NONLINEAR ZERO DYNAMICS IN CONTROL SYSTEMS

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Contents

Summary

This chapter reviews a central issue in modern control system design, the zero or internal dynamics. Current techniques in both linear and nonlinear control often emphasize explicit utilization of the plant model in the control system design with some form of model inversion. This frequently admits tunable controller parameters with the important property of explicit physical significance. The subsystem generated by nulling the system outputs is called the zero dynamics, which captures the nature of this form of invertibility. Model inversion can be hindered by unstable zero dynamics, leading to a nonminimum phase (NMP) system. The stability of this key subsystem governs both the feasibility of some controller designs, and provides a baseline for minimum achievable closed-loop performance. Zero dynamics are the nonlinear generalization of the properties of zeros in a linear plant model, and must be assessed for each nonlinear control system. Some typical formulations of zero dynamics arising from different viewpoints, or methods of controller design, are covered to provide a feel for the nature of the problem. There are fairly general analytic solutions for many classes of nonlinear minimum phase systems. However, many important technological systems, including bioreactors and air or land vehicles, fall into the class of nonlinear NMP systems. These are generally more difficult for controller design, and rarely possess completely general solutions. A brief review of important model-based nonlinear control structures is included since their design is intimately connected with the zero dynamics of the system. Some interesting NMP systems are discussed, with a few approximate-analytic methods of design, to highlight the difficulties that stem from the existence of unstable zero dynamics.

1. Introduction

Modern technological systems, so ubiquitous in society today, continue to expand in their usages and requirements. Virtually all such systems possess degrees of nonlinear behavior. To extract the best performance out of a given system, often by maximizing energy efficiency while minimizing losses and wastage, sophisticated model-based nonlinear controllers (MB-NLC) are increasingly used, such as exact/feedback linearization control (ELC), backstepping, and differential flatness-based control, among others. They tend to have the advantage of being derived from explicit physical laws. Such controllers inherit parameters related to the physical model, a desirable design feature. They are found today in diverse technologies including biotechnology, aerospace, chemical processes, mechatronics, and power systems.

The generic design of most of the MB-NLC's involves some form of model inversion. The feasibility of this inversion is largely determined by the stability properties of a special dynamical subsystem generated by nulling the relevant outputs of the control system, the *zero dynamics*. If the zero dynamics system is stable, the system is said to be minimum phase (MP). Otherwise it is said to be nonminimum phase (NMP).

Analytic controller synthesis algorithms such as the ELC and others mentioned above tend to assume that the system is minimum phase. There exist important results indicating that the system being MP is a *necessary* condition to design analytic control laws that achieve high performance, namely asymptotic or even exact tracking, for an important class of systems. Philosophically, the negation of this condition by NMP systems suggests approximate or specific solutions to the controller design problem, some of which will be demonstrated here. NMP systems may impose basal performance limitations, which is evident if a similar "ideal" MP system can be compared. Many technologically significant systems turn out to be NMP, and accordingly pose a significant obstacle to effective controller design.

Stability of the zero dynamics is a key issue in the design for any control law synthesis for a nonlinear controlled system (NLCS), and must always be explicitly checked before commencing controller design. Unstable zero dynamics, in the form of NMP systems, today continue to constitute an open control systems research problem, and has led to several new innovative control laws. Some of these are covered briefly to indicate the evolving state of the art, and to demonstrate the style of approaches to this broad-based and important set of systems.

2. Nonlinear Control System Paradigms

There is a range of powerful methods available with various modifications, both generalizations and specializations, available for designing model-based nonlinear controllers (MB-NLC). A detailed exposition is both beyond the scope of this chapter, as well as too voluminous. In place of that, some of the more important methods, particularly exact/feedback linearizing control (ELC), backstepping, differential flatness-based control (DFC), and variable structure control (VSC) will be described briefly with an emphasis on design (rather than analytic properties). These methods share some common features. All are function topologically oriented, finite dimensional (ordinary differential or differential-algebraic model-based), relying on possibly iterative and at least partially linearizing geometric transformations rooted in differential geometry or algebra.

In this chapter it will be assumed that there is full access to state information. For ease of notation and simplicity, all systems will be assumed to be single-input/single-output (SISO), unless otherwise mentioned. Short sketches of some relevant NLCS design paradigms are covered here. Each method has a definite and strong dependence on the behavior of the zero dynamics subsystem. The connection to zero dynamics is mentioned only briefly in this section, and will be expanded upon appropriately in the next section.

A fairly general SISO nonlinear control system (NLCS) n-dimensional model can be given by

$$\begin{aligned} \dot{x} &= f(x,u) \\ y &= h(x), \end{aligned} \tag{1}$$

where $x \in \mathcal{X} \subset \Re^n$ is the state, $u \in \Re^1$ the control input, and $y \in \Re^1$ is the (selected) system output. $\mathcal{X}$ denotes the state manifold. Typically, the system descriptor functions are assumed to be sufficiently smooth, with the state map $f \in C^{\infty}(\Re^n \times \Re^1, \Re^n)$, and the output map $h \in C^{\infty}(\Re^n, \Re^1)$.

2.1. Exact / Feedback Linearizing Control

Many systems can be described or approximated well by an important substructure of the general NLCS of Eq. (1) called an *input-affine* form

$$\begin{aligned} \dot{x} &= f(x) + g(x)u(t) \\ y &= h(x). \end{aligned} \tag{2}$$

In this sketch we emphasize the *input-output linearization* version of ELC, which has less stringent analytic requirements compared to the closely related but more difficult to calculate state space linearization version (which includes the analytical solution of a system of partial differential equations). With similar functional assumptions to the general NLCS (1), and additional restrictions particularly certain involutivity conditions, ELC design is implementable. Without loss of generality, assume $x = 0$ is an equilibrium point of Eq. (2), i.e. $f(0) = 0$.

Consider an input-affine NLCS of the form (2). The relative degree of the system is said to be $r \le n$ at a point x^o if

$$L_g L_f^k h(x) = 0, k < r-1, \forall x \in B(x^o)$$
$$L_g L_f^{r-1} h(x^o) \ne 0,$$

where

$$L_g h(x) = \frac{\partial h}{\partial x} g(x); \quad L_g L_f h(x) = \frac{\partial L_f h(x)}{\partial x} g(x) = L_g (L_f h(x))$$
$$L_f^k h(x) = \frac{\partial L_f^{k-1} h(x)}{\partial x} f(x) = L_f (L_f^{k-1} h(x))$$

denote the appropriate system Lie derivative operations used extensively in ELC.
Let system (2) have a relative degree $r \le n$ in a certain neighborhood $\mathcal{X}_0$ of $x = 0$. Consider a stable linear differential operator $\gamma(\rho)$ of degree r ($\rho \equiv \frac{d}{dt}$)

$$\gamma(\rho) = \gamma_r \rho^r + \gamma_{r-1} \rho^{r-1} + \ldots + 1. \quad (3)$$

Then the system output can be written as

$$\gamma(\rho) y(t) = b(x) + a(x) u(t) \quad (4)$$

with $a(x) \ne 0$, $b(x)$ suitable ELC-derived functions which arise from appropriate state variable coordinate transformations of the form

$$z = \Phi(x) = (h(x), h^{(1)}(x), \ldots, h^{(r-1)}(x), \phi_{r+1}(x), \ldots, \phi_n(x)), \quad (5)$$

where $h^{(i)}(x)$ denotes the i^{th} time derivative of the output map $h(x(t))$, and has a special functional form due to the structure of system (2). These functions are intimately related to the system zero dynamics. Relevant functional specifications are given separately in Section 3.1. The ELC control input is then given by

$$u(t) = \frac{\upsilon^*(t) - b(x)}{a(x)}, \quad (6)$$

which transforms the original nonlinear system into a linear one:

$$\gamma(\rho)y(t) = \upsilon^*(t), \quad (7)$$

where $\upsilon^*(t)$ is an external signal, usually some form of linear control given by

$$\upsilon^*(t) = L[e(t)], \quad (8)$$

where L is a linear operator representing the control law, and $e(t) = y(t) - y_R(t)$ is the reference model/trajectory tracking error. Feasibility of ELC design implicitly ensures stability, in the sense of a system that is stabilizable by external linear control.
ELC structure and design represent a key point of departure from linear systems, and provide a sound basis for constructing extended and generalized NLCS models and control designs.

2.2. Backstepping Control

Backstepping is an iterative analytic controller design procedure for systems with models of special structure, similar to the form of Eq. (2). It can be used on such systems to find an output z that has a passivity property, i.e. demonstrates a relative degree 1, and allows stable system inversion (and hence zero dynamics). A system with input $u(t)$ and output $z(t)$ is *passive* if there exists a nonnegative storage function $U(x(t),t)$ with the property

$$\int_0^T u(t)z(t)dt \geq U(x(T),T) - U(x(0),0). \quad (9)$$

In practice passivity is a very strong requirement. An appropriate passive system can be rendered asymptotically stable by the simple output feedback $u(t) = -z(t)$. The recursive backstepping design generates a passive system given a system in a specific form.

Some manipulations and simplifications can often allow the state description to be rewritten in a special input-linear form, convenient for backstepping design. Let $\vec{x}_i$ denote the first i components of x. A vector or matrix function ϕ is said to be lower triangular dependent (strictly) on x if the i^{th} row is a function at most of $\vec{x}_i$ $(\vec{x}_{i-1})$. Consider the system

$$\dot{x} = Fx + \tilde{f}(x) + gu \ , \quad (10)$$

where $\tilde{f}(x)$ is lower triangular dependent on x, F is a zero matrix save for 1's on the upper diagonal, and g is a zero vector save for the last entry being 1. It is said to be in "strict feedback form". Somewhat analogous to ELC design, the backstepping algorithm starts by defining an invertible coordinate transformation generating an auxiliary state variable $z = \Phi_z(x)$ of the affine form

$$z = x + F^T \phi(x) \tag{11}$$

and an auxiliary input $\upsilon = \Phi_\upsilon(x,u)$ of the affine form

$$\upsilon = u + g^T \phi(x)\,, \tag{12}$$

where ϕ is lower triangular dependent on x, and to be determined. This generates a transformed system with the same structure as Eq. (10)

$$\dot{z} = Fz + \tilde{\phi}(z) + g\upsilon \tag{13}$$

and leads to the backstepping transformation function

$$\phi(x) = \tilde{f}(x) + F^T\left[\nabla_x \phi(x)\right](Fx + \tilde{f}(x)) - \tilde{\phi}(x + F^T\phi(x))\,, \tag{14}$$

which depends on the selection of an appropriate lower triangular dependent structure for $\tilde{\phi}(z)$. Since ϕ is lower triangular dependent on x, Eq. (14) is used to construct it row by row.

This yields the core backstepping design for strict feedback systems. It is a purely structural design, which does not necessarily insure stability. Stability can be enforced by selecting

$$\tilde{\phi}(z) = -(F^T + C(z,t))z \tag{15}$$

with $C(z,t)$ lower triangular dependent on z and positive semidefinite but otherwise arbitrary. Selection of the Lyapunov function $V(z) = \frac{1}{2} z^T z$ then demonstrates stability for this system.

2.3. Differentially Flat Control

Consider the general NLCS of Eq. (1) in multi-input/multi-output (MIMO) form with $u \in \Re^m$, where $f(0,0) = 0$ and

$$\text{rank}\left[\frac{\partial f}{\partial u}(0,0)\right] = m.$$

Many such systems are expressible in *differentially flat* form, i.e. with some outputs such that states and inputs can be expressed in terms of those outputs and a finite number of their derivatives.

Flatness is not a generic property of a control system. Requirements for a differentially flat controller (DFC) design are summarized in the model

$$\begin{aligned} \dot{x} &= f(x,u) \\ y &= h_Y(x) \\ z &= h_Z(x,\bar{u}^{(\gamma)}) \\ x &= \psi_X(\bar{z}^{(\alpha)}) \\ u &= \psi_U(\bar{z}^{(\beta)}) \end{aligned} \tag{16}$$

for some maps ψ_X and ψ_U. Here y are the tracking outputs, and z the flat outputs. The notation

$$\bar{p}^{(\delta)} = (p, \dot{p}, \ldots, p^{(\delta)}) \tag{17}$$

denotes a system flag which stacks some system input, or output, and their derivatives up to order δ. This is equivalent to finding a regular *endogenous* dynamic feedback compensator of the form

$$\begin{aligned} \dot{w} &= a(x,w,\upsilon) \\ u &= b(x,w,\upsilon)(w \in \Re^q, \upsilon \in \Re^m), \end{aligned} \tag{18}$$

where $a(0,0,0)=0$ and $b(0,0,0)=0$. Regularity implies the invertibility of (18) with input υ and output u; and a diffeomorphism

$$\xi = \Phi(x,w)(\xi \in \Re^{n+q}), \tag{19}$$

which transform the dynamic feedback system of Eqs. (18)-(19) into a controllable linear system of the form $\dot{\xi} = F\xi + G\upsilon$. An additional linear invertible transformation (and at most a static state feedback) converts this linear system into the Brunovsky canonical form

$$\begin{aligned} z_1^{(\nu_1)} &= v_1 \\ &\vdots \\ z_m^{(\nu_m)} &= v_m \end{aligned} \tag{20}$$

where $v_1, \ldots, v_m$ are the controllability indices. Then $Z = (z_1, \ldots, z_1^{(\nu_1 - 1)}, \ldots, z_m, \ldots, z_m^{(\nu_m - 1)})$ is another basis for ξ-space, or there exists an invertible $(n+q)\times(n+q)$ matrix T such that $Z = T\xi$. Therefore $Z = T\Phi(x,w)$ and invertibility implies

$$(x,w) = \Phi^{-1}(T^{-1}Z). \tag{21}$$

$z = (z_1, \ldots, z_m)$ is precisely the *desired* flat output. From Eqs. (18), (20) and (21), where

$u = b(\Phi^{-1}(T^{-1}Z),\upsilon)$ and $v_i = z_i^{(v_i)}$, x and u can be expressed as analytic functions of this flat output and a finite number of its derivatives namely the appropriate maps ψ_X and ψ_U as shown in Eq. (16). The dynamic feedback is endogenous if and only if the converse also holds, namely that the flat output z can be expressed as an analytic function of x, u and a finite number of its derivatives, or

$$z = h_Z(x,\bar{u}^{(\gamma)}) \tag{22}$$

as in Eq. (16). Eqs. (21) and (22) imply that w can be expressed as a function of $(x,\bar{u}^{(\kappa)})$ for some integer κ, hence the dynamic extension is endogenous.

A key benefit of flat systems is the recovery of system and input information without explicit integration of the system equations. DFC design differs somewhat from the methods described above in Sections 2.1 and 2.2. First, a set of flat outputs has to be verified by the designer. Unlike the constructive topological methods of backstepping and ELC, currently there are no well defined checkability conditions to find flat outputs, a current limitation of this powerful technique.

Given a verified flat output, however, produces several desirable features for DFC design. A flat system is equivalent to a linear system, via at most a dynamic feedback. This feature bears a strong conceptual resemblance to the ELC structure, with the advantage of allowing stabilization across an entire trajectory, rather than in the local region of an equilibrium point. A unique and powerful feature of flat systems is the generation of an explicit two-degree-of-freedom (2-dof) trajectory-driven controller design. Given a predetermined reference output trajectory y_R, an open-loop feedforward control u_R can be generated via the state-input/flat flag maps ψ_X and ψ_U given in Eq. (16). If the model is sufficiently good, and there are no significant noise and disturbance signals, the trajectory will be tracked exactly.

Outline of a feedforward-feedback 2-*dof DFC design* To compensate for noise, disturbances and model mismatch, an auxiliary linear feedback compensator is designed as follows. Let the system vector $(x_R(t), y_R(t), u_R(t))$ be an instance of the feedforward reference system trajectory $(\vec{x}_R, \vec{y}_R, \vec{u}_R)$ generated as above with $(x(t), y(t), u(t))$ instantaneous values of the system vector. In general $u(t) \neq u_R(t)$. The instantaneous actual and reference flat outputs are reconstructed as

$$\begin{aligned} z(t) &= h_Z(x(t),\bar{u}^{(\gamma)}(t)) \\ z_R(t) &= h_Z(x_R(t),\bar{u}_R^{(\gamma)}(t)). \end{aligned} \tag{23}$$

Define $e_Z(t) = z(t) - z_R(t)$ as the *flat output* tracking error. Then in analogy to the ELC design procedure, choose

$$\upsilon_i(t) = L_i[e_{Z,i}(t)];\ i = 1,\ldots,m, \tag{24}$$

where $\{L_i\}_{i=1}^{m}$ are SISO linear operators representing linearized feedback control laws for the decoupled linear system (20). Eq. (20) is used to construct the new flat output $z(t)$. Finally the overall control input is calculated using the map (16)

$$u(t) = \psi_U\left(\overline{z}^{(\beta)}(t); e_Z(t)\right), \tag{25}$$

which can be expressed as a sum of feedback and feedforward controllers or $u = u_{fb} + u_{ff}$. Here $u_{ff} = u_R$, so that the effective feedback correction is $u_{fb} = u - u_R$. In the case of perfect tracking, $e_Z = 0$, so that $u \equiv u_R$. This formulation requires the number of selected flat outputs to be the same as the number of inputs. When the flat and tracking outputs do not coincide, a zero dynamics is generated. The implications of this are discussed further in Section 4.5.

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Further Reading

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Biographical Sketches

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