

The Control Handbook
Second Edition

CONTROL SYSTEM ADVANCED METHODS



Edited by

William S. Levine

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Control System Advanced Methods

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The Control Handbook

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Edited by

William S. Levine

University of Maryland

College Park, MD, USA

Control System Fundamentals

Control System Applications

Control System Advanced Methods

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Edited by

William S. Levine

University of Maryland

College Park, MD, USA



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Preface to the Second Edition

As you may know, the first edition of *The Control Handbook* was very well received. Many copies were sold and a gratifying number of people took the time to tell me that they found it useful. To the publisher, these are all reasons to do a second edition. To the editor of the first edition, these same facts are a modest disincentive. The risk that a second edition will not be as good as the first one is real and worrisome. I have tried very hard to insure that the second edition is at least as good as the first one was. I hope you agree that I have succeeded.

I have made two major changes in the second edition. The first is that all the *Applications* chapters are new. It is simply a fact of life in engineering that once a problem is solved, people are no longer as interested in it as they were when it was unsolved. I have tried to find especially inspiring and exciting applications for this second edition.

Secondly, it has become clear to me that organizing the *Applications* book by academic discipline is no longer sensible. Most control applications are interdisciplinary. For example, an automotive control system that involves sensors to convert mechanical signals into electrical ones, actuators that convert electrical signals into mechanical ones, several computers and a communication network to link sensors and actuators to the computers does not belong solely to any specific academic area. You will notice that the applications are now organized broadly by application areas, such as automotive and aerospace.

One aspect of this new organization has created a minor and, I think, amusing problem. Several wonderful applications did not fit into my new taxonomy. I originally grouped them under the title Miscellaneous. Several authors objected to the slightly pejorative nature of the term “miscellaneous.” I agreed with them and, after some thinking, consulting with literate friends and with some of the library resources, I have renamed that section “Special Applications.” Regardless of the name, they are all interesting and important and I hope you will read those articles as well as the ones that did fit my organizational scheme.

There has also been considerable progress in the areas covered in the *Advanced Methods* book. This is reflected in the roughly two dozen articles in this second edition that are completely new. Some of these are in two new sections, “Analysis and Design of Hybrid Systems” and “Networks and Networked Controls.”

There have even been a few changes in the *Fundamentals*. Primarily, there is greater emphasis on sampling and discretization. This is because most control systems are now implemented digitally.

I have enjoyed editing this second edition and learned a great deal while I was doing it. I hope that you will enjoy reading it and learn a great deal from doing so.

William S. Levine

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The MathWorks, Inc.

3 Apple Hill Drive

Natick, MA, 01760-2098 USA

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E-mail: info@mathworks.com

Web: www.mathworks.com

Acknowledgments

The people who were most crucial to the second edition were the authors of the articles. It took a great deal of work to write each of these articles and I doubt that I will ever be able to repay the authors for their efforts. I do thank them very much.

The members of the advisory/editorial board for the second edition were a very great help in choosing topics and finding authors. I thank them all. Two of them were especially helpful. Davor Hrovat took responsibility for the automotive applications and Richard Braatz was crucial in selecting the applications to industrial process control.

It is a great pleasure to be able to provide some recognition and to thank the people who helped bring this second edition of *The Control Handbook* into being. Nora Konopka, publisher of engineering and environmental sciences for Taylor & Francis/CRC Press, began encouraging me to create a second edition quite some time ago. Although it was not easy, she finally convinced me. Jessica Vakili and Kari Budyk, the project coordinators, were an enormous help in keeping track of potential authors as well as those who had committed to write an article. Syed Mohamad Shajahan, senior project executive at Techset, very capably handled all phases of production, while Richard Tressider, project editor for Taylor & Francis/CRC Press, provided direction, oversight, and quality control. Without all of them and their assistants, the second edition would probably never have appeared and, if it had, it would have been far inferior to what it is.

Most importantly, I thank my wife Shirley Johannesen Levine for everything she has done for me over the many years we have been married. It would not be possible to enumerate all the ways in which she has contributed to each and everything I have done, not just editing this second edition.

William S. Levine

Editorial Board

Frank Allgöwer

Institute for Systems Theory and
Automatic Control
University of Stuttgart
Stuttgart, Germany

Tamer Başar

Department of Electrical and
Computer Engineering
University of Illinois at Urbana–Champaign
Urbana, Illinois

Richard Braatz

Department of Chemical Engineering
Massachusetts Institute of Technology
Cambridge, Massachusetts

Christos Cassandras

Department of Manufacturing Engineering
Boston University
Boston, Massachusetts

Davor Hrovat

Research and Advanced Engineering
Ford Motor Company
Dearborn, Michigan

Naomi Leonard

Department of Mechanical and
Aerospace Engineering
Princeton University
Princeton, New Jersey

Masayoshi Tomizuka

Department of Mechanical
Engineering
University of California, Berkeley
Berkeley, California

Mathukumalli Vidyasagar

Department of Bioengineering
The University of Texas at Dallas
Richardson, Texas

Editor

William S. Levine received B.S., M.S., and Ph.D. degrees from the Massachusetts Institute of Technology. He then joined the faculty of the University of Maryland, College Park where he is currently a research professor in the Department of Electrical and Computer Engineering. Throughout his career he has specialized in the design and analysis of control systems and related problems in estimation, filtering, and system modeling. Motivated by the desire to understand a collection of interesting controller designs, he has done a great deal of research on mammalian control of movement in collaboration with several neurophysiologists.

He is co-author of *Using MATLAB to Analyze and Design Control Systems*, March 1992. Second Edition, March 1995. He is the coeditor of *The Handbook of Networked and Embedded Control Systems*, published by Birkhauser in 2005. He is the editor of a series on control engineering for Birkhauser. He has been president of the IEEE Control Systems Society and the American Control Council. He is presently the chairman of the SIAM special interest group in control theory and its applications.

He is a fellow of the IEEE, a distinguished member of the IEEE Control Systems Society, and a recipient of the IEEE Third Millennium Medal. He and his collaborators received the Schroers Award for outstanding rotorcraft research in 1998. He and another group of collaborators received the award for outstanding paper in the *IEEE Transactions on Automatic Control*, entitled “Discrete-Time Point Processes in Urban Traffic Queue Estimation.”

Contributors

Eyad H. Abed

Department of Electrical Engineering
and the Institute for Systems Research
University of Maryland
College Park, Maryland

Andrew G. Alleyne

Department of Mechanical Science
and Engineering
University of Illinois at Urbana–Champaign
Urbana, Illinois

Albert N. Andry, Jr.

Teledyne Reynolds Group
Los Angeles, California

Panos J. Antsaklis

Department of Electrical Engineering
University of Notre Dame
Notre Dame, Indiana

Murat Arcak

Department of Electrical Engineering
and Computer Sciences
University of California, Berkeley
Berkeley, California

Karl J. Åström

Department of Automatic Control
Lund Institute of Technology
Lund, Sweden

Michael Athans

Department of Electrical Engineering
and Computer Science
Massachusetts Institute of
Technology
Cambridge, Massachusetts

J. Baillieul

Department of Electrical & Computer
Engineering
Boston University
Boston, Massachusetts

Gary J. Balas

Department of Aerospace Engineering
and Mechanics
University of Minnesota
Minneapolis, Minnesota

Simone Baldi

Department of Systems and Computer Science
University of Florence
Florence, Italy

Kira L. Barton

Department of Mechanical Science
and Engineering
University of Illinois at Urbana–Champaign
Urbana, Illinois

Sorin Bengea

Control Systems Group
United Technologies Research Center
East Hartford, Connecticut

Franco Blanchini

Department of Mathematics and
Computer Science
University of Udine
Udine, Italy

Anthony M. Bloch

Department of Mathematics
University of Michigan
Ann Arbor, Michigan

Julio H. Braslavsky

School of Electrical Engineering and
Computer Science
The University of Newcastle
Callaghan, New South Wales, Australia

Douglas A. Bristow

Department of Mechanical and
Aerospace Engineering
Missouri University of Science and Technology
Rolla, Missouri

Christopher I. Byrnes

Department of Systems Sciences
and Mathematics
Washington University
St. Louis, Missouri

Alan Chao

Laboratory for Information and
Decision Systems
Massachusetts Institute of Technology
Cambridge, Massachusetts

David W. Clarke

Department of Engineering Science
University of Oxford
Oxford, United Kingdom

Fabrizio Dabbene

Institute for Electronics,
Engineering, Information and
Telecommunications–National
Research Council
Polytechnic University of Turin
Turin, Italy

Munther A. Dahleh

Laboratory for Information and
Decision Systems
Massachusetts Institute of Technology
Cambridge, Massachusetts

Edward J. Davison

Department of Electrical and
Computer Engineering
University of Toronto
Toronto, Ontario, Canada

Raymond A. DeCarlo

Department of Electrical and
Computer Engineering
Purdue University
West Lafayette, Indiana

David F. Delchamps

School of Electrical and Computer
Engineering
Cornell University
Ithaca, New York

Maria Domenica Di Benedetto

Department of Electrical Engineering
University of L'Aquila
L'Aquila, Italy

Joel Douglas

Department of Electrical Engineering
and Computer Science
Massachusetts Institute of Technology
Cambridge, Massachusetts

T.E. Duncan

Department of Mathematics
University of Kansas
Lawrence, Kansas

Sergey V. Drakunov

Department of Physical Sciences
Embry-Riddle Aeronautical University
Daytona Beach, Florida

Jay A. Farrell

Department of Electrical Engineering
University of California, Riverside
Riverside, California

Bruce Francis

Department of Electrical and
Computer Engineering
University of Toronto
Toronto, Ontario, Canada

Randy A. Freeman

University of California, Santa Barbara
Santa Barbara, California

James S. Freudenberg

Department of Electrical Engineering and
Computer Science
University of Michigan
Ann Arbor, Michigan

Bernard Friedland

Department of Electrical and
Computer Engineering
New Jersey Institute of Technology
Newark, New Jersey

T.T. Georgiou

Department of Electrical Engineering
University of Minnesota
Minneapolis, Minnesota

Simon Grocott

Space Engineering Research Center
Massachusetts Institute of Technology
Cambridge, Massachusetts

John A. Gubner

College of Engineering
University of Wisconsin–Madison
Madison, Wisconsin

Vijay Gupta

Department of Electrical Engineering
University of Notre Dame
Notre Dame, Indiana

Tore Hägglund

Department of Automatic Control
Lund Institute of Technology
Lund, Sweden

Fumio Hamano

Department of Electrical Engineering
California State University
Long Beach, California

Constantine H. Houppis

Department of Electrical and Computer
Engineering
Air Force Institute of Technology
Wright-Patterson Air Force Base, Ohio

Petros Ioannou

Department of Electrical
Engineering
University of Southern California
Los Angeles, California

Alberto Isidori

Department of Computer Science and
Systemics
Sapienza University of Rome
Rome, Italy

Ulf T. Jönsson

Department of Mathematics
Royal Institute of Technology
Stockholm, Sweden

V. Jurdjevic

Department of Mathematics
University of Toronto
Toronto, Ontario, Canada

Edward W. Kamen

School of Electrical and Computer
Engineering
Georgia Institute of Technology
Atlanta, Georgia

Chung-Yao Kao

Department of Electrical
Engineering
National Sun Yat-Sen University
Kaohsiung, Taiwan

Hassan K. Khalil

Department of Electrical & Computer
Engineering
Michigan State University
East Lansing, Michigan

Petar V. Kokotović

Department of Electrical and Computer
Engineering
University of California, Santa Barbara
Santa Barbara, California

Miroslav Krstić

Department of Mechanical and
Aerospace Engineering
University of California, San Diego
San Diego, California

Vladimír Kučera

Czech Technical University
Prague, Czech Republic
and

Institute of Information Theory and Automation
Academy of Sciences
Prague, Czech Republic

P.R. Kumar

Department of Electrical and Computer
Engineering and Coordinated Science
Laboratory
University of Illinois at Urbana–Champaign
Urbana, Illinois

Alex A. Kurzhanskiy

Department of Electrical Engineering
and Computer Sciences
University of California, Berkeley
Berkeley, California

Françoise Lamnabhi-Lagarrigue

Laboratory of Signals and Systems
National Center for Scientific Research
Supélec
Gif-sur-Yvette, France

J.E. Lagnese

Department of Mathematics
Georgetown University
Washington, D.C.

A.J. Laub

Electrical Engineering Department
University of California, Los Angeles
Los Angeles, California

Jay H. Lee

Korean Advanced Institute of
Science and Technology
Daejeon, South Korea

B. Lehman

Department of Electrical & Computer
Engineering
Northeastern University
Boston, Massachusetts

G. Leugering

Faculty for Mathematics, Physics, and
Computer Science
University of Bayreuth
Bayreuth, Germany

William S. Levine

Department of Electrical & Computer
Engineering
University of Maryland
College Park, Maryland

Frank L. Lewis

Automation and Robotics Research
Institute
The University of Texas at Arlington
Arlington, Texas

Hai Lin

Department of Electrical and
Computer Engineering
National University of Singapore
Singapore

Shu-Jun Liu

Department of Mathematics
Southeast University
Nanjing, China

Lennart Ljung

Department of Electrical
Engineering
Linköping University
Linköping, Sweden

Douglas P. Looze

Department of Electrical and
Computer Engineering
University of Massachusetts
Amherst
Amherst, Massachusetts

Kenneth A. Loparo

Department of Electrical Engineering
and Computer Science
Case Western Reserve University
Cleveland, Ohio

Leonard Lublin

Space Engineering Research Center
Massachusetts Institute of Technology
Cambridge, Massachusetts

Kevin M. Lynch

Mechanical Engineering Department
Northwestern University
Evanston, Illinois

Lorenzo Marconi

Department of Electronics, Computer
Sciences and Systems
University of Bologna
Bologna, Italy

Nuno C. Martins

Department of Electrical & Computer
Engineering
University of Maryland
College Park, Maryland

Alexandre Megretski

Laboratory for Information and
Decision Systems
Massachusetts Institute of Technology
Cambridge, Massachusetts

Richard H. Middleton

The Hamilton Institute
National University of Ireland,
Maynooth
Maynooth, Ireland

Manfred Morari

Swiss Federal Institute of Technology
Zurich, Switzerland

Kirsten Morris

Department of Applied Mathematics
University of Waterloo
Waterloo, Ontario, Canada

Andy Packard

Department of Mechanical Engineering
University of California, Berkeley
Berkeley, California

B. Pasik-Duncan

Department of Mathematics
University of Kansas
Lawrence, Kansas

Kevin M. Passino

Department of Electrical Engineering
The Ohio State University
Columbus, Ohio

Stephen D. Patek

Laboratory for Information and
Decision Systems
Massachusetts Institute of Technology
Cambridge, Massachusetts

R.V. Patel

Department of Electrical and
Computer Engineering
University of Western Ontario
London, Ontario, Canada

Marios M. Polycarpou

Department of Electrical and
Computer Engineering
University of Cyprus
Nicosia, Cyprus

L. Praly

Systems and Control Centre
Mines Paris Institute of Technology
Paris, France

Jörg Raisch

Department of Electrical Engineering and
Computer Science
Technical University of Berlin
Berlin, Germany

Anders Rantzer

Department of Automatic Control
Lund University
Lund, Sweden

Mahmut Reyhanoglu

Physical Sciences Department
Embry-Riddle Aeronautical University
Daytona Beach, Florida

Michael C. Rotkowitz

Department of Electrical and
Electronic Engineering
The University of Melbourne
Melbourne, Australia

David L. Russell

Department of Mathematics
Virginia Polytechnic Institute and
State University
Blacksburg, Virginia

Carsten Scherer

Department of Mathematics
University of Stuttgart
Stuttgart, Germany

Jacquelin M.A. Scherpen

Faculty of Mathematics and Natural
Sciences
University of Groningen
Groningen, the Netherlands

Thomas I. Seidman

Department of Mathematics and
Statistics
University of Maryland, Baltimore County
Baltimore, Maryland

M.E. Sezer

Department of Electrical and
Electronics Engineering
Bilkent University
Ankara, Turkey

Eliezer Y. Shapiro (deceased)

HR Textron
Valencia, California

Adam Shwartz

Department of Electrical Engineering
Technion-Israel Institute of Technology
Haifa, Israel

D.D. Šiljak

Department of Electrical Engineering
Santa Clara University
Santa Clara, California

Andrey Smyshlyaev

Department of Mechanical and
Aerospace Engineering
University of California, San Diego
La Jolla, California

Kenneth M. Sobel

Department of Electrical Engineering
The City College of New York
New York, New York

Torsten Söderström

Department of Information Technology
Uppsala University
Uppsala, Sweden

Eduardo D. Sontag

Department of Mathematics
Rutgers University
New Brunswick, New Jersey

A.R. Teel

Department of Electrical Engineering
University of Minnesota
Minneapolis, Minnesota

Roberto Tempo

Institute for Electronics, Engineering,
Information and Telecommunications-
National Research Council
Polytechnic University of Turin
Turin, Italy

Alberto Tesi

Department of Systems and
Computer Science
University of Florence
Florence, Italy

Kasemsak Uthaichana

Department of Electrical Engineering
Chiang Mai University
Chiang Mai, Thailand

P.M. Van Dooren

Department of Mathematical Engineering
Catholic University of Leuven
Leuven, Belgium

Pravin Varaiya

Department of Electrical Engineering
and Computer Sciences
University of California, Berkeley
Berkeley, California

Rafael Vazquez

Department of Aerospace Engineering
University of Seville
Seville, Spain

Draguna Vrabie

Automation and Robotics Research Institute
The University of Texas at Arlington
Arlington, Texas

Hua O. Wang

United Technologies Research Center
East Hartford, Connecticut

Siep Weiland

Department of Electrical Engineering
Eindhoven University of Technology
Eindhoven, the Netherlands

Trevor Williams

Department of Aerospace Engineering
and Engineering Mechanics
University of Cincinnati
Cincinnati, Ohio

Stephen Yurkovich

Department of Electrical Engineering
The Ohio State University
Columbus, Ohio

S.H. Żak

School of Electrical and
Computer Engineering
Purdue University
West Lafayette, Indiana

Milos Žefran

Department of Electrical Engineering
University of Illinois at Chicago
Chicago, Illinois

Dmitry Zenkov

Department of Mathematics
North Carolina State University
Raleigh, North Carolina

I

Analysis Methods for MIMO Linear Systems

1

Numerical and Computational Issues in Linear Control and System Theory*

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A.J. Laub

University of California, Los Angeles

R.V. Patel

University of Western Ontario

P.M. Van Dooren

Catholic University of Leuven

1.1 Introduction

This chapter provides a survey of various aspects of the numerical solution of selected problems in linear systems, control, and estimation theory. Space limitations preclude an exhaustive survey and extensive list of references. The interested reader is referred to [1,4,10,14] for sources of additional detailed information.

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Many of the problems considered in this chapter arise in the study of the “standard” linear model

$$\dot{x}(t) = Ax(t) + Bu(t), \quad (1.1)$$

$$y(t) = Cx(t) + Du(t). \quad (1.2)$$

Here, $x(t)$ is an n -vector of states, $u(t)$ is an m -vector of controls or inputs, and $y(t)$ is a p -vector of outputs. The standard discrete-time analog of Equations 1.1 and 1.2 takes the form

$$x_{k+1} = Ax_k + Bu_k, \quad (1.3)$$

$$y_k = Cx_k + Du_k. \quad (1.4)$$

Of course, considerably more elaborate models are also studied, including time-varying, stochastic, and nonlinear versions of the above, but these are not discussed in this chapter. In fact, the above linear models are usually derived from linearizations of nonlinear models regarding selected nominal points.

The matrices considered here are, for the most part, assumed to have real coefficients and to be small (of order a few hundred or less) and dense, with no particular exploitable structure. Calculations for most problems in classical single-input, single-output control fall into this category. Large sparse matrices or matrices with special exploitable structures may significantly involve different concerns and methodologies than those discussed here.

The systems, control, and estimation literature is replete with *ad hoc* algorithms to solve the computational problems that arise in the various methodologies. Many of these algorithms work quite well on some problems (e.g., “small-order” matrices) but encounter numerical difficulties, often severe, when “pushed” (e.g., on larger order matrices). The reason for this is that little or no attention has been paid to the way algorithms perform in “finite arithmetic,” that is, on a finite word length digital computer.

A simple example by Moler and Van Loan [14, p. 649]* illustrates a typical pitfall. Suppose it is desired to compute the matrix e^A in single precision arithmetic on a computer which gives six decimal places of precision in the fractional part of floating-point numbers. Consider the case

$$A = \begin{bmatrix} -49 & 24 \\ -64 & 31 \end{bmatrix}$$

and suppose the computation is attempted with the Taylor series formula

$$e^A = \sum_{k=0}^{+\infty} \frac{1}{k!} A^k. \quad (1.5)$$

This is easily coded and it is determined that the first 60 terms in the series suffice for the computation, in the sense that the terms for $k \geq 60$ of the order 10^{-7} no longer add anything significant to the sum. The resulting answer is

$$\begin{bmatrix} -22.2588 & -1.43277 \\ -61.4993 & -3.47428 \end{bmatrix}.$$

Surprisingly, the true answer is (correctly rounded)

$$\begin{bmatrix} -0.735759 & 0.551819 \\ -1.47152 & 1.10364 \end{bmatrix}.$$

What happened here was that the intermediate terms in the series became very large before the factorial began to dominate. The 17th and 18th terms, for example, are of the order of 10^7 but of opposite signs so

* The page number indicates the location of the appropriate reprint in [14].

that the less significant parts of these numbers, while significant for the final answer, are “lost” because of the finiteness of the arithmetic.

For this particular example, various fixes and remedies are available. However, in more realistic examples, one seldom has the luxury of having the “true answer” available so that it is not always easy to simply inspect or test a computed solution and determine that it is erroneous. Mathematical analysis (truncation of the series, in the example above) alone is simply not sufficient when a problem is analyzed or solved in finite arithmetic (truncation of the arithmetic). Clearly, a great deal of care must be taken.

The finiteness inherent in representing real or complex numbers as floating-point numbers on a digital computer manifests itself in two important ways: floating-point numbers have only finite precision and finite range. The degree of attention paid to these two considerations distinguishes many reliable algorithms from more unreliable counterparts.

The development in systems, control, and estimation theory of stable, efficient, and reliable algorithms that respect the constraints of finite arithmetic began in the 1970s and still continues. Much of the research in numerical analysis has been directly applicable, but there are many computational issues in control (e.g., the presence of hard or structural zeros) where numerical analysis does not provide a ready answer or guide. A symbiotic relationship has developed, especially between numerical linear algebra and linear system and control theory, which is sure to provide a continuing source of challenging research areas.

The abundance of numerically fragile algorithms is partly explained by the following observation:

If an algorithm is amenable to “easy” manual calculation, it is probably a poor method if implemented in the finite floating-point arithmetic of a digital computer.

For example, when confronted with finding the eigenvalues of a 2×2 matrix, most people would find the characteristic polynomial and solve the resulting quadratic equation. But when extrapolated as a general method for computing eigenvalues and implemented on a digital computer, this is a very poor procedure for reasons such as roundoff and overflow/underflow. The preferred method now would generally be the double Francis QR algorithm (see [17] for details) but few would attempt that manually, even for very small-order problems.

Many algorithms, now considered fairly reliable in the context of finite arithmetic, are not amenable to manual calculations (e.g., various classes of orthogonal similarities). This is a kind of converse to the observation quoted above. Especially in linear system and control theory, we have been too easily tempted by the ready availability of closed-form solutions and numerically naive methods to implement those solutions. For example, in solving the initial value problem

$$\dot{x}(t) = Ax(t); \quad x(0) = x_0, \quad (1.6)$$

it is not at all clear that one should explicitly compute the intermediate quantity e^{tA} . Rather, it is the vector $e^{tA}x_0$ that is desired, a quantity that may be computed by treating Equation 1.6 as a system of (possibly stiff) differential equations and using an implicit method for numerically integrating the differential equation. But such techniques are definitely not attractive for manual computation.

The awareness of such numerical issues in the mathematics and engineering community has increased significantly in the last few decades. In fact, some of the background material well known to numerical analysts has already filtered down to undergraduate and graduate curricula in these disciplines. This awareness and education has affected system and control theory, especially linear system theory. A number of numerical analysts were attracted by the wealth of interesting numerical linear algebra problems in linear system theory. At the same time, several researchers in linear system theory turned to various methods and concepts from numerical linear algebra and attempted to modify them in developing reliable algorithms and software for specific problems in linear system theory. This cross-fertilization has been greatly enhanced by the widespread use of software packages and by developments over the last couple of decades in numerical linear algebra. This process has already begun to have a significant impact on the future directions and development of system and control theory, and on applications, as evident

from the growth of computer-aided control system design (CACSD) as an intrinsic tool. Algorithms implemented as mathematical software are a critical “inner” component of a CACSD system.

In the remainder of this chapter, we survey some results and trends in this interdisciplinary research area. We emphasize numerical aspects of the problems/algorithms, which is why we also spend time discussing appropriate numerical tools and techniques. We discuss a number of control and filtering problems that are of widespread interest in control.

Before proceeding further, we list here some notations to be used:

$\mathbb{F}^{n \times m}$	the set of all $n \times m$ matrices with coefficients in the field \mathbb{F} (\mathbb{F} is generally \mathbb{R} or \mathbb{C})
A^T	the transpose of $A \in \mathbb{R}^{n \times m}$
A^H	the complex-conjugate transpose of $A \in \mathbb{C}^{n \times m}$
A^+	the Moore–Penrose pseudoinverse of A
$\ A\ $	the spectral norm of A (i.e., the matrix norm subordinate to the Euclidean vector norm: $\ A\ = \max_{\ x\ _2=1} \ Ax\ _2$)
$\text{diag}(a_1, \dots, a_n)$	the diagonal matrix $\begin{bmatrix} a_1 & & 0 \\ & \ddots & \\ 0 & & a_n \end{bmatrix}$
$\Lambda(A)$	the set of eigenvalues $\lambda_1, \dots, \lambda_n$ (not necessarily distinct) of $A \in \mathbb{F}^{n \times n}$
$\lambda_i(A)$	the i th eigenvalue of A
$\Sigma(A)$	the set of singular values $\sigma_1, \dots, \sigma_m$ (not necessarily distinct) of $A \in \mathbb{F}^{n \times m}$
$\sigma_i(A)$	the i th singular value of A

Finally, let us define a particular number to which we make frequent reference in the following. The *machine epsilon* or *relative machine precision* is defined, roughly speaking, as the smallest positive number ϵ that, when added to 1 on our computing machine, gives a number greater than 1. In other words, any machine representable number δ less than ϵ gets “rounded off” when (floating-point) added to 1 to give exactly 1 again as the rounded sum. The number ϵ , of course, varies depending on the kind of computer being used and the precision of the computations (single precision, double precision, etc.). But the fact that such a positive number ϵ exists is entirely a consequence of finite word length.

1.2 Numerical Background

In this section, we give a very brief discussion of two concepts fundamentally important in numerical analysis: *numerical stability* and *conditioning*. Although this material is standard in textbooks such as [8], it is presented here for completeness and because the two concepts are frequently confused in the systems, control, and estimation literature.

Suppose we have a mathematically defined problem represented by f which acts on data d belonging to some set of data \mathcal{D} , to produce a solution $f(d)$ in a solution set \mathcal{S} . These notions are kept deliberately vague for expository purposes. Given $d \in \mathcal{D}$, we desire to compute $f(d)$. Suppose d^* is some approximation to d . If $f(d^*)$ is “near” $f(d)$, the problem is said to be well conditioned. If $f(d^*)$ may potentially differ greatly from $f(d)$ even when d^* is near d , the problem is said to be ill-conditioned. The concept of “near” can be made precise by introducing norms in the appropriate spaces. We can then define the condition of the problem f with respect to these norms as

$$\kappa[f(d)] = \lim_{\delta \rightarrow 0} \sup_{d_2(d, d^*) = \delta} \left[\frac{d_1(f(d), f(d^*))}{\delta} \right], \quad (1.7)$$

where $d_i(\cdot, \cdot)$ are distance functions in the appropriate spaces. When $\kappa[f(d)]$ is infinite, the problem of determining $f(d)$ from d is *ill-posed* (as opposed to *well-posed*). When $\kappa[f(d)]$ is finite and *relatively large* (or *relatively small*), the problem is said to be *ill-conditioned* (or *well-conditioned*).

A simple example of an ill-conditioned problem is the following. Consider the $n \times n$ matrix

$$A = \begin{bmatrix} 0 & 1 & 0 & \cdot & \cdot & \cdot & 0 \\ \cdot & \cdot & \cdot & \cdot & \cdot & \cdot & \cdot \\ \cdot & & \cdot & \cdot & \cdot & \cdot & \cdot \\ \cdot & & & \cdot & \cdot & \cdot & 0 \\ \cdot & & & & \cdot & \cdot & 1 \\ 0 & \cdot & \cdot & \cdot & \cdot & \cdot & 0 \end{bmatrix},$$

with n eigenvalues at 0. Now, consider a small perturbation of the data (the n^2 elements of A) consisting of adding the number 2^{-n} to the first element in the last (n th) row of A . This perturbed matrix then has n distinct eigenvalues $\lambda_1, \dots, \lambda_n$ with $\lambda_k = 1/2 \exp(2k\pi j/n)$, where $j := \sqrt{-1}$. Thus, we see that this small perturbation in the data has been magnified by a factor on the order of 2^n resulting in a rather large perturbation in solving the problem of computing the eigenvalues of A . Further details and related examples can be found in [9,17].

Thus far, we have not mentioned how the problem f above (computing the eigenvalues of A in the example) was to be solved. Conditioning is a function solely of the problem itself. To solve a problem numerically, we must implement some numerical procedures or algorithms which we denote by f^* . Thus, given d , $f^*(d)$ is the result of applying the algorithm to d (for simplicity, we assume d is “representable”; a more general definition can be given when some approximation d^{**} to d must be used). The algorithm f^* is said to be numerically (backward) stable if, for all $d \in \mathcal{D}$, there exists $d^* \in \mathcal{D}$ near d so that $f^*(d)$ is near $f(d^*)$, ($f(d^*)$ = the exact solution of a nearby problem). If the problem is well-conditioned, then $f(d^*)$ is near $f(d)$ so that $f^*(d)$ is near $f(d)$ if f^* is numerically stable. In other words, f^* does not introduce any more sensitivity to perturbation than is inherent in the problem. Example 1.1 further illuminates this definition of stability which, on a first reading, can seem somewhat confusing.

Of course, one cannot expect a stable algorithm to solve an ill-conditioned problem any more accurately than the data warrant, but an unstable algorithm can produce poor solutions even to well-conditioned problems. Example 1.2, illustrates this phenomenon. There are thus two separate factors to consider in determining the accuracy of a computed solution $f^*(d)$. First, if the algorithm is stable, $f^*(d)$ is near $f(d^*)$, for some d^* , and second, if the problem is well conditioned, then, as above, $f(d^*)$ is near $f(d)$. Thus, $f^*(d)$ is near $f(d)$ and we have an “accurate” solution.

Rounding errors can cause unstable algorithms to give disastrous results. However, it would be virtually impossible to account for every rounding error made at every arithmetic operation in a complex series of calculations. This would constitute a *forward* error analysis. The concept of *backward* error analysis based on the definition of numerical stability given above provides a more practical alternative. To illustrate this, let us consider the singular value decomposition (SVD) of an arbitrary $m \times n$ matrix A with coefficients in \mathbb{R} or \mathbb{C} [8] (see also Section 1.3.3),

$$A = U \Sigma V^H. \tag{1.8}$$

Here U and V are $m \times m$ and $n \times n$ unitary matrices, respectively, and Σ is an $m \times n$ matrix of the form

$$\Sigma = \left[\begin{array}{c|c} \Sigma_r & 0 \\ \hline 0 & 0 \end{array} \right]; \quad \Sigma_r = \text{diag}\{\sigma_1, \dots, \sigma_r\} \tag{1.9}$$

with the *singular values* σ_i positive and satisfying $\sigma_1 \geq \sigma_2 \geq \dots \geq \sigma_r > 0$. The computation of this decomposition is, of course, subject to rounding errors. Denoting computed quantities by an overbar, for some

error matrix E_A ,

$$\bar{A} = A + E_A = \overline{U\Sigma V^H}. \quad (1.10)$$

The computed decomposition thus corresponds exactly to a *perturbed* matrix \bar{A} . When using the SVD algorithm available in the literature [8], this perturbation can be bounded by

$$\|E_A\| \leq \pi \epsilon \|A\|, \quad (1.11)$$

where ϵ is the machine precision and π is some quantity depending on the dimensions m and n , but reasonably close to 1 (see also [14, p. 74]). Thus, the *backward error* E_A induced by this algorithm has roughly the same norm as the *input error* E_i resulting, for example, when reading the data A into the computer. Then, according to the definition of numerical stability given above, when a bound such as that in Equation 1.11 exists for the error induced by a numerical algorithm, the algorithm is said to be *backward stable* [17]. Note that backward stability does not guarantee any bounds on the errors in the result \bar{U} , $\bar{\Sigma}$, and \bar{V} . In fact, this depends on how perturbations in the data (namely, $E_A = \bar{A} - A$) affect the resulting decomposition (namely, $E_U = \bar{U} - U$, $E_\Sigma = \bar{\Sigma} - \Sigma$, and $E_V = \bar{V} - V$). This is commonly measured by the condition $\kappa[f(A)]$.

Backward stability is a property of an algorithm, and the condition is associated with a problem and the specific data for that problem. The errors in the result depend on the stability of the algorithm used and the condition of the problem solved. A *good* algorithm should, therefore, be backward stable because the size of the errors in the result is then mainly due to the condition of the problem, not to the algorithm. An unstable algorithm, on the other hand, may yield a large error even when the problem is well conditioned.

Bounds of the type Equation 1.11 are obtained by an error analysis of the algorithm used, and the condition of the problem is obtained by a sensitivity analysis; for example, see [9,17].

We close this section with two simple examples to illustrate some of the concepts introduced.

Example 1.1:

Let x and y be two floating-point computer numbers and let $fl(x * y)$ denote the result of multiplying them in floating-point computer arithmetic. In general, the product $x * y$ requires more precision to be represented exactly than was used to represent x or y . But for most computers

$$fl(x * y) = x * y(1 + \delta), \quad (1.12)$$

where $|\delta| < \epsilon$ (= relative machine precision). In other words, $fl(x * y)$ is $x * y$ correct to within a unit in the last place. Another way to write Equation 1.12 is as follows:

$$fl(x * y) = x(1 + \delta)^{1/2} * y(1 + \delta)^{1/2}, \quad (1.13)$$

where $|\delta| < \epsilon$. This can be interpreted as follows: the computed result $fl(x * y)$ is the exact product of the two slightly perturbed numbers $x(1 + \delta)^{1/2}$ and $y(1 + \delta)^{1/2}$. The slightly perturbed data (not unique) may not even be representable as floating-point numbers. The representation of Equation 1.13 is simply a way of accounting for the roundoff incurred in the algorithm by an initial (small) perturbation in the data.

Example 1.2:

Gaussian elimination with no pivoting for solving the linear system of equations

$$Ax = b \tag{1.14}$$

is known to be numerically unstable; see for example [8] and Section 1.3. The following data illustrate this phenomenon. Let

$$A = \begin{bmatrix} 0.0001 & 1.000 \\ 1.000 & -1.000 \end{bmatrix}, \quad b = \begin{bmatrix} 1.000 \\ 0.000 \end{bmatrix}.$$

All computations are carried out in four-significant-figure decimal arithmetic. The “true answer” $x = A^{-1}b$ is

$$\begin{bmatrix} 0.9999 \\ 0.9999 \end{bmatrix}.$$

Using row 1 as the “pivot row” (i.e., subtracting $10,000 \times$ row 1 from row 2) we arrive at the equivalent triangular system

$$\begin{bmatrix} 0.0001 & 1.000 \\ 0 & -1.000 \times 10^4 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} = \begin{bmatrix} 1.000 \\ -1.000 \times 10^4 \end{bmatrix}.$$

The coefficient multiplying x_2 in the second equation should be $-10,001$, but because of roundoff, becomes $-10,000$. Thus, we compute $x_2 = 1.000$ (a good approximation), but back substitution in the equation

$$0.0001x_1 = 1.000 - fl(1.000 * 1.000)$$

yields $x_1 = 0.000$. This extremely bad approximation to x_1 is the result of numerical instability. The problem, it can be shown, is quite well conditioned.

1.3 Fundamental Problems in Numerical Linear Algebra

In this section, we give a brief overview of some of the fundamental problems in numerical linear algebra that serve as building blocks or “tools” for the solution of problems in systems, control, and estimation.

1.3.1 Linear Algebraic Equations and Linear Least-Squares Problems

Probably the most fundamental problem in numerical computing is the calculation of a vector x which satisfies the linear system

$$Ax = b, \tag{1.15}$$

where $A \in \mathbb{R}^{n \times n}$ (or $\mathbb{C}^{n \times n}$) and has rank n . A great deal is now known about solving Equation 1.15 in finite arithmetic both for the general case and for a large number of special situations, for example, see [8,9].

The most commonly used algorithm for solving Equation 1.15 with general A and small n (say $n \leq 1000$) is Gaussian elimination with some sort of pivoting strategy, usually “partial pivoting.” This amounts to factoring some permutation of the rows of A into the product of a unit lower triangular matrix L and an upper triangular matrix U . The algorithm is effectively stable, that is, it can be proved that the computed solution is near the exact solution of the system

$$(A + E)x = b \tag{1.16}$$

with $|e_{ij}| \leq \phi(n) \gamma \beta \epsilon$, where $\phi(n)$ is a modest function of n depending on details of the arithmetic used, γ is a “growth factor” (which is a function of the pivoting strategy and is usually—but not always—small), β behaves essentially like $\|A\|$, and ϵ is the machine precision. In other words, except for moderately pathological situations, E is “small”—on the order of $\epsilon \|A\|$.

The following question then arises. If, because of rounding errors, we are effectively solving Equation 1.16 rather than Equation 1.15, what is the relationship between $(A + E)^{-1}b$ and $A^{-1}b$? To answer this question, we need some elementary perturbation theory and this is where the notion of condition number arises. A condition number for Equation 1.15 is given by

$$\kappa(A) := \|A\| \|A^{-1}\|. \quad (1.17)$$

Simple perturbation results can show that perturbation in A and/or b can be magnified by as much as $\kappa(A)$ in the computed solution. Estimating $\kappa(A)$ (since, of course, A^{-1} is unknown) is thus a crucial aspect of assessing solutions of Equation 1.15 and the particular estimating procedure used is usually the principal difference between competing linear equation software packages. One of the more sophisticated and reliable condition estimators presently available is implemented in LINPACK [5] and its successor LAPACK [2]. LINPACK and LAPACK also feature many codes for solving Equation 1.14 in case A has certain special structures (e.g., banded, symmetric, or positive definite).

Another important class of linear algebra problems, and one for which codes are available in LINPACK and LAPACK, is the linear least-squares problem

$$\min \|Ax - b\|_2, \quad (1.18)$$

where $A \in \mathbb{R}^{m \times n}$ and has rank k , with (in the simplest case) $k = n \leq m$, for example, see [8]. The solution of Equation 1.18 can be written formally as $x = A^+b$. The method of choice is generally based on the QR factorization of A (for simplicity, let $\text{rank}(A) = n$)

$$A = QR, \quad (1.19)$$

where $R \in \mathbb{R}^{n \times n}$ is upper triangular and $Q \in \mathbb{R}^{m \times n}$ has orthonormal columns, that is, $Q^T Q = I$. With special care and analysis, the case $k < n$ can also be handled similarly. The factorization is effected through a sequence of Householder transformations H_i applied to A . Each H_i is symmetric and orthogonal and of the form $I - 2uu^T/u^T u$, where $u \in \mathbb{R}^m$ is specially chosen so that zeros are introduced at appropriate places in A when it is premultiplied by H_i . After n such transformations,

$$H_n H_{n-1} \dots H_1 A = \begin{bmatrix} R \\ 0 \end{bmatrix},$$

from which the factorization Equation 1.19 follows. Defining c and d by

$$\begin{bmatrix} c \\ d \end{bmatrix} := H_n H_{n-1} \dots H_1 b,$$

where $c \in \mathbb{R}^n$, it is easily shown that the least-squares solution x of Equation 1.18 is given by the solution of the linear system of equations

$$Rx = c. \quad (1.20)$$

The above algorithm is numerically stable and, again, a well-developed perturbation theory exists from which condition numbers can be obtained, this time in terms of

$$\kappa(A) := \|A\| \|A^+\|.$$

Least-squares perturbation theory is fairly straightforward when $\text{rank}(A) = n$, but is considerably more complicated when A is rank deficient. The reason for this is that, although the inverse is a continuous

function of the data (i.e., the inverse is a continuous function in a neighborhood of a nonsingular matrix), the pseudoinverse is discontinuous. For example, consider

$$A = \begin{bmatrix} 1 & 0 \\ 0 & 0 \end{bmatrix} = A^+$$

and perturbations

$$E_1 = \begin{bmatrix} 0 & 0 \\ \delta & 0 \end{bmatrix} \quad \text{and} \quad E_2 = \begin{bmatrix} 0 & 0 \\ 0 & \delta \end{bmatrix}$$

with δ being small. Then

$$(A + E_1)^+ = \begin{bmatrix} \frac{1}{1 + \delta^2} & \frac{\delta}{1 + \delta^2} \\ 0 & 0 \end{bmatrix},$$

which is close to A^+ but

$$(A + E_2)^+ = \begin{bmatrix} 1 & 0 \\ 0 & \frac{1}{\delta} \end{bmatrix},$$

which gets arbitrarily far from A^+ as δ is decreased toward 0.

In lieu of Householder transformations, Givens transformations (elementary rotations or reflections) may also be used to solve the linear least-squares problem [8]. Givens transformations have received considerable attention for solving linear least-squares problems and systems of linear equations in a parallel computing environment. The capability of introducing zero elements selectively and the need for only local interprocessor communication make the technique ideal for “parallelization.”

1.3.2 Eigenvalue and Generalized Eigenvalue Problems

In the algebraic eigenvalue/eigenvector problem for $A \in \mathbb{R}^{n \times n}$, one seeks nonzero solutions $x \in \mathbb{C}^n$ and $\lambda \in \mathbb{C}$, which satisfy

$$Ax = \lambda x. \tag{1.21}$$

The classic reference on the numerical aspects of this problem is Wilkinson [17]. A briefer textbook introduction is given in [8].

Quality mathematical software for eigenvalues and eigenvectors is available; the EISPACK [7,15] collection of subroutines represents a pivotal point in the history of mathematical software. The successor to EISPACK (and LINPACK) is LAPACK [2], in which the algorithms and software have been restructured to provide high efficiency on vector processors, high-performance workstations, and shared memory multiprocessors.

The most common algorithm now used to solve Equation 1.21 for general A is the QR algorithm of Francis [17]. A shifting procedure enhances convergence and the usual implementation is called the double-Francis-QR algorithm. Before the QR process is applied, A is initially reduced to upper Hessenberg form A_H ($a_{ij} = 0$ if $i - j \geq 2$). This is accomplished by a finite sequence of similarities of the Householder form discussed above. The QR process then yields a sequence of matrices orthogonally similar to A and converging (in some sense) to a so-called quasi-upper triangular matrix S also called the real Schur form (RSF) of A . The matrix S is block upper triangular with 1×1 diagonal blocks corresponding to real eigenvalues of A and 2×2 diagonal blocks corresponding to complex-conjugate pairs of eigenvalues. The quasi-upper triangular form permits all arithmetic to be real rather than complex as would be necessary for convergence to an upper triangular matrix. The orthogonal transformations from both the Hessenberg reduction and the QR process may be accumulated in a single orthogonal transformation U so that

$$U^T A U = R \tag{1.22}$$

compactly represents the entire algorithm. An analogous process can be applied in the case of symmetric A , and considerable simplifications and specializations result.

Closely related to the QR algorithm is the QZ algorithm for the generalized eigenvalue problem

$$Ax = \lambda Mx, \quad (1.23)$$

where $A, M \in \mathbb{R}^{n \times n}$. Again, a Hessenberg-like reduction, followed by an iterative process, is implemented with orthogonal transformations to reduce Equation 1.23 to the form

$$QAZy = \lambda QMZy, \quad (1.24)$$

where QAZ is quasi-upper triangular and QMZ is upper triangular. For a review and references to results on stability, conditioning, and software related to Equation 1.23 and the QZ algorithm, see [8]. The generalized eigenvalue problem is both theoretically and numerically more difficult to handle than the ordinary eigenvalue problem, but it finds numerous applications in control and system theory [14, p. 109].

1.3.3 The Singular Value Decomposition and Some Applications

One of the basic and most important tools of modern numerical analysis, especially numerical linear algebra, is the SVD. Here we make a few comments about its properties and computation as well as its significance in various numerical problems.

Singular values and the SVD have a long history, especially in statistics and numerical linear algebra. These ideas have found applications in the control and signal processing literature, although their use there has been overstated somewhat in certain applications. For a survey of the SVD, its history, numerical details, and some applications in systems and control theory, see [14, p. 74].

The fundamental result was stated in Section 1.2 (for the complex case). The result for the real case is similar and is stated below.

Theorem 1.1:

Let $A \in \mathbb{R}^{m \times n}$ with $\text{rank}(A) = r$. Then there exist orthogonal matrices $U \in \mathbb{R}^{m \times m}$ and $V \in \mathbb{R}^{n \times n}$ so that

$$A = U\Sigma V^T, \quad (1.25)$$

where

$$\Sigma = \begin{bmatrix} \Sigma_r & 0 \\ 0 & 0 \end{bmatrix}$$

and $\Sigma_r = \text{diag} \{ \sigma_1, \dots, \sigma_r \}$ with $\sigma_1 \geq \dots \geq \sigma_r > 0$.

The proof of Theorem 1.1 is straightforward and can be found, for example, in [8]. Geometrically, the theorem says that bases can be found (separately) in the domain and codomain spaces of a linear map with respect to which the matrix representation of the linear map is diagonal. The numbers $\sigma_1, \dots, \sigma_r$, together with $\sigma_{r+1} = 0, \dots, \sigma_n = 0$, are called the singular values of A , and they are the positive square roots of the eigenvalues of $A^T A$. The columns $\{u_k, k = 1, \dots, m\}$ of U are called the left singular vectors of A (the orthonormal eigenvectors of AA^T), while the columns $\{v_k, k = 1, \dots, n\}$ of V are called the right singular vectors of A (the orthonormal eigenvectors of $A^T A$). The matrix A can then be written (as a dyadic expansion) also in terms of the singular vectors as follows:

$$A = \sum_{k=1}^r \sigma_k u_k v_k^T.$$

The matrix A^T has m singular values, the positive square roots of the eigenvalues of AA^T . The r [= $\text{rank}(A)$] nonzero singular values of A and A^T are, of course, the same. The choice of $A^T A$ rather than

AA^T in the definition of singular values is arbitrary. Only the nonzero singular values are usually of any real interest and their number, given the SVD, is the rank of the matrix. Naturally, the question of how to distinguish nonzero from zero singular values in the presence of rounding error is a nontrivial task.

It is not generally advisable to compute the singular values of A by first finding the eigenvalues of $A^T A$, tempting as that is. Consider the following example, where μ is a real number with $|\mu| < \sqrt{\epsilon}$ (so that $fl(1 + \mu^2) = 1$, where $fl(\cdot)$ denotes floating-point computation). Let

$$A = \begin{bmatrix} 1 & 1 \\ \mu & 0 \\ 0 & \mu \end{bmatrix}.$$

Then

$$fl(A^T A) = \begin{bmatrix} 1 & 1 \\ 1 & 1 \end{bmatrix}.$$

So we compute $\hat{\sigma}_1 = \sqrt{2}$, $\hat{\sigma}_2 = 0$ leading to the (erroneous) conclusion that the rank of A is 1. Of course, if we could compute in infinite precision, we would find

$$A^T A = \begin{bmatrix} 1 + \mu^2 & \mu \\ \mu & 1 + \mu^2 \end{bmatrix}$$

with $\sigma_1 = \sqrt{2 + \mu^2}$, $\sigma_2 = |\mu|$ and thus $\text{rank}(A) = 2$. The point is that by working with $A^T A$ we have unnecessarily introduced μ^2 into the computations. The above example illustrates a potential pitfall in attempting to form and solve the normal equations in a linear least-squares problem and is at the heart of what makes square root filtering so attractive numerically. Very simplistically, square root filtering involves working directly on an “ A -matrix,” for example, updating it, as opposed to updating an “ $A^T A$ -matrix.”

Square root filtering is usually implemented with the QR factorization (or some closely related algorithm) as described previously rather than SVD. Moreover, critical information may be lost irrecoverably by simply forming $A^T A$.

Returning now to the SVD, two features of this matrix factorization make it attractive in finite arithmetic: first, it can be computed in a numerically stable way, and second, singular values are well conditioned. Specifically, there is an efficient and numerically stable algorithm by Golub and Reinsch [8] which works directly on A to give the SVD. This algorithm has two phases. In the first phase, it computes orthogonal matrices U_1 and V_1 so that $B = U_1^T A V_1$ is in bidiagonal form, that is, only the elements on its diagonal and first superdiagonal are nonzero. In the second phase, the algorithm uses an iterative procedure to compute orthogonal matrices U_2 and V_2 so that $U_2^T B V_2$ is diagonal and nonnegative. The SVD defined in Equation 1.25 is then $\Sigma = U^T B V$, where $U = U_1 U_2$ and $V = V_1 V_2$. The computed U and V are orthogonal approximately to the working precision, and the computed singular values are the exact σ_i 's for $A + E$, where $\|E\|/\|A\|$ is a modest multiple of ϵ . Fairly sophisticated implementations of this algorithm can be found in [5,7]. The well-conditioned nature of the singular values follows from the fact that if A is perturbed to $A + E$, then it can be proved that

$$\|\sigma_i(A + E) - \sigma_i(A)\| \leq \|E\|.$$

Thus, the singular values are computed with small absolute error although the relative error of sufficiently small singular values is not guaranteed to be small.

It is now acknowledged that the singular value decomposition is the most generally reliable method of determining rank numerically (see [14, p. 589] for a more elaborate discussion). However, it is considerably more expensive to compute than, for example, the QR factorization which, with column pivoting [5], can usually give equivalent information with less computation. Thus, while the SVD is a useful theoretical tool, its use for actual computations should be weighed carefully against other approaches.

The problem of numerical determination of rank is now well understood. The essential idea is to try to determine a “gap” between “zero” and the “smallest nonzero singular value” of a matrix A . Since the computed values are exact for a matrix near A , it makes sense to consider the ranks of all matrices in some δ -ball (with respect to the spectral norm $\|\cdot\|$, say) around A . The choice of δ may also be based on measurement errors incurred in estimating the coefficients of A , or the coefficients may be uncertain because of rounding errors incurred in a previous computation. However, even with SVD, numerical determination of rank in finite arithmetic is a difficult problem.

That other methods of rank determination are potentially unreliable is demonstrated by the following example. Consider the Ostrowski matrix $A \in \mathbb{R}^{n \times n}$ whose diagonal elements are all -1 , whose upper triangle elements are all $+1$, and whose lower triangle elements are all 0. This matrix is clearly of rank n , that is, is invertible. It has a good “solid” upper triangular shape. All of its eigenvalues (-1) are well away from zero. Its determinant $(-1)^n$ is definitely not close to zero. But this matrix is, in fact, very nearly singular and becomes more nearly so as n increases. Note, for example, that

$$\begin{bmatrix} -1 & +1 & \cdots & \cdots & +1 \\ 0 & \ddots & \ddots & & \vdots \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ & & & \ddots & \ddots \\ \vdots & & & \ddots & \ddots & +1 \\ 0 & \cdots & \cdots & 0 & -1 \end{bmatrix} \begin{bmatrix} 1 \\ 2^{-1} \\ \vdots \\ 2^{-n+1} \end{bmatrix} = \begin{bmatrix} -2^{-n+1} \\ -2^{-n+1} \\ \vdots \\ -2^{-n+1} \end{bmatrix} \rightarrow \begin{bmatrix} 0 \\ 0 \\ \vdots \\ 0 \end{bmatrix} \quad (n \rightarrow +\infty).$$

Moreover, adding 2^{-n+1} to every element in the first column of A gives an exactly singular matrix. Arriving at such a matrix by, say, Gaussian elimination would give no hint as to the near singularity. However, it is easy to check that $\sigma_n(A)$ behaves as 2^{-n+1} . A corollary for control theory is that eigenvalues do not necessarily give a reliable measure of “stability margin.” It is useful to note that in this example of an invertible matrix, the crucial quantity, $\sigma_n(A)$, which measures nearness to singularity, is simply $1/\|A^{-1}\|$, and the result is familiar from standard operator theory. There is nothing intrinsic about singular values in this example and, in fact, $\|A^{-1}\|$ might be more cheaply computed or estimated in other matrix norms.

Because rank determination, in the presence of rounding error, is a nontrivial problem, the same difficulties naturally arise in any problem equivalent to, or involving, rank determination, such as determining the independence of vectors, finding the dimensions of certain subspaces, etc. Such problems arise as basic calculations throughout systems, control, and estimation theory. Selected applications are discussed in more detail in [14, p. 74] and in [1,4,10].

Finally, let us close this section with a brief example illustrating a totally inappropriate use of SVD. The rank condition

$$\text{rank } [B, AB, \dots, A^{n-1}B] = n \tag{1.26}$$

for the controllability of Equation 1.1 is too well known. Suppose

$$A = \begin{bmatrix} 1 & \mu \\ 0 & 1 \end{bmatrix}, \quad B = \begin{bmatrix} 1 \\ \mu \end{bmatrix}$$

with $|\mu| < \sqrt{\epsilon}$. Then

$$f[B, AB] = \begin{bmatrix} 1 & 1 \\ \mu & \mu \end{bmatrix},$$

and now even applying SVD, the erroneous conclusion of uncontrollability is reached. Again the problem is in just forming AB ; not even SVD can come to the rescue after that numerical *faux pas*.

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1 1.6 Concluding Remarks

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Among these measures are the quadratic norm condition number of the closed-loop modal matrix and

the sum of the squares of the quadratic norms of the left eigenvectors. Burrows et al. [4] have proposed a

stability augmentation system for a well-behaved light aircraft by using eigenstructure assignment with an

optimization which minimizes the condition number of the closed-loop modal matrix. Such an approach

may sometimes produce an acceptable controller. In contrast to eigensensitivity, Doyle and Stein [7]

have characterized the stability robustness of a multi-input, multi-output system by the minimum of the

smallest singular value of the return difference matrix at the plant input or output. Gavito and Collins [8]

have proposed an eigenstructure assignment design in which a constraint is placed on the minimum of

the smallest singular value of the return difference matrix at the inputs of both an L-1011 aircraft and

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books on the subject are [10,15,19,29,31]. For a comprehensive treatment of decentralized control theory,

For further information on vector Lyapunov functions and stability analysis of large scale intercon

nected systems, see the survey papers [22,33], and books [16,17].

Adaptive decentralized control has been of widespread recent interest, see [2,9,21,23,28,30,36]. "73648_C026" – #24

Robustness of decentralized control to both structured and unstructured perturbations has been one of

the central issues in the control of large scale systems. For the background of robustness issues in control,

which are relevant to decentralized control, see [18,30]. For new and interesting results on the subject,

see [4-6,8].

There is a number of papers devoted to design of decentralized control via parameter space optimiza

Overlapping decentralized control and the inclusion principle are surveyed in [30]. Useful extensions

were presented in [13]. The concept of overlapping is basic to reliable control under controller failures

using multiple decentralized controllers [30]. For more information about this area, see [7,12,32,35].

In a recent development [25], it has been shown how optimal decentralized control of large scale inter

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More information on the design and analysis of MRAC for linear, time-varying plants can be found in the following monograph:

6. Linear Time-Varying Systems: Control and Adaptation, by K. Tsakalis and P. Ioannou, Prentice Hall, Englewood Cliffs, NJ, 1993.

More information on the design and analysis of robust MRAC can be found in the books 1, 2, 3, and 6 given above.

Details on MRAC for discrete-time plants can be found in the book 4 by Landau given above and in:

7. Adaptive Filtering Prediction and Control by G. Goodwin and K. Sin, Prentice Hall, Englewood Cliffs, NJ, 1984.

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For Further Information

Our presentation of the asymptotic methods is based on the textbook by Khalil [1]. For further information

on this topic, the reader is referred to Chapters 10 and 11 of Khalil's book. Chapter 10 covers the

perturbation method and averaging, and Chapter 11 covers the singular perturbation method. Proofs of

the results stated here are given in the book.

The discussion of model order reduction is based on Chapter 1 of Kokotovic et al. [2]. This book gives a

broader view of the use of singular perturbation methods in systems and control. Modeling two timescale "73648_C039" - #13

systems in the singularly perturbed form is discussed in Chapters 1 and 2 of Kokotovic [2] and Chapter 11

of Khalil [1].

For a broader view of the averaging method, the reader may consult Sanders, J.A., Verhulst, F. and

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For Further Information

While the books and articles referenced above (particularly, Passino, 2005) should provide the reader

with an introduction to the area of intelligent control, there are other sources that may also be useful. For

instance, there are many relevant conferences including: (1) IEEE International Symposium on Intelligent

Control, (2) American Control Conference, (3) IEEE Conference on Decision and Control, (4) IEEE Confer

ence on Control Applications, and (5) IEEE International Conference on Systems, Man, and Cybernetics.

In addition, there are many conferences on fuzzy systems, expert systems, genetic algorithms, and neural

networks where applications to control are often studied. There are many journals that cover the topic

of intelligent control including: (1) IEEE Control Systems Magazine, (2) IEEE Trans. on Control Systems

Technology, (3) IEEE Trans. on Systems, Man, and Cybernetics, (4) IEEE Trans. on Fuzzy Systems, (5) IEEE

Trans. on Neural Networks, (6) Engineering Applications of Artificial Intelligence, (7) Journal of Intelligent

and Robotic Systems, (8) Applied Artificial Intelligence, and (9) Journal of Intelligent and Fuzzy Systems.

There are many other journals on expert systems, neural networks, genetic algorithms, and fuzzy systems

where applications to control can often be found. The professional societies most active in intelligent

control are the IEEE Control Systems Society, International Federation on Automatic Control, and the

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System

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Further Reading

For background on probability theory, especially conditional expectation, we recommend [2].

For linear SDEs driven by orthogonal-increments processes, we recommend the very readable text by

Davis [4].

For SDEs driven by continuous martingales, there is the more advanced book by Karatzas and

Shreve [7].

For SDEs driven by right-continuous martingales, the theory becomes considerably more complicated.

However, the tutorial paper by Segall [9], which compares discrete-time and continuous-time results, is

very readable. Also, Chapter 6 of [10] is accessible.

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18

\mathcal{H}_2 (LQG) and \mathcal{H}_∞ Control

Leonard Lublin

Massachusetts Institute of Technology

Simon Grocott

Massachusetts Institute of Technology

Michael Athans

Massachusetts Institute of Technology

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18.1 Introduction

The fundamentals of output feedback \mathcal{H}_2 (linear quadratic Gaussian or LQG) and \mathcal{H}_∞ controllers, which are the primary synthesis tools available for linear time-invariant systems, are presented in an analogous and tutorial fashion without rigorous mathematics. Since \mathcal{H}_2 and \mathcal{H}_∞ syntheses are carried out in the modern control design paradigm, a review of the paradigm is presented, along with the definitions of the \mathcal{H}_2 and \mathcal{H}_∞ norms and the methods used to compute them. The state-space formulae for the optimal controllers, under less restrictive assumptions than are usually found in the literature, are provided in an analogous fashion to emphasize the similarities between them. Rather than emphasizing the derivation of the controllers, we elaborate on the physical interpretation of the results and how one uses frequency weights to design \mathcal{H}_∞ and \mathcal{H}_2 controllers. Finally, a simple disturbance rejection design for the longitudinal motion of an aircraft is provided to illustrate the similarities and differences between \mathcal{H}_∞ and \mathcal{H}_2 controller synthesis.

18.2 The Modern Paradigm

\mathcal{H}_2 and \mathcal{H}_∞ syntheses are carried out in the modern control paradigm. In this paradigm both performance and robustness specifications can be incorporated in a common framework along with the controller synthesis. In the modern paradigm, all of the information about a system is cast into the generalized block diagram shown in Figure 18.1 [1–3]. The generalized plant, P , which is assumed to be linear and time-invariant throughout this article contains all the information a designer would like to incorporate into the synthesis of the controller, K . System dynamics, models of the uncertainty in the system's dynamics, frequency weights to influence the controller synthesis, actuator dynamics, sensor dynamics, and implementation hardware dynamics from amplifiers, and analog-to-digital and digital-to-analog converters are all included in P . The inputs and outputs of P are, in general, vector valued signals. The sensor measurements that are used by the feedback controller are denoted y , and the inputs generated

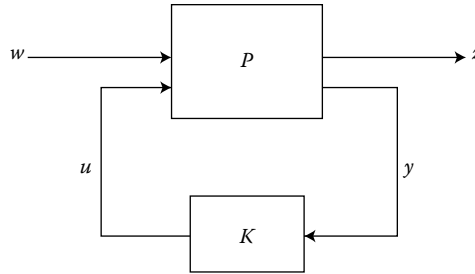


FIGURE 18.1 Generalized block diagram of the modern paradigm.

by the controller are denoted u . The components of w are all the exogenous inputs to the system. Typically these consist of disturbances, sensor noise, reference commands, and fictitious signals that drive frequency weights and models of the uncertainty in the dynamics of the system. The components of z are all the variables we wish to control. These include the performance variables of interest, tracking errors between reference signals and plant outputs, and the actuator signals which cannot be arbitrarily large and fast.

The general control problem in this framework is to synthesize a controller that will keep the size of the performance variables, z , small in the presence of the exogenous signals, w . For a classical disturbance rejection problem, z would contain the performance variables we wish to keep small in the presence of the disturbances contained in w that would tend to drive z away from zero. Hence, the disturbance rejection performance would depend on the “size” of the closed-loop transfer function from w to z , which we shall denote as $T_{zw}(s)$. This is also true for a command following control problem in which z would contain the tracking error that we would like to keep small in the presence of the commands in w that drive the tracking error away from zero.

Clearly then, the “size” of $T_{zw}(s)$ influences the effect that the exogenous signals in w have on z . Thus, in this framework, we seek controllers that minimize the “size” of the closed-loop transfer function $T_{zw}(s)$. Given that $T_{zw}(s)$ is a transfer function matrix, it is necessary to use appropriate norms to quantify its size. The two most common and physically meaningful norms that are used to classify the “size” of $T_{zw}(s)$ are the \mathcal{H}_2 and \mathcal{H}_∞ norms. As such, we seek controllers that minimize either the \mathcal{H}_2 or \mathcal{H}_∞ norm of $T_{zw}(s)$ in the modern control paradigm.

18.2.1 System Norms

Here we define and discuss the \mathcal{H}_2 and \mathcal{H}_∞ norms of the linear, time-invariant, stable system with transfer function matrix

$$G(s) = C(sI - A)^{-1}B$$

This notation is meant to be general, and the reader should not think of $G(s)$ as only the actuator to sensor transfer function of a system. Realize that $G(s)$ is a system and thus requires an appropriate norm to classify its size. By a norm, we mean a positive, scalar number that is a measure of the size of $G(s)$ over all points in the complex s -plane. This is quite different from, for example, the maximum singular value of a matrix, $\sigma_{\max}[A]$, which is a norm that classifies the size of the matrix A .

The \mathcal{H}_2 Norm

Definition 18.1: \mathcal{H}_2 Norm

The \mathcal{H}_2 norm of $G(s)$, denoted $\|G\|_2$, is defined as

$$\|G\|_2 = \left(\frac{1}{2\pi} \int_{-\infty}^{\infty} \text{trace} [G(j\omega)G^*(j\omega)] d\omega \right)^{\frac{1}{2}} = \left(\frac{1}{2\pi} \int_{-\infty}^{\infty} \sum_{i=1}^r \sigma_i^2 [G(j\omega)] d\omega \right)^{\frac{1}{2}}$$

where σ_i denotes the i th singular value, $G^*(j\omega)$ is the complex conjugate transpose of $G(j\omega)$, and r is the rank of $G(j\omega)$.

The \mathcal{H}_2 norm has an attractive, physically meaningful interpretation. If we consider $G(s)$ to be the transfer function matrix of a system driven by independent, zero mean, unit intensity white noise, u , then the sum of the variances of the outputs y is exactly the square of the \mathcal{H}_2 norm of $G(s)$. That is

$$\mathbb{E} [y^T(t)y(t)] = \|G(s)\|_2^2 \quad (18.1)$$

The \mathcal{H}_2 norm of $G(s)$ thus gives a precise measure of the “power” or signal strength of the output of a system driven with unit intensity white noise. Note that in the scalar case $\sqrt{\mathbb{E}[y^T(t)y(t)]}$ is the RMS or root mean squared value for $y(t)$ so the \mathcal{H}_2 norm specifies the RMS value of $y(t)$. A well-known fact for stochastic systems is that the mean squared value of the outputs can be computed by solving the appropriate Lyapunov equation [4]. As such, a state space procedure for computing the \mathcal{H}_2 norm of $G(s)$ is as follows [2].

Computing the \mathcal{H}_2 Norm

If L_c denotes the controllability Gramian of (A, B) and L_o the observability Gramian of (A, C) , then

$$AL_c + L_c A^T + BB^T = 0 \quad A^T L_o + L_o A + C^T C = 0$$

and

$$\|G\|_2 = \left[\text{trace}(CL_c C^T) \right]^{\frac{1}{2}} = \left[\text{trace}(B^T L_o B) \right]^{\frac{1}{2}}$$

Note that this procedure for computing the \mathcal{H}_2 norm involves the solution of linear Lyapunov equations and can be done without iteration.

The \mathcal{H}_∞ Norm

Definition 18.2: \mathcal{H}_∞ Norm

The \mathcal{H}_∞ norm of $G(s)$, denoted $\|G\|_\infty$, is defined as

$$\|G\|_\infty = \sup_{\omega} \sigma_{\max}[G(j\omega)]$$

In this definition “sup” denotes the supremum or least upper bound of the function $\sigma_{\max}[G(j\omega)]$, and thus the \mathcal{H}_∞ norm of $G(s)$ is nothing more than the maximum value of $\sigma_{\max}[G(j\omega)]$ over all frequencies ω . The supremum must be used in the definition since, strictly speaking, the maximum of $\sigma_{\max}[G(j\omega)]$ may not exist even though $\sigma_{\max}[G(j\omega)]$ is bounded from above.

\mathcal{H}_∞ norms also have a physically meaningful interpretation when considering the system $y(s) = G(s)u(s)$. Recall that when the system is driven with a unit magnitude sinusoidal input at a specific frequency, $\sigma_{\max}[G(j\omega)]$ is the largest possible output size for the corresponding sinusoidal output. Thus, the \mathcal{H}_∞ norm is the largest possible amplification over all frequencies of a unit sinusoidal input. That is, it classifies the greatest increase in energy that can occur between the input and output of a given system. A state space procedure for calculating the \mathcal{H}_∞ norm is as follows.

Computing the \mathcal{H}_∞ Norm

Let $\|G\|_\infty = \gamma_{\min}$. For the transfer function $G(s) = C(sI - A)^{-1}B$ with A stable and $\gamma > 0$, $\|G\|_\infty < \gamma$ if and only if the Hamiltonian matrix

$$H = \begin{bmatrix} A & \frac{1}{\gamma^2}BB^T \\ -C^TC & -A^T \end{bmatrix}$$

has no eigenvalues on the $j\omega$ -axis. This fact lets us compute a bound, γ , on $\|G\|_\infty$ such that $\|G\|_\infty < \gamma$. So to find γ_{\min} , select a $\gamma > 0$ and test if H has eigenvalues on the $j\omega$ -axis. If it does, increase γ . If it does not, decrease γ and recompute the eigenvalues of H . Continue until γ_{\min} is calculated to within the desired tolerance.

The iterative computation of the \mathcal{H}_∞ norm, which can be carried out efficiently using a bisection search over γ , is to be expected given that by definition we must search for the largest value of $\sigma_{\max}[G(j\omega)]$ over all frequencies.

Note, the \mathcal{H}_2 norm is not an induced norm, whereas the \mathcal{H}_∞ norm is. Thus, the \mathcal{H}_2 norm does not obey the submultiplicative property of induced norms. That is, the \mathcal{H}_∞ norm satisfies

$$\|G_1G_2\|_\infty \leq \|G_1\|_\infty\|G_2\|_\infty$$

but the \mathcal{H}_2 norm does not have the analogous property. This fact makes synthesizing controllers that minimize $\|T_{zw}(s)\|_\infty$ attractive when one is interested in directly shaping loops to satisfy norm bounded robustness tests*. On the other hand, given the aforementioned properties of the \mathcal{H}_2 norm, synthesizing controllers that minimize $\|T_{zw}(s)\|_2$ is attractive when the disturbances, w , are stochastic in nature. In fact, \mathcal{H}_2 controllers are nothing more than linear quadratic Gaussian (LQG) controllers so the vast amount of insight into the well-understood LQG problem can be readily applied to \mathcal{H}_2 synthesis.

Example 18.1:

In this example the \mathcal{H}_2 and \mathcal{H}_∞ norms are calculated for the simple four-spring, four-mass system shown in Figure 18.2. The equations of motion for this system can be found in Example 17.2 in Chapter 17. The system has force inputs on the second and fourth masses along with two sensors that provide a measure of the displacement of these masses. The singular values of the transfer function from the inputs to outputs, which we denote by $G(s)$, are shown in Figure 18.3. The \mathcal{H}_∞ norm of the system is equal to the peak of $\sigma_1 = 260.4$, and the \mathcal{H}_2 norm of the system is equal to the square root of the sum of the areas under the square of each of the singular values, 14.5. Note that when considering the \mathcal{H}_2 norm, observing the log log plot of the transfer function can be very deceiving, since the integral is of σ_j , not $\log(\sigma_j)$, over ω , not $\log \omega$.

As pointed out in the example, the differences between the \mathcal{H}_∞ and \mathcal{H}_2 norms for a system $G(s)$ are best viewed in the frequency domain from a plot of the singular values of $G(j\omega)$. Specifically, the \mathcal{H}_∞ norm is the peak value of $\sigma_{\max}[G(j\omega)]$ while the \mathcal{H}_2 norm is related to the area underneath the singular values of $G(j\omega)$. For a more in-depth treatment of these norms the reader is referred to [1,2,5,6].

* See Chapter 20 for a detailed exposition of this concept.

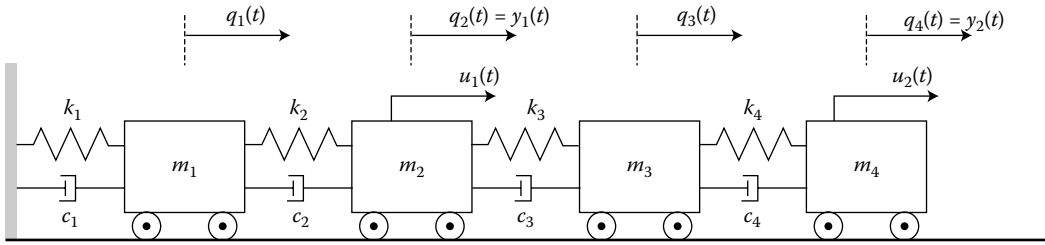


FIGURE 18.2 Mass, spring, dashpot system from Example 18.1. For the example $k_i = m_i = 1 \forall i$, and $c_i = 0.05 \forall i$.

18.3 Output Feedback \mathcal{H}_∞ and \mathcal{H}_2 Controllers

Given that all the information a designer would like to include in the controller synthesis is incorporated into the system P , the synthesis of \mathcal{H}_2 and \mathcal{H}_∞ controllers is quite straightforward. In this respect all of the design effort is focused on defining P . Below, we discuss how to define P using frequency weights to meet typical control system specifications. Here we simply present the formulas for the controllers.

All the formulas will be based on the following state-space realization of P ,

$$P := \left[\begin{array}{c|cc} A & B_1 & B_2 \\ \hline C_1 & D_{11} & D_{12} \\ C_2 & D_{21} & D_{22} \end{array} \right]$$

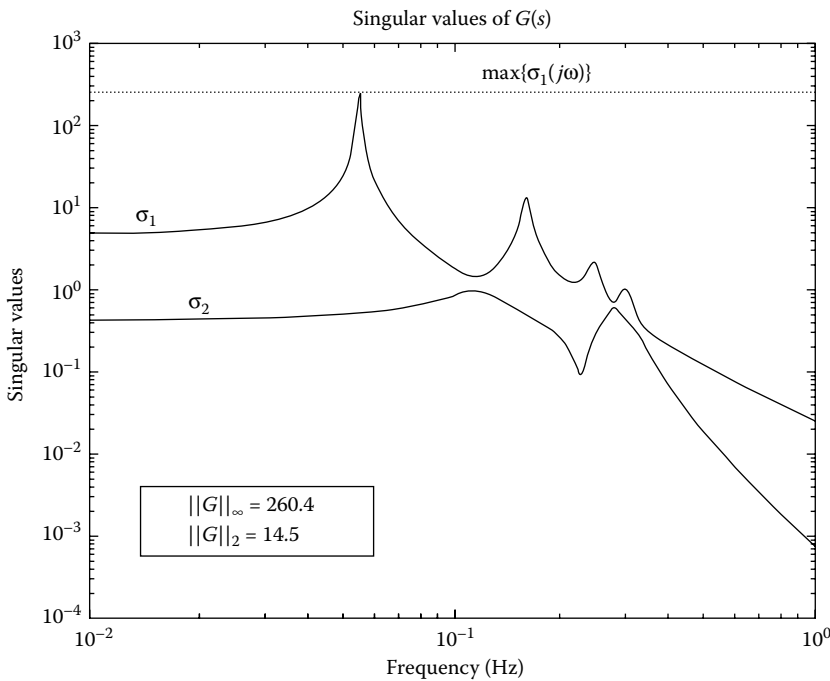


FIGURE 18.3 Singular values of the transfer function between the inputs and outputs of the mass, spring system shown in Figure 18.2.

This notation is a shorthand representation for the system of equations

$$\dot{x}(t) = Ax(t) + B_1w(t) + B_2u(t) \quad (18.2)$$

$$z(t) = C_1x(t) + D_{11}w(t) + D_{12}u(t) \quad (18.3)$$

$$y(t) = C_2x(t) + D_{21}w(t) + D_{22}u(t) \quad (18.4)$$

Additionally, the following assumptions concerning the allowable values for the elements of P are made.

Assumptions on P

$$1. \quad D_{11} = 0 \quad (A.1)$$

$$2. \quad [A \quad B_2] \text{ is stabilizable} \quad (A.2)$$

$$3. \quad [A \quad C_2] \text{ is detectable} \quad (A.3)$$

$$4. \quad V = \begin{bmatrix} B_1 \\ D_{21} \end{bmatrix} \begin{bmatrix} B_1^T & D_{21}^T \end{bmatrix} := \begin{bmatrix} V_{xx} & V_{xy} \\ V_{xy}^T & V_{yy} \end{bmatrix} \geq 0 \quad \text{with } V_{yy} > 0 \quad (A.4)$$

$$5. \quad R = \begin{bmatrix} C_1^T \\ D_{12}^T \end{bmatrix} \begin{bmatrix} C_1 & D_{12} \end{bmatrix} := \begin{bmatrix} R_{xx} & R_{xu} \\ R_{xu}^T & R_{uu} \end{bmatrix} \geq 0 \quad \text{with } R_{uu} > 0 \quad (A.5)$$

Assumption A.1 ensures that none of the disturbances feed through to the performance variables which is necessary for \mathcal{H}_2 synthesis but may be removed for \mathcal{H}_∞ synthesis (see [7] for details.) Assumptions A.2 and A.3 are needed to guarantee the existence of a stabilizing controller while the remaining assumptions are needed to guarantee the existence of positive semidefinite solutions to the Riccati equations associated with the optimal controllers.

Theorem 18.1: \mathcal{H}_2 Output Feedback

Assuming that $w(t)$ is a unit intensity white noise signal, $E[w(t)w^T(\tau)] = I\delta(t - \tau)$, the unique, stabilizing, optimal controller which minimizes the \mathcal{H}_2 norm of $T_{zw}(s)$ is

$$K_2 := \left[\begin{array}{c|c} \frac{A + B_2F_2 + L_2C_2 + L_2D_{22}F_2}{F_2} & \begin{array}{c} -L_2 \\ 0 \end{array} \end{array} \right] \quad (18.5)$$

where

$$\begin{aligned} F_2 &= -R_{uu}^{-1} \left(R_{xu}^T + B_2^T X_2 \right) \\ L_2 &= - \left(Y_2 C_2^T + V_{xy} \right) V_{yy}^{-1} \end{aligned} \quad (18.6)$$

and X_2 and Y_2 are the unique, positive semidefinite solutions to the following Riccati equations

$$0 = X_2 A_r + A_r^T X_2 + R_{xx} - R_{xu} R_{uu}^{-1} R_{xu}^T - X_2 B_2 R_{uu}^{-1} B_2^T X_2 \quad (18.7)$$

$$0 = A_e Y_2 + Y_2 A_e^T + V_{xx} - V_{xy} V_{yy}^{-1} V_{xy}^T - Y_2 C_2^T V_{yy}^{-1} C_2 Y_2 \quad (18.8)$$

where

$$A_r = \left(A - B_2 R_{uu}^{-1} R_{xu}^T \right) \quad \text{and} \quad A_e = \left(A - V_{xy} V_{yy}^{-1} C_2 \right)$$

Theorem 18.2: \mathcal{H}_∞ Output Feedback [8]

Assuming that $w(t)$ is a bounded \mathcal{L}_2 signal, $\int_{-\infty}^{\infty} w^T(t)w(t) dt < \infty$, a stabilizing controller which satisfies $\|T_{zw}(j\omega)\|_\infty < \gamma$ is

$$K_\infty := \left[\begin{array}{c|c} A & -Z_\infty L_\infty \\ \hline F_\infty & 0 \end{array} \right] \quad (18.9)$$

where

$$A_\infty = A + (B_1 + L_\infty D_{21}) W_\infty + B_2 F_\infty + Z_\infty L_\infty C_2 + Z_\infty L_\infty D_{22} F_\infty$$

where

$$F_\infty = -R_{uu}^{-1} (R_{xu}^T + B_2^T X_\infty) \quad W_\infty = \frac{1}{\gamma^2} B_1^T X_\infty$$

$$L_\infty = - (Y_\infty C_2^T + V_{xy}) V_{yy}^{-1} \quad Z_\infty = \left(I - \frac{1}{\gamma^2} Y_\infty X_\infty \right)^{-1}$$

and X_∞ and Y_∞ are the solutions to the following Riccati equations

$$0 = X_\infty A_r + A_r^T X_\infty + R_{xx} - R_{xu} R_{uu}^{-1} R_{xu}^T - X_\infty \left(B_2 R_{uu}^{-1} B_2^T - \frac{1}{\gamma^2} B_1 B_1^T \right) X_\infty \quad (18.10)$$

$$0 = A_e Y_\infty + Y_\infty A_e^T + V_{xx} - V_{xy} V_{yy}^{-1} V_{xy}^T - Y_\infty \left(C_2^T V_{yy}^{-1} C_2 - \frac{1}{\gamma^2} C_1^T C_1 \right) Y_\infty \quad (18.11)$$

that satisfy the following conditions

1. $X_\infty \geq 0$
2. The Hamiltonian matrix for Equation 18.10,

$$\begin{bmatrix} A - B_2 R_{uu}^{-1} R_{xu}^T & -B_2 R_{uu}^{-1} B_2^T + \frac{1}{\gamma^2} B_1 B_1^T \\ -R_{xx} + R_{xu} R_{uu}^{-1} R_{xu}^T & - (A - B_2 R_{uu}^{-1} R_{xu}^T)^T \end{bmatrix}$$

has no $j\omega$ -axis eigenvalues, or equivalently $A + B_1 W_\infty + B_2 F_\infty$ is stable

3. $Y_\infty \geq 0$
4. The Hamiltonian matrix for Equation 18.11,

$$\begin{bmatrix} \left(A - V_{xy} V_{yy}^{-1} C_2 \right)^T & -C_2^T V_{yy}^{-1} C_2 + \frac{1}{\gamma^2} C_1^T C_1 \\ -V_{xx} + V_{xy} V_{yy}^{-1} V_{xy}^T & -A + V_{xy} V_{yy}^{-1} C_2 \end{bmatrix}$$

has no $j\omega$ -axis eigenvalues, or equivalently $A + L_\infty C_2 + \frac{1}{\gamma^2} Y_\infty C_1^T C_1$ is stable

5. $\rho(Y_\infty X_\infty) < \gamma^2$, where $\rho(\cdot) = \max_i |\lambda_i(\cdot)|$ is the spectral radius

The (sub)optimal central \mathcal{H}_∞ controller which minimizes $\|T_{zw}\|_\infty$ to within the desired tolerance is K_∞ with γ equal to the smallest value of $\gamma > 0$ that satisfies conditions 1 to 5.

Unlike the \mathcal{H}_2 controller, the \mathcal{H}_∞ controller presented here is not truly optimal. Since there is no closed-form, state-space solution to the problem of minimizing the infinity norm of a multiple-input,

multiple-output (MIMO) transfer function matrix $T_{zw}(s)$, the connections between the mini-max optimization problem

$$\inf_u \sup_w \int_0^\infty [z^T(t)z(t) - \gamma^2 w^T(t)w(t)] dt \tag{18.12}$$

and \mathcal{H}_∞ optimization are used to arrive at the constructive approach for synthesizing suboptimal \mathcal{H}_∞ controllers given in Theorem 18.2. In fact, satisfying the conditions 1 to 5 of Theorem 18.2 is analogous to finding a saddle point of the optimization problem (Equation 18.12), and the search for γ_{\min} is analogous to finding the global minimum over all the possible saddle points. As such, any value of $\gamma > \gamma_{\min}$ will also satisfy conditions 1 to 5 of Theorem 18.2, and thus produce a stabilizing controller. Such controllers are neither \mathcal{H}_2 nor \mathcal{H}_∞ optimal. Since in the limit as $\gamma \rightarrow \infty$ the equations from Theorem 18.2 reduce to the equations for the \mathcal{H}_2 optimal controller, controllers with values of γ between γ_{\min} and infinity provide a trade off between \mathcal{H}_∞ and \mathcal{H}_2 performance. Along these lines, it is also worth noting that there is a rich theory for mixed $\mathcal{H}_2/\mathcal{H}_\infty$ controllers that minimize the \mathcal{H}_2 norm of $T_{zw}(s)$ subject to additional \mathcal{H}_∞ constraints. See [9–11] for details.

The value of $w(t)$ that maximizes the cost in Equation 18.12 is known as the worst case disturbance, as it seeks to maximize the detrimental effect the disturbances have on the system. In this regard, \mathcal{H}_∞ controllers provide optimal disturbance rejection to worst case disturbance, whereas the \mathcal{H}_2 controllers provide optimal disturbance rejection to stochastic disturbances.

Both \mathcal{H}_2 and \mathcal{H}_∞ controllers are observer-based compensators [2], which can be seen from their block diagrams, shown in Figures 18.4 and 18.5. The regulator gains F_2 and F_∞ arise from synthesizing the full-state feedback controller, which minimizes the appropriate size of $z^T(t)z(t)$ constrained by the system dynamics Equation 18.2. Then the control law is formed by applying these regulator gains to an estimate of the states $x(t)$. The states, $x(t)$, are estimated using the noisy measurements of $y(t)$ from Equation 18.4, and L_2 and $Z_\infty L_\infty$ are the corresponding filter gains of the estimators.

In particular, F_2 is the full-state feedback LQR gain that minimizes the quadratic cost

$$J_{LQ} = E \left\{ \lim_{\tau \rightarrow \infty} \frac{1}{\tau} \int_0^\tau [z^T(t)z(t)] dt \right\}$$

constrained by the dynamics of Equation 18.2, and L_2 is the Kalman filter gain from estimating the states x based on the measurements $y(t)$. Under the assumption that $z(t)$ is an ergodic process*

$$J_{LQ} = \lim_{\tau \rightarrow \infty} \frac{1}{\tau} \int_0^\tau z^T(t)z(t)dt = E [z^T(t)z(t)] = \|T_{zw}\|_2^2 \tag{18.13}$$

and this is exactly why \mathcal{H}_2 synthesis is nothing more than LQG control.

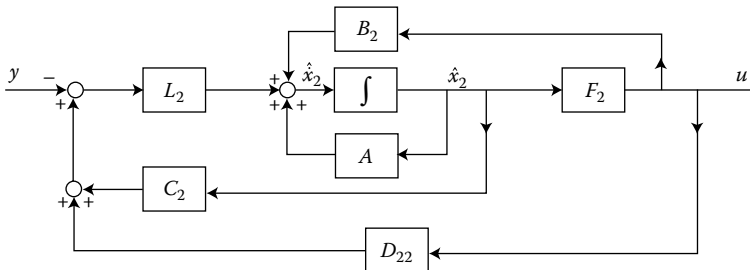


FIGURE 18.4 Block diagram of K_2 from Equation 18.9. Note, the Kalman Filter estimate of the states $x(t)$ from Equation 18.2, $\hat{x}_2(t)$, are the states of K_2 .

* Assuming $z(t)$ is ergodic implies that its mean can be computed from the time average of a measurement of $z(t)$ as $t \rightarrow \infty$ [4].

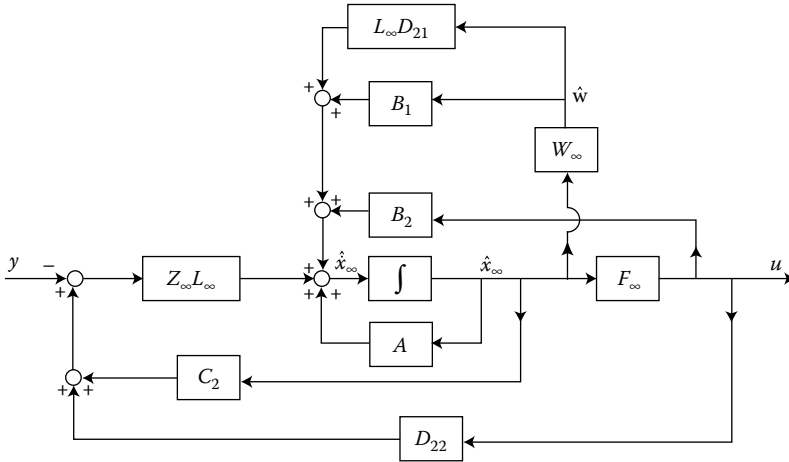


FIGURE 18.5 Block diagram of K_∞ from Equation 18.9. Note, the \mathcal{H}_∞ optimal estimate of the states $x(t)$ from Equation 18.2, $\hat{x}_\infty(t)$, are the states of K_∞ , and $\hat{w}(t)$ is an estimate of the worst case disturbance.

Analogously, F_∞ is the full-state feedback \mathcal{H}_∞ control gain that results from optimizing the mini-max cost of Equation 18.12, and W_∞ is the full-state feedback gain that produces the worst case disturbance which maximizes the cost of Equation 18.12*. Unlike the Kalman filter in the \mathcal{H}_2 controller, the \mathcal{H}_∞ optimal estimator must estimate the states of P in the presence of the worst case disturbance which is evident from the block diagram of K_∞ shown in Figure 18.5 [12]. This is why the filter gain of the \mathcal{H}_∞ optimal estimator, $Z_\infty L_\infty$, is coupled to the regulator portion of the problem through X_∞ from Equation 18.10.

Since the \mathcal{H}_2 controller is an LQG controller, the closed-loop poles of $T_{zw}(s)$ separate into the closed-loop poles of the regulator, $\text{eig}(A - B_2 F_2)$, and estimator, $\text{eig}(A - L_2 C_2)$. A consequence of this separation property is that the \mathcal{H}_2 Riccati equations (Equations 18.7 and 18.8) can be solved directly without iteration. Since the worst case disturbance must be taken into consideration when synthesizing the \mathcal{H}_∞ optimal estimator, the regulator and estimator problems in the \mathcal{H}_∞ synthesis are coupled. Thus, the \mathcal{H}_∞ controller does not have a separation structure that is analogous to that of the \mathcal{H}_2 controller. In addition, the \mathcal{H}_∞ Riccati equations (Equations 18.10 and 18.11) are further coupled through the γ parameter, and we must iterate over the value of γ to find solutions of the \mathcal{H}_∞ Riccati equations that satisfy conditions 1 to 5 of Theorem 18.2.

Note that in the literature the following set of additional, simplifying assumptions on the values of the elements of P are often made to arrive at less complicated sets of equations for the optimal \mathcal{H}_∞ and \mathcal{H}_2 controllers [6,13,14].

Additional Assumptions on P

1. $D_{22} = 0$ (No control feed-through term)
2. $C_1^T D_{12} = 0$ (No cross penalty on control and state)
3. $B_1 D_{21}^T = 0$ (Uncorrelated process and sensor noise)
4. $D_{12}^T D_{12} = I$ (Unity penalty on every control)
5. $D_{21}^T D_{21} = I$ (Unit intensity sensor noise on every measurement)

* See the section on \mathcal{H}_∞ Full State Feedback in Chapter 17 for details.

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V

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Further Reading

Contributions to the development of the Lyapunov design methodology for systems with no uncertainties include [1,9,10,13,18]. A good introduction to Lyapunov redesign can be found [12]. Corless [3] has recently surveyed various methods for the design of robust controllers for nonlinear systems using *quadratic* Lyapunov functions. Details of the recursive design presented in Section 49.4 can be found in [6, 14,16]. The state-space techniques discussed in this chapter can be combined with nonlinear input/output techniques to obtain more advanced designs (see Chapter 44). Finally, when the uncertain nonlinearities are given by *constant* parameters multiplying *known* nonlinearities, adaptive control techniques apply (see Chapter 53).

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Further Reading

Detailed discussions of bifurcation and chaos are available in many excellent books (e.g., [6,10,11,18,19,28,30,32,34,35,43,46,48]). These books also discuss a variety of interesting applications. Many examples of bifurcations in mechanical systems are given in [52]. There are also several journals devoted to bifurcations and chaos. Of particular relevance to engineers are *Nonlinear Dynamics*, the *International Journal of Bifurcation and Chaos*, and the journal *Chaos, Solitons and Fractals*.

The book [45] discusses feedforward control in the context of “trimming” an aircraft using its nonlinear equations of motion and the available controls. Bifurcation and chaos in flight dynamics are discussed in [8]. Lucid explanations on specific uses of washout filters in aircraft control systems are given in [15, pp. 474–475], [7, pp. 144–146], [39, pp. 946–948 and pp. 1087–1095], and [45, pp. 243–246 and p. 276]. The book [31] discussed applications of nonlinear dynamics in biology and population dynamics.

Computational issues related to bifurcation analysis are addressed in [43]. Classification of bifurcations as safe or dangerous is discussed in [32,43,48,49].

The edited book [14] contains interesting articles on research needs in applications of bifurcations and chaos.

The article [3] contains a large number of references on bifurcation control, related work on stabilization, and applications of these techniques. The review papers [9,44] address control of chaos methods. In particular, [44] includes a discussion of use of sensitive dependence on initial conditions to direct trajectories to targets. The book [35] includes articles on control of chaos, detection of chaos in time series, chaotic data analysis, and potential applications of chaos in communication systems. The book [32] also contains discussions of control of bifurcations and chaos, and of analysis of chaotic data.

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Further Reading

Other introductions to fuzzy control are given in [2–5]. The book [2] is available for a free download (as a .pdf file) from <http://www.ece.osu.edu/~passino/>. To study other applications of fuzzy control the reader should consult the references in the papers and books cited above. There are several conferences that have sessions or papers that study various aspects of fuzzy control. For instance, see the *Proceedings of FuzzIEEE* (in the World Congress on Computational Intelligence), *IEEE International Symposium on Intelligent Control*, *Proceedings of the IEEE Conference on Decision and Control*, the *American Control Conference*, and many others. The best journals to consult are the *IEEE Transactions on Fuzzy Systems*, *Fuzzy Sets and Systems*, and the *IEEE Transactions on Systems, Man, and Cybernetics*. Also, there are many short courses that are given on the topic of fuzzy control.

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Further Reading

For discounted dynamic programming, the classic reference is Blackwell [1]. For the positive cost case, we refer the reader to Strauch [2]. For the negative cost case, we refer the reader to Blackwell [3,4] and Ornstein [5]. For the average cost case, we refer the reader to Blackwell [6] for early fundamental work, and to Hernandez-Lerma and Lasserre [7] and the references contained there for recent developments. For a study of the measurability issues which arise when considering uncountable sets, we refer the reader to Bertsekas and Shreve [8]. For continuous time stochastic control of diffusions we refer the reader to Lions [9], and to Crandall, Ishii and Lions [10] for a guide to viscosity solutions.

For the several ways in which dynamic programming can be employed, see Bellman [11].

Some recent textbooks which cover dynamic programming are Bertsekas [12], Kumar and Varaiya [13], and Ross [14].

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XII

Networks and Networked Controls

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