1,$$

where $D_\text{c} = \{\hat{\theta} : (\phi^\text{T}(k, \hat{\theta}))(\theta - \hat{\theta}) = 0\}$. If, furthermore, $\phi^\text{T}(k, \hat{\theta})(\theta - \hat{\theta}) = 0$ has a unique solution (richness condition) then the condition $H_\text{r}'(z^{-1})$ given by (2.36) be strictly positive real matrix implies that

$$\text{Prob}\{\lim_{k\to\infty} \hat{\theta}(k+1) = \theta\} = 1.$$

2.4.4 Frequency Distribution of the Asymptotic Bias

For the case of the closed-loop methods, the formula for the estimated parameter vector when number of data $N \to \infty$ is

$$\hat{\theta}^* = \arg\min_{\hat{\theta}\in\mathcal{D}} \int_{-\pi}^{\pi} |S_\text{yp}|^2[|S_\text{yu} - \hat{S}_\text{yu}|^2|\hat{S}_\text{yp}|^2\phi_\text{w} + |H|^2\phi_\text{e}]\text{d}\omega,$$

where $S_\text{yp}(\omega)$ corresponds to the true output sensitivity transfer function matrix between the output disturbances vector p and the plant outputs vector y, $\hat{S}_\text{yp}(\omega)$ is the estimated output sensitivity transfer function matrix, $\phi_\text{w}(\omega)$ corresponds to the spectral density of the external excitation signals vector w, and the $\phi_\text{e}(\omega)$ is the spectral density of the noise vector e. This formula shows that the noise does not affect the parameter estimation.

2.5 SIMULATION EXPERIMENTS FOR SISO SYSTEMS

The following state-space representation of the test plant $\mathcal{S}$ is considered:

$$\begin{aligned} A_0(\theta) &= \left[\begin{array}{c|c} -\theta_1 + \theta_2 & -\theta_3 + \theta_4 \\ \hline 1 & 0 \end{array}\right], \quad B_0(\theta) = \left[\begin{array}{c} 1 \\ \hline 0 \end{array}\right], \\ C_0(\theta) &= \left[\begin{array}{c|c} \theta_1 + \theta_2 + \theta_3 + \theta_4 & -\theta_3 \end{array}\right], \end{aligned}$$

for the parameter vector defined by:

$$\theta = [\theta_1,\ \theta_2,\ \theta_3,\ \theta_4]^\text{T}.$$

Since the state-space matrices are given in a canonical controller form, one can immediately determine the vector of transfer function coefficients $\Gamma(\theta)$ and its Jacobian matrix $\Gamma_\theta'(\theta)$:

$$\Gamma(\theta) = \begin{bmatrix} \theta_1 - \theta_2 \\ \theta_3 - \theta_4 \\ \theta_1 + \theta_2 + \theta_3 + \theta_4 \\ -\theta_3 \end{bmatrix}, \qquad \Gamma_\theta'(\theta) = \begin{bmatrix} 1 & -1 & 0 & 0 \\ 0 & 0 & 1 & -1 \\ 1 & 1 & 1 & 1 \\ 0 & 0 & -1 & 0 \end{bmatrix}.$$

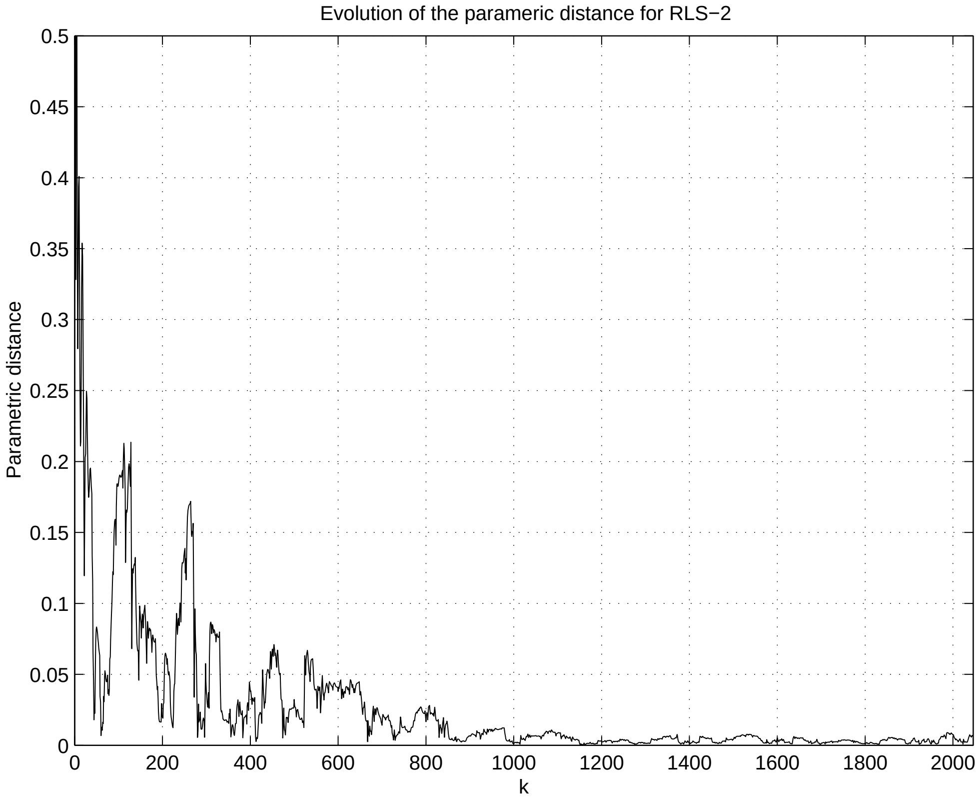

Figure 2.1: Evolution of the parametric distance using the RLS-2 algorithm.

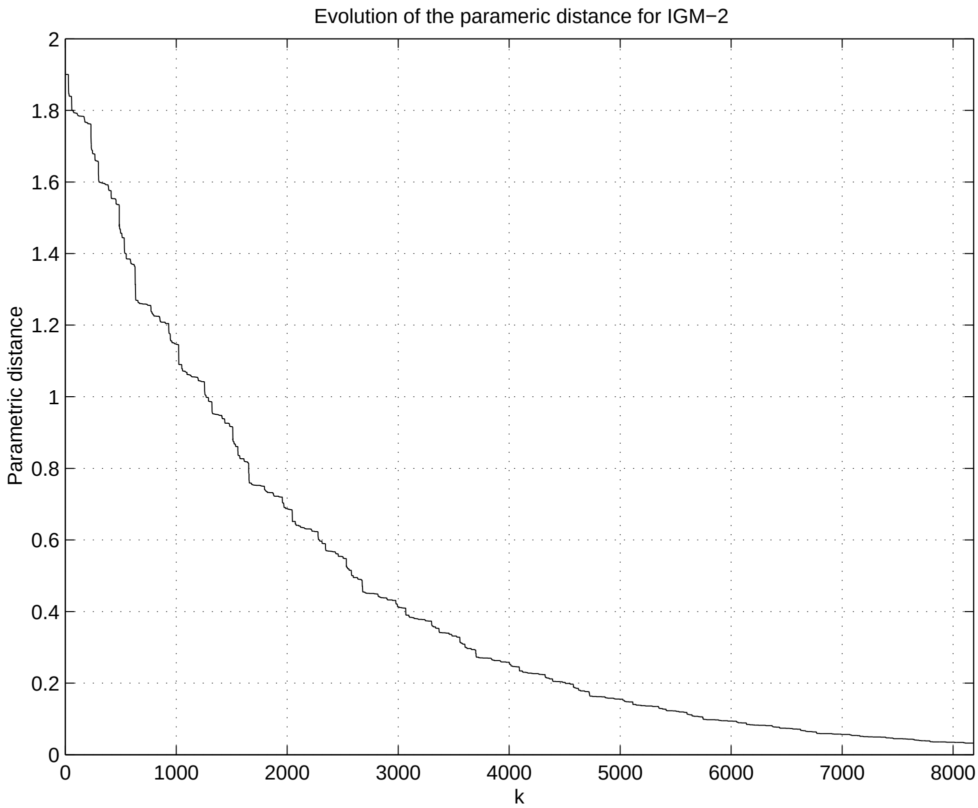

Figure 2.2: Evolution of the parametric distance using the IGM-2 algorithm.

Figure 2.1 shows that the convergence speed of the RLS-2 algorithm remains unbeaten. The RLS-2 algorithm offers the best estimation in this case. Figure 2.1 shows the same experiment using the IGM-2 algorithm and it suffers from a lower convergence speed.

3 THE DVD PLAYER SYSTEM DESCRIPTION

A diagram for the optical pick-up unit control of the DVD player is illustrated in fig. 3.1.

The focus error signal is given by:

$$e_{\mathrm{F}}(\Delta z) = (V_{\mathrm{A}} + V_{\mathrm{C}}) - (V_{\mathrm{B}} + V_{\mathrm{D}}), \tag{3.1}$$

where V_{A}, V_{B}, V_{C}, V_{D} are voltages from the quadrants A, B, C, D of the photodetector and Δz is the vertical spot position error. The focus error signal e_{F} is feed back to the servo system, to control the actuator fine displacement along the vertical direction z.

The photodetector provides a non-linear bipolar focus error characteristic $e_{\mathrm{F}}(\Delta z)$, usually called *S-curve*, which is used to determine if the laser spot is correctly focused on the disk information layer.

3.1 MODELLING PROBLEM DESCRIPTION

Despite the S-curve practical importance, no simple analytical or numerical model of the S-curve is available to our knowledge at present. These models could be important for more sophisticated control system design and for testing its functionality (normal playing mode, start-up procedure, calibration, error detection, offset compensation, *etc.*).

3.2 MODELLING OF THE FOCUS ERROR SIGNAL GENERATION

The main idea of model creation is expressed by the following sentences: Since the whole optical system contains cylindrical lens (causing the astigmatism), we can assume that *the whole optical system is separated into two subsystems with lenses, which are easier to describe.* The separation of the whole optical system into the two orthogonal planes yz and xz allows to use the theory of a system formed by two centered thin lenses, [BW87].

The analytical and numerical model have been created. The analytical model uses uniform intensity distribution of the laser beam on the photodetector while the numerical model uses Gaussian intensity distribution.

Fig. 3.2 shows the focus error characteristic $e_{\mathrm{F}}(\Delta z)$. The curve fitting procedure, described above, has been applied to obtain the focus error characteristics of two created mathematical models $\hat{e}_{\mathrm{F}}(\Delta z, \theta)$ that are shown in fig. 3.2. They are indicated by the solid and dot-dashed lines. The "nominal" focus error characteristic $e_{\mathrm{F}}(\Delta z)$ obtained from preprocessing is indicated by the dashed line there.

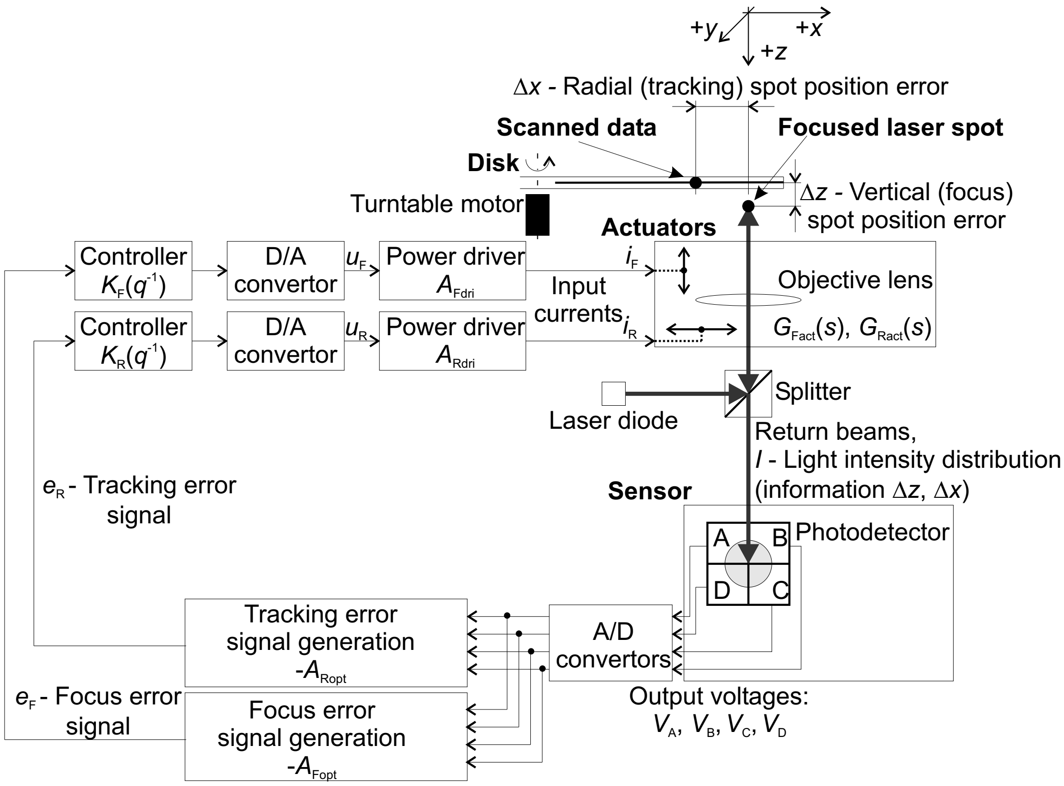

Figure 3.1: Control of the DVD player, the focus and tracking loop, and main physical components.

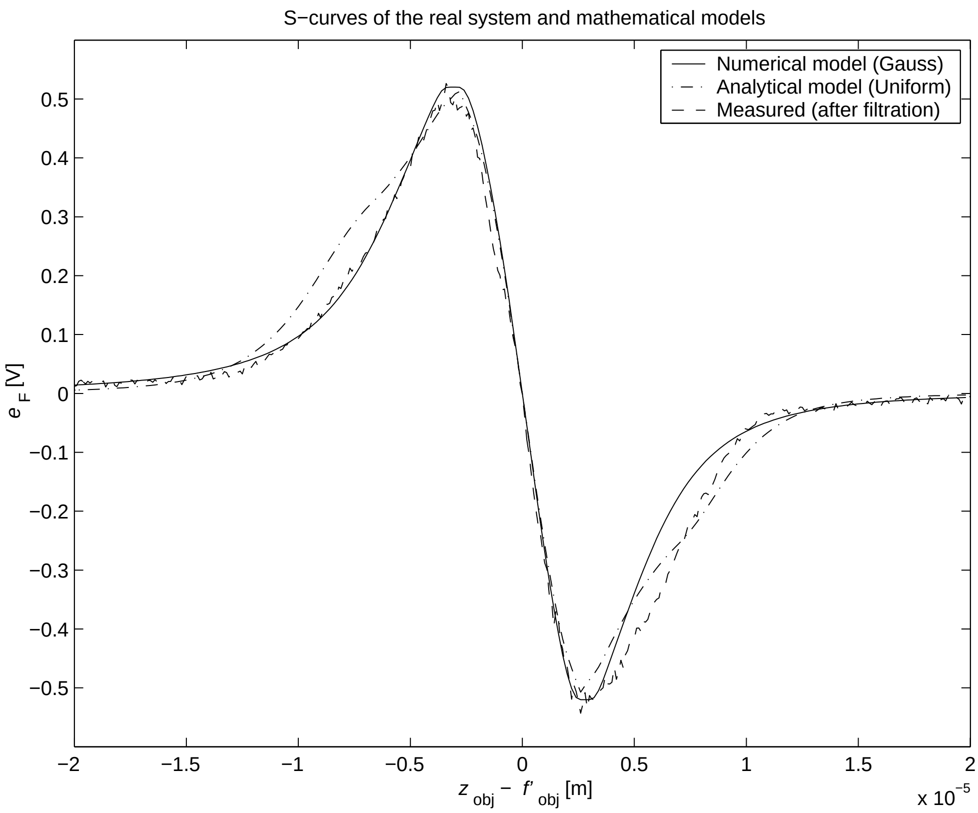

Figure 3.2: S-curves of the real system and mathematical models.

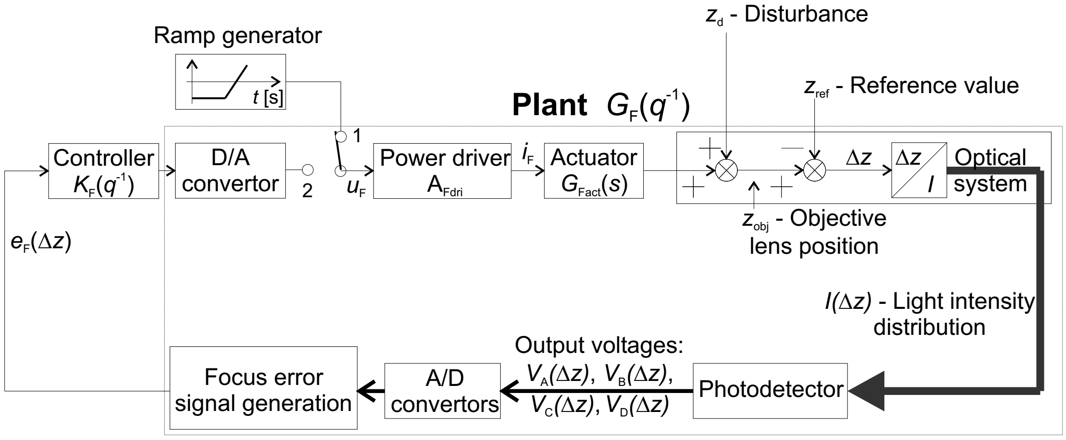

Figure 3.3: Block diagram of the focus control loop, focus processing procedure to automatically close the focus loop.

3.3 POSSIBLE APPLICATIONS OF PHOTODETECTOR CHARACTERISTICS

3.3.1 Simulation of Start-up Procedure

Figs. 3.4 and 3.5 present the focus error signal from simulation/real data measuring of the focus processing procedure to automatically close the focus loop. The simulated focus error signal $e_{\mathrm{F}}(t)$, see fig. 3.4, has been obtained from realistic DVD simulator using the developed models. Its block diagram is shown in fig. 3.3. The disturbance z_{d} was set to zero during simulation experiment, *i.e.* $z_{\mathrm{d}} = 0$.

This procedure is currently used in DVD players and a good coincidence can be observed between simulated and measured focus error signals $e_{\mathrm{F}}(t)$. It is clear that the common gain approximation of the S-curve, widely used for controller design, does not allow testing of this start-up procedure that is in fact always executed at least at the beginning of DVD/CD disk playing.

3.3.2 Feedback Focus Control Loop

The second application of the models is the same as in linear gain approximation, (in *lock-on range*), for focus control loop, like in fig. 3.3. Nevertheless, the designed models are more realistic. Moreover, they are also usable in the small/long jump procedure while the linear gain approximation cannot be used here.

3.3.3 New Photodetector Characteristics Design

The third application of models could be in new photodetector design, since the presented modelling describes which parameters influence the whole S-curve shape.

3.4 CONTROL PROBLEM DESCRIPTION

In the present study, the aim is to find a fixed low-order controller being able to improve the eccentricity suppression in the focus/radial control loops for DVD play-

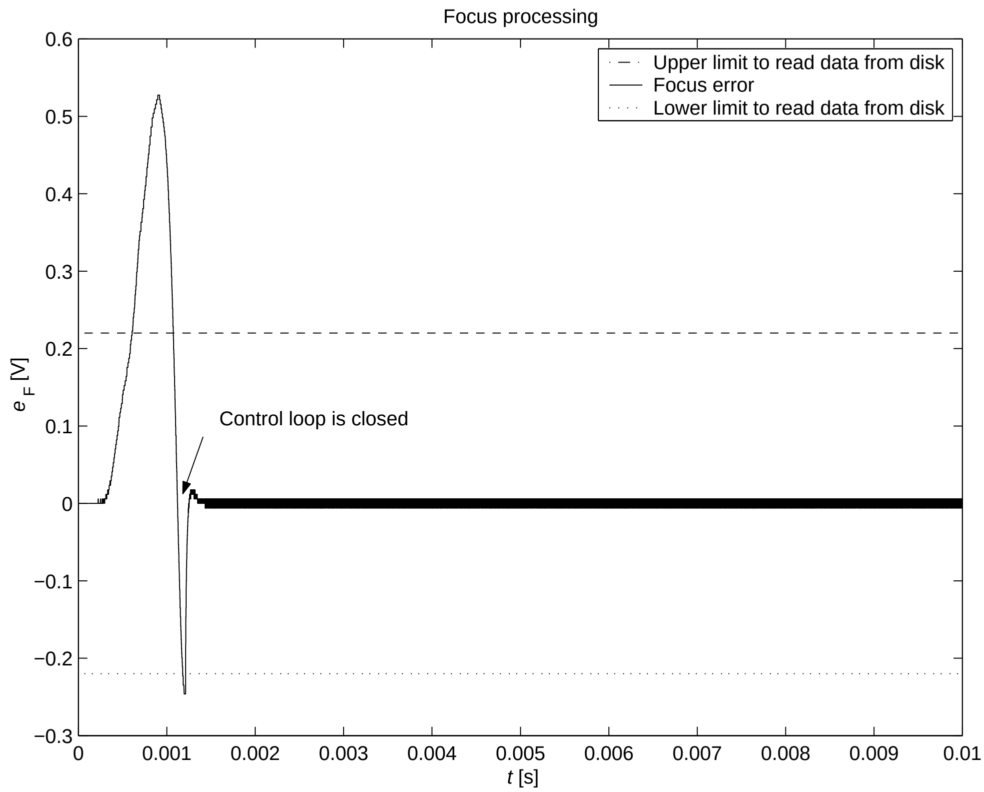

Figure 3.4: Focus processing procedure to automatically close the focus loop obtained from simulation where the created analytical model of the S-curve has been included.

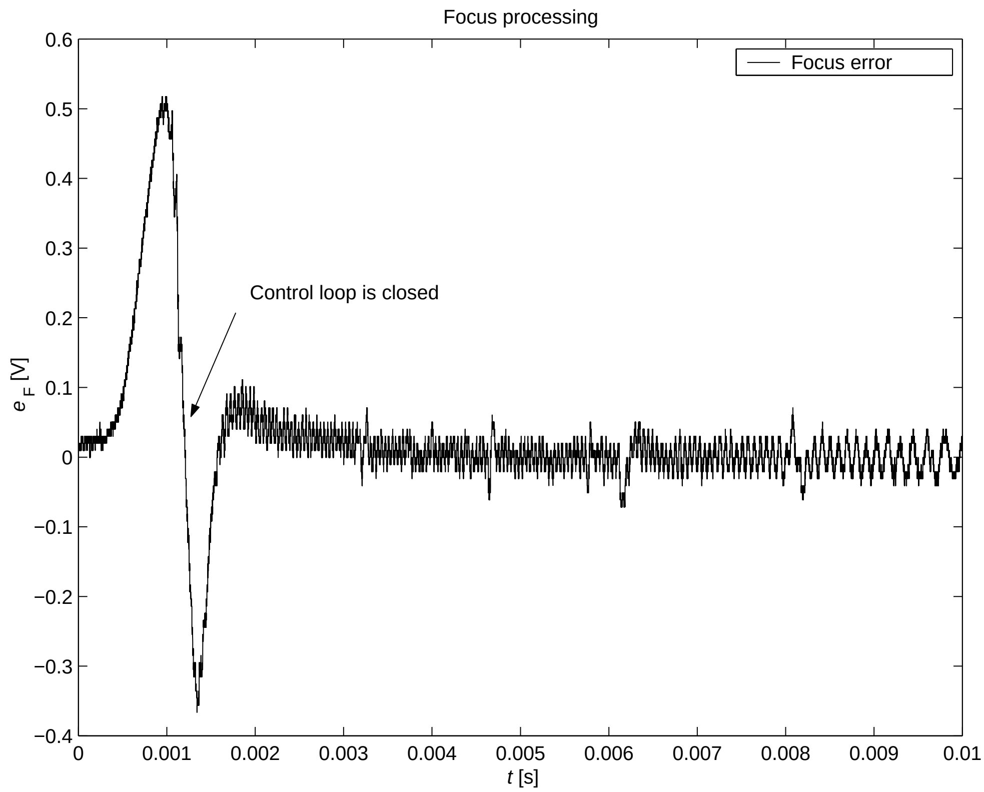

Figure 3.5: Focus processing procedure to automatically close the focus loop obtained by measurement on a real DVD-video player from STMicroelectronics.

ers. It is shown that this can be achieved with pole placement design, followed by controller order reduction. Pole placement method is here adapted to realize repetitive disturbance rejection in a certain bandwidth.

4 CONTROL SYSTEM DESIGN

4.1 COMBINED POLE PLACEMENT/SENSITIVITY FUNCTION SHAPING

The standard digital control configuration obtained with polynomial RS controller, see *e.g.* [LLM97], is presented in fig. 4.1. Part T of RST structure has been omitted because the control design in the focus/tracking loop of DVD/CD players only deals with the disturbance rejection problem.

The linear time invariant model of the plant is in general described by the transfer function:

$$\begin{aligned} G(q^{-1}) &= \frac{q^{-d}\,B(q^{-1})}{A(q^{-1})} \\ &= \frac{q^{-d}\left(b_1\,q^{-1} + \cdots + b_{n_B}\,q^{-n_B}\right)}{1 + a_1\,q^{-1} + \cdots + a_{n_A}\,q^{-n_A}}, \end{aligned} \tag{4.1}$$

where q^{-1} is the backward time shift operator, d is the pure time delay, T_s is the sampling period and $f_\mathrm{s} = 1/T_\mathrm{s}$ is the sampling frequency.

Pole placement method has been used to design the RS controller which has the following transfer function:

$$\begin{aligned} K(q^{-1}) &= \frac{R(q^{-1})}{S(q^{-1})} = \frac{R^{'}(q^{-1})\,H_\mathrm{R}(q^{-1})}{S^{'}(q^{-1})\,H_\mathrm{S}(q^{-1})} \\ &= \frac{r_0 + r_1\,q^{-1} + \cdots + r_{n_\mathrm{R}}\,q^{-n_\mathrm{R}}}{1 + s_1\,q^{-1} + \cdots + s_{n_\mathrm{S}}\,q^{-n_\mathrm{S}}}, \end{aligned} \tag{4.2}$$

where $H_\mathrm{R}(q^{-1})$ and $H_\mathrm{S}(q^{-1})$ denote the fixed parts of the controller (either imposed by the design or introduced in order to shape the sensitivity functions) and $R^{'}(q^{-1})$, $S^{'}(q^{-1})$ are the solutions of the Bezout equation:

$$AS^{'}H_\mathrm{S} + BR^{'}H_\mathrm{R} = P, \tag{4.3}$$

where P represents the characteristic polynomial (closed-loop poles).

The sensitivity functions play a crucial role in the robustness analysis of the closed-loop system with respect to modelling errors. These functions are *shaped* in order to assure *nominal performance* for the rejection of the disturbances and the stability of the closed-loop system in the presence of model mismatch.

The *output sensitivity function* $S_\mathrm{yp}(q^{-1})$ is the transfer function between the output disturbance $p(t)$ and the plant output $y(t)$. It is given by expression:

$$S_\mathrm{yp}(q^{-1}) = \frac{y(t)}{p(t)} = \frac{AS^{'}H_S}{P}\,. \tag{4.4}$$

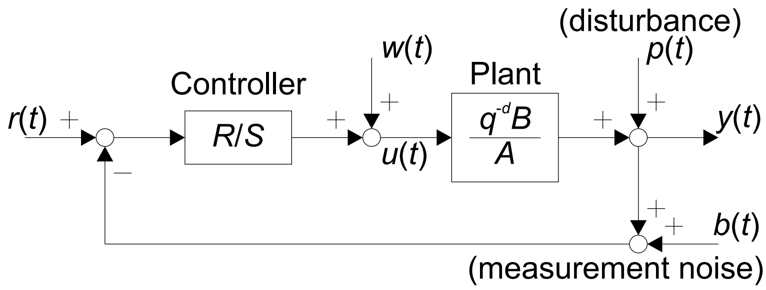

Figure 4.1: Closed-loop system with RS controller.

The other sensitivity functions are defined in the same way as follows: The *input sensitivity function* $S_{\rm up}(q^{-1})$ is the transfer function between the output disturbance $p(t)$ and the plant input $u(t)$. The *output sensitivity function* with respect to an input disturbance $S_{\rm yw}(q^{-1})$ is the transfer function between the input disturbance $w(t)$ and the plant output $y(t)$. The *complementary sensitivity function* $S_{\rm yr}(q^{-1})$ is the transfer function between the reference $r(t)$ and the plant output $y(t)$. This complementary sensitivity function with a negative sign is called the *noise sensitivity function* $S_{\rm yb}(q^{-1})$.

In our case not only robustness (the modulus margin ΔM, delay margin $\Delta\tau$ and phase margin $\Delta\phi$) but also the performances specifications have to be checked. Therefore the sensitivity function shaping is a useful tool to the controller design in case of DVD/CD players.

4.2 CONTROLLER ORDER REDUCTION

One useful methodology, that has been used here, is the balanced reduction method in state space domain which is using a Gramian of the balanced state-space realization of the reachable, observable, stable system, [SP96]. If the system is normalized properly, small elements in the balanced Gramian indicate states that can be removed to reduce the controller to lower order.

By applying this algorithm, a 3rd order controller has been obtained, *i.e.* $n_{\rm R} = 3$, $n_{\rm S} = 3$ in (4.2), which has been implemented in discrete-time on the industrial benchmark.

4.3 FOCUS LOOP: REAL-TIME MEASUREMENTS

The experimental results are shown only for the 3rd order RS controller $K_{\rm RS3}$ and the 3rd order actual (standard) controller $K_{\rm act}$ because the 4th order controller is not implementable into a DSP controller structure in DVD-video player benchmark.

In fig. 4.2 the measured and the simulated output sensitivity function magnitudes $|S_{\rm yp}|$, obtained for the 3rd order controller $K_{\rm RS3}$ are presented. Here, as term of comparison, we present also the measured and the simulated output sensitivity function magnitudes computed with the actual implemented controller $K_{\rm act}$.

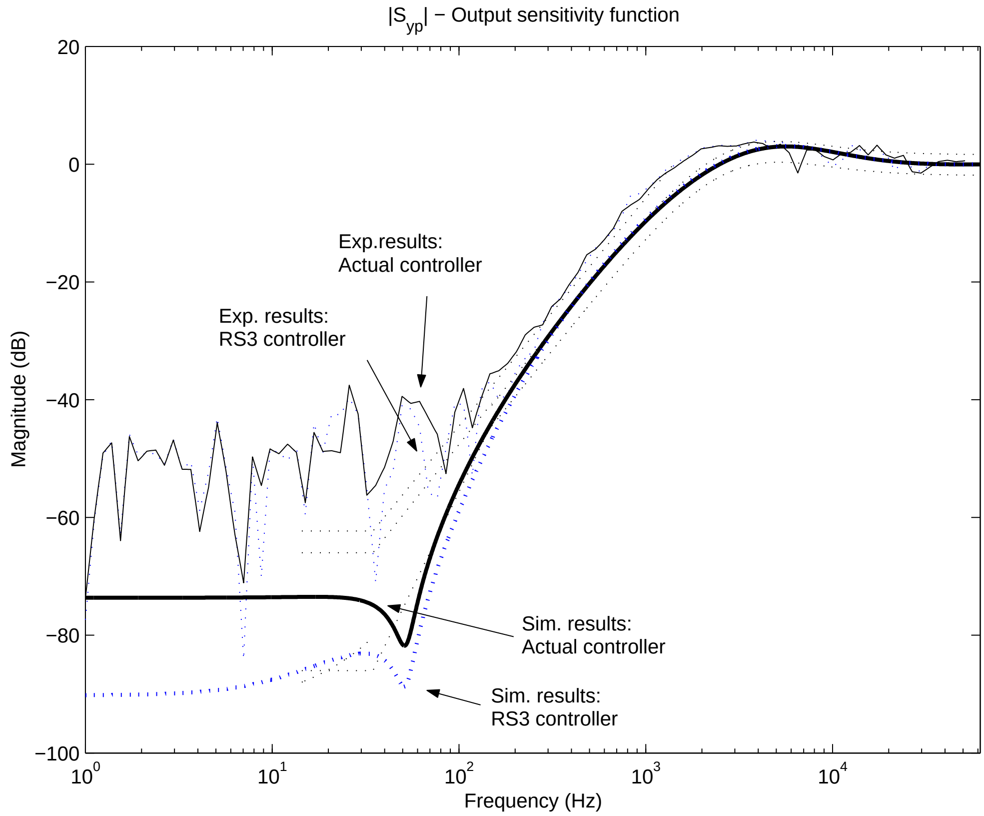

Figure 4.2: The measured and the simulated magnitude of the output sensitivity function obtained for the 3rd order controller K_{RS3} and the actual controller K_{RS3}, focus loop.

From this figure it can be seen that closed-loop measurements are very ill-conditioned at low frequencies since the high attenuation level needed at low frequencies to suppress external disturbances reduces the output Signal to Noise Ratio (SNR). This is true when both designed and actual controllers are used. This explains why, in fig. 4.2, the measured frequency responses of $|S_{\mathrm{yp}}|$, obtained for both controllers, lie above the specification requirements up to $100\,$Hz, and matches with the simulated curves obtained with the same controllers only after this frequency.

Finally, in figs. 4.3 and 4.4 the measured Power Spectral Densities (PSD) of the focus error signal e_{F}, obtained for the 3rd order controller K_{RS3} and the actual controller implemented in the current industrial solution K_{act}, are presented. Measurements have been acquired by using the worst-case disk (test disk having nominal vertical deviation at the disk outer edge $z_{\mathrm{d_max}} = 0.5\,$mm and nominal eccentricity of $x_{\mathrm{d_max}} \approx 0\,\mu$m) and for two different disk rotational frequencies of about $f_{\mathrm{rot}} = 15\,$Hz and $f_{\mathrm{rot}} = 33\,$Hz. From these figures it appear that the 3rd order designed controller K_{RS3} provides better level of periodic disturbance rejection than the actual implemented (standard) controller K_{act}, for these rotational frequencies.

These results also point out that the obtained improvements are still influenced by disk rotational frequency f_{rot} and its harmonics.

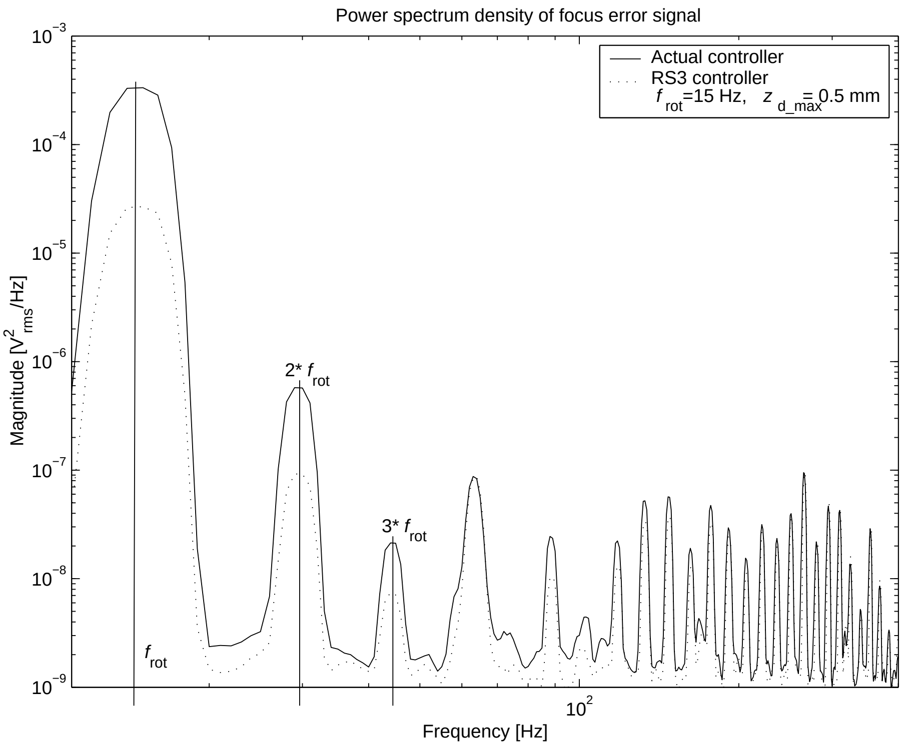

Figure 4.3: The measured power spectrum density of the focus error signal e_F for the 3rd order designed controller K_RS3 and actual implemented controller K_act. The test disk has very small disk eccentricity $x_\mathrm{d\_max}$, but with high disk vertical deviation at the disk outer edge $z_\mathrm{d\_max} = 0.5\,\mathrm{mm}$, $f_\mathrm{rot} = 15\,\mathrm{Hz}$.

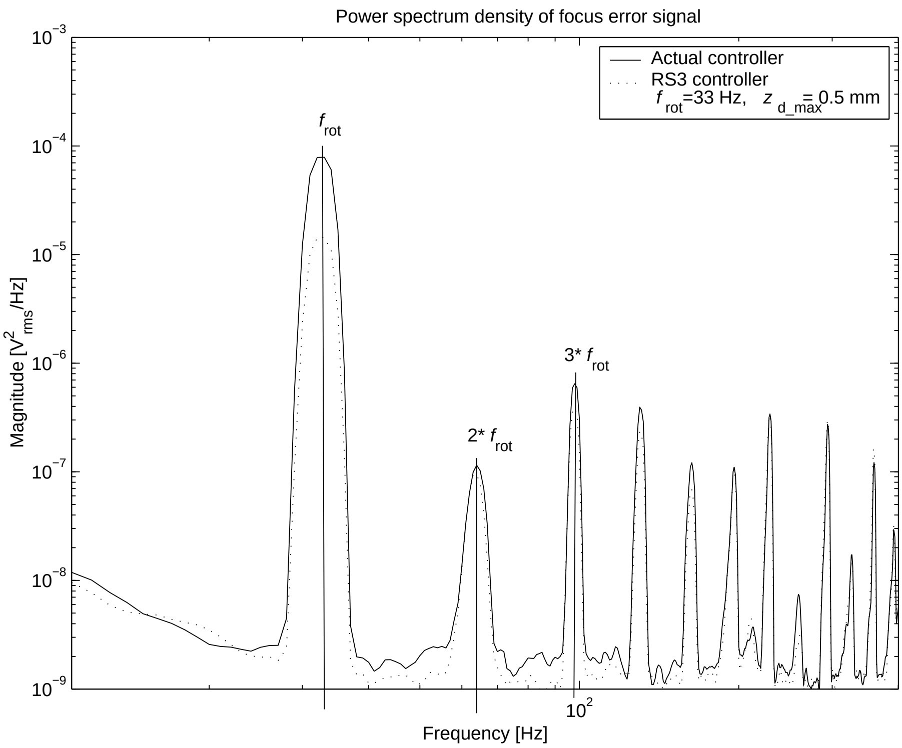

Figure 4.4: The same as in fig. 4.3 but for $f_\mathrm{rot} = 33\,\mathrm{Hz}$.

5 CONCLUSIONS

Part I - Identification

- Several novel recursive parameter adaptation algorithms have been presented to identify various state-space discrete-time plant models with a given structure (parametrization). They belong to the class of output-error algorithms, and can be interpreted as a recursive pseudolinear regression. The RLS-2 algorithm incorporates a numerical optimization component represented by computation of a matrix pseudoinverse. On the other hand, the IGM-1, IGM-2 and GM-2 algorithms, which all use the gradient technique to minimize an one-step quadratic criterion, have the advantage of a closed analytic form.
- In general, the RLS-2 algorithm possesses the fastest convergence. The strength of the IGM-1 and IGM-2 algorithms lies in robustness, *i.e.* in their insensitivity to the realization of the disturbance noise and to various initial settings (initial parameter vector estimate, adaptation gain settings). One can conclude from the verification that RLS-2 algorithm offers the best properties from these algorithms.
- Sufficient conditions for stability in a deterministic environment and convergence in a stochastic environment are always related to the actual state-space model and its parametrization. Only the particular case of parametrization, where the transformation function is linear towards the parameter vector, has been analyzed in more detail. The properties of the algorithms have been shown for this parametrization and they are related to a positive real condition on a sensitivity-type function. This condition can be relaxed by data filtering or adding a proportional adaptation. It can be concluded that the canonical state-space representation is the best theoretical case.
- It has been shown that developed algorithms give on one hand a bias distribution which is not influenced by noise and, on the other hand, contain an implicit frequency weighting filter which is matched with a robust performance control criterion. These properties make these algorithms a suitable tool for control relevant identification.

Part II - DVD player: Modelling and Control

- A new approach has been developed to model the DVD players properties. An analytical and a numerical model of the focus error signal generation have been developed, based on the astigmatic method and the opto-geometrical analysis.
- To estimate the unknown model parameters, a curve fitting method is applied, using measured data from an industrial DVD-video player.
- The performance of the analytical model is inferior in comparison with the numerical model but this model is important because it is a *complete analytical model*. This advantage is most important from the identification point of view

where the non-measurable parameters (or hardly measurable ones) could be estimated much faster than in the case of the numerical model. Then, the analytical model is more useful in the focus closed loop of the DVD player. Moreover, any analytical model of the S-curve $e_{\mathrm{F}}(\Delta z)$ significantly saves simulation time.

- Comparison with the real data, acquired from an industrial DVD-video player during start-up procedure, illustrates the quality of the analytical model.
- A new designed control system based on a combined pole placement/sensitivity function shaping methodology is proposed for construction of a low-complexity controller, being able to achieve an enhanced focus/radial following performance and repetitive disturbance rejection. Controller order reduction is performed to allow its practical implementation.
- An uncertainty model set, based on a parametric description, is considered to analyze how the variations of the plant physical parameters influence performance and robustness of the achieved solution. A simple robustness analysis demonstrates that the designed closed-loops remain stable for large uncertainty of the physical actuator parameters and the performance specifications are met.
- Final comparison of the existing and designed controllers illustrates that new controllers provide better system performance and robustness than the existing controllers. It is clear that we have obtained the limit of the controller order reduction for the imposed performance and implementation constraints.

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ABSTRACT

This thesis concerns two main parts: The first part deals with the algorithms for identification of state-space systems with a given structure in closed-loop and the second part treats the DVD modelling and control.

- The recursive algorithms for identification of a plant state-space model operating in closed-loop are presented. It is shown that these algorithms give an unbiased estimation of the plant model parameters in the presence of noise when the plant model is in the model set and the partial parametrization is applied. The bias distribution analysis shows that the inherent filtering effect in this algorithm makes it suitable tool for control relevant identification. The results are also extended for closed-loop identification of MIMO systems.

- An analytical and a numerical photodetector models are developed. The influence of model parameters on the focus error signal is discussed. Identified model quality is illustrated by a comparison with the real focus error signal. These models will be used to improve a future control performance.

  A methodology useful for designing focus and radial control loops that may be used for a large quantity of future DVD players is presented. This methodology is based on a pole placement with sensitivity function calibration which allows the improved performances and robustness compared to the actual methods used in industry, without the increased complexity of controllers.

ABSTRAKT

Tato disertační práce obsahuje dvě hlavní části: První část se zabývá identifikačními algoritmy v uzavřené smyčce pro systémy popsané stavovou reprezentací s danou strukturou. Druhá část pojednává modelování a řízení DVD přehrávačů.

- V práci jsou prezentovány rekurzivní algoritmy pro identifikaci modelů soustav, popsaných stavovou reprezentací, v uzavřené smyčce. Je ukázáno, že tyto algoritmy poskytují přesný odhad parametrů modelu soustavy v přítomnosti šumu, pokud se model soustavy nachází v třídě identifikovaných modelů a je použita patřičná parametrizace transformační funkce. Analýza rozložení chyby odhadu ve frekvenční oblasti ukazuje, že díky filtraci zahrnuté v tomto algoritmu, je tento algoritmus vhodným nástrojem pro identifikaci modelů sloužících pro návrh regulátoru. Výsledky jsou aplikovatelné pro identifikaci MIMO systémů v uzavřené smyčce.

- Byl nalezen analytický a numerický model fotodetektoru, ke kterým je provedena analýza vlivu parametrů modelů na vznik chybového signálu pro zaostřování. Srovnání reálného chybového signálu se signálem získaného z modelů demonstruje přesnost identifikovaných modelů. Oba modely lze použít ke zlepšení kvality řídicích algoritmů.

  V práci je ukázána metoda vhodná pro návrh regulátorů v regulační smyčce pro zaostřování a pro sledování stopy na optickém disku v radiálním směru. Metodu lze také použít pro nové generace DVD přehrávačů. Metoda je založena na pricipu umístění pólů spolu s kalibrací citlivostních funkcí. Aplikací této metody bylo dosaženo zlepšení kvality a robustnosti řízení ve srovnání se současnými metodami používanými v průmyslu, bez nároku na vyšší složitost regulátoru.