New Approaches in Automation and Robotics

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Edited by
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I-Tech

Published by I-Tech Education and Publishing

I-Tech Education and Publishing
Vienna
Austria

www.i-techonline.com
Additional copies can be obtained from:
publication@ars-journal.com

First published May 2008
Printed in Croatia

A catalogue record for this book is available from the Austrian Library.
Automation and Robotics, New Approaches, Edited by Harald Aschemann
p. cm.
ISBN 978-3-902613-26-4
1. Automation and Robotics. 2. New Approaches. I. Harald Aschemann

Preface

The book at hand on "New Approaches in Automation and Robotics" offers in 22 chapters a collection of recent developments in automation, robotics as well as control theory. It is dedicated to researchers in science and industry, students, and practicing engineers, who wish to update and enhance their knowledge on modern methods and innovative applications.

The authors and editor of this book wish to motivate people, especially undergraduate students, to get involved with the interesting field of robotics and mechatronics. We hope that the ideas and concepts presented in this book are useful for your own work and could contribute to problem solving in similar applications as well. It is clear, however, that the wide area of automation and robotics can only be highlighted at several spots but not completely covered by a single book.

The editor would like to thank all the authors for their valuable contributions to this book. Special thanks to Editors in Chief of International Journal of Advanced Robotic Systems for their effort in making this book possible.

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1

A Model Reference Based 2-DOF Robust Observer-Controller Design Methodology

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1. Introduction

As it is well known, standard feedback control is based on generating the control signal u by processing the error signal, $e = r - y$, that is, the difference between the reference input and the actual output. Therefore, the input to the plant is

$$u = K(r - y) \tag{1}$$

It is well known that in such a scenario the design problem has one degree of freedom (1-DOF) which may be described in terms of the stable Youla parameter (Vidyasagar, 1985). The error signal in the 1-DOF case, see figure 1, is related to the external input r and d by means of the sensitivity function $S \doteq (1 + P_o K)^{-1}$, i.e., $e = S(r - d)$.

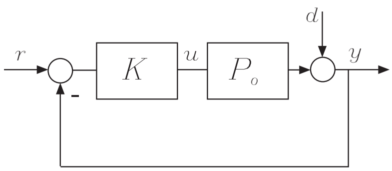

Fig. 1. Standard 1-DOF control system.

Disregarding the sign, the reference r and the disturbance d have the same effect on the error e. Therefore, if r and d vary in a similar manner the controller K can be chosen to minimize e in some sense. Otherwise, if r and d have different nature, the controller has to be chosen to provide a good trade-off between the command tracking and the disturbance rejection responses. This compromise is inherent to the nature of 1-DOF control schemes. To allow independent controller adjustments for both r and d, additional controller blocks have to be introduced into the system as in figure 2.
Two-degree-of-freedom (2-DOF) compensators are characterized by allowing a separate processing of the reference inputs and the controlled outputs and may be characterized by means of two stable Youla parameters. The 2-DOF compensators present the advantage of a complete separation between feedback and reference tracking properties (Youla & Bongiorno, 1985): the feedback properties of the controlled system are assured by a feedback

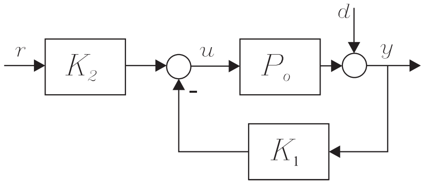

Fig. 2. Standard 2-DOF control configuration.

controller, i.e., the first degree of freedom; the reference tracking specifications are addressed by a prefilter controller, i.e., the second degree of freedom, which determines the open-loop processing of the reference commands. So, in the 2-DOF control configuration shown in figure 2 the reference r and the measurement y, enter the controller separately and are independently processed, i.e.,

$$u = K\begin{bmatrix} r \\ y \end{bmatrix} = K_2 r - K_1 y \tag{2}$$

As it is pointed out in (Vilanova & Serra, 1997), classical control approaches tend to stress the use of feedback to modify the systems' response to commands. A clear example, widely used in the literature of linear control, is the usage of reference models to specify the desired properties of the overall controlled system (Astrom & Wittenmark, 1984). What is specified through a reference model is the desired closed-loop system response. Therefore, as the system response to a command is an open-loop property and robustness properties are associated with the feedback (Safonov et al., 1981), no stability margins are guaranteed when achieving the desired closed-loop response behaviour.

A 2-DOF control configuration may be used in order to achieve a control system with both a performance specification, e.g., through a reference model, and some guaranteed stability margins. The approaches found in the literature are mainly based on optimization problems which basically represent different ways of setting the Youla parameters characterizing the controller (Vidyasagar, 1985), (Youla & Bongiorno, 1985), (Grimble, 1988), (Limebeer et al., 1993).

The approach presented in (Limebeer et al., 1993) expands the role of H_∞ optimization tools in 2-DOF system design. The 1-DOF loop-shaping design procedure (McFarlane & Glover, 1992) is extended to a 2-DOF control configuration by means of a parameterization in terms of two stable Youla parameters (Vidyasagar, 1985), (Youla & Bongiorno, 1985). A feedback controller is designed to meet robust performance requirements in a manner similar as in the 1-DOF loop-shaping design procedure and a prefilter controller is then added to the overall compensated system to force the response of the closed-loop to follow that of a specified reference model. The approach is carried out by assuming uncertainty in the normalized coprime factors of the plant (Glover & McFarlane, 1989). Such uncertainty description allows a formulation of the $\mathcal{H}_\infty$ robust stabilization problem providing explicit formulae.

A frequency domain approach to model reference control with robustness considerations was presented in (Sun et al., 1994). The design approach consists of a nominal design part plus a modelling error compensation component to mitigate errors due to uncertainty.

However, the approach inherits the restriction to minimum-phase plants from the Model Reference Adaptive Control theory in which it is based upon.
In this chapter we present a 2-DOF control configuration based on a right coprime factorization of the plant. The presented approach, similar to that in (Pedret C. et al., 2005), is not based on setting the two Youla parameters arbitrarily, with internal stability being the only restriction. Instead,

1. An observer-based feedback control scheme is designed to guarantee robust stability. This is achieved by means of solving a constrained $\mathcal{H}_\infty$ optimization using the right coprime factorization of the plant in an active way.
2. A prefilter controller is added to improve the open-loop processing of the robust closed-loop. This is done by assuming a reference model capturing the desired input-output relation and by solving a model matching problem for the prefilter controller to make the overall system response resemble as much as possible that of the reference model.

The chapter is organized as follows: section 2 introduces the Observer-Controller configuration used in this work within the framework of stabilizing control laws and the Youla parameterization for the stabilizing controllers. Section 3 reviews the generalized control framework and the concept of $\mathcal{H}_\infty$ optimization based control. Section 4 displays the proposed 2-DOF control configuration and describes the two steps in which the associated design is divided. In section 5 the suggested methodology is illustrated by a simple example. Finally, Section 6 closes the chapter summarizing its content and drawing some conclusions.

2. Stabilizing control laws and the Observer-Controller configuration

This section is devoted to introduce the reader to the celebrated Youla parameterization, mentioned throughout the introduction. This result gives all the control laws that attain closed-loop stability in terms of two stable but otherwise free parameters. In order to do so, first a basic review of the factorization framework is given and then the Observer-Controller configuration used in this chapter is presented within the aforementioned framework. The Observer-Controller configuration constitutes the basis for the control structure presented in this work.

2.1 The factorization framework

A short introduction to the so-called factorization or fractional approach is provided in this section. The central idea is to factor a transfer function of a system, not necessarily stable, as a ratio of two stable transfer functions. The factorization framework will constitute the foundations for the analysis and design in subsequent sections. The treatment in this section is fairly standard and follows (Vilanova, 1996), (Vidyasagar, 1985) or (Francis, 1987).

2.1.2 Coprime factorizations over $\mathcal{RH}_\infty$

A usual way of representing a scalar system is as a rational transfer function of the form

$$P_o(s) = \frac{n(s)}{m(s)} \tag{3}$$

where $n(s)$ and $m(s)$ are polynomials and (3) is called polynomial fraction representation of $P_o(s)$. Another way of representing $P_o(s)$ is as the product of a stable transfer function and a transfer function with stable inverse, i.e.,

$$P_o(s) = N(s)M^{-1}(s) \tag{4}$$

where $N(s), M(s) \in \mathcal{RH}_\infty$, the set of stable and proper transfer functions.

In the Single-Input Single-Output (SISO) case, it is easy to get a fractional representation in the polynomial form (3). Let $\delta(s)$ be a Hurwitz polynomial such that $\deg \delta(s) = \deg m(s)$ and set

$$N(s) = \frac{n(s)}{\delta(s)} \qquad M(s) = \frac{m(s)}{\delta(s)} \tag{5}$$

The factorizations to be used will be of a special type called Coprime Factorizations. Two polynomials $n(s)$ and $m(s)$ are said to be coprime if their greatest common divisor is 1 (no common zeros). It follows from Euclid's algorithm – see for example (Kailath, 1980) – that $n(s)$ and $m(s)$ are coprime iff there exists polynomials $x(s)$ and $y(s)$ such that the following identity is satisfied:

$$x(s)m(s) + y(s)n(s) = 1 \tag{6}$$

Note that if z is a common zero of $n(s)$ and $m(s)$ then $x(z)m(z) + y(z)n(z) = 0$ and therefore $n(s)$ and $m(s)$ are not coprime. This concept can be readily generalized to transfer functions $N(s), M(s), X(s), Y(s)$ in $\mathcal{RH}_\infty$. Two transfer functions $M(s), N(s)$ in $\mathcal{RH}_\infty$ are coprime when they do not share zeros in the right half plane. Then it is always possible to find $X(s), Y(s)$ in $\mathcal{RH}_\infty$ such that $X(s)M(s) + Y(s)N(s) = 1$.

When moving to the multivariable case, we also have to distinguish between right and left coprime factorizations since we lose the commutative property present in the SISO case.

The following definitions tackle directly the multivariable case.

Definition 1. (Bezout Identity) Two stable matrix transfer functions N_r and M_r are right coprime if and only if there exist stable matrix transfer functions X_r and Y_r such that

$$\begin{bmatrix} X_r & Y_r \end{bmatrix} \begin{bmatrix} M_r \\ N_r \end{bmatrix} = X_r M_r + Y_r N_r = I \tag{7}$$

Similarly, two stable matrix transfer functions N_l and M_l are left coprime if and only if there exist stable matrix transfer functions X_l and Y_l such that

$$\begin{bmatrix} M_l & N_l \end{bmatrix} \begin{bmatrix} X_l \\ Y_l \end{bmatrix} = M_l X_l + N_l Y_l = I \tag{8}$$

The matrix transfer functions X_r, Y_r (X_l, Y_l) belonging to $\mathcal{RH}_\infty$ are called right (left) Bezout complements.

Now let $P_o(s)$ be a proper real rational transfer function. Then,

Definition 2. A right (left) coprime factorization, abbreviated RCF (LCF), is a factorization $P_o(s) = N_r M_r^{-1}$ ($P_o(s) = M_l^{-1} N_l$), where N_r, M_r (N_l, M_l) are right (left) coprime over $\mathcal{RH}_\infty$.

With the above definitions, the following theorem arises to provide right and left coprime factorizations of a system given in terms of a state-space realization. Let us suppose that

$$P_o(s) \doteq \left[\begin{array}{c|c} A & B \\ \hline C & D \end{array}\right] \tag{9}$$

is a minimal stabilisable and detectable state-space realization of the system $P_o(s)$.

Theorem 1. Define

$$\begin{gathered}\begin{bmatrix} M_r & -Y_l \\ N_r & X_l \end{bmatrix} \doteq \left[\begin{array}{c|cc} A+BF & B & -L \\ \hline F & I & 0 \\ C+DF & -D & I \end{array}\right] \\ \begin{bmatrix} X_r & Y_r \\ -N_l & M_l \end{bmatrix} \doteq \left[\begin{array}{c|cc} A+LC & -(B+LD) & -L \\ \hline F & I & 0 \\ C & -D & I \end{array}\right]\end{gathered} \tag{10}$$

where F and L are such that $A+BF$ and $A+LC$ are stable. Then, $P_o(s) = N_r(s) M_r^{-1}(s)$ ($P_o(s) = M_l^{-1}(s) N_l(s)$) is a RCF (LCF).

Proof. The theorem is demonstrated by substituting (1.10) into equation (1.7).

Standard software packages can be used to compute appropriate F and L matrices numerically for achieving that the eigenvalues of $A+BF$ are those in the vector

$$p_F = \begin{bmatrix} p_{F_1} \cdots p_{F_n} \end{bmatrix}^T \tag{11}$$

Similarly, the eigenvalues of $A+LC$ can be allocated in accordance to the vector

$$p_L = \begin{bmatrix} p_{L_1} \cdots p_{L_n} \end{bmatrix}^T \tag{12}$$

By performing this pole placement, we are implicitly making active use of the degrees of freedom available for building coprime factorizations. Our final design of section 4 will make use of this available freedom for trying to meet all the controller specifications.

2.2 The Youla parameterization and the Observer-Controller configuration

A control law is said to be stabilizing if it provides internal stability to the overall closed-loop system, which means that we have Bounded-Input-Bounded-Output (BIBO) stability between every input-output pair of the resulting closed-loop arrangement. For instance, if we consider the general control law $u = K_2 r - K_1 y$ in figure 3a internal stability amounts to being stable all the entries in the mapping $\left(r, d_i, d_o\right) \to \left(u, y\right)$.

Let us reconsider the standard 1-DOF control law of figure 1 in which $u = K(r - y)$. For this particular case, the following theorem gives a parameterization of all the stabilizing control laws.

Theorem 2. (1-DOF Youla parameterization) For a given plant $P = N_r M_r^{-1}$, let $C_{stab}(P)$ denote the set of stabilizing 1-DOF controllers K_1, that is,

$$C_{stab}(P) \doteq \left\{K_1 : \text{the control law } u = K_1(r - y) \text{ is stabilizing}\right\}. \tag{13}$$

The set $C_{stab}(P)$ can be parameterized by

$$C_{stab}(P) = \left\{ \frac{X_r + M_r Q_y}{Y_r - N_r Q_y} : Q_y \in \mathcal{RH}_\infty \right\} \tag{14}$$

As it was pointed out in the introduction of this chapter, the standard feedback control configuration of figure 1 lacks the possibility of offering independent processing of disturbance rejection and reference tracking. So, the controller has to be designed for providing closed-loop stability and a good trade-off between the conflictive performance objectives. For achieving this independence of open-loop and closed-loop properties, we added the extra block K_2 (the prefilter) to figure 1, leading to the standard 2-DOF control scheme in figure 2. Now the control law is of the form

$$u = K_2 r - K_1 y \tag{15}$$

where K_1 and K_2 are to be chosen to provide closed-loop stability and meet the performance specifications. This control law is the most general stabilizing linear time invariant control law since it includes all the external inputs (y and r) in u.

Because of the fact that two compensator blocks are needed for expressing u according to (15), 2-DOF compensators are also referred to as two-parameter compensators. It is worth emphasizing that (15) represents the most general feedback compensation scheme and that, for example, there is no *three-parameter compensator*.

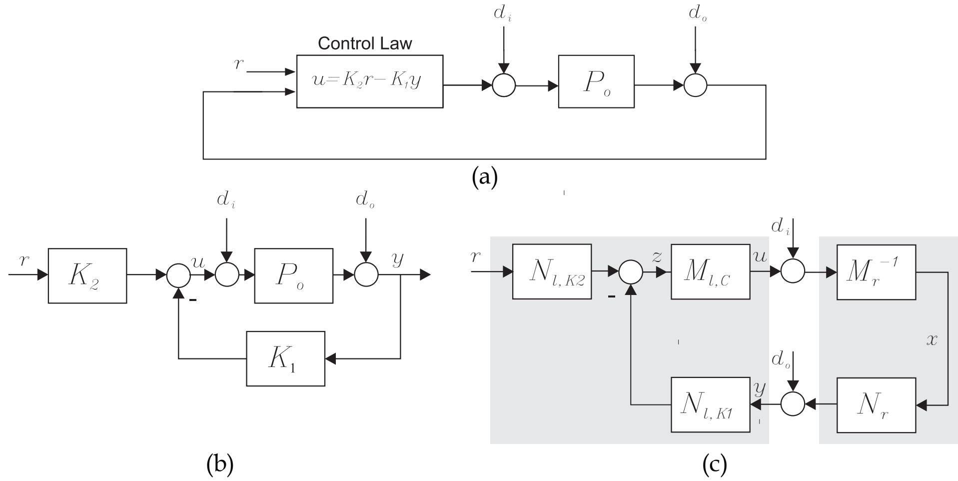

Fig. 3. (a) 2-DOF control diagram. (b) An unfeasible implementation of the 2-DOF control law $u = K_2 r - K_1 y$. (c) A feasible implementation of the control law $u = K_2 r - K_1 y$.

It is evident that if we make $K_1 = K_2 = K$, then we have $u = K(r - y)$ and recover the standard 1-DOF feedback configuration (1 parameter compensator) of figure 1. Once we have designed K_1 and K_2 , equation (15) simply gives a control law but it says nothing about the actual implementation of it, see (Wilfred, W.K. et al., 2007). For instance, in figure 3b we can see one possible implementation of the control law given by (15) which is a direct translation of the equation into a block diagram. It should be noted that this implementation is not valid when K_2 is unstable, since this block acts in an open-loop fashion and this would result in an unstable overall system, in spite of the control law being a stabilizing one. To circumvent this problem we can make use of the previously presented factorization framework and proceed as follows: define $C = [K_1 \quad K_2]$ and let $K_1 = M_{l,C}{}^{-1} N_{l,K1}$ and $K_2 = M_{l,C}{}^{-1} N_{l,K2}$ such that $(M_{l,C}, [N_{l,K1} \quad N_{l,K2}])$ is a LCF of C . Once $C = [K_1 \quad K_2]$ has been factorized as suggested, the control action in (15) can be implemented as shown in figure 3c. In this figure the plant has been right-factored as $N_r M_r{}^{-1}$. It can be shown that the mapping $(r, d_i, d_o) \rightarrow (z_1, z_2, u, y)$ remains stable (necessary for internal stability) if and only if so it does the mapping $(r, d_i, d_o) \rightarrow (u, y)$. The following theorem states when the system depicted in figure 3c is internally stable.

Theorem 3. The system of figure 3c is internally stable if and only if

$$R^{-1} := M_{l,C} M_r + N_{l,K2} N_r \in \mathcal{RH}_\infty, \quad R \in \mathcal{RH}_\infty \tag{16}$$

We can proceed now to announce the 2-DOF Youla Paramaterization.

Theorem 4. (2-DOF Youla parameterization) For a given plant $P = N_r M_r^{-1}$, let $C_{stab}(P)$ denote the set of stabilizing 2-DOF controllers $C = [K_1 \quad K_2]$, that is,

$$C_{stab}(P) \doteq \{C = [K_1, K_2] : \text{the control law } u = K_2 r - K_1 y) \text{ is stabilizing}\}. \tag{17}$$

The set $C_{stab}(P)$ can be parameterized as follows

$$C_{stab}(P) \doteq \left\{ \left(\frac{X_r + M_r Q_y}{Y_r - N_r Q_y} \quad , \quad \frac{Q_r}{Y_r - N_r Q_y} \right) : Q_y, Q_r \in RH_\infty \right\} \tag{18}$$

Proof. Based on theorem 2, it follows that the transfer function R will satisfy theorem 3 if and only if $M_{l,C}^{-1} N_{l,K1}$ equals $(Y_r - Q_y N_r)^{-1}(X_r + Q_y M_r)$ for some Q_y in $\mathcal{RH}_\infty$ such that $|Y_r - Q_y N_r| \neq 0$. Moreover, R is independent of N_{l,K_1}. This leads at once to (18).
Following with figure 3c, let us assume that we take

$$N_{l,K1} = 1, \;\; N_{l,K2} = K_1 X_r, \;\; M_{l,C} = 1 + K_1 Y_r \tag{19}$$

where $K_1 \in \mathcal{RH}_\infty$. Then the two-parameter compensator can be redrawn as shown in figure 4a. For reasons that will become clear later on, this particular two-parameter compensator is referred to as the Observer-Controller scheme.

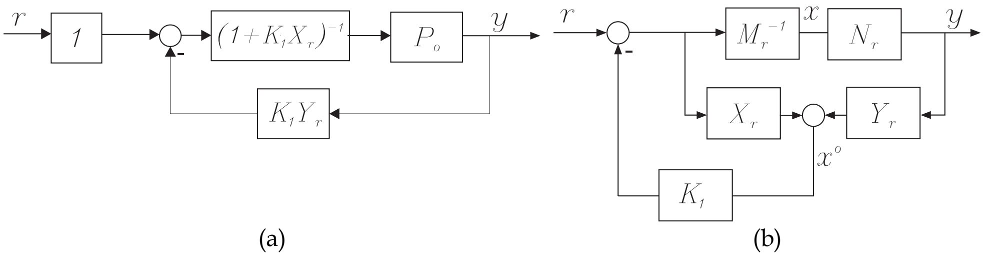

Fig. 4. (a) Observer-Controller in two blocks form. (b) Observer-Controller in three blocks form where $P_o = N_r M_r^{-1}$ is a RCF.

Applying theorem 3 for the particular case at hand the stability condition for the system of figure 4a reduces to

$$R^{-1} = (1 + K_1 X_r) M_r + K_1 Y_r N_r = M_r + K_1 \in \mathcal{RH}_\infty, \;\; R \in \mathcal{RH}_\infty \tag{20}$$

It can be verified that the relation between r and y is given by $N_r R$. In order to T_{yr} being stable, we have to require R to be stable. On the other hand, R^{-1} is given by $M_r + K_1$ which is stable having chosen K_1 stable. Choosing such an R for our design the stability requirements for the overall system to be internally stable are satisfied.

It is easy to see that figure 4a can be rearranged as in figure 4b, where the plant appears in right-factored form ($P_o = N_r M_r^{-1}$). Now it is straightforward to notice that the relation between x and x_o is given by

$$x^o = (X_r M_r + Y_r N_r)x = x \tag{21}$$

where the Bezout identity applies. This way, the X_r and Y_r blocks can be though of as an observer for the fictitious signal x appearing in the middle of the RCF. So, feeding back the observation of x lets to place the close-loop eigenvalues at prescribed locations since the achieved input to output relations is given by $y = N_r R r$ and the stable poles of both N_r and R are freely assignable. This may remind of a basic result coming from state-space control theory associated with observed state feedback: assuming a minimal realization of the plant, state feedback using observers let you change the dynamics of the plant by moving the closed-loop poles of the resulting control system to desired positions in the left half plane. Let us assume the following situation for the figure 4b

$$P = \frac{b}{a}, \quad M_r = \frac{a}{p_K}, \quad N_r = \frac{b}{p_K}, \quad X_r = \frac{n_x}{p_L}, \quad Y_r = \frac{n_y}{p_L} \tag{22}$$

Now let us take K_1 to be of the form

$$K_1 = \frac{m}{p_K} \tag{23}$$

being m an arbitrary polynomial in s of degree n-1. With p_K and p_L we refer here to monic polynomials in s having as roots the entries of the vectors in (11) and (12), respectively .The dependence of s has been dropped to simplify the notation. By choosing this stable K_1 the relation between the input r and the output y remains as follows

$$T_{yr} = \frac{b}{a+m} \tag{24}$$

So we have achieved a reallocation of the closed-loop poles leaving the zeros of the plant unaltered, as it happens in the context of state-space theory when one makes use of observed state feedback.

What follows is intended to fully understand the relationship between the scheme of figure 4 and conventional state-feedback controllers. For this purpose, we will remind here results