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THE R-S-T DIGITAL CONTROLLER DESIGN AND APPLICATIONS

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Abstract: The two degrees of freedom R-S-T digital controller is becoming a standard for computer control in industry. The paper presents the methodology for the design of the R-S-T controller which involves identification of the plant model from data combined with a robust control design. The performances of the controller can be further enhanced by plant model identification in closed loop and re-tuning of the controller. For large parameter variations adaptation has to be considered for preserving the performances. Software packages are available for the design, implementation and commissioning of the R-S-T digital controllers. The methodology is illustrated by its application to the control of deposited zinc in hot-dip galvanizing at SOLLAC (Florange, France)

Keywords: system identification, digital control, robust control, adaptation, software tools

1. INTRODUCTION

A “good” control system has in general an important economic impact in industry. Fig.1.1 illustrates the histogram of a controlled variable of a “poor” control and for a “good” control.

If the variance of the controlled variable is high, a significant number of measurements are far from the desired value. In a large number of applications, a minimal acceptable value is imposed (ex.: the humidity of the paper, the depth of the coating, etc...) and poor quality of the control will require the choice of a higher value for the reference. As a consequence, more energy or material will be needed and the direct consequence is the increase in production costs.

If one has a “good” controller which significantly reduces the variance of the controlled variable around the reference value, this will on the one hand improve the quality and, on the other hand, will allow a reduction of the reference value. This leads in general to energy and material savings which corresponds to the reduction in production costs.

Therefore, the impact of a “good” control is:

- 1) Improvement of the quality of the products
- 2) Energy and material savings.

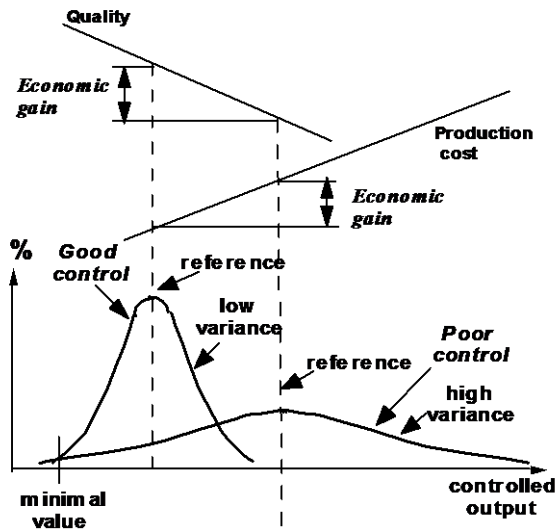


Fig. 1.1. Histograms for good and poor control

However it is important that the investment return for control improvements can be clearly evaluated in order to justify the investment.

The question is : How to improve the investment return for high performance control systems ?

The answers to this question are :

- 1) Reduction of the design cost.
- 2) Reduction of the implementation cost (including commissioning).
- 3) Achievement of the desired performances.

Therefore, it is necessary to develop an efficient design and implementation methodology. This development has to be considered in the context of computer control which is now generalized in industry. All the advantages and peculiarities of using computers for control have to be taken into account. Among these aspects, those of system identification and the introduction of a standard form for a digital controller (the R-S-T controller) play a crucial role.

Fig.1.2 summarizes the basic principles for control system design. In order to design and to tune a good controller one needs :

- 1) To specify the desired control loop performances.
- 2) To know the dynamic model of the plant to be controlled (this can be obtained from real data by identification).
- 3) To possess a suitable controller design methodology compatible with desired performances and the corresponding plant model.
- 4) To have a procedure for controller validation and on site re-tuning
- 5) To have appropriate software packages with real-time capabilities for data acquisition, system identification, control design and on site commissioning.

This paper will present such a methodology for the design and application of the R-S-T digital controllers. The methodology will be illustrated by its application to the control of the deposited zinc in hot-dip galvanizing at SOLLAC (Florange, France).

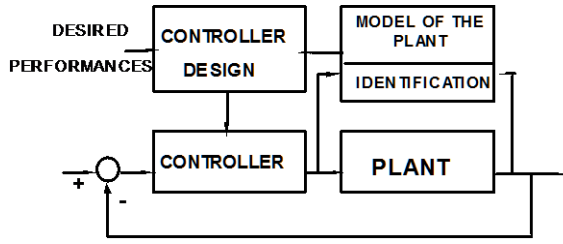


Fig. 1.2. Principle of controller design.

2. IDENTIFICATION OF DISCRETE-TIME MODELS FOR INDUSTRIAL PROCESSES

Fig.2.1 illustrates the appropriate setting for a computer control system. The set D.A.C.+ plant + A.D.C. is interpreted as a discretized system whose control input is the sequence $\{u(t)\}$ generated by the computer, the output being the sequence $\{y(t)\}$ resulting from the A/D conversion of the system output $y(\tau)$. The discretized plant is characterized by a discrete-time model which should be identified.

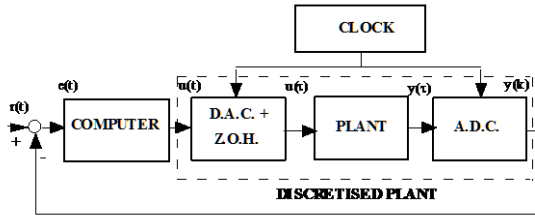


Fig. 2.1. Computer control system.

D.A.C.: Digital to Analog Converter

Z.O.H.: Zero Order Hold

A.D.C.: Analog to Digital Converter

Note that the sampling frequency is selected in accordance with the bandwidth of the continuous-time plant and more specifically in accordance with the desired bandwidth of the closed loop. The basic rule is :

$$f_s = (6 \text{ to } 25) f_B^{CL}$$

where f_s is the sampling frequency and f_B^{CL} is the desired bandwidth of the closed loop.

The discrete-time model of the plant to be controlled is described in the time-domain by:

$$y(t) = -\sum_{i=1}^{n_A} a_i y(t-i) + \sum_{i=1}^{n_B} b_i u(t-d-i)$$

where d is the integer number of sampling periods contained in the time-delay of the plant, t is the normalized discrete time (0, 1, 2, 3,...) and it corresponds to the discrete time divided by the sampling periods T_s . The discrete time model of the plant to be controlled can alternatively be represented by its pulse transfer operator $H(q^{-1})$:

$$y(t) = H(q^{-1}) u(t)$$

$H(q^{-1})$ is defined by:

$$G(q^{-1}) = \frac{q^d B(q^{-1})}{A(q^{-1})} = \frac{q^{d-1} B^*(q^{-1})}{A(q^{-1})}$$

where q^{-1} is the backward shift operator ($q^{-1}y(t) = y(t-1)$) and

$$A(q^{-1}) = 1 + a_1 q^{-1} + \dots + a_{n_A} q^{-n_A}$$

$$B(q^{-1}) = b_1 q^{-1} + \dots + b_{n_B} q^{-n_B} = q^{-1} B^*(q^{-1})$$

For models with constant parameters replacing q by z (the complex variable) in the expression of the pulse transfer operator gives the pulse transfer function.

The principle of the identification of the discrete-time models is illustrated in fig. 2.2.

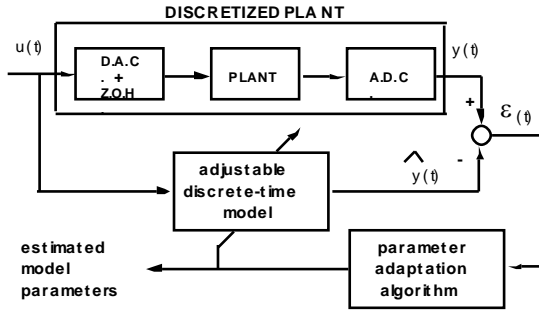


Fig. 2.2. Parameter estimation of discrete-time models.

A discrete-time model with adjustable parameters is implemented on the computer. The error between the system output at instant t , $y(t)$ and the output predicted by the model $\hat{y}(t)$ (known as the prediction error) is used by a parameter adaptation algorithm which, at each sampling instant, will modify the model parameters in order to minimize this error. The input is in general a very low-level pseudo-random binary sequence generated by the computer (sequence of rectangular pulses with randomly variable duration). Once the model is obtained, an objective validation can be made by carrying out statistical tests on the prediction error $\varepsilon(t)$ and the predicted output $\hat{y}(t)$. The validation test enables the best algorithm for the estimation of the parameters.

This approach provides much more accurate models than the methods based on step response or frequency response. In addition it requires an input signal of much lower magnitude than those used for step or frequency response.

The identification methodology includes four steps:

- 1) Input-output data acquisition around an operating point using as input in general a centered pseudo-random binary sequence (PRBS) of small magnitude,
- 2) Estimation (choice) of the model complexity (structure),
- 3) Estimation of the model parameters,
- 4) Validation of the identified model (structure and values of the parameters).

One of the important facts to be emphasized is that the plant measurements are generally noisy. Unfortunately, no unique parameter estimation method exists which may be used successfully for all the types of disturbances such that the estimated parameters are always unbiased.

Therefore, a good identification of a plant model requires in general the use of an interactive system featuring various parameter estimation methods and corresponding validation techniques.

For a more detailed discussion see (Landau, 1990), (Ljung, 1987).

3. THE R-S-T DIGITAL CONTROLLER

The canonical structure of the R-S-T digital controller is represented in Fig. 3.1.

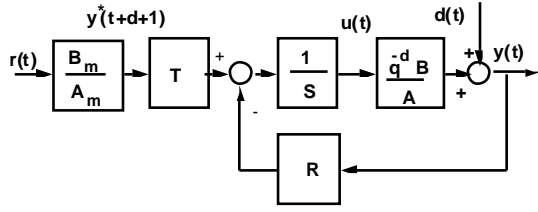


Fig. 3.1. The R-S-T canonical structure of a digital controller.

This structure has two degrees of freedom, i.e., the digital filters R and S are designed in order to achieve the desired regulation performances and the digital filter T is designed afterwards in order to achieve the desired tracking performances. This structure allows achievement of different performances in tracking and regulation.

The case of a controller operating on the regulation error (which does not allow independent specification of tracking and regulation performances) corresponds to $T = R$. Digital PID can also be represented under this form leading to particular choices of R, S and T.

The equation of the R-S-T canonical controller is given by :

$$S(q^{-1}) u(t) + R(q^{-1}) y(t) = T(q^{-1}) y^*(t+d+1)$$

where $u(t)$ and $y(t)$ are the input and output of the plant and $y^*(t+d+1)$ is the desired tracking trajectory which is either generated by a tracking reference model (B_m/A_m) or stored in the computer memory.

The polynomials $R(q^{-1})$, $S(q^{-1})$, $T(q^{-1})$ have the form:

$$\begin{aligned} R(q^{-1}) &= r_0 + r_1 q^{-1} \dots + r_{n_R} q^{-n_R} \\ S(q^{-1}) &= s_0 + s_1 q^{-1} \dots + s_{n_S} q^{-n_S} \quad (\text{often } s_0 = 1) \\ T(q^{-1}) &= t_0 + t_1 q^{-1} \dots + t_{n_T} q^{-n_T} \end{aligned}$$

The corresponding time domain expression of the control law is given by ($s_0 = 1$):

$$u(t) = - \sum_{i=1}^{n_S} s_i u(t-i) - \sum_{i=0}^{n_R} r_i y(t-i) + \sum_{i=0}^{n_T} t_i y^*(t+d+1-i)$$

The closed-loop control transfer operator (between $r(t)$ and $y(t)$) is given by :

$$\begin{aligned} H_{CL}(q^{-1}) &= \frac{q^{-d} T(q^{-1}) B(q^{-1})}{A(q^{-1}) S(q^{-1}) + q^{-d} B(q^{-1}) R(q^{-1})} \\ &= \frac{q^{-d} T(q^{-1}) B(q^{-1})}{P(q^{-1})} \end{aligned}$$

and the behaviour with respect to an output disturbance is given by the output sensitivity function :

$$\begin{aligned} S_{yp}(q^{-1}) &= \frac{A(q^{-1}) S(q^{-1})}{A(q^{-1}) S(q^{-1}) + q^{-d} B(q^{-1}) R(q^{-1})} \\ &= \frac{A(q^{-1}) S(q^{-1})}{P(q^{-1})} \end{aligned}$$

where $P(q^{-1})$ defines the desired closed loop poles (regulation behaviour).

The input sensitivity function which reflects the effects of an output disturbance upon the plant input is given by:

$$\text{Sup}(q^{-1}) = - \frac{A(q^{-1}) R(q^{-1})}{P(q^{-1})}$$

In general, the desired closed loop poles are specified under the form :

$$P(q^{-1}) = P_D(q^{-1}) P_F(q^{-1})$$

where $P_D(q^{-1})$ specifies the desired dominant poles of the closed loop and $P_F(q^{-1})$ specifies the auxiliary poles of the closed loop.

Once the closed loop poles have been defined, solving the equation :

$$P(q^{-1}) = A(q^{-1}) S(q^{-1}) + q^{-d} B(q^{-1}) R(q^{-1})$$

allows determination of $S(q^{-1})$ and $R(q^{-1})$ which will assure the desired closed loop poles.

Let the degrees of polynomials $A(q^{-1})$ and $B(q^{-1})$ be defined by:

$$n_A = \deg A(q^{-1}) ; \quad n_B = \deg B(q^{-1})$$

then, the above equation has a unique solution (assuming that $A(q^{-1})$ and $B(q^{-1})$ do not have common factors) for:

$$n_P = \deg P(q^{-1}) \leq n_A + n_B + d - 1$$

$$n_S = \deg S(q^{-1}) = n_B + d - 1$$

$$n_R = \deg R(q^{-1}) = n_A - 1$$

in which :

$$S(q^{-1}) = 1 + s_1 q^{-1} + \dots + s_{n_S} q^{-n_S} = 1 + q^{-1} S^*(q^{-1})$$

$$R(q^{-1}) = r_0 + r_1 q^{-1} + \dots + r_{n_R} q^{-n_R}$$

However, in general the polynomials $R(q^{-1})$ and $S(q^{-1})$ of the controller may contain for various reasons some prespecified fixed parts. For this reason it is convenient to factorize the polynomials $R(q^{-1})$ and $S(q^{-1})$ as follows :

$$R(q^{-1}) = H_R(q^{-1}) R'(q^{-1})$$

$$S(q^{-1}) = H_S(q^{-1}) S'(q^{-1})$$

where $H_R(q^{-1})$ and $H_S(q^{-1})$ are prespecified polynomials. These polynomials are defined by the performance specifications (ex : integrator in the controller) and by robustness considerations. For example, the introduction of an integrator in the controller requires to take $H_S(q^{-1}) = 1 - q^{-1}$.

Many control strategies can be applied to the design of the R-S-T controller by an appropriate reformulation. See for example (Landau, 1990), (Astrom & Wittenmark, 1990), (Landau *et al.*, 1997), as well as the special issue of *European Journal of Control* vol.1, n°2, 1995, dedicated to a robust control benchmark.

However, the pole placement can be considered as the basic design technique and most of the various designs can be related to it.

Pole placement

This control strategy can be used for plant models of any order with or without time delay featuring stable or unstable zeros. The only assumption is that the polynomials $A(q^{-1})$ and $B(q^{-1})$ characterizing the plant model do not have common factors.

The controller polynomials $S(q^{-1}) = H_S(q^{-1}) S'(q^{-1})$ and $R(q^{-1}) = H_R(q^{-1}) R'(q^{-1})$ are obtained by solving the equation :

$$P(q^{-1}) = A(q^{-1}) H_S(q^{-1}) S'(q^{-1}) + q^{-d-1} B^*(q^{-1}) H_R(q^{-1}) R'(q^{-1})$$

where $P(q^{-1})$ defines the desired closed loop poles.

The $T(q^{-1})$ polynomial is chosen as :

$$T(q^{-1}) = P(q^{-1}) / B(1)$$

($B(1) = B(q^{-1})$ for $q = 1$) and the tracking behaviour is described by the equation :

$$y(t) = q^{d+1} \frac{B^*(q^{-1})}{B(1)} y^*(t+d+1).$$

In other words one follows the desired trajectory filtered through the plant zeros. For more details see (Landau, 1990), (Landau, 1993).

Note that unstable discrete-time zeros occur when a fractional delay larger than half of the sampling period is present or when a high sampling frequency is used for continuous-time models having a difference of degree between denominator and numerator greater or equal to 2 (Åström & Wittenmark, 1990), (Landau, 1990). Both phenomena lead to the conclusion that the sampling frequency has to be chosen as low as possible but in accordance with the desired closed-loop performances.

Relationship with other control strategies

For plant models with stable zeros, one can use the strategy called « tracking and regulation with independent objectives », which can be viewed as a particular case of pole placement strategy, where the desired closed loop poles contains the zeros of the plant, i.e.,

$$P(q^{-1}) = P_D(q^{-1}) \cdot B^*(q^{-1}) \cdot P_F(q^{-1})$$

« Internal model control » (Morari, 1989) corresponds to pole placement strategy where the desired closed loop poles contains the poles of the plant model, i.e.,

$$P(q^{-1}) = A(q^{-1}) \cdot P_F(q^{-1})$$

Digital PID can be designed using pole placement. In this case, the orders of the plant model are limited to $n_A \leq 2$, $n_B \leq 2$, $d=0$.

Predictive control strategies and LQ control strategy using an appropriate formulation of the criterion to be minimized lead to a R-S-T controller and can be viewed as an approximation of the pole placement in the sense of a certain quadratic criterion (Landau *et al.*, 1997).

4. ROBUSTNESS

Four indicators are generally used to express the robustness of the design in terms of the minimal distance with respect to the critical point $[-1, j0]$ in the Nyquist plane. These indicators are the gain margin (ΔG), the phase mar-

gain ($\Delta\phi$), the modulus margin (ΔM) and the delay margin ($\Delta\tau$). The modulus margin and the delay margin are the most interesting for applications. Fig. 4.1 illustrates the modulus, gain and phase margins.

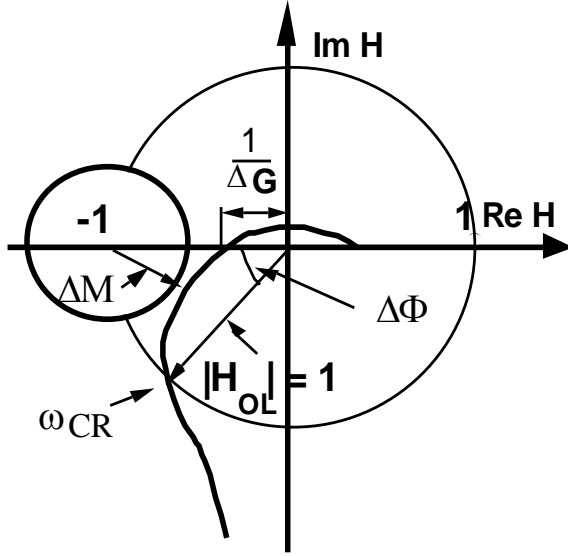


Fig. 4.1. Modulus, gain and phase margins.

The modulus margin is the minimal distance between the critical point $[-1, j0]$ and the Nyquist plot of the open loop transfer function :

$$H_{OL}(z^{-1}) = z^{-d} B(z^{-1}) R(z^{-1}) / A(z^{-1}) S(z^{-1}).$$

The modulus margin ΔM is defined as the radius of a circle centered in $[-1, j0]$ and tangent to the Nyquist plot of $H_{OL}(e^{-jw})$ (see fig. 4.1).

It results that :

$$\begin{aligned} \Delta M &= |1 + H_{OL}(e^{-jw})|_{\min}. \\ &= |S_{yp}^{-1}(e^{-jw})|_{\min}. \\ &= (|S_{yp}(e^{-jw})|_{\max})^{-1} \end{aligned}$$

In other words, the modulus margin corresponds to the inverse of the H_{∞} norm of the output sensitivity function.

Minimization of the H_{∞} norm of $S_{yp}(e^{-jw})$ will maximize the modulus margin.

To obtain the modulus margin, it is therefore sufficient to simply plot the frequency characteristics of the modulus (gain) of the output sensitivity function in dB. In this case :

$$\begin{aligned} \Delta M \text{ dB} &= (|S_{yp}(e^{-jw})|_{\max})^{-1} \text{ dB} \\ &= -|S_{yp}(e^{-jw})|_{\max} \text{ dB} \end{aligned}$$

Note that a given value of the modulus margin will guarantee certain phase and gain margins while the converse is not true (systems with good phase and gain margins can pass very close to the critical point).

The delay margin is the maximal additional delay which will be tolerated in the open loop system without causing the instability of the closed loop system. If the Nyquist plot of the open loop system intersect the unit circle at several cross-over frequencies ω_{ci}^i , characterized by the corresponding phase margins $\Delta\phi_i$, the delay margin of the system is defined as :

$$\Delta \tau = \min_i \frac{\Delta \phi_i}{\omega_{cr}^i}$$

Typical values of these robustness indicators are :

- modulus margin : $\Delta M \geq 0.5$ (-6 dB)
- delay margin : $\Delta \tau \geq T_S$ (sampling period)
- gain margin : $\Delta G \geq 2$ (6 dB)
- phase margin : $30^\circ \leq \Delta \phi \leq 60^\circ$

Note that $\Delta M \geq 0.5$ implies $\Delta G \geq 2$ and $\Delta \phi > 29^\circ$ (the converse is not necessarily true).

Sensitivity functions are related to the robust stability of the closed loop with respect to plant model uncertainties, see (Doyle *et al.*, 1992). Bounds on the magnitude of the frequency dependent model uncertainties convert on upper constraints upon the modulus of the various sensitivity functions. These constraints, as well as the ones on the modulus margin and delay margin, can be translated into desired templates for the sensitivity functions. Conversely, imposed constraints on the sensitivity functions can be translated into tolerated model uncertainties.

Typical templates for S_{yp} and S_{up} are shown in Fig. 4.2. The upper and the lower bounds on the output sensitivity function S_{yp} in the high frequencies region come from the translation of the delay margin constraints in the frequency domain (Landau, 1995).

The input sensitivity function reflects both tolerance with respect to additive uncertainties and the activity of the input in the presence of disturbances (high values in certain frequency regions indicate low model uncertainty tolerance and important stress on the actuator).

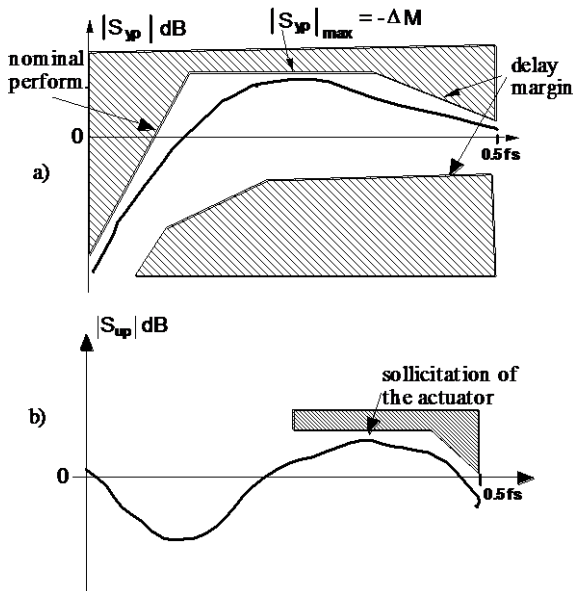


Fig. 4.2 Templates for the sensitivity functions
a) output sensitivity function
b) input sensitivity function

A design methodology which combines pole placement with the shaping of the sensitivity function has been developed in order to both assure performances and robustness. See (Landau, 1993), (Landau *et al.*, 1996).

5. IDENTIFICATION IN CLOSED LOOP

In a number of practical situations, it may be not possible to operate the plant in open loop in order to do system identification. Such situations are encountered for example when the plant contain an integrator or when important drifts of the operating point may occurs during input/output data acquisition. In a number of other situations, a controller exists already and for various reasons it can not be disconnected. Therefore, techniques for plant identification in closed loop operation should be used.

Significant progress in this area has been made in the last few years. New algorithms dedicated to the identification in closed-loop have been developed. For a detailed presentation see (Van den Hof & Schrama, 1995), (Landau & Karimi, 1997), (Landau *et al*, 1997).

A basic scheme for plant model identification in closed-loop is shown in Fig. 5.1 (Landau & Karimi, 1997). The upper part represents the true closed loop system and the lower part corresponds to an adjustable predictor for the closed loop re-parametrized in terms of a known fixed controller and of an adjustable plant model. The error between the system output and the closed-loop predictor output (called closed-loop output error) is used by a parameter adaptation algorithm which will drive the parameters of the estimated plant model in order to minimize the error between the two closed-loop systems. In other terms, the model obtained in closed-loop will allow to better predict the behaviour of the closed loop system.

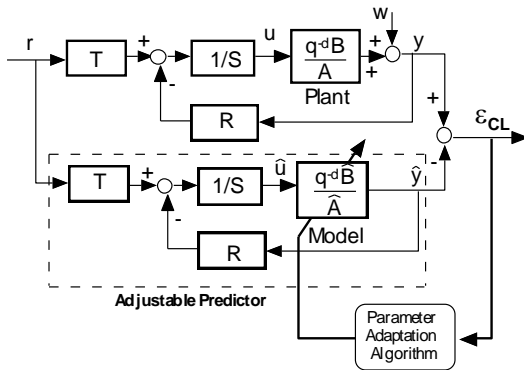


Fig. 5.1 Plant model identification in closed loop operation.

6. CONTROLLER VALIDATION AND ON-LINE RETUNING

As indicated in Section 1, validation of the designed controller is a key issues for assessing the effective performances of the closed loop system. Identification of the closed loop will allow to compare the achieved performances with the designed ones by comparing the achieved and the designed closed loop poles. It will also allow to assess the robustness of the control schemes by comparing the designed and the achieved sensitivity function in the frequency domain.

However, with the same data acquired in closed loop, one can also identify a new plant model. The new model identified in closed loop is then used for re-tuning of the controller.

This procedure is used for two purposes:

- improvement of a previous design
- controller maintenance

Examples can be found in (Zhang *et al*, 1995), (Langer & Landau, 1996), (Landau *et al*, 1997).

7. ADAPTATION

When “system identification” plus “robust control design” does not allow to obtain a single linear controller giving acceptable performances for the all range of operating points, because of the too wider variations of the dynamic characteristics of the plant, one has to consider the “adaptation” of the controller.

The term “adaptation” (adaptive control) refers to a set of techniques for the automatic tuning of the controller in real time in order to maintain the desired performances when the plant parameters vary.

One can distinguish two basic adaptive control techniques :

- 1) “closed loop” adaptive control (Fig. 7.1)
- 2) “open loop” adaptive control (Fig. 7.2)

The “closed loop” adaptive control system combines in general a real-time identification algorithm with the computation of the controller in real time based on the current estimation of the plant model and the desired performances.

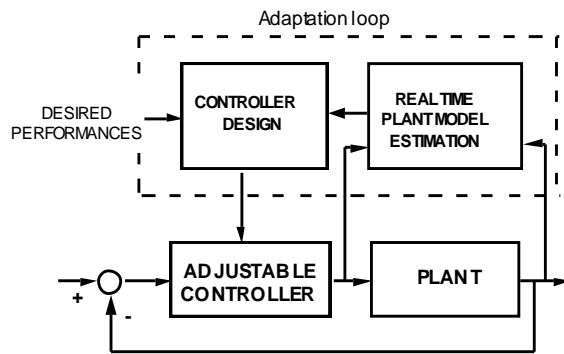


Fig.7.1 “Closed loop” adaptive control.

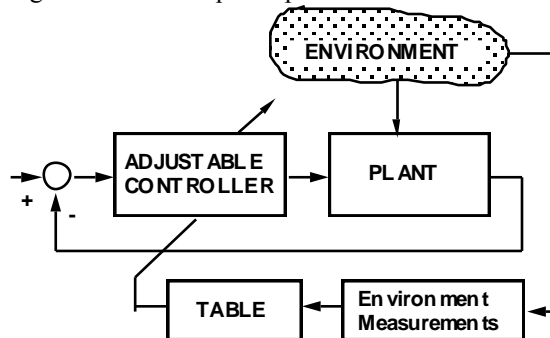


Fig. 7.2. “Open loop” adaptation.

However, in a number of applications, the characteristics of the plant dynamic model depend upon a set of measured variables which define an operating point. In this case, one can use an “open loop” adaptive control (Fig. 7.2). The range of operating points is divided in a number of operation intervals. For each interval, a relevant operating point is selected and a corresponding controller is designed based on an identified model. This controller assures the desired performances for all the operating points located in the interval. The corresponding controllers are stored in a table. When the plant will operate in a certain point, the corresponding values of the controller parameters will be used according to the table.

8. HOT - DIP GALVANIZING AT SOLLAC (FLORANGE)

The objective of the galvanizing line is to obtain galvanized steel with formability, surface quality and weldability equivalent to uncoated cold rolled steel. The variety of products is very large in terms of deposited zinc thickness and steel strip thickness. The deposited zinc may vary between 50 to 350 g/m² (each side) and the strip speed may vary from 30 to 180 m/mn.

The most important part of the process is the hot-dip galvanizing. The principle of the hot-dip galvanizing is illustrated in Fig. 8.1. Preheated steel strip is passed through a bath of liquid zinc and then rises vertically out of the bath through the stripping "air knives" which remove the excess zinc. The remaining zinc on the strip surface solidifies before it reaches the rollers, which guide the finished product. The measurement of the deposited zinc can be made only on the cooled finished strip. The effect of air knives depends on the air pressure, the distance between the air knives and the strip, and the speed of the strip. Nonlinear static models have been developed for computing the appropriate pressure, distance and speed for a given value of the desired deposited zinc.

The objective of the control is to assure a good uniformity of the deposited zinc whilst guaranteeing a minimum value of the deposited zinc per unit area. Tight control (i.e., small variance of the controlled variable) will allow a more uniform coating and a reduction of the average quantity of deposited zinc per unit area. As a consequence, in addition to quality improvement, a tight control of the deposited zinc per unit area has an important commercial impact since the average consumption for a modern galvanizing line is of the order of 40 tons per day (price \approx 1 500 USD/ton).

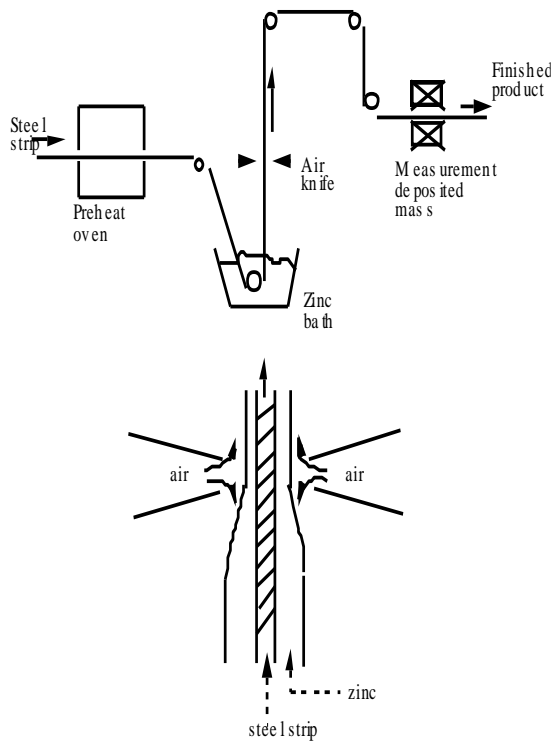


Fig. 8.1. Process description.

9. MODEL OF THE PROCESS

For analysis of the process the model originally proposed by Harvey and Carton (1974) and completed by Jacobs (1991) can be used :

$$m = KD\sqrt{\frac{V}{P}} + \zeta_m$$

where m is the deposited mass per unit area, K is a constant of proportionality, D is the distance between the air knives and the strip, P is the air pressure and V is the strip speed. ζ_m accounts for unpredictable effects and/or modelling errors. At SOLLAC Ste Agathe, the control variable is the air pressure.

A linearized model around an operating point (P_0 , V_0 , D_0) can be obtained using a standard Taylor series expansion for variations of pressure (ΔP), speed (ΔV) and distance (ΔD). It has the form :

$$m = KD_0 \sqrt{\frac{V_0}{P_0}} + \alpha \Delta D + \beta \Delta V - \mu \Delta P + \xi_m \quad ; \quad \alpha, \beta, \mu > 0$$

It can be seen that using the pressure as the control variable one can compensate for the disturbances created by variations of distance and speed as well as by the term ξ_m .

The pressure in the air knives is regulated through a pressure loop, which can be approximated by a first order system. The delay of the process will depend linearly on the speed. Therefore, a continuous-time linear dynamic model relating variations of the pressure to variations of the deposited mass, of the form :

$$H(s) = \frac{G e^{-s\tau}}{1 + sT} \quad ; \quad \tau = \frac{L}{V}$$

can be considered, where L is the distance between the air knives and the transducers and V is the strip speed. When discretizing this model, the major difficulty comes from the variable time-delay. In order to obtain a controller with a fixed number of parameters, the delay of the discrete-time model should remain constant. Therefore, the sampling period is tied to the strip speed using the formula :

$$T_s = \frac{\frac{L}{V} + \delta}{d} \quad ; \quad (d = \text{integer})$$

where d is an additional small time-delay due to implementation and d is the discrete-time delay (integer). The corresponding linearized discrete-time model will be of the form :

$$H(q^{-1}) = \frac{q^d (b_1 q^{-1})}{1 + a_1 q^{-1}}$$

The fractional delay (which corresponds to the presence of an additional term $b_2 q^{-2}$) is negligible because of the way the sampling period T_s is selected and this was confirmed by the model identification procedure. However, the parameters of the model, given above, will depend on the distance D and on the speed V .

10. IDENTIFICATION OF THE DISCRETE-TIME PLANT MODEL

The process is formed by the air pressure control loop and the coating process. The control input to the process is the reference of the air pressure control loop, the output of the process is the measured deposited mass per unit area (see Fig.10.1).

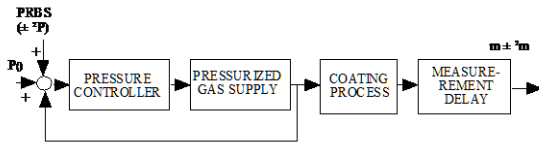


Fig. 10.1. Process block-diagram.

The PC used for data acquisition identification and control is connected to the process through an industrial network. The identification has been done with the coating process operating in open loop.

The sampling frequency has been chosen in each operation point, in order to have the discrete-time delay $d = 7$. The data acquisition is illustrated in Fig.10.2. First an analog anti-aliasing filter is used before a high frequency

sampling is done (a multiple of the desired sampling frequency). A digital anti-aliasing filter is inserted between the two samplers.

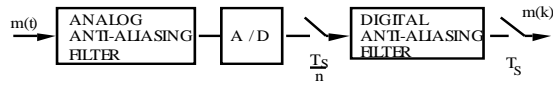


Fig.10.2. Data acquisition.

The input used was a P.R.B.S. (Pseudo Random Binary Sequence) of a magnitude of $\pm 4\%$ with respect to the static pressure (P_0). The P.R.B.S. was generated by a shift register with $N = 5$ cells and a clock frequency equal to half of the sampling frequency (length of the sequence: 64). 100 to 160 (average: 128) measurements have been used for the various identifications made in different regions of operation. The choice made for the P.R.B.S. allowed at least one full sequence to be sent for each experiment and yielded the largest pulse width ($10 T_s$) comparable with the rise time of the process. As both two sides of the steel strip have to be galvanised, and because the position and physical realisation of the two actuators are not symmetric, both "front" side and "back" side models have been identified.

As no unique identification method gives unbiased result for all types of disturbances, the following recursive identification methods have been used (Landau, 1990) :

- Recursive Least Squares
- Extended Least Squares
- Recursive Maximum Likelihood
- Instrumental Variable with Auxiliary Model
- Output Error
- Generalized Least Squares

For validation of the identified models and comparison of the models obtained with the different methods, the cross correlation between the predicted output (using an output error predictor) and the output error has been used (Landau, 1990). (For the first four methods also, the whiteness test on the prediction error has been used for validation and comparison). Systematically « output error » provided the best results for this application.

It was observed that a significant variability of the parameters occurs with the change of the operating points. This required to split the operation of the plant into several regions. However a variability of the parameters is observed even within a region of operation at a constant distance and relatively small speed variations. One of the causes is the imperfect measurement of the strip/air knives distance. This variability will require a robust control design.

11. CONTROLLER DESIGN AND ADAPTATION

The "tracking and regulation with independent objectives" (which in this case is equivalent with the poles placement since the model does not have finite zeros) has been used.

The robust control design using an identified model and based on the shaping of the sensitivity function allowed to obtain a modulus margin $> -6\text{dB}$, and a delay margin $> 2T_s$. These robustness margins assure satisfactory performances in a region of operation despite of the variability of the model.

In order to assure satisfactory performances for all regions of operation an "open-loop adaptation" technique has been considered. The open-loop adaptation is made with respect to :

- steel strip speed,
- distance between the air knives and the steel strip.

The strip speed directly affects the sampling period according to the relationship :

$$T_s = \frac{L + \delta}{V}$$

where δ is the equivalent time-delay of the industrial network and of the programmable controller used for pressure regulation. The speed range and the distance range have been split into 3 regions giving a total of 9 operating regions. For each of these operating regions an identification has been performed and controllers based on the identified model have been computed and stored in a table.

Anti wind-up procedures have been used for the implementation of the controller and a smooth transfer from open-loop to closed-loop operation has also been assured. For a detailed presentation of this application see (Fenot *et al*, 1993).

12. RESULTS

Fig. 12.1 shows one of the typical results obtained when one of the sides is under digital regulation and the other side is under computer aided manual control (the operator has on display a moving short-time history of the deposited zinc and applied pressure).

A reduction of the dispersion of coating is noticed when closed-loop digital control is used. This provides a better quality finished product (extremely important in the automotive industry, for example).

The average quantity of deposited zinc is reduced by 3% when closed-loop digital control is used, still guaranteeing the specifications for minimum zinc deposit. Taking into account the line production and the price of the zinc this corresponds to an annual saving over 350 000 USD. The closed-loop operation also reduces the task of the operator thereby creating better working conditions.

13. CONCLUSIONS

A methodology for the design and tuning of R-S-T digital controllers has been presented. These controllers have already been used in a significant number of industrial applications (Rolland & Landau, 1991). The economic impact of the improved performances have largely justified their use. Software packages are available for the design, implementation and commissioning of the R-S-T digital controllers.

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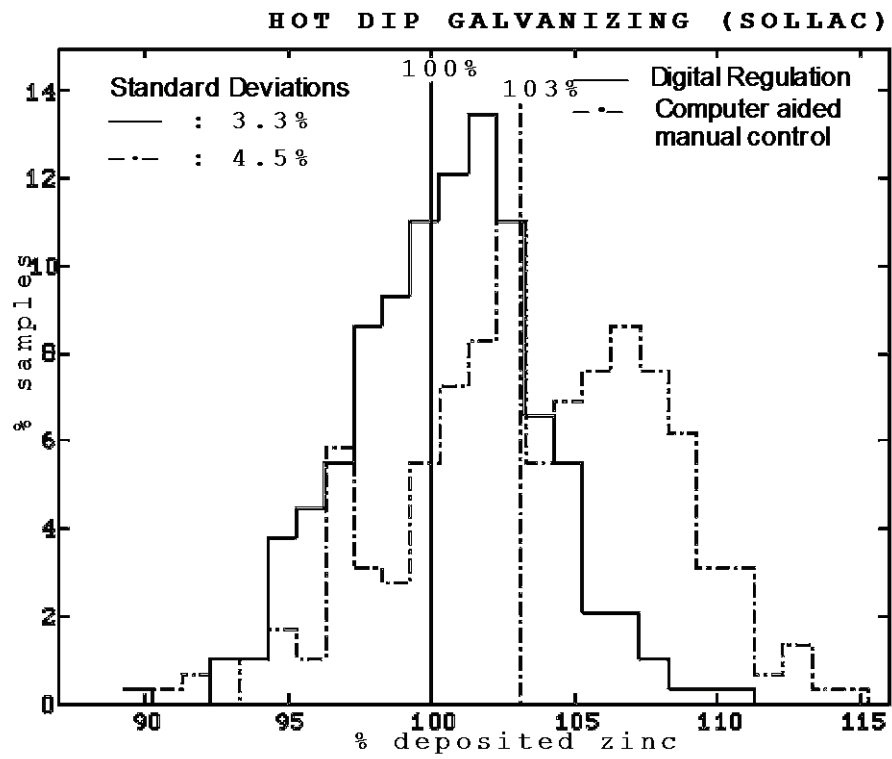


Fig. 12.1. Typical performance of the digital regulation of the deposit zinc.