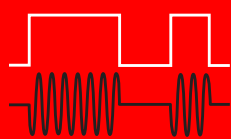


MICROELECTRONIC CIRCUIT DESIGN





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RICHARD C. JAEGER

Auburn University

TRAVIS N. BLALOCK

University of Virginia

BENJAMIN J. BLALOCK

University of Tennessee, Knoxville

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Product Developer: *Erin Kamm*

Marketing Manager: *Lisa Granger*

Content Project Managers: *Jane Mohr and Samantha Donisi*

Buyer: *Sandy Ludovissy*

Designer: *David Hash*

Content Licensing Specialist: *Beth Cray*

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TO

To Joan, my loving wife and life-long partner

—Richard C. Jaeger

In memory of my father, Professor Theron Vaughn Blalock,
an inspiration to me and to the countless students whom he
mentored both in electronic design and in life.

—Travis N. Blalock

To my family, for their love, support, and inspiration.

—Benjamin J. Blalock

BRIEF CONTENTS

Preface xx

Chapter-by-Chapter Summary xxiii

PART ONE

SOLID-STATE ELECTRONICS AND DEVICES

- 1 Introduction to Electronics 3
- 2 Solid-State Electronics 42
- 3 Solid-State Diodes and Diode Circuits 73
- 4 Bipolar Junction Transistors 146
- 5 Field-Effect Transistors 212

PART TWO

ANALOG ELECTRONICS

- 6 Introduction to Amplifiers 313
- 7 The Transistor as an Amplifier 352
- 8 Transistor Amplifier Building Blocks 424
- 9 Amplifier Frequency Response 544

PART THREE

OPERATIONAL AMPLIFIERS AND FEEDBACK

- 10 Ideal Operational Amplifiers 649
- 11 Nonideal Operational Amplifiers and Feedback Amplifier Stability 683
- 12 Operational Amplifier Applications 781

- 13 Differential Amplifiers and Operational Amplifier Design 866

- 14 Analog Integrated Circuit Design Techniques 946

- 15 Transistor Feedback Amplifiers and Oscillators 1030

SUPPLEMENTAL CHAPTERS (E-Book Only)

DIGITAL ELECTRONICS

- S6 Introduction to Digital Electronics S6-3

- S7 Complementary MOS (CMOS) Logic Design S7-1

- S8 MOS Memory Circuits S8-1

- S9 Bipolar Logic Circuits S9-1

APPENDICES

- A Standard Discrete Component Values 1109

- B Solid-State Device Models and SPICE Simulation Parameters 1112

- C Two-Port Review 1117

- D Physical Constants and Transistor Model Summary 1120

- Index 1123

CONTENTS

Preface

xx

Chapter-by-Chapter Summary

xxiii

CHAPTER 1

INTRODUCTION TO ELECTRONICS 3

1.1

A Brief History of Electronics: From Vacuum Tubes to Giga-Scale Integration

4

1.2

Classification of Electronic Signals

8

1.2.1

Digital Signals

8

1.2.2

Analog Signals

9

1.2.3

A/D and D/A Converters—Bridging the Analog and Digital Domains

10

1.3

Notational Conventions

12

1.4

Problem-Solving Approach

13

1.5

Important Concepts from Circuit Theory

15

1.5.1

Voltage and Current Division

15

1.5.2

Thévenin and Norton Circuit Representations

16

1.6

Frequency Spectrum of Electronic Signals

21

1.7

Amplifiers

22

1.7.1

Ideal Operational Amplifiers

23

1.7.2

Amplifier Frequency Response

26

1.8

Element Variations in Circuit Design

26

1.8.1

Mathematical Modeling of Tolerances

27

1.8.2

Worst-Case Analysis

27

1.8.3

Monte Carlo Analysis

29

1.8.4

Temperature Coefficients

32

1.9

Numeric Precision

34

Summary

34

Key Terms

35

References

36

Additional Reading

36

Problems

37

CHAPTER 2

SOLID-STATE ELECTRONICS 42

2.1

Solid-State Electronic Materials

44

2.2

Covalent Bond Model

45

2.3

Drift Currents and Mobility in Semiconductors

48

2.3.1

Drift Currents

48

2.3.2

Mobility

49

2.3.3

Velocity Saturation

49

2.4

Resistivity of Intrinsic Silicon

50

2.5

Impurities in Semiconductors

51

2.5.1

Donor Impurities in Silicon

52

2.5.2

Acceptor Impurities in Silicon

52

2.6

Electron and Hole Concentrations in Doped Semiconductors

52

2.6.1

n -Type Material ($N_D > N_A$)

53

2.6.2

p -Type Material ($N_A > N_D$)

54

2.7

Mobility and Resistivity in Doped Semiconductors

55

2.8

Diffusion Currents

59

2.9

Total Current

60

2.10

Energy Band Model

61

2.10.1

Electron—Hole Pair Generation in an Intrinsic Semiconductor

61

2.10.2

Energy Band Model for a Doped Semiconductor

62

2.10.3

Compensated Semiconductors

62

2.11

Overview of Integrated Circuit Fabrication

64

Summary

67

Key Terms

68

Additional Reading

69

Problems

69

CHAPTER 3

SOLID-STATE DIODES AND DIODE CIRCUITS 73

3.1

The pn Junction Diode

74

3.1.1

pn Junction Electrostatics

74

3.1.2

Internal Diode Currents

78

3.2

The i - v Characteristics of the Diode

79

3.3

The Diode Equation: A Mathematical Model for the Diode

81

3.4

Diode Characteristics Under Reverse, Zero, and Forward Bias

84

3.4.1

Reverse Bias

84

3.4.2

Zero Bias

84

3.4.3

Forward Bias

85

3.5	Diode Temperature Coefficient	87		
3.6	Diodes Under Reverse Bias	87		
3.6.1	Saturation Current in Real Diodes	88		
3.6.2	Reverse Breakdown	90		
3.6.3	Diode Model for the Breakdown Region	91		
3.7	<i>pn</i> Junction Capacitance	91		
3.7.1	Reverse Bias	91		
3.7.2	Forward Bias	92		
3.8	Schottky Barrier Diode	93		
3.9	SPICE Model and Layout for a Diode	93		
3.9.1	Diode Layout	95		
3.10	Diode Circuit Analysis	96		
3.10.1	Load-Line Analysis	97		
3.10.2	Analysis Using the Mathematical Model for the Diode	98		
3.10.3	The Ideal Diode Model	102		
3.10.4	Constant Voltage Drop Model	104		
3.10.5	Model Comparison and Discussion	105		
3.11	Multiple-Diode Circuits	106		
3.12	Analysis of Diodes Operating in the Breakdown Region	109		
3.12.1	Load-Line Analysis	109		
3.12.2	Analysis with the Piecewise Linear Model	109		
3.12.3	Voltage Regulation	110		
3.12.4	Analysis Including Zener Resistance	111		
3.12.5	Line and Load Regulation	112		
3.13	Half-Wave Rectifier Circuits	113		
3.13.1	Half-Wave Rectifier with Resistor Load	113		
3.13.2	Rectifier Filter Capacitor	114		
3.13.3	Half-Wave Rectifier with <i>RC</i> Load	115		
3.13.4	Ripple Voltage and Conduction Interval	116		
3.13.5	Diode Current	118		
3.13.6	Surge Current	119		
3.13.7	Peak-Inverse-Voltage (PIV) Rating	120		
3.13.8	Diode Power Dissipation	121		
3.13.9	Half-Wave Rectifier with Negative Output Voltage	121		
3.14	Full-Wave Rectifier Circuits	123		
3.14.1	Full-Wave Rectifier with Negative Output Voltage	124		
3.15	Full-Wave Bridge Rectification	124		
3.16	Rectifier Comparison and Design Tradeoffs	125		
3.17	Dynamic Switching Behavior of the Diode	129		
3.18	Photo Diodes, Solar Cells, and Light-Emitting Diodes	130		
3.18.1	Photo Diodes and Photodetectors	130		
3.18.2	Power Generation from Solar Cells	131		
3.18.3	Light-Emitting Diodes (LEDs)	132		
	<i>Summary</i>	133		
	<i>Key Terms</i>	134		
	<i>Reference</i>	135		
	<i>Additional Reading</i>	135		
	<i>Problems</i>	135		
	CHAPTER 4			
	BIPOLAR JUNCTION TRANSISTORS	146		
4.1	Physical Structure of the Bipolar Transistor	147		
4.2	The Transport Model for the <i>npn</i> Transistor	148		
4.2.1	Forward Characteristics	149		
4.2.2	Reverse Characteristics	151		
4.2.3	Complete Transport Model Equations for Arbitrary Bias Conditions	152		
4.3	The <i>pnp</i> Transistor	154		
4.4	Equivalent Circuit Representations for the Transport Models	156		
4.4.1	Another Look at the Forward-Active Region	157		
4.5	The <i>i-v</i> Characteristics of the Bipolar Transistor	157		
4.5.1	Output Characteristics	157		
4.5.2	Transfer Characteristics	158		
4.6	The Operating Regions of the Bipolar Transistor	159		
4.7	Transport Model Simplifications	160		
4.7.1	Simplified Model for the Cutoff Region	160		
4.7.2	Model Simplifications for the Forward-Active Region	162		
4.7.3	Diodes in Bipolar Integrated Circuits	168		
4.7.4	Simplified Model for the Reverse-Active Region	169		
4.7.5	Modeling Operation in the Saturation Region	172		
4.8	Nonideal Behavior of the Bipolar Transistor	175		
4.8.1	Junction Breakdown Voltages	175		
4.8.2	Minority-Carrier Transport in the Base Region	175		
4.8.3	Base Transit Time	176		
4.8.4	Diffusion Capacitance	178		
4.8.5	Frequency Dependence of the Common-Emitter Current Gain	179		
4.8.6	The Early Effect and Early Voltage	179		
4.8.7	Modeling the Early Effect	180		
4.8.8	Origin of the Early Effect	180		
4.9	Transconductance	181		
4.10	Bipolar Technology and SPICE Model	182		
4.10.1	Qualitative Description	182		
4.10.2	Spice Model Equations	183		
4.10.3	High-Performance Bipolar Transistors	185		

4.11	Practical Bias Circuits for the BJT	185
4.11.1	Four-Resistor Bias Network	187
4.11.2	Design Objectives for the Four-Resistor Bias Network	189
4.11.3	Iterative Analysis of the Four-Resistor Bias Circuit	193
4.12	Tolerances in Bias Circuits	194
4.12.1	Worst-Case Analysis	194
4.12.2	Monte Carlo Analysis	196
	<i>Summary</i>	199
	<i>Key Terms</i>	201
	<i>References</i>	202
	<i>Additional Readings</i>	202
	<i>Problems</i>	202
CHAPTER 5		
FIELD-EFFECT TRANSISTORS 212		
5.1	Characteristics of the MOS Capacitor	214
5.1.1	Accumulation Region	214
5.1.2	Depletion Region	214
5.1.3	Inversion Region	215
5.2	The NMOS Transistor	216
5.2.1	Qualitative i - v Behavior of the NMOS Transistor	217
5.2.2	Triode Region Characteristics of the NMOS Transistor	218
5.2.3	On Resistance	220
5.2.4	Transconductance	222
5.2.5	Saturation of the i - v Characteristics	222
5.2.6	Mathematical Model in the Saturation (Pinch-Off) Region	224
5.2.7	Transconductance in Saturation	225
5.2.8	Transconductance Efficiency in Saturation	225
5.2.9	Channel-Length Modulation	226
5.2.10	Transfer Characteristics and Depletion-Mode MOSFETs	227
5.2.11	Body Effect or Substrate Sensitivity	228
5.3	PMOS Transistors	230
5.4	MOSFET Circuit Symbols	232
5.5	MOS Transistor Symmetry	233
5.5.1	The One-Transistor Dram Cell	233
5.5.2	Data Storage in the 1-T Cell	234
5.5.3	Reading Data from the 1-T Cell	235
5.6	CMOS Technology	238
5.6.1	CMOS Voltage Transfer Characteristics	240
5.7	CMOS Latchup	242
5.8	Capacitances in MOS Transistors	244
5.8.1	NMOS Transistor Capacitances in the Triode Region	244
5.8.2	Capacitances in the Saturation Region	247
5.8.3	Capacitances in Cutoff	247
5.9	MOSFET Modeling in SPICE	247
5.10	MOS Transistor Scaling	249
5.10.1	Drain Current	250
5.10.2	Gate Capacitance	250
5.10.3	Circuit and Power Densities	250
5.10.4	Power-Delay Product	251
5.10.5	Cutoff Frequency	251
5.10.6	High Field Limitations	252
5.10.7	The Unified MOS Transistor Model, Including High Field Limitations	253
5.10.8	Subthreshold Conduction	254
5.11	All Region Modeling	255
5.11.1	Interpolation Model	255
5.11.2	Interpolation Model in the Saturation Region	255
5.11.3	Transconductance Efficiency	256
5.12	MOS Transistor Fabrication and Layout Design Rules	258
5.12.1	Minimum Feature Size and Alignment Tolerance	258
5.12.2	MOS Transistor Layout	259
5.12.3	CMOS Inverter Layout	261
5.13	Advanced CMOS Technologies	262
5.14	Biasing the NMOS Field-Effect Transistor	265
5.14.1	Why Do We Need Bias?	265
5.14.2	Four-Resistor Biasing	267
5.14.3	Constant Gate-Source Voltage Bias	271
5.14.4	Graphical Analysis for the Q-Point	271
5.14.5	Analysis Including Body Effect	272
5.14.6	Analysis Using the Unified Model	274
5.14.7	NMOS Circuit Analysis Comparisons	276
5.14.8	Two-Resistor Bias	276
5.15	Biasing the PMOS Field-Effect Transistor	276
5.16	Biasing the CMOS Inverter as an Amplifier	279
5.17	The CMOS Transmission Gate	280
5.18	The Junction Field-Effect Transistor (JFET)	282
5.18.1	The JFET with Bias Applied	283
5.18.2	JFET Channel with Drain-Source Bias	283
5.18.3	n -Channel JFET i - v Characteristics	285
5.18.4	The p -Channel JFET	286
5.18.5	Circuit Symbols and JFET Model Summary	286
5.18.6	JFET Capacitances	287
5.19	JFET Modeling in SPICE	288
5.20	Biasing the JFET and Depletion-Mode MOSFET	289
	<i>Summary</i>	291
	<i>Key Terms</i>	293
	<i>References</i>	294
	<i>Additional Readings</i>	295
	<i>Problems</i>	296

CHAPTER 6**INTRODUCTION TO AMPLIFIERS 313**

- 6.1 An Example of an Analog Electronic System 314
- 6.2 Amplification 315
 - 6.2.1 Voltage Gain 316
 - 6.2.2 Current Gain 317
 - 6.2.3 Power Gain 317
 - 6.2.4 Location of the Amplifier 317
 - 6.2.5 The Decibel Scale 318
- 6.3 Two-Port Models for Amplifiers 321
 - 6.3.1 The g -Parameters 321
- 6.4 Mismatched Source and Load Resistances 325
- 6.5 The Differential Amplifier 328
 - 6.5.1 Differential Amplifier Voltage Transfer Characteristic 329
 - 6.5.2 Voltage Gain 329
- 6.6 Distortion in Amplifiers 331
- 6.7 Differential Amplifier Model 332
- 6.8 Amplifier Frequency Response 334
 - 6.8.1 Bode Plots 334
 - 6.8.2 The Low-Pass Amplifier 335
 - 6.8.3 The High-Pass Amplifier 338
 - 6.8.4 Band-Pass Amplifiers 341
- Summary 344*
- Key Terms 345*
- References 345*
- Additional Reading 345*
- Problems 345*

CHAPTER 7**THE TRANSISTOR AS AN AMPLIFIER 352**

- 7.1 The Transistor as an Amplifier 353
 - 7.1.1 The BJT Amplifier 354
 - 7.1.2 The MOSFET Amplifier 355
- 7.2 Coupling and Bypass Capacitors 356
- 7.3 Circuit Analysis Using dc and ac Equivalent Circuits 358
 - 7.3.1 Menu for dc and ac Analysis 358
- 7.4 Introduction to Small-Signal Modeling 362
 - 7.4.1 Graphical Interpretation of the Small-Signal Behavior of the Diode 362
 - 7.4.2 Small-Signal Modeling of the Diode 363
- 7.5 Small-Signal Models for Bipolar Junction Transistors 365
 - 7.5.1 The Hybrid- π Model 367
 - 7.5.2 Graphical Interpretation of the Transconductance 368

- 7.5.3 Small-Signal Current Gain 368
- 7.5.4 The Intrinsic Voltage Gain of the BJT 369
- 7.5.5 Equivalent Forms of the Small-Signal Model 370
- 7.5.6 Simplified Hybrid- π Model 371
- 7.5.7 Definition of a Small Signal for the Bipolar Transistor 371
- 7.5.8 Small-Signal Model for the *pnp* Transistor 373
- 7.5.9 ac Analysis versus Transient Analysis in SPICE 374
- 7.6 The Common-Emitter (C-E) Amplifier 374
 - 7.6.1 Terminal Voltage Gain 374
 - 7.6.2 Input Resistance 376
 - 7.6.3 Signal Source Voltage Gain 376
- 7.7 Important Limits and Model Simplifications 376
 - 7.7.1 A Design Guide for the Common-Emitter Amplifier 377
 - 7.7.2 Upper Bound on the Common-Emitter Gain 378
 - 7.7.3 Small-Signal Limit for the Common-Emitter Amplifier 378
- 7.8 Small-Signal Models for Field-Effect Transistors 381
 - 7.8.1 Small-Signal Model for the MOSFET 381
 - 7.8.2 Intrinsic Voltage Gain of the MOSFET 383
 - 7.8.3 Definition of Small-Signal Operation for the MOSFET 384
 - 7.8.4 Body Effect in the Four-Terminal MOSFET 385
 - 7.8.5 Small-Signal Model for the PMOS Transistor 386
 - 7.8.6 Small-Signal Modeling for MOS Transistors in Weak Inversion 387
 - 7.8.7 Small-Signal Model for the Junction Field-Effect Transistor 387
- 7.9 Summary and Comparison of the Small-Signal Models of the BJT and FET 388
- 7.10 The Common-Source (C-S) Amplifier 391
 - 7.10.1 Common-Source Terminal Voltage Gain 392
 - 7.10.2 Signal Source Voltage Gain for the Common-Source Amplifier 392
 - 7.10.3 A Design Guide for the Common-Source Amplifier 393
 - 7.10.4 Small-Signal Limit for the Common-Source Amplifier 394
 - 7.10.5 Input Resistances of the Common-Emitter and Common-Source Amplifiers 396

7.10.6	Common-Emitter and Common-Source Output Resistances	398	8.4	Noninverting Amplifiers—Common-Base and Common-Gate Circuits	464
7.10.7	Comparison of the Three Amplifier Examples	404	8.4.1	Terminal Voltage Gain and Input Resistance	466
7.11	Common-Emitter and Common-Source Amplifier Summary	405	8.4.2	Signal Source Voltage Gain	467
7.11.1	Guidelines for Neglecting the Transistor Output Resistance	405	8.4.3	Input Signal Range	468
7.12	Amplifier Power and Signal Range	406	8.4.4	Resistance at the Collector and Drain Terminals	468
7.12.1	Power Dissipation	406	8.4.5	Current Gain	469
7.12.2	Signal Range	407	8.4.6	Overall Input and Output Resistances for the Noninverting Amplifiers	469
	<i>Summary</i>	410	8.4.7	C-B/C-G Amplifier Summary	473
	<i>Key Terms</i>	411	8.5	Amplifier Prototype Review and Comparison	474
	<i>Reference</i>	411	8.5.1	The BJT Amplifiers	474
	<i>Problems</i>	412	8.5.2	The FET Amplifiers	476
CHAPTER 8			8.6	Common-Source Amplifiers Using MOS Transistor Loads	479
TRANSISTOR AMPLIFIER BUILDING BLOCKS			479	8.6.1	Voltage Gain Estimate
8.1	Amplifier Classification	425	8.6.2	Detailed Analysis	480
8.1.1	Signal Injection and Extraction—The BJT	425	8.6.3	Alternative Loads	481
8.1.2	Signal Injection and Extraction—The FET	426	8.6.4	Input and Output Resistances	482
8.1.3	Common-Emitter (C-E) and Common-Source (C-S) Amplifiers	427	8.7	Coupling and Bypass Capacitor Design	485
8.1.4	Common-Collector (C-C) and Common-Drain (C-D) Topologies	428	8.7.1	Common-Emitter and Common-Source Amplifiers	485
8.1.5	Common-Base (C-B) and Common-Gate (C-G) Amplifiers	430	8.7.2	Common-Collector and Common-Drain Amplifiers	489
8.1.6	Small-Signal Model Review	431	8.7.3	Common-Base and Common-Gate Amplifiers	492
8.2	Inverting Amplifiers—Common-Emitter and Common-Source Circuits	431	8.7.4	Setting Lower Cutoff Frequency f_L	495
8.2.1	The Common-Emitter (C-E) Amplifier	432	8.8	Amplifier Design Examples	496
8.2.2	Common-Emitter Example Comparison	444	8.8.1	Monte Carlo Evaluation of the Common-Base Amplifier Design	505
8.2.3	The Common-Source Amplifier	445	8.9	Multistage ac-Coupled Amplifiers	510
8.2.4	Small-Signal Limit for the Common-Source Amplifier	448	8.9.1	A Three-Stage ac-Coupled Amplifier	511
8.2.5	Common-Emitter and Common-Source Amplifier Characteristics	452	8.9.2	Voltage Gain	513
8.2.6	C-E/C-S Amplifier Summary	453	8.9.3	Input Resistance	514
8.2.7	Equivalent Transistor Representation of the Generalized C-E/C-S Transistor	453	8.9.4	Signal Source Voltage Gain	514
8.3	Follower Circuits—Common-Collector and Common-Drain Amplifiers	454	8.9.5	Output Resistance	515
8.3.1	Terminal Voltage Gain	456	8.9.6	Current and Power Gain	516
8.3.2	Input Resistance	457	8.9.7	Input Signal Range	516
8.3.3	Signal Source Voltage Gain	457	8.9.8	Estimating the Lower Cutoff Frequency of the Multistage Amplifier	520
8.3.4	Follower Signal Range	458	8.10	Introduction to dc-Coupled Amplifiers	520
8.3.5	Follower Output Resistance	458	8.10.1	A dc-Coupled Three-Stage Amplifier	522
8.3.6	Current Gain	460	8.10.2	Two Transistor dc-Coupled Amplifiers	523
8.3.7	C-C/C-D Amplifier Summary	460		<i>Summary</i>	525
				<i>Key Terms</i>	527
				<i>Additional Reading</i>	527
				<i>Problems</i>	527

CHAPTER 9**AMPLIFIER FREQUENCY RESPONSE 544**

- 9.1 Amplifier Frequency Response 545
 - 9.1.1 Low-Frequency Response 546
 - 9.1.2 Estimating ω_L in the Absence of a Dominant Pole 546
 - 9.1.3 High-Frequency Response 549
 - 9.1.4 Estimating ω_H in the Absence of a Dominant Pole 549
 - 9.2 Direct Determination of the Low-Frequency Poles and Zeros—The Common-Source Amplifier 550
 - 9.3 Estimation of ω_L Using the Short-Circuit Time-Constant Method 555
 - 9.3.1 Estimate of ω_L for the Common-Emitter Amplifier 556
 - 9.3.2 Estimate of ω_L for the Common-Source Amplifier 560
 - 9.3.3 Estimate of ω_L for the Common-Base Amplifier 561
 - 9.3.4 Estimate of ω_L for the Common-Gate Amplifier 562
 - 9.3.5 Estimate of ω_L for the Common-Collector Amplifier 563
 - 9.3.6 Estimate of ω_L for the Common-Drain Amplifier 563
 - 9.4 Transistor Models at High Frequencies 564
 - 9.4.1 Frequency-Dependent Hybrid- Π Model for the Bipolar Transistor 564
 - 9.4.2 Modeling C_π and C_μ in SPICE 565
 - 9.4.3 Unity-Gain Frequency f_T 565
 - 9.4.4 High-Frequency Model for the FET 568
 - 9.4.5 Modeling C_{GS} and C_{GD} in SPICE 569
 - 9.4.6 Channel Length Dependence of f_T 569
 - 9.4.7 Limitations of the High-Frequency Models 571
 - 9.5 Base and Gate Resistances in the Small-Signal Models 571
 - 9.5.1 Effect of Base and Gate Resistances on Midband Amplifiers 572
 - 9.6 High-Frequency Common-Emitter and Common-Source Amplifier Analysis 573
 - 9.6.1 The Miller Effect 575
 - 9.6.2 Common-Emitter and Common-Source Amplifier High-Frequency Response 577
 - 9.6.3 Direct Analysis of the Common-Emitter Transfer Characteristic 579
 - 9.6.4 Poles of the Common-Emitter Amplifier 580
 - 9.6.5 Dominant Pole for the Common-Source Amplifier 583
 - 9.6.6 Estimation of ω_H Using the Open-Circuit Time-Constant Method 585
 - 9.6.7 Common-Source Amplifier with Source Degeneration Resistance 586
 - 9.6.8 Poles of the Common-Emitter with Emitter Degeneration Resistance 588
 - 9.7 Common-Base and Common-Gate Amplifier High-Frequency Response 591
 - 9.8 Common-Collector and Common-Drain Amplifier High-Frequency Response 593
 - 9.9 Single-Stage Amplifier High-Frequency Response Summary 596
 - 9.9.1 Amplifier Gain-Bandwidth (GBW) Limitations 596
 - 9.10 Frequency Response of Multistage Amplifiers 597
 - 9.10.1 Differential Amplifier 597
 - 9.10.2 The Common-Collector/Common-Base Cascade 599
 - 9.10.3 High-Frequency Response of the Cascode Amplifier 600
 - 9.10.4 Cutoff Frequency for the Current Mirror 601
 - 9.10.5 Three-Stage Amplifier Example 602
 - 9.11 Introduction to Radio Frequency Circuits 610
 - 9.11.1 Radio Frequency Amplifiers 611
 - 9.11.2 The Shunt-Peaked Amplifier 611
 - 9.11.3 Single-Tuned Amplifier 613
 - 9.11.4 Use of a Tapped Inductor—the Auto Transformer 615
 - 9.11.5 Multiple Tuned Circuits—Synchronous and Stagger Tuning 617
 - 9.11.6 Common-Source Amplifier with Inductive Degeneration 618
 - 9.12 Mixers and Balanced Modulators 622
 - 9.12.1 Introduction to Mixer Operation 622
 - 9.12.2 A Single-Balanced Mixer 623
 - 9.12.3 The Differential Pair as a Single-Balanced Mixer 624
 - 9.12.4 A Double-Balanced Mixer 626
 - 9.12.5 The Jones Mixer—a Double-Balanced Mixer/Modulator 628
- Summary* 632
Key Terms 633
References 633
Problems 634

CHAPTER 10

IDEAL OPERATIONAL AMPLIFIERS 649

- 10.1 Ideal Differential and Operational Amplifiers 650
 - 10.1.1 Assumptions for Ideal Operational Amplifier Analysis 651
 - 10.2 Analysis of Circuits Containing Ideal Operational Amplifiers 651
 - 10.2.1 The Inverting Amplifier 652
 - 10.2.2 The Transresistance Amplifier—a Current-to-Voltage Converter 655
 - 10.2.3 The Noninverting Amplifier 657
 - 10.2.4 The Unity-Gain Buffer, or Voltage Follower 659
 - 10.2.5 The Summing Amplifier 662
 - 10.2.6 The Difference Amplifier 664
 - 10.3 Frequency Dependent Feedback 666
 - 10.3.1 An Active Low-Pass Filter 667
 - 10.3.2 An Active High-Pass Filter 670
 - 10.3.3 The Integrator 671
 - 10.3.4 The Differentiator 675
- Summary 675*
Key Terms 676
References 677
Additional Reading 677
Problems 677

CHAPTER 11

NONIDEAL OPERATIONAL AMPLIFIERS AND FEEDBACK AMPLIFIER STABILITY 683

- 11.1 Classic Feedback Systems 684
 - 11.1.1 Closed-Loop Gain Analysis 685
 - 11.1.2 Gain Error 685
- 11.2 Analysis of Circuits Containing Nonideal Operational Amplifiers 686
 - 11.2.1 Finite Open-Loop Gain 686
 - 11.2.2 Nonzero Output Resistance 689
 - 11.2.3 Finite Input Resistance 693
 - 11.2.4 Summary of Nonideal Inverting and Noninverting Amplifiers 697
- 11.3 Series and Shunt Feedback Circuits 698
 - 11.3.1 Feedback Amplifier Categories 698
 - 11.3.2 Voltage Amplifiers—Series-Shunt Feedback 699
 - 11.3.3 Transimpedance Amplifiers—Shunt-Shunt Feedback 699

- 11.3.4 Current Amplifiers—Shunt-Series Feedback 699
- 11.3.5 Transconductance Amplifiers—Series-Series Feedback 699
- 11.4 Unified Approach to Feedback Amplifier Gain Calculations 699
 - 11.4.1 Closed-Loop Gain Analysis 700
 - 11.4.2 Resistance Calculations Using Blackman’s Theorem 700
- 11.5 Series-Shunt Feedback—Voltage Amplifiers 700
 - 11.5.1 Closed-Loop Gain Calculation 701
 - 11.5.2 Input Resistance Calculations 701
 - 11.5.3 Output Resistance Calculations 702
 - 11.5.4 Series-Shunt Feedback Amplifier Summary 703
- 11.6 Shunt-Shunt Feedback—Transresistance Amplifiers 707
 - 11.6.1 Closed-Loop Gain Calculation 707
 - 11.6.2 Input Resistance Calculations 708
 - 11.6.3 Output Resistance Calculations 708
 - 11.6.4 Shunt-Shunt Feedback Amplifier Summary 709
- 11.7 Series-Series Feedback—Transconductance Amplifiers 712
 - 11.7.1 Closed-Loop Gain Calculation 713
 - 11.7.2 Input Resistance Calculation 713
 - 11.7.3 Output Resistance Calculation 714
 - 11.7.4 Series-Series Feedback Amplifier Summary 714
- 11.8 Shunt-Series Feedback—Current Amplifiers 716
 - 11.8.1 Closed-Loop Gain Calculation 717
 - 11.8.2 Input Resistance Calculation 717
 - 11.8.3 Output Resistance Calculation 718
 - 11.8.4 Shunt-Series Feedback Amplifier Summary 718
- 11.9 Finding the Loop Gain Using Successive Voltage and Current Injection 721
 - 11.9.1 Simplifications 724
- 11.10 Distortion Reduction through the Use of Feedback 724
- 11.11 DC Error Sources and Output Range Limitations 725
 - 11.11.1 Input-Offset Voltage 725
 - 11.11.2 Offset-Voltage Adjustment 727
 - 11.11.3 Input-Bias and Offset Currents 728
 - 11.11.4 Output Voltage and Current Limits 730
- 11.12 Common-Mode Rejection and Input Resistance 733
 - 11.12.1 Finite Common-Mode Rejection Ratio 733
 - 11.12.2 Why Is CMRR Important? 734
 - 11.12.3 Voltage-Follower Gain Error Due to CMRR 737

11.12.4	Common-Mode Input Resistance	740
11.12.5	An Alternate Interpretation of CMRR	741
11.12.6	Power Supply Rejection Ratio	741
11.13	Frequency Response and Bandwidth of Operational Amplifiers	743
11.13.1	Frequency Response of the Noninverting Amplifier	745
11.13.2	Inverting Amplifier Frequency Response	748
11.13.3	Using Feedback to Control Frequency Response	750
11.13.4	Large-Signal Limitations—Slew Rate and Full-Power Bandwidth	752
11.13.5	Macro Model for Operational Amplifier Frequency Response	753
11.13.6	Complete Op Amp Macro Models in SPICE	754
11.13.7	Examples of Commercial General-Purpose Operational Amplifiers	754
11.14	Stability of Feedback Amplifiers	755
11.14.1	The Nyquist Plot	755
11.14.2	First-Order Systems	756
11.14.3	Second-Order Systems and Phase Margin	757
11.14.4	Step Response and Phase Margin	758
11.14.5	Third-Order Systems and Gain Margin	761
11.14.6	Determining Stability from the Bode Plot	762
	<i>Summary</i>	766
	<i>Key Terms</i>	768
	<i>References</i>	769
	<i>Problems</i>	769

CHAPTER 12

OPERATIONAL AMPLIFIER APPLICATIONS 781

12.1	Cascaded Amplifiers	782
12.1.1	Two-Port Representations	782
12.1.2	Amplifier Terminology Review	784
12.1.3	Frequency Response of Cascaded Amplifiers	787
12.2	The Instrumentation Amplifier	795
12.3	Active Filters	798
12.3.1	Low-Pass Filter	798
12.3.2	A High-Pass Filter with Gain	802
12.3.3	Band-Pass Filter	804
12.3.4	Sensitivity	806
12.3.5	Magnitude and Frequency Scaling	807
12.4	Switched-Capacitor Circuits	808
12.4.1	A Switched-Capacitor Integrator	808
12.4.2	Noninverting SC Integrator	810
12.4.3	Switched-Capacitor Filters	812

12.5	Digital-to-Analog Conversion	815
12.5.1	D/A Converter Fundamentals	815
12.5.2	D/A Converter Errors	816
12.5.3	Digital-to-Analog Converter Circuits	818
12.6	Analog-to-Digital Conversion	822
12.6.1	A/D Converter Fundamentals	823
12.6.2	Analog-to-Digital Converter Errors	824
12.6.3	Basic A/D Conversion Techniques	825
12.7	Oscillators	836
12.7.1	The Barkhausen Criteria for Oscillation	836
12.7.2	Oscillators Employing Frequency-Selective RC Networks	837
12.8	Nonlinear Circuit Applications	841
12.8.1	A Precision Half-Wave Rectifier	841
12.8.2	Nonsaturating Precision-Rectifier Circuit	842
12.9	Circuits Using Positive Feedback	844
12.9.1	The Comparator and Schmitt Trigger	844
12.9.2	The Astable Multivibrator	846
12.9.3	The Monostable Multivibrator or One Shot	847
	<i>Summary</i>	851
	<i>Key Terms</i>	853
	<i>Additional Reading</i>	854
	<i>Problems</i>	854

CHAPTER 13

DIFFERENTIAL AMPLIFIERS AND OPERATIONAL AMPLIFIER DESIGN 866

13.1	Differential Amplifiers	867
13.1.1	Bipolar and MOS Differential Amplifiers	867
13.1.2	dc Analysis of the Bipolar Differential Amplifier	868
13.1.3	Transfer Characteristic for the Bipolar Differential Amplifier	870
13.1.4	ac Analysis of the Bipolar Differential Amplifier	871
13.1.5	Differential-Mode Gain and Input and Output Resistances	872
13.1.6	Common-Mode Gain and Input Resistance	874
13.1.7	Common-Mode Rejection Ratio (CMRR)	876
13.1.8	Analysis Using Differential- and Common-Mode Half-Circuits	877
13.1.9	Biasing with Electronic Current Sources	880
13.1.10	Modeling the Electronic Current Source in SPICE	881

13.1.11	dc Analysis of the MOSFET Differential Amplifier	881	14.2.4	Altering the BJT Current Mirror Ratio	954
13.1.12	Differential-Mode Input Signals	884	14.2.5	Multiple Current Sources	955
13.1.13	Small-Signal Transfer Characteristic for the MOS Differential Amplifier	885	14.2.6	Buffered Current Mirror	956
13.1.14	Common-Mode Input Signals	885	14.2.7	Output Resistance of the Current Mirrors	957
13.1.15	Model for Differential Pairs	886	14.2.8	Two-Port Model for the Current Mirror	958
13.2	Evolution to Basic Operational Amplifiers	890	14.2.9	The Widlar Current Source	960
13.2.1	A Two-Stage Prototype for an Operational Amplifier	891	14.2.10	The MOS Version of the Widlar Source	963
13.2.2	Improving the Op Amp Voltage Gain	896	14.2.11	MOS Widlar Source in Weak Inversion	963
13.2.3	Darlington Pairs	897	14.3	High-Output-Resistance Current Mirrors	964
13.2.4	Output Resistance Reduction	898	14.3.1	The Wilson Current Sources	965
13.2.5	A CMOS Operational Amplifier Prototype	902	14.3.2	Output Resistance of the Wilson Source	966
13.2.6	BiCMOS Amplifiers	904	14.3.3	Cascode Current Sources	967
13.2.7	All Transistor Implementations	904	14.3.4	Output Resistance of the Cascode Sources	968
13.3	Output Stages	906	14.3.5	Regulated Cascode Current Source	969
13.3.1	The Source Follower—a Class-A Output Stage	906	14.3.6	Current Mirror Summary	970
13.3.2	Efficiency of Class-A Amplifiers	907	14.4	Reference Current Generation	973
13.3.3	Class-B Push-Pull Output Stage	908	14.5	Supply-Independent Biasing	974
13.3.4	Class-AB Amplifiers	910	14.5.1	A V_{BE} -Based Reference	974
13.3.5	Class-AB Output Stages for Operational Amplifiers	911	14.5.2	The Widlar Source	974
13.3.6	Short-Circuit Protection	911	14.5.3	Power-Supply-Independent Bias Cell	975
13.3.7	Transformer Coupling	913	14.5.4	A Supply-Independent MOS Reference Cell	976
13.4	Electronic Current Sources	916	14.6	The Bandgap Reference	978
13.4.1	Single-Transistor Current Sources	917	14.7	The Current Mirror as an Active Load	982
13.4.2	Figure of Merit for Current Sources	917	14.7.1	CMOS Differential Amplifier with Active Load	982
13.4.3	Higher Output Resistance Sources	918	14.7.2	Bipolar Differential Amplifier with Active Load	989
13.4.4	Current Source Design Examples	919	14.8	Active Loads in Operational Amplifiers	993
	Summary	927	14.8.1	CMOS Op-Amp Voltage Gain	993
	Key Terms	928	14.8.2	dc Design Considerations	994
	References	928	14.8.3	Bipolar Operational Amplifiers	996
	Additional Reading	929	14.8.4	Input Stage Breakdown	997
	Problems	929	14.9	The $\mu A741$ Operational Amplifier	998
			14.9.1	Overall Circuit Operation	998
			14.9.2	Bias Circuitry	999
			14.9.3	dc Analysis of the 741 Input Stage	1000
			14.9.4	ac Analysis of the 741 Input Stage	1003
			14.9.5	Voltage Gain of the Complete Amplifier	1004
			14.9.6	The 741 Output Stage	1008
			14.9.7	Output Resistance	1010
			14.9.8	Short-Circuit Protection	1010
			14.9.9	Summary of the $\mu A741$ Operational Amplifier Characteristics	1010
CHAPTER 14					
ANALOG INTEGRATED CIRCUIT DESIGN TECHNIQUES 946					
14.1	Circuit Element Matching	947			
14.2	Current Mirrors	948			
14.2.1	dc Analysis of the MOS Transistor Current Mirror	949			
14.2.2	Changing the MOS Mirror Ratio	951			
14.2.3	dc Analysis of the Bipolar Transistor Current Mirror	952			

14.10 The Gilbert Analog Multiplier 1011

*Summary 1013**Key Terms 1014**References 1015**Additional Readings 1015**Problems 1015***CHAPTER 15****TRANSISTOR FEEDBACK AMPLIFIERS
AND OSCILLATORS 1030**

15.1 Basic Feedback System Review 1031

15.1.1 Closed-Loop Gain 1031

15.1.2 Closed-Loop Impedances 1032

15.1.3 Feedback Effects 1032

15.2 Feedback Amplifier Analysis at Midband 1034

15.2.1 Closed-Loop Gain 1034

15.2.2 Input Resistance 1035

15.2.3 Output Resistance 1035

15.2.4 Offset Voltage Calculation 1036

15.3 Feedback Amplifier Circuit Examples 1037

15.3.1 Series-Shunt Feedback—Voltage
Amplifiers 103715.3.2 Differential Input Series-Shunt Voltage
Amplifier 104215.3.3 Shunt-Shunt Feedback—Transresistance
Amplifiers 104515.3.4 Series-Series Feedback—
Transconductance Amplifiers 105115.3.5 Shunt-Series Feedback—Current
Amplifiers 1054

15.4 Review of Feedback Amplifier Stability 1057

15.4.1 Closed-Loop Response of the
Uncompensated Amplifier 1058

15.4.2 Phase Margin 1059

15.4.3 Higher-Order Effects 1063

15.4.4 Response of the Compensated
Amplifier 1064

15.4.5 Small-Signal Limitations 1066

15.5 Single-Pole Operational Amplifier
Compensation 1066

15.5.1 Three-Stage Op-Amp Analysis 1067

15.5.2 Transmission Zeros in FET Op
Amps 1069

15.5.3 Bipolar Amplifier Compensation 1070

15.5.4 Slew Rate of the Operational
Amplifier 107115.5.5 Relationships between Slew
Rate and Gain-Bandwidth
Product 1072

15.6 High-Frequency Oscillators 1081

15.6.1 The Colpitts Oscillator 1082

15.6.2 The Hartley Oscillator 1083

15.6.3 Amplitude Stabilization In LC
Oscillators 1084

15.6.4 Negative Resistance in Oscillators 1084

15.6.5 Negative G_m Oscillator 1085

15.6.6 Crystal Oscillators 1087

15.6.7 Ring Oscillators 1090

15.6.8 Positive Feedback and Latchup 1091

*Summary 1094**Key Terms 1096**Additional Readings 1096**Problems 1096***CHAPTER 16****INTRODUCTION TO DIGITAL ELECTRONICS 16-3**

16.1 Ideal Logic Gates 16-5

16.2 Logic Level Definitions and
Noise Margins 16-5

16.2.1 Logic Voltage Levels 16-7

16.2.2 Noise Margins 16-7

16.2.3 Logic Gate Design Goals 16-8

16.3 Dynamic Response of Logic Gates 16-9

16.3.1 Rise Time and Fall Time 16-9

16.3.2 Propagation Delay 16-10

16.3.3 Power-Delay Product 16-10

16.4 Review of Boolean Algebra 16-11

16.5 NMOS Logic Design 16-13

16.5.1 NMOS Inverter with Resistive Load 16-14

16.5.2 Design of the W/L Ratio of M_s 16-15

16.5.3 Load Resistor Design 16-16

16.5.4 Load-Line Visualization 16-16

16.5.5 On-Resistance of the Switching
Device 16-18

16.5.6 Noise Margin Analysis 16-19

16.5.7 Calculation of V_{IL} and V_{OH} 16-1916.5.8 Calculation of V_{IH} and V_{OL} 16-2016.5.9 Resistor Load Inverter Noise
Margins 16-20

16.5.10 Load Resistor Problems 16-21

16.6 Transistor Alternatives to the Load
Resistor 16-2216.6.1 The NMOS Saturated Load
Inverter 16-2316.6.2 NMOS Inverter with a Linear Load
Device 16-3116.6.3 NMOS Inverter with a Depletion-Mode
Load 16-32

16.7 NMOS Inverter Summary and Comparison 16-35

16.8 Impact of Velocity Saturation on Static
Inverter Design 16-36

S6.8.1	Switching Transistor Design	S6-36
S6.8.2	Load Transistor Design	S6-36
S6.8.3	Velocity Saturation Impact	
	Summary	S6-37
S6.9	NMOS NAND and NOR Gates	S6-37
S6.9.1	NOR Gates	S6-38
S6.9.2	NAND Gates	S6-39
S6.9.3	NOR and NAND Gate Layouts	
	in NMOS Depletion-Mode	
	Technology	S6-40
S6.10	Complex NMOS Logic Design	S6-41
S6.11	Power Dissipation	S6-46
S6.11.1	Static Power Dissipation	S6-46
S6.11.2	Dynamic Power Dissipation	S6-47
S6.11.3	Power Scaling in MOS Logic	
	Gates	S6-48
S6.12	Dynamic Behavior of MOS Logic Gates	S6-49
S6.12.1	Capacitances in Logic Circuits	S6-50
S6.12.2	Dynamic Response of the NMOS	
	Inverter with a Resistive Load	S6-51
S6.12.3	Comparison of NMOS Inverter	
	Delays	S6-56
S6.12.4	Impact of Velocity Saturation on	
	Inverter Delays	S6-57
S6.12.5	Scaling Based upon Reference	
	Circuit Simulation	S6-57
S6.12.6	Ring Oscillator Measurement of	
	Intrinsic Gate Delay	S6-58
S6.12.7	Unloaded Inverter Delay	S6-58
S6.13	PMOS Logic	S6-61
S6.13.1	PMOS Inverters	S6-61
S6.13.2	NOR and NAND Gates	S6-63
	Summary	S6-64
	Key Terms	S6-66
	References	S6-67
	Additional Reading	S6-67
	Problems	S6-67

CHAPTER S7

COMPLEMENTARY MOS (CMOS) LOGIC DESIGN S7-1

S7.1	CMOS Inverter Technology	S7-2
S7.1.1	CMOS Inverter Layout	S7-4
S7.2	Static Characteristics of the CMOS Inverter	S7-4
S7.2.1	CMOS Voltage Transfer	
	Characteristics	S7-5
S7.2.2	Noise Margins for the CMOS	
	Inverter	S7-7
S7.3	Dynamic Behavior of the CMOS Inverter	S7-9
S7.3.1	Propagation Delay Estimate	S7-9
S7.3.2	Rise and Fall Times	S7-11
S7.3.3	Performance Scaling	S7-11

S7.3.4	Impact of Velocity Saturation on CMOS	
	Inverter Delays	S7-13
S7.3.5	Delay of Cascaded Inverters	S7-14
S7.4	Power Dissipation and Power	
	Delay Product in CMOS	S7-15
S7.4.1	Static Power Dissipation	S7-15
S7.4.2	Dynamic Power Dissipation	S7-16
S7.4.3	Power-Delay Product	S7-17
S7.5	CMOS NOR and NAND Gates	S7-19
S7.5.1	CMOS NOR Gate	S7-19
S7.5.2	CMOS NAND Gates	S7-22
S7.6	Design of Complex Gates in CMOS	S7-23
S7.7	Minimum Size Gate Design and	
	Performance	S7-29
S7.8	Cascade Buffers	S7-31
S7.8.1	Cascade Buffer Delay Model	S7-31
S7.8.2	Optimum Number of Stages	S7-32
S7.9	The CMOS Transmission Gate	S7-34
S7.10	Bistable Circuits	S7-35
S7.10.1	The Bistable Latch	S7-35
S7.10.2	RS Flip-Flop	S7-38
S7.10.3	The D-Latch Using Transmission	
	Gates	S7-39
S7.10.4	A Master-Slave D Flip-Flop	S7-39
S7.11	CMOS Latchup	S7-39
	Summary	S7-44
	Key Terms	S7-45
	References	S7-46
	Problems	S7-46

CHAPTER S8

MOS MEMORY CIRCUITS S8-1

S8.1	Random-Access Memory (RAM)	S8-2
S8.1.1	Random-Access Memory (RAM)	
	Architecture	S8-2
S8.1.2	A 256-Mb Memory Chip	S8-3
S8.2	Static Memory Cells	S8-4
S8.2.1	Memory Cell Isolation and	
	Access—The 6-T Cell	S8-4
S8.2.2	The Read Operation	S8-5
S8.2.3	Writing Data into the	
	6-T Cell	S8-9
S8.3	Dynamic Memory Cells	S8-11
S8.3.1	The One-Transistor Cell	S8-12
S8.3.2	Data Storage in the 1-T Cell	S8-12
S8.3.3	Reading Data from the 1-T Cell	S8-14
S8.3.4	The Four-Transistor Cell	S8-15
S8.4	Sense Amplifiers	S8-17
S8.4.1	A Sense Amplifier for the 6-T Cell	S8-17
S8.4.2	A Sense Amplifier for the 1-T Cell	S8-19
S8.4.3	The Boosted Wordline Circuit	S8-20
S8.4.4	Clocked CMOS Sense	
	Amplifiers	S8-21

- S8.5 Address Decoders S8-23
 - S8.5.1 NOR Decoder S8-23
 - S8.5.2 NAND Decoder S8-23
 - S8.5.3 Pass-Transistor Column Decoder S8-25
- S8.6 Read-Only Memory (ROM) S8-26
- S8.7 Flash Memory S8-29
 - S8.7.1 Floating Gate Technology S8-29
 - S8.7.2 NOR Circuit Implementations S8-32
 - S8.7.3 NAND Implementations S8-32
- Summary* S8-34
- Key Terms* S8-35
- References* S8-36
- Additional Readings* S8-36
- Problems* S8-36

- CHAPTER S9**
- BIPOLAR LOGIC CIRCUITS S9-1**
 - S9.1 The Current Switch (Emitter-Coupled Pair) S9-2
 - S9.1.1 Mathematical Model for Static Behavior of the Current Switch S9-2
 - S9.1.2 Current Switch Analysis for $v_i > V_{REF}$ S9-4
 - S9.1.3 Current Switch Analysis for $v_i < V_{REF}$ S9-5
 - S9.2 The Emitter-Coupled Logic (ECL) Gate S9-5
 - S9.2.1 ECL Gate with $v_i = V_H$ S9-6
 - S9.2.2 ECL Gate with $v_i = V_L$ S9-7
 - S9.2.3 Input Current of the ECL Gate S9-7
 - S9.2.4 ECL Summary S9-7
 - S9.3 Noise Margin Analysis for the ECL Gate S9-8
 - S9.3.1 V_{IL} , V_{OH} , V_{IH} , and V_{OL} S9-8
 - S9.3.2 Noise Margins S9-9
 - S9.4 Current Source Implementation S9-10
 - S9.5 The ECL OR-NOR Gate S9-12
 - S9.6 The Emitter Follower S9-14
 - S9.6.1 Emitter Follower with a Load Resistor S9-15
 - S9.7 “Emitter Dotting” or “Wired-OR” Logic S9-17
 - S9.7.1 Parallel Connection of Emitter-Follower Outputs S9-18
 - S9.7.2 The Wired-OR Logic Function S9-18
 - S9.8 ECL Power-Delay Characteristics S9-18
 - S9.8.1 Power Dissipation S9-18
 - S9.8.2 Gate Delay S9-20
 - S9.8.3 Power-Delay Product S9-21
 - S9.9 Positive ECL (PECL) S9-22
 - S9.10 Current-Mode Logic S9-23
 - S9.10.1 CML Logic Gates S9-23
 - S9.10.2 CML Logic Levels S9-24
 - S9.10.3 V_{EE} Supply Voltage S9-24
 - S9.10.4 Higher-Level CML S9-25
 - S9.10.5 CML Power Reduction S9-26
 - S9.10.6 Source-Coupled FET Logic (SCFL) S9-26
 - S9.11 The Saturating Bipolar Inverter S9-29
 - S9.11.1 Static Inverter Characteristics S9-29
 - S9.11.2 Saturation Voltage of the Bipolar Transistor S9-30
 - S9.11.3 Load-Line Visualization S9-32
 - S9.11.4 Switching Characteristics of the Saturated BJT S9-33
 - S9.12 Transistor-Transistor Logic S9-36
 - S9.12.1 TTL Inverter Analysis for $v_i = V_L$ S9-36
 - S9.12.2 Analysis for $v_i = V_H$ S9-38
 - S9.12.3 Power Consumption S9-39
 - S9.12.4 TTL Propagation Delay and Power-Delay Product S9-39
 - S9.12.5 TTL Voltage Transfer Characteristic and Noise Margins S9-40
 - S9.12.6 Fanout Limitations of Standard TTL S9-40
 - S9.13 Logic Functions in TTL S9-40
 - S9.13.1 Multi-Emitter Input Transistors S9-41
 - S9.13.2 TTL NAND Gates S9-41
 - S9.13.3 Input Clamping Diodes S9-42
 - S9.14 Schottky-Clamped TTL S9-43
 - S9.15 Comparison of the Power-Delay Products of ECL and TTL S9-44
 - S9.16 BiCMOS Logic S9-44
 - S9.16.1 BiCMOS Buffers S9-45
 - S9.16.2 BiNMOS Inverters S9-47
 - S9.16.3 BiCMOS Logic Gates S9-48
- Summary* S9-49
- Key Terms* S9-50
- Reference* S9-51
- Additional Reading* S9-51
- Problems* S9-51

- A Standard Discrete Component Values 1109
- B Solid-State Device Models and SPICE Simulation Parameters 1112
- C Two-Port Review 1117
- D Physical Constants and Transistor Model Summary 1120

- Index 1123

Electronics in Action

Chapter 1

Cellular Phone Evolution 8
 Player Characteristics 20
 A Familiar Electronic System—The Cellular Phone 25

Chapter 2

CCD Cameras 63
 Lab-on-a-chip 66

Chapter 3

The PTAT Voltage and Electronic Thermometry 89
 The SPICE Circuit Simulation Program An IEEE Global
 History Network Milestone 94
 AM Demodulation 122
 Power Cubes and Cell Phone Chargers 128
 Solar Energy 132

Chapter 4

The Bipolar Transistor PTAT Cell 169
 Optical Isolators 174

Chapter 5

MOS Memory 237
 CMOS Camera on a Chip 245
 FinFET Technologies 263
 Thermal Inkjet Printers 281

Chapter 6

Player Characteristics 320
 Laptop Computer Touchpad 327

Chapter 7

Direct Measurement of Intrinsic Gain ($\mu_f = g_m r_o$) 390
 Electric Guitar Distortion Circuits 408

Chapter 8

Noise in Electronic Circuits 454
 Revisiting the CMOS Imager Circuitry 465
 Humbucker Guitar Pickup 521

Chapter 9

Graphic Equalizer 584
 RF Network Transformations 612
 Noise Factor, Noise Figure, and Minimum
 Detectable Signal 621
 Passive Diode Mixers 627
 The Jones Mixer 631

Chapter 10

Fiber Optic Receiver 656
 Digital-to-Analog Converter (DAC) Circuits 663
 Dual-Ramp or Dual-Slope Analog-to-Digital
 Converters (ADCs) 672

Chapter 11

Three-Terminal IC Voltage Regulators 706
 Fiber Optic Receiver 712
 Low Voltage Differential Signaling (LVDS) 735
 Offset Voltage, Bias Current, and CMRR
 Measurement 742

Chapter 12

CMOS Navigation Chip Prototype for Optical Mice 797
 Band-Pass Filters in BFSK Reception 806
 Body Sensor Networks 813
 Sample-and-Hold Circuits 835
 An AC Voltmeter 843
 Numerically Controlled Oscillators and
 Direct Digital Synthesis 848
 Function Generators 850

Chapter 13

Limiting Amplifiers for Optical Communications 887
 Class-D Audio Amplifiers 915
 Medical Ultrasound Imaging 925

Chapter 14

The PTAT Voltage 962
 G_m -C Integrated Filters 988

Chapter 15

A Transresistance Amplifier Implementation 1050
 Fully Differential Design 1057
 A MEMS Oscillator 1092

Supplemental Chapters (E-Book Only)

Chapter S6

Silicon Art S6-42
 MEMS-Based Computer Projector S6-62

Chapter S7

CMOS—The Enabler for Handheld Technologies S7-18
 AND-OR-INVERT Gates in a Standard Cell
 Library S7-30
 A Second Look at Noise Margins S7-36
 Advanced CMOS Technologies S7-42

Chapter S8

Field Programmable Gate Arrays (FPGAs) S8-11
 Flash Memory Growth S8-31

Chapter S9

Electronics for Optical Communications S9-27

PREFACE

Through study of this text, the reader will develop a comprehensive understanding of the basic techniques of modern analog electronic circuit design. Even though most readers may not ultimately be engaged in the design of integrated circuits (ICs) themselves, a thorough understanding of the internal circuit structure of ICs is prerequisite to avoiding many pitfalls that prevent the effective and reliable application of integrated circuits in system design.

The writing integrates the authors' extensive industrial backgrounds in precision analog and digital design with their many years of experience in the classroom. A broad spectrum of topics is included, and material can easily be selected to satisfy either a two-semester or three-quarter sequence in electronics.

In order to reduce the length, cost, and weight of the text, the digital electronics chapters from earlier editions have been included as supplemental chapters in the e-book version of the textbook that is available in Connect.

IN THIS EDITION

This edition continues to update the material to achieve improved readability and accessibility to the student. In addition to general material updates, a number of specific changes have been included.

The five chapters of Part One have been reorganized to improve material flow. Chapter 4, "Bipolar Junction Transistors" now follows directly after the diode chapter, and "Field-Effect Transistors" becomes Chapter 5. A new low-power, low-voltage, and weak inversion thread begins in Part One. Chapter 5 specifically introduces the behavior and modeling of the FET in the moderate and weak inversion regions, and this thread continues throughout Parts Two and Three.

Other important elements include:

- At least 30 percent revised or new problems.

- Updated PowerPoint slides are available from the authors at www.JaegerBlalock.com or Connect.

- Popular digital features can be found through McGraw Hill Education's Connect platform, details of which can be found later in the Preface.

- The structured problem-solving approach continues throughout the examples.

- Popular Electronics in Action features have been revised and expanded to include IEEE Societies, Historical Development of SPICE, Body Sensor Networks, Jones Mixer, Advanced CMOS Technology, Fully Differential Amplifiers, and DACs and ADCs to name a few.

Chapter openers enhance the reader's understanding of historical developments in electronics. Design notes highlight important ideas that the circuit designer should remember. The Internet is viewed as an integral extension of the text.

Features of the book are outlined below.

- The Structured Problem-Solving Approach is used throughout the examples.

- Electronics in Action features in each chapter.

- Chapter openers highlighting developments in the field of electronics.

- Design Notes and emphasis on practical circuit design.

- Broad use of SPICE throughout the text, examples, and problems.

- Integrated treatment of device modeling in SPICE.

- Numerous Exercises, Examples, and Design Examples.

- Large number of problems.

- Integrated web materials.




Part Two consists of Chapters 6 through 9 and begins with an overview of general amplifier characteristics, followed by small-signal modeling of transistors and comprehensive discussion of classical single-stage amplifier design including frequency response.

The first three chapters of Part Three focus on ideal and nonideal operational amplifiers, including feedback and amplifier stability. The last three chapters concentrate on analog integrated circuit design and design techniques.

Design remains a difficult issue in educating engineers. The use of the well-defined problem-solving methodology presented in this text can significantly enhance the students ability to understand issues related to design. The design examples assist in building an understanding of the design process.

Methods for making design estimates and decisions are stressed throughout the analog portion of the text. Expressions for amplifier behavior are simplified beyond the standard hybrid- π model expressions whenever appropriate. For example, the expression for the voltage gain of an amplifier in most texts is simply written as $|A_v| = g_m R_L$, which tends to hide the power supply voltage as the fundamental design variable. Rewriting this expression in approximate form as $g_m R_L \cong 10V_{CC}$ for the BJT, or $g_m R_L \cong V_{DD}$ for the FET, explicitly displays the dependence of amplifier design on the choice of power supply voltage and provides a simple first-order design estimate for the voltage gain of the common-emitter and common-source amplifiers. The gain advantage of the BJT stage is also clear. These approximation techniques and methods for performance estimation are included as often as possible. Comparisons and design tradeoffs between the properties of BJTs and FETs are included throughout Part Three.

Worst-case and Monte-Carlo analysis techniques are introduced at the end of the first chapter. These are not topics traditionally included in undergraduate courses. However, the ability to design circuits in the face of wide component tolerances and variations is a key component of electronic circuit design, and the design of circuits using standard components and tolerance assignment are discussed in examples and included in many problems.

Specific design problems, computer problems, and SPICE problems are included at the end of each chapter. Design problems are indicated by , computer problems are indicated by , and SPICE problems are indicated by . The problems are keyed to the topics in the text with the more difficult or time-consuming problems indicated by * and **. An Instructor's Manual containing solutions to all the problems is available to instructors from the authors. In addition, the graphs and figures are available as

PowerPoint files and can be retrieved on the Instructor's Resources section of Connect, along with various web materials referenced in the textbook for students. Instructor notes are available as PowerPoint slides.

To access the Instructor Resources through Connect, you must first contact your McGraw Hill Learning Technology Representative to obtain a password. If you do not know your McGraw Hill representative, please go to www.mhhe.com/rep, to find your representative.

Once you have your password, please go to connect.mheducation.com, and log in. Click on the course for which you are using *Microelectronic Circuit Design, 6e*. If you have not added a course, click "Add Course," and select "Engineering-Electrical and Computer" from the drop-down menu. Select this textbook and click "Next."

Once you have added the course, click on the "Library" link, and then click "Instructor Resources."

We want to thank the large number of people who have had an impact on the material in this text and on its preparation. Our students have helped immensely in polishing the manuscript and have managed to survive the many revisions of the manuscript. Our department heads, J. D. Irwin and Mark Nelms of Auburn University, N. Sidiropoulos of the University of Virginia and Gregory Peterson of the University of Tennessee, have always been highly supportive of faculty efforts to develop improved texts.

We want to thank all reviewers, including the following:

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We are also thankful for inspiration from the classic text *Applied Electronics* by J. F. Pierce and T. J. Paulus. Professor Travis Blalock Learned Electronics from Professor Pierce many years ago and still appreciates many of the analytical techniques employed in their long out-of-print text.

Those familiar with Professor Don Pederson's "Yellow Peril" will see its influence throughout this text. Shortly after Professor Jaeger became Professor Art

Brodersen's student at the University of Florida, he was fortunate to be given a copy of Pederson's book to study from cover to cover.

Finally, we want to thank the team at McGraw Hill, including Theresa Collins and Erin Kamm, Product Developers; Jane Mohr, Content Project Manager; Lisa Granger, Marketing Manager; and Sadika Rehman, Full-Service Project Manager.

In developing this text, we have attempted to integrate our industrial backgrounds in analog and digital design with many years of experience in the classroom. We hope

we have at least succeeded to some extent. Constructive suggestions and comments will be appreciated.

Richard C. Jaeger
Auburn University

Travis N. Blalock
University of Virginia

Benjamin J. Blalock
University of Tennessee, Knoxville

CHAPTER-BY-CHAPTER SUMMARY

PART ONE—SOLID-STATE ELECTRONICS AND DEVICES

Chapter 1 provides a historical perspective on the field of electronics beginning with vacuum tubes and advancing to Tera-scale integration and its impact on the global economy. Chapter 1 also provides a classification of electronic signals and a review of some important tools from network analysis, including the ideal operational amplifier. Because developing a good problem-solving methodology is of such import to an engineer's career, the comprehensive Structured Problem Solving Approach is used to help students develop their problem solving skills. The structured approach is discussed in detail in the first chapter and used in the subsequent examples in the text. Component tolerances and variations play an extremely important role in practical circuit design, and Chapter 1 closes with introductions to tolerances, temperature coefficients, worst-case design, and Monte Carlo analysis.

Chapter 2 discusses semiconductor materials including the covalent-bond and energy-band models of semiconductors. The chapter includes material on intrinsic carrier density, electron and hole populations, n - and p -type material, and impurity doping. Mobility, resistivity, and carrier transport by both drift and diffusion are included as topics. Velocity saturation is discussed, as well as an introductory discussion of micro-electronic fabrication.

Chapter 3 introduces the structure and i - v characteristics of solid-state diodes. Discussions of Schottky diodes, variable capacitance diodes, photo-diodes, solar cells, and LEDs are also included. This chapter introduces the concepts of device modeling and the use of different levels of modeling to achieve various approximations to reality. The SPICE model for the diode is discussed. The concepts of bias, operating point, and load-line are all introduced, and iterative mathematical solutions are also used to find the operating point with MATLAB and spreadsheets. Diode applications in rectifiers are discussed in detail and a discussion of the dynamic switching characteristics of diodes is also presented.

Chapter 4 introduces the bipolar junction transistor and presents a heuristic development of the transport (simplified Gummel-Poon) model of the BJT based upon superposition. The various regions of operation are discussed in detail. Common-emitter and common-base current gains are defined, and base transit-time, diffusion capacitance, and cutoff frequency are all discussed. Bipolar technology and physical structure are introduced. The four-resistor bias circuit is discussed in detail. The SPICE model for the BJT and SPICE model parameters are also discussed in Chapter 4.

Chapter 5 discusses MOS and junction field-effect transistors, starting with a qualitative description of the MOS capacitor. Models are developed for the FET i - v characteristics, and a complete discussion of the regions of operation of the device is presented. Body effect is included. MOS transistor performance limits—including scaling, cut-off frequency, and subthreshold conduction—are discussed as well as basic Λ -based layout methods. Biasing circuits and load-line analysis are presented. The concept of velocity saturation from Chapter 2 is reinforced with the addition of the unified MOS model of Rabaey and Chandrakasan to Chapter 5. FET SPICE models and model parameters are discussed in Chapter 5. In the 6th edition, the discussion of moderate and weak inversion is expanded, and a low voltage/weak inversion thread continues through the rest of the text.

PART TWO—ANALOG ELECTRONICS

Chapter 6 provides a succinct introduction to analog electronics. The concepts of voltage gain, current gain, and power gain are developed using two-port circuit models. Much care has been taken to be consistent in the use of the notation that defines these quantities as well as in the use of dc, ac, and total signal notation throughout the book. Bode plots are reviewed and amplifiers are classified by frequency response. MATLAB is utilized as a tool for producing Bode plots. SPICE simulation using built-in SPICE models is introduced.

Chapter 7 begins the general discussion of linear amplification using the BJT and FET as C-E and C-S amplifiers. Biasing for linear operation and the concept of small-signal modeling are both introduced, and small-signal models of the diode, BJT, and FET are all developed. The limits for small-signal operation are all carefully defined. The use of coupling and bypass capacitors and inductors to separate the ac and dc designs is explored. The important $10V_{CC}$ and V_{DD} design estimates for the voltage gain of the C-E and C-S amplifiers are introduced, and the role of the transistor's intrinsic gain in bounding circuit performance is discussed. The role of Q-point design on power dissipation and signal range is also introduced.

Chapter 8 proceeds with an in-depth comparison of the characteristics of single-transistor amplifiers, including small-signal amplitude limitations. Appropriate points for signal injection and extraction are identified, and amplifiers are classified as inverting amplifiers (C-E, C-S), noninverting amplifiers (C-B, C-G), and followers (C-C, C-D). The treatment of MOS and bipolar devices is merged from Chapter 8 on, and design tradeoffs between the use of the BJT and the FET in amplifier circuits is an important thread that is followed through all of Part Two. A detailed discussion of the design of coupling and bypass capacitors and the role of these capacitors in controlling the low frequency response of amplifiers appears in this chapter.

Chapter 9 discusses the frequency response of analog circuits. The behavior of each of the three categories of single-stage amplifiers (C-E/C-S, C-B/C-G, and C-C/C-D) is discussed in detail, and BJT behavior is contrasted with that of the FET. The frequency response of the transistor is discussed, and the high frequency, small-signal models are developed for both the BJT and FET. Miller multiplication is used to obtain estimates of the lower and upper cutoff frequencies of complex multistage amplifiers. Gain-bandwidth products and gain-bandwidth tradeoffs in design are discussed. Cascode amplifier frequency response, and tuned amplifiers are included in this chapter. The important short-circuit and open-circuit time-constant techniques for estimating the dominant low- and high-frequency poles are covered in detail.

Because of the renaissance and pervasive use of RF circuits, Chapter 9 includes an introductory section on RF amplifiers, including shunt peaked and tuned amplifiers. A discussion of gate resistance in FETs mirrors that of base resistance in the BJT. The discussion of the impact of the frequency-dependent current gain of the FET includes both the input and output impedances of the source follower configuration. Material on mixers includes passive and active single- and double-balanced mixers and the widely used Jones Mixer.

Chapter 10 reviews classic ideal operational amplifier circuits that include the inverting, noninverting, summing, and difference amplifiers as well as the integrator, differentiator, and low-pass and high-pass filters.

Chapter 11 focuses on a comprehensive discussion of the characteristics and limitations of real operational amplifiers, including the effects of finite gain and input resistance, nonzero output resistance, input offset voltage, input bias and offset currents, output voltage and current limits, finite bandwidth, and common-mode rejection. A consistent loop-gain analysis approach is used to study the four classic feedback configurations, and Blackman's theorem is utilized to find input and output resistances of closed-loop amplifiers. The important successive voltage and current injection technique for finding loop-gain is included in Chapter 11. Stability of first-, second-, and third-order systems is discussed, and the concepts of phase and gain margin are introduced. Relationships between Nyquist and Bode techniques are explicitly discussed. A section concerning the relationship between phase margin and time domain response is included. The macro model concept is introduced and the discussion of SPICE simulation of op-amp circuits using various levels of models continues in Chapter 11.

Chapter 12 covers a wide range of operational amplifier applications that include multistage amplifiers, the instrumentation amplifier, and continuous time and discrete time active filters. Cascade amplifiers are investigated including a discussion of the bandwidth of multistage amplifiers. An introduction to D/A and A/D converters appears in this chapter. The Barkhausen criterion for oscillation are presented and followed by a discussion of op-amp-based sinusoidal oscillators. High frequency oscillators are discussed in Chapter 15. Nonlinear circuits applications including rectifiers, Schmitt triggers, and multivibrators conclude the material in Chapter 12.

Chapter 13 explores the design of multistage direct coupled amplifiers. An evolutionary approach to multistage op amp design is used. MOS and bipolar differential amplifiers are first introduced. Subsequent addition of a second gain stage and then an output stage convert the differential amplifiers into simple op amps. Class A, B, and AB operations are defined. Electronic current sources are designed and used for biasing of the basic operational amplifiers. Discussion of important FET-BJT design tradeoffs are included wherever appropriate. Additional low voltage/weak inversion problems have been added to Chapters 13, 14, and 15.

Chapter 14 introduces techniques that are of particular import in integrated circuit design. A variety of current mirror circuits are introduced and applied in bias circuits and as active loads in operational amplifiers. A wealth of circuits and analog design techniques are explored through the detailed analysis of the classic 741 operational amplifier. The Brokaw bandgap reference and Gilbert analog multiplier as well as the MOS weak inversion reference are introduced in Chapter 14.

Chapter 15 presents detailed examples of feedback as applied to transistor amplifier circuits. The loop-gain analysis approach introduced in Chapter 11 is used to find the closed-loop gain of various amplifiers, and Blackman's theorem is utilized to find input and output resistances of closed-loop amplifiers.

Amplifier stability is also discussed in Chapter 15, and Nyquist diagrams and Bode plots (with MATLAB) are used to explore the phase and gain margin of amplifiers. Basic single-pole op-amp compensation is discussed, and the unity gain-bandwidth product is related to amplifier

slew rate. Design of op-amp compensation to achieve a desired phase margin is presented. The discussion of transistor oscillator circuits includes the classic Colpitts, Hartley, and negative G_m configurations. Crystal oscillators, ring oscillators and a discussion of positive feedback in flip-flops are also included.

The Digital Electronics chapters from the fifth edition are now included as supplemental chapters in the e-book version of this text, which is available to users of this edition through Connect.

Four Appendices include tables of standard component values (Appendix A), summary of the device models and sample SPICE parameters (Appendix B), review of two-port networks (Appendix C), and Physical Constants and Transistor Model Summary (Appendix D). Data sheets for representative solid-state devices and operational amplifiers are available via the Internet. A table in Appendix C helps relate various two-port parameters that often appear in specification sheets to the FET and BJT model parameters that appear in the text.



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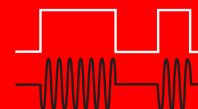
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PART ONE

SOLID-STATE ELECTRONICS AND DEVICES



CHAPTER 1

INTRODUCTION TO ELECTRONICS 3

CHAPTER 2

SOLID-STATE ELECTRONICS 42

CHAPTER 3

SOLID-STATE DIODES AND DIODE CIRCUITS 73

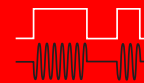
CHAPTER 4

BIPOLAR JUNCTION TRANSISTORS 146

CHAPTER 5

FIELD-EFFECT TRANSISTORS 212

1



INTRODUCTION TO ELECTRONICS

- 1.1 A Brief History of Electronics: From Vacuum Tubes to Giga-Scale Integration 4
- 1.2 Classification of Electronic Signals 8
- 1.3 Notational Conventions 12
- 1.4 Problem-Solving Approach 13
- 1.5 Important Concepts from Circuit Theory 15
- 1.6 Frequency Spectrum of Electronic Signals 21
- 1.7 Amplifiers 22
- 1.8 Element Variations in Circuit Design 26
- 1.9 Numeric Precision 34
- Summary 34
- Key Terms 35
- References 36
- Additional Reading 36
- Problems 37

- Present a brief history of electronics
- Quantify the explosive development of integrated circuit technology
- Discuss initial classification of electronic signals
- Review important notational conventions and concepts from circuit theory
- Introduce methods for including tolerances in circuit analysis
- Present the problem-solving approach used in this text

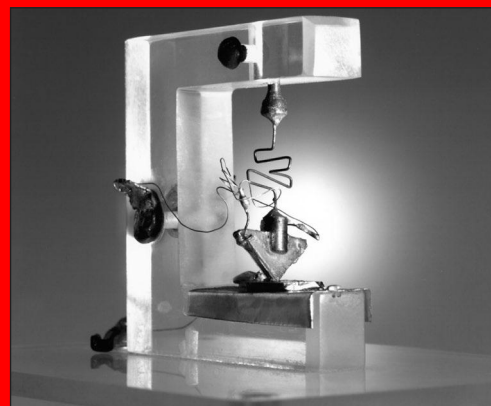
November 2022 is the 75th anniversary of the 1947 discovery of the bipolar transistor by John Bardeen and Walter Brattain at Bell Laboratories, a seminal event that marked the beginning of the semiconductor age (see Figs. 1.1 and 1.2). The invention of the transistor and the subsequent development of microelectronics have done more to shape the modern era than any other event. The transistor and microelectronics have reshaped how business is transacted, machines are designed, information moves, wars are fought, people interact, and countless other areas of our lives.

This textbook develops the basic operating principles and design techniques governing the behavior of the devices and circuits that form the backbone of much of the infrastructure of our modern world. This knowledge will enable students who aspire to design and create the next generation of this technological revolution to build



John Bardeen, William Shockley, and Walter Brattain in Brattain's laboratory in 1948.

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The first germanium bipolar transistor.

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a solid foundation for more advanced design courses. In addition, students who expect to work in some other technology area will learn material that will help them understand microelectronics, a technology that will continue to have impact on how their chosen field develops. This understanding will enable them to fully exploit microelectronics in their own technology area. Now let us return to our short history of the transistor.

After the discovery of the transistor, it was but a few months until William Shockley developed a theory that described the operation of the bipolar junction transistor.

Only 10 years later, in 1956, Bardeen, Brattain, and Shockley received the Nobel Prize in physics for the discovery of the transistor.

In June 1948 Bell Laboratories held a major press conference to announce the discovery. In 1952 Bell Laboratories, operating under legal consent decrees, made licenses for the transistor available for the modest fee of \$25,000 plus future royalty payments. About this time, Gordon Teal, another member of the solid-state group, left Bell Laboratories to work on the transistor at Geophysical

Services, Inc., which subsequently became Texas Instruments (TI). There he made the first silicon transistors, and TI marketed the first all-transistor radio. Another early licensee of the transistor was Tokyo Tsushin Kogyo, which became the Sony Company in 1955. Sony subsequently sold a transistor radio with a marketing strategy based on the idea that everyone could now have a personal radio; thus was launched the consumer market for transistors. A very interesting account of these and other developments can be found in [1, 2] and their references.

Activity in electronics began more than a century ago with the first radio transmissions in 1895 by Marconi, and these experiments were followed after only a few years by the invention of the first electronic amplifying device, the triode vacuum tube. In this period, electronics—loosely defined as the design and application of electron devices—has had such a significant impact on our lives that we often overlook just how pervasive electronics has really become. One measure of the degree of this impact can be found in the gross domestic product (GDP) of the world. In 2020 the world GDP was approximately U.S. \$90 trillion, and of this total more than 15 percent was directly traceable to electronics [3–5].

We commonly encounter electronics in the form of cellular phones, radios, televisions, and audio equipment, but electronics can be found even in seemingly mundane appliances such as vacuum cleaners, washing machines, and refrigerators. Wherever one looks in industry, electronics is found. The corporate world obviously depends heavily on data processing systems to manage its operations. In fact, it is hard to see how the computer industry could have evolved without the use of its own products. In addition, the design process depends ever more heavily on computer-aided design (CAD) systems, and manufacturing relies on electronic systems for process control—in petroleum refining, automobile tire production, food processing, power generation, and so on.

1.1 A BRIEF HISTORY OF ELECTRONICS: FROM VACUUM TUBES TO GIGA-SCALE INTEGRATION

Because most of us have grown up with electronic products all around us, we often lose perspective of how far the industry has come in a relatively short time. At the beginning of the twentieth century, there were no commercial electron devices, and transistors were not invented until the late 1940s! Explosive growth was triggered by first the commercial availability of the bipolar transistor in the late 1950s, and then the realization of the integrated circuit (IC) in 1961. Since that time, signal processing using electron devices and electronic technology has become a pervasive force in our lives.

Table 1.1 lists a number of important milestones in the evolution of the field of electronics. The Age of Electronics began in the early 1900s with the invention of the first electronic two-terminal devices, called **diodes**. The **vacuum diode**, or diode **vacuum tube**, was invented by Fleming in 1904; in 1906 Pickard created a diode by forming a point contact to a silicon crystal. (Our study of electron devices begins with the introduction of the solid-state diode in Chapter 3.)

DeForest's invention of the three-element vacuum tube known as the **triode** was an extremely important milestone. The addition of a third element to a diode enabled electronic amplification to take place with good isolation between the input and output ports of the device.

Milestones in Electronics

YEAR	EVENT
1874	Ferdinand Braun invents the solid-state rectifier.
1884	American Institute of Electrical Engineers (AIEE) formed.
1895	Marconi makes first radio transmissions.
1904	Fleming invents diode vacuum tube—Age of Electronics begins.
1906	Pickard creates solid-state point-contact diode (silicon).
1906	Deforest invents triode vacuum tube (audion).
1910–1911	“Reliable” tubes fabricated.
1912	Institute of Radio Engineers (IRE) founded.
1907–1927	First radio circuits developed from diodes and triodes.
1920	Armstrong invents super heterodyne receiver.
1925	TV demonstrated.
1925	Lilienfeld files patent application on the field-effect device.
1927–1936	Multigrid tubes developed.
1933	Armstrong invents FM modulation.
1935	Heil receives British patent on a field-effect device.
1940	Radar developed during World War II—TV in limited use.
1947	Bardeen, Brattain, and Shockley at Bell Laboratories invent bipolar transistors.
1950	First demonstration of color TV.
1952	Shockley describes the unipolar field-effect transistor.
1952	Commercial production of silicon bipolar transistors begins at Texas Instruments.
1952	Ian Ross and George Dacey demonstrate the junction field-effect transistor.
1956	Bardeen, Brattain, and Shockley receive Nobel Prize for invention of bipolar transistors.
1958	Integrated circuit developed simultaneously by Kilby at Texas Instruments and Noyce and Moore at Fairchild Semiconductor.
1961	First commercial digital IC available from Fairchild Semiconductor.
1963	AIEE and IRE merge to become the Institute of Electrical and Electronic Engineers (IEEE)
1967	First semiconductor RAM (64 bits) discussed at the IEEE International Solid-State Circuits Conference (ISSCC).
1968	First commercial IC operational amplifier—the μ A709—introduced by Fairchild Semiconductor.
1970	One-transistor dynamic memory cell invented by Dennard at IBM.
1970	Low-loss optical fiber invented.
1971	4004 microprocessor introduced by Intel.
1972	First 8-bit microprocessor—the 8008—introduced by Intel.
1973	Martin Cooper demonstrated a prototype of Motorola’s handheld mobile phone.
1974	First commercial 1-kilobit memory chip developed.
1974	8080 microprocessor introduced.
1978	First 16-bit microprocessor developed.
1984	Megabit memory chip introduced.
1985	Flash memory introduced at ISSCC.
1987	Erbium doped, laser-pumped optical fiber amplifiers demonstrated.
1995	Experimental gigabit memory chip presented at the IEEE ISSCC.
2000	Alferov, Kilby, and Kromer share the Nobel Prize in physics for optoelectronics, invention of the integrated circuit, and heterostructure devices, respectively.
2007	Fert and Grünberg share the Nobel Prize in physics for the discovery of giant magnetoresistance.
2009	Kao shares one-half of the 2009 Nobel Prize in physics for fiber optic communication using light with Boyle and Smith for invention of the Charge-Coupled Device (CCD).
2010	Geim and Novoselov share the Nobel Prize in physics for groundbreaking experiments regarding the two-dimensional material graphene.
2014	Akasaki, Amano, and Nakamura share the Nobel Prize in physics for the invention of efficient blue light-emitting diodes, which has enabled bright and energy-saving white light sources.
2018	Ten billion transistor integrated circuit chip presented at ISSCC.
2019	Goodenough, Whittingham, and Yoshino share the Nobel Prize in chemistry for the development of lithium-ion batteries.

Silicon-based three-element devices now form the basis of virtually all electronic systems. Fabrication of tubes that could be used reliably in circuits followed the invention of the triode by a few years and enabled rapid circuit innovation. Amplifiers and oscillators were developed that significantly improved radio transmission and reception. Armstrong invented the super heterodyne receiver in 1920 and FM modulation in 1933. Electronics developed rapidly during World War II, with great advances in the field of radio communications and the development of radar. Although first demonstrated in 1930, television did not begin to come into widespread use until the 1950s.

An important event in electronics occurred in 1947, when John Bardeen, Walter Brattain, and William Shockley at Bell Telephone Laboratories invented the **bipolar transistor**.¹ Although field-effect devices had actually been conceived by Lilienfeld in 1925, Heil in 1935, and Shockley in 1952 [2], the technology to produce such devices on a commercial basis did not yet exist. Bipolar devices, however, were rapidly commercialized.

Then in 1958, the nearly simultaneous invention of the **integrated circuit (IC)** by Kilby at Texas Instruments and Noyce and Moore at Fairchild Semiconductor produced a new technology that would profoundly change our lives. The miniaturization achievable through IC technology made available complex electronic functions with high performance at low cost. The attendant characteristics of high reliability, low power, and small physical size and weight were additional important advantages.

In 2000, Jack St. Clair Kilby received a share of the Nobel Prize for the invention of the integrated circuit. In the mind of the authors, this was an exceptional event as it represented one of the first awards to an electronic technologist.

Most of us have had some experience with personal computers, and nowhere is the impact of the integrated circuit more evident than in the area of digital electronics. For example, 4-gigabit (Gb) dynamic memory chips, similar to those in Fig. 1.3(c), contain more than 4 billion transistors. A 128-Gb flash memory chip stores 2 or 3 bits per memory cell using multilevel storage techniques and has more than 17 billion transistors in the memory array alone, not counting address decoding and sensing circuitry. Creating this much memory using individual vacuum tubes [depicted in Fig. 1.3(a)] or even discrete transistors [shown in Fig. 1.3(b)] would be almost inconceivable (see Prob. 1.9).

Levels of Integration

The dramatic progress of integrated circuit miniaturization is shown graphically in Figs. 1.4 and 1.5. The complexities of memory chips and microprocessors have grown exponentially with time. In over four decades since 1970, the number of transistors on a microprocessor chip has increased by a factor of 10 million as depicted in Fig. 1.4. Similarly, memory density has grown by a factor of more than 10 million from a 64-bit chip in 1968 to the announcement of 32-Gb chip production in 2018.

Since the commercial introduction of the integrated circuit, these increases in density have been achieved through a continued reduction in the minimum line width, or **minimum feature size**, that can be defined on the surface of the integrated circuit (see Fig. 1.5). Today most corporate semiconductor laboratories around the world are actively working on deep submicron processes with feature sizes below 10 nm—less than one five-thousandth the diameter of a human hair.

As the miniaturization process has continued, a series of commonly used abbreviations has evolved to characterize the various levels of integration. Prior to the invention of the integrated circuit, electronic systems were implemented in discrete form. Early ICs, with fewer than

¹ The term **transistor** is said to have originated as a contraction of “transfer resistor,” based on the voltage-controlled resistance of the characteristics of the MOS transistor.

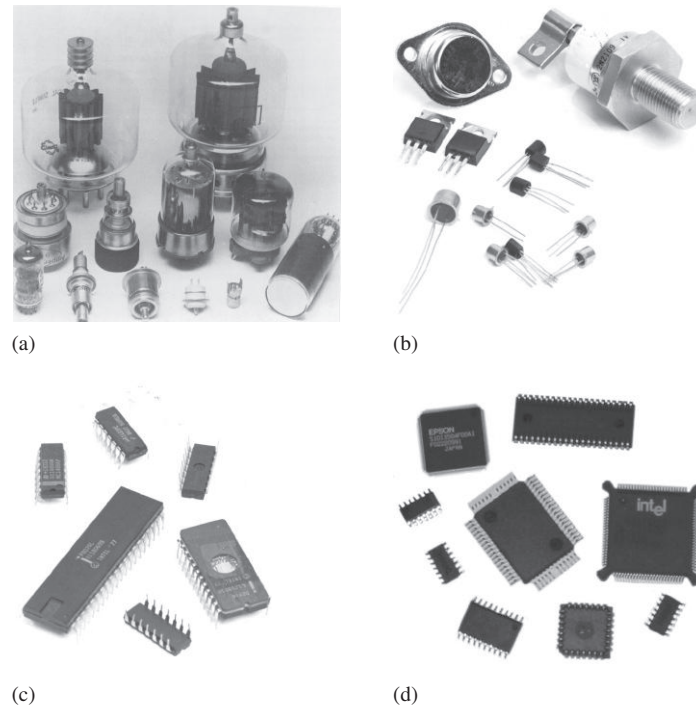


Figure 1.3 Comparison of (a) vacuum tubes, (b) individual transistors, (c) integrated circuits in dual-in-line packages (DIPs), and (d) ICs in surface mount packages.

Source: (a) Courtesy of ARRL Handbook for Radio Amateurs, 1992; (b, c, and d) Richard Jaeger

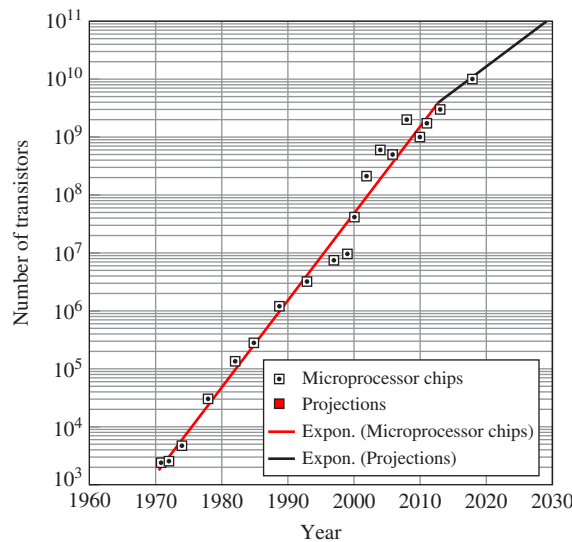


Figure 1.4 Microprocessor complexity versus time.

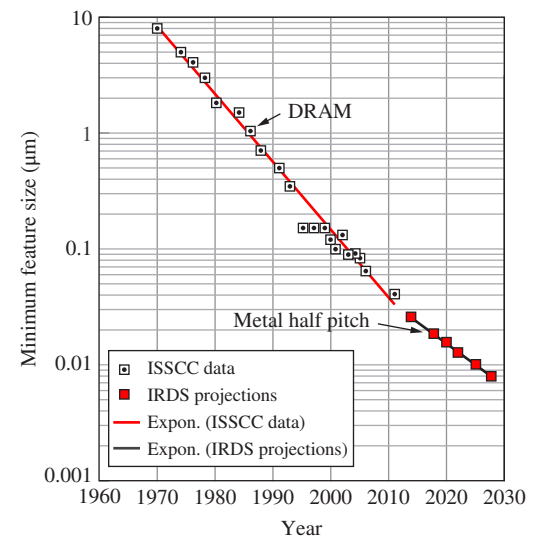


Figure 1.5 DRAM feature size versus year.

100 components, were characterized as **small-scale integration**, or **SSI**. As density increased, circuits became identified as **medium-scale integration** (**MSI**, 100–1000 components/chip), **large-scale integration** (**LSI**, 10^3 – 10^4 components/chip), and **very-large-scale integration** (**VLSI**, 10^4 – 10^9 components/chip). Today discussions focus on **giga-scale integration** (**GSI**, above 10^9 components/chip) and beyond.

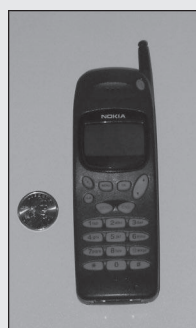
ELECTRONICS IN ACTION

Cellular Phone Evolution

The impact of technology scaling is ever present in our daily lives. One example appears visually in the pictures of cellular phone evolution below. Early mobile phones were often large and had to be carried in a relatively large pouch (hence the term “bag phone”). The next generation of analog phones could easily fit in your hand, but they had poor battery life caused by their analog communications technology. Implementations of fourth- and fifth-generation digital cellular technology are considerably smaller and have much longer battery life. As IC density increased, additional functions such as high-function cameras, GPS, Bluetooth, and Wifi were integrated with the digital phone.



(a)



(b)



(c)

A decade of cellular phone evolution: (a) early Uniden “bag phone,” (b) Nokia analog phone, and (c) Apple iPhone.
Source: (a and b) Richard Jaeger; (c) Yalcin Sonat/Shutterstock

Cell phones also represent excellent examples of the application of **mixed-signal** integrated circuits that contain both analog and digital circuitry on the same chip. ICs in the cell phone contain analog radio-frequency receiver and transmitter circuitry, analog-to-digital and digital-to-analog converters, CMOS logic and memory, power conversion circuits, imaging chips, accelerometers, and more.

1.2 CLASSIFICATION OF ELECTRONIC SIGNALS

The signals that electronic devices are designed to process can be classified into two broad categories: analog and digital. **Analog signals** can take on a continuous range of values, and thus represent continuously varying quantities; purely **digital signals** can appear at only one of several discrete levels. Examples of these types of signals are described in more detail in the next two subsections, along with the concepts of digital-to-analog and analog-to-digital conversion, which make possible the interface between the two systems.

1.2.1 DIGITAL SIGNALS

When we speak of digital electronics, we are most often referring to electronic processing of **binary digital signals**, or signals that can take on only one of two discrete amplitude levels as illustrated in Fig. 1.6. The status of binary systems can be represented by two symbols: a logical 1 is assigned to represent one level, and a logical 0 is assigned to the second level.² The two

² This assignment facilitates the use of Boolean algebra, reviewed in Chapter S6 of the eBook.

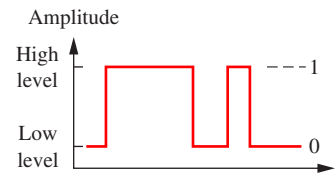


Figure 1.6 A time-varying binary digital signal.

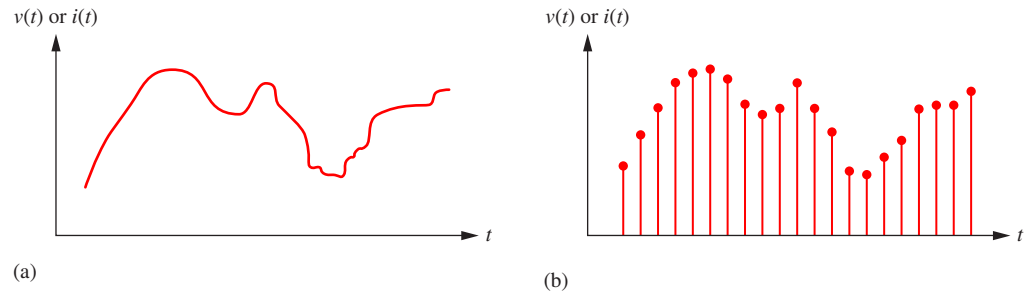


Figure 1.7 (a) A continuous analog signal; (b) sampled data version of signal in (a).

logic states generally correspond to two separate voltages— V_H and V_L —representing the high and low amplitude levels, and a number of voltage ranges are in common use. Although $V_H = 5$ V and $V_L = 0$ V represented the primary standard for many years, these have given way to lower voltage levels because of power consumption and semiconductor device limitations. Systems employing $V_H = 3.3$, down to 1 V or less with $V_L = 0$ V, are now used in many types of electronics.

However, binary voltage levels can also be negative or even bipolar. One high-performance logic family called ECL uses $V_H = -0.8$ V and $V_L = -2.0$ V, and the early standard RS-422 and RS-232 communication links between a small computer and its peripherals used $V_H = +12$ V and $V_L = -12$ V. In addition, the time-varying binary signal in Fig. 1.6 could equally well represent the amplitude of a current or that of an optical signal being transmitted down a fiber in an optical digital communication system. Recent USB and similar standards returned to the use of a single positive supply voltage.

Detailed discussion of logic circuits that were included in earlier editions can now be found in Chapters S6–S9 of the e-book. These include PMOS, NMOS, and CMOS logic,³ which use field-effect transistors, and the TTL and ECL families, which are based on bipolar transistors.

1.2.2 ANALOG SIGNALS

Although quantities such as electronic charge and electron spin or the position of a switch are discrete, much of the physical world is really analog in nature. Our senses of vision, hearing, smell, taste, and touch are all analog processes. Analog signals directly represent variables such as temperature, humidity, pressure, light intensity, or sound—all of which may take on any value, typically within some finite range. In practice, classification of digital and analog signals is largely one of perception. If we look at a digital signal similar to the one in Fig. 1.6 with an oscilloscope, we find that it actually makes a continuous transition between the high and low levels. The signal cannot make truly abrupt transitions between two levels. Designers of high-speed digital systems soon realize that they are really dealing with analog signals. The time-varying voltage or current plotted in Fig. 1.7(a) could be the electrical representation of temperature, flow rate, or pressure versus time, or the continuous audio output from a microphone. Some analog transducers

³ For now, let us accept these initials as proper names without further definition. The details of each of these circuits are developed in Chapters S6–S9 of the eBook.

produce output *voltages* in the range of 0 to 5 or 0 to 10 V, whereas others are designed to produce an output *current* that ranges between 4 and 20 mA. At the other extreme, signals detected by a radio antenna can be as small as a fraction of a microvolt.

To process the information contained in these analog signals, electronic circuits are used to selectively modify the amplitude, phase, and frequency content of the signals. In addition, significant increases in the voltage, current, and power level of the signal are usually needed. All these modifications to the signal characteristics are achieved using various forms of amplifiers, and Parts Two and Three of this text provide an in-depth discussion of the analysis and design of a wide range of amplifiers using operational amplifiers and bipolar and field-effect transistors.

1.2.3 A/D AND D/A CONVERTERS—BRIDGING THE ANALOG AND DIGITAL DOMAINS

For analog and digital systems to be able to operate together, we must be able to convert signals from analog to digital form and vice versa. We sample the input signal at various points in time as in Fig. 1.7(b) and convert or quantize its amplitude into a digital representation. The quantized value can be represented in binary form or can be a decimal representation as given by the display on a digital multimeter. The electronic circuits that perform these translations are called analog-to-digital (A/D) and digital-to-analog (D/A) converters.

Digital-to-Analog Conversion

The **digital-to-analog converter**, often referred to as a **D/A converter** or **DAC**, provides an interface between the digital signals of computer systems and the continuous signals of the analog world. The D/A converter takes digital information, most often in binary form, as input and generates an output voltage or current that may be used for electronic control or analog information display. In the DAC in Fig. 1.8(a), an n -bit binary input word (b_1, b_2, \dots, b_n) is treated as a binary fraction and multiplied by a full-scale reference voltage V_{FS} to set the output of the D/A converter. The behavior of the DAC can be expressed mathematically as

$$v_O = (b_1 2^{-1} + b_2 2^{-2} + \dots + b_n 2^{-n}) V_{FS} \quad \text{for } b_i \in \{1, 0\} \quad (1.1)$$

Examples of typical values of the full-scale voltage V_{FS} are 1, 2, 5, 5.12, 10, and 10.24 V. The smallest voltage change that can occur at the output takes place when the **least significant bit** b_n , or **LSB**, in the digital word changes from a 0 to a 1. This minimum voltage change is given by

$$V_{LSB} = 2^{-n} V_{FS} \quad (1.2)$$

At the other extreme, b_1 is referred to as the **most significant bit**, or **MSB**, and has a weight of one-half V_{FS} .

The **resolution of a converter** is typically specified in terms of the number of digital bits (i.e., 8-, 10-, 12-, 14- or 16-bit resolution, and so on).

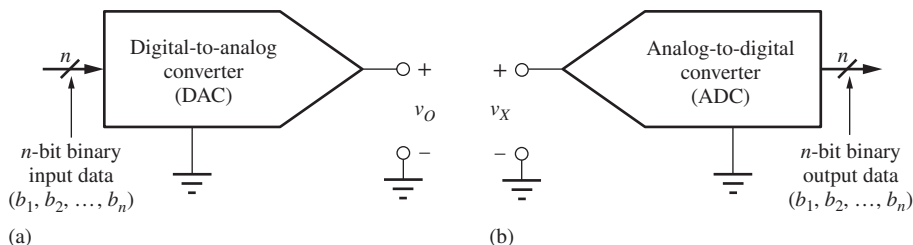


Figure 1.8 Block diagram representation for (a) a D/A converter and (b) an A/D converter.

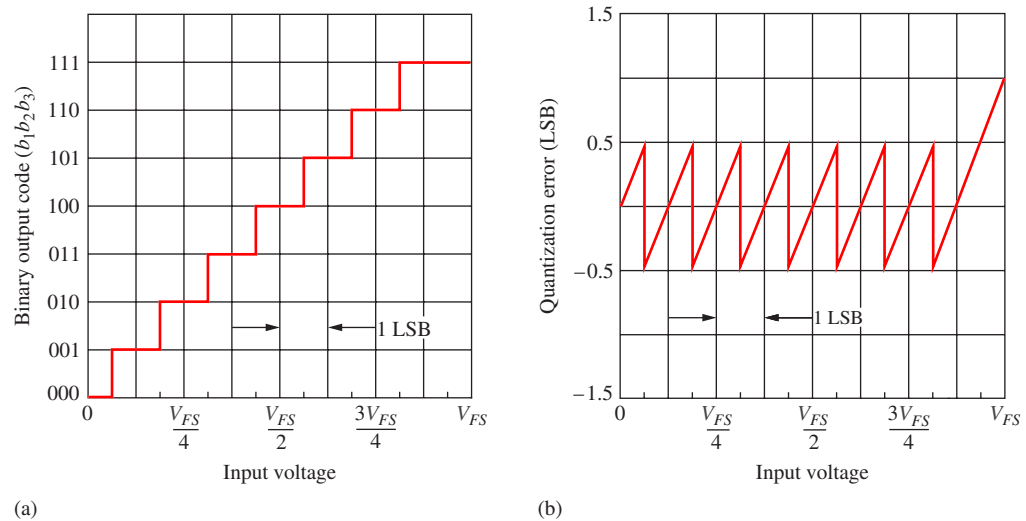


Figure 1.9 (a) Input–output relationship and (b) quantization error for 3-bit ADC.

A 10-bit D/A converter has $V_{FS} = 5.12$ V. What is the output voltage for a binary input code of (1100010001)? What is V_{LSB} ? What is the size of the MSB?

3.925 V; 5 mV; 2.56 V

Analog-to-Digital Conversion

The **analog-to-digital converter (A/D converter or ADC)** is used to transform analog information in electrical form into digital data. The ADC in Fig. 1.8(b) takes an unknown continuous analog input signal, usually a voltage v_X , and converts it into an n -bit binary number that can be easily manipulated by a computer. The n -bit number is a binary fraction representing the ratio between the unknown input voltage v_X and the converter's full-scale voltage V_{FS} .

For example, the input–output relationship for an ideal 3-bit A/D converter is shown in Fig. 1.9(a). As the input increases from zero to full scale, the output digital code word stair-steps from 000 to 111.⁴ The output code is constant for an input voltage range equal to 1 LSB of the ADC. Thus, as the input voltage increases, the output code first underestimates and then overestimates the input voltage. This error, called **quantization error**, is plotted against input voltage in Fig. 1.9(b).

For a given output code, we know only that the value of the input voltage lies somewhere within a 1-LSB quantization interval. For example, if the output code of the 3-bit ADC is 100, corresponding to a voltage $V_{FS}/2$, then the input voltage can be anywhere between $\frac{7}{16}V_{FS}$ and $\frac{9}{16}V_{FS}$, a range of $V_{FS}/8$ V or 1 LSB. From a mathematical point of view, the ADC circuitry in Fig. 1.8(b) picks the values of the bits in the binary word to minimize the magnitude of the quantization error v_e between the unknown input voltage v_X and the nearest quantized voltage level:

$$v_e = |v_X - (b_12^{-1} + b_22^{-2} + \cdots + b_n2^{-n})V_{FS}| \quad (1.3)$$

⁴ The binary point is understood to be to the immediate left of the digits of the code word. As the code word stair-steps from 000 to 111, the binary fraction steps from 0.000 to 0.111.

(a) An 8-bit A/D converter has $V_{FS} = 5$ V. What is the digital output code word for an input of 1.2 V? What is the voltage range corresponding to 1 LSB of the converter? (b) Repeat for $V_{FS} = 5.12$ V.

001111101; 19.5 mV; 001111100; 20.0 mV

1.3 NOTATIONAL CONVENTIONS

In many circuits we will be dealing with both dc and time-varying values of voltages and currents. The following standard notation will be used to keep track of the various components of an electrical signal. Total quantities will be represented by lowercase letters with capital subscripts, such as v_T and i_T in Eq. (1.4). The dc components are represented by capital letters with capital subscripts as, for example, V_{DC} and I_{DC} in Eq. (1.4); changes or variations from the dc value are represented by signal components v_{sig} and i_{sig} . Total quantities are then given by

$$v_T = V_{DC} + v_{sig} \quad \text{or} \quad i_T = I_{DC} + i_{sig} \quad (1.4)$$

As examples, the total base-emitter voltage v_{BE} of a transistor and the total drain current i_D of a field-effect transistor are written as

$$v_{BE} = V_{BE} + v_{be} \quad \text{and} \quad i_D = I_D + i_d \quad (1.5)$$

Unless otherwise indicated, the equations describing a given network will be written assuming a consistent set of units: volts, amperes, and ohms. For example, the equation $5 \text{ V} = (10,000 \text{ } \Omega)I_1 + 0.6 \text{ V}$ may be written as $5 = 10,000I_1 + 0.6$.

The fourth upper/lowercase combination, such as V_{be} or I_d , is reserved for the amplitude of a sinusoidal signal's phasor representation as defined in Sec. 1.5.

Suppose the voltage at a circuit node is described by

$$v_A = (5 \sin 2000\pi t + 4 + 3 \cos 1000\pi t) \text{ V}$$

What are the expressions for V_A and v_a ?

$$V_A = 4 \text{ V}; v_a = (5 \sin 2000\pi t + 3 \cos 1000\pi t) \text{ V}$$

Resistance and Conductance Representations

In the circuits throughout this text, resistors will be indicated symbolically as R_x or r_x , and the values will be expressed in Ω , $\text{k}\Omega$, $\text{M}\Omega$, and so on. During analysis, however, it may be more convenient to work in terms of conductance with the following convention:

$$G_x = \frac{1}{R_x} \quad \text{and} \quad g_\pi = \frac{1}{r_\pi} \quad (1.6)$$

For example, conductance G_x always represents the reciprocal of the value of R_x , and g_π represents the reciprocal of r_π . The values next to a resistor symbol will always be expressed in terms of resistance (Ω , $\text{k}\Omega$, $\text{M}\Omega$).

Dependent Sources

In electronics, **dependent** (or **controlled**) **sources** are used extensively. Four types of dependent sources are summarized in Fig. 1.10, in which the standard diamond shape is used for controlled sources. The **voltage-controlled current source (VCCS)**, **current-controlled current source (CCCS)**, and **voltage-controlled voltage source (VCVS)** are used routinely in this text to model transistors and amplifiers or to simplify more complex circuits. Only the **current-controlled voltage source (CCVS)** sees limited use.

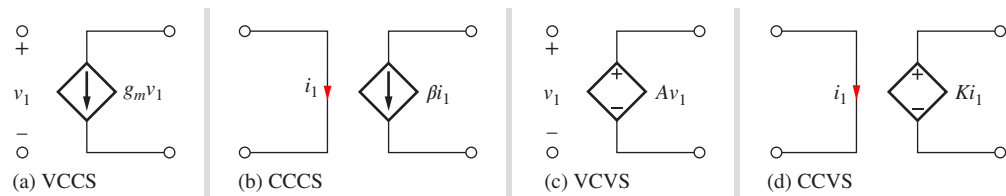


Figure 1.10 Controlled sources: (a) voltage-controlled current source (VCCS); (b) current-controlled current source (CCCS); (c) voltage-controlled voltage source (VCVS); (d) current-controlled voltage source (CCVS).

1.4 PROBLEM-SOLVING APPROACH

Solving problems is a centerpiece of an engineer's activity. As engineers, we use our creativity to find new solutions to problems that are presented to us. A well-defined **problem-solving approach** is essential. The examples in this text highlight an approach that can be used in all facets of your career, as a student and as an engineer in industry. The method is outlined in the following nine steps:

1. State the **problem** as clearly as possible.
2. List the **known information and given data**.
3. Define the **unknowns** that must be found to solve the problem.
4. List your **assumptions**. You may discover additional assumptions as the analysis progresses.
5. Develop an **approach** from a group of possible alternatives.
6. Perform an **analysis** to find a solution to the problem. As part of the analysis, be sure to draw the circuit and label the variables.
7. **Check the results**. Has the problem been solved? Is the math correct? Have all the unknowns been found? Have the assumptions been satisfied? Do the results satisfy simple consistency checks?
8. **Evaluate the solution**. Is the solution realistic? Can it be built? If not, repeat steps 4–7 until a satisfactory solution is obtained.
9. **Computer-aided analysis**. SPICE and other computer tools are highly useful to check the results and to see if the solution satisfies the problem requirements. Compare the computer results to your hand results.

To begin solving a problem, we must try to understand its details. The first four steps, which attempt to clearly define the problem, can be the most important part of the solution process. Time spent in understanding, clarifying, and defining the problem can save much time and frustration.

The first step is to write down a statement of the problem. The original problem description may be quite vague; we must try to understand the problem as well as, or even better than, the individual who posed the problem. As part of this focus on understanding the problem, we list the information that is known and unknown. Problem-solving errors can often be traced to imprecise definition of the unknown quantities. For example, it is very important for analysis to draw the circuit properly and to clearly label voltages and currents on our circuit diagrams.

Often there are more unknowns than constraints, and we need engineering judgment to reach a solution. Part of our task in studying electronics is to build up the background for selecting between various alternatives. Along the way, we often need to make approximations and assumptions that simplify the problem or form the basis of the chosen approach. It is important to state these assumptions, so that we can be sure to check their validity at the end. Throughout this text you will encounter opportunities to make assumptions. Most often, you should make assumptions that simplify your computational effort yet still achieve useful results.

The exposition of the known information, unknowns, and assumptions helps us not only to better understand the problem but also to think about various alternative solutions. We must choose the approach that appears to have the best chance of solving the problem. There may be more than one satisfactory approach. Each person will view the problem somewhat differently, and the approach that is clearest to one individual may not be the best for another. Pick the one that seems best to you. As part of defining the approach, be sure to think about what computational tools are available to assist in the solution, including MATLAB®, Mathcad®, spreadsheets, SPICE, and your calculator.

Once the problem and approach are defined as clearly as possible, then we can perform any analysis required and solve the problem. After the analysis is completed we need to check the results. A number of questions should be resolved. First, have all the unknowns been found? Do the results make sense? Are they consistent with each other? Are the results consistent with assumptions used in developing the approach to the problem?

Then we need to evaluate the solution. Are the results viable? For example, are the voltage, current, and power levels reasonable? Can the circuit be realized with reasonable yield with real components? Will the circuit continue to function within specifications in the face of significant component variations? Is the cost of the circuit within specifications? If the solution is not satisfactory, we need to modify our approach and assumptions and attempt a new solution. An iterative solution is often required to meet the specifications in realistic design situations. SPICE and other computer tools are highly useful for checking results and ensuring that the solution satisfies the problem requirements.

The solutions to the examples in this text have been structured following the problem-solving approach introduced here. Although some examples may appear trivial, the power of the structured approach increases as the problem becomes more complex.

WHAT ARE REASONABLE NUMBERS?

Part of our “check of results” should be to decide if the answer is “reasonable” and makes sense. Over time we must build up an understanding of what numbers are reasonable. Most solid-state devices that we will encounter are designed to operate from voltages ranging from a battery voltage of less than 1 V on the low end to no more than 40–50 V⁵ at the high end. Typical power supply voltages will range from less than 1 V to 20 V or so, and typical resistance values encountered will range from a few ohms up to many GΩ.

Based on our knowledge of dc circuits, we should expect that the voltages in our circuits not exceed the power supply voltages. For example, if a circuit is operating from +8 and –5-V supplies, all of our calculated dc voltages should be between –5 and +8 V. In addition, the peak-to-peak amplitude of an ac signal should not exceed 13 V, the difference of the two supply voltages. With a 10-V supply, the maximum current that can go through a 100-Ω resistor is 100 mA; the current through a 10-MΩ resistor can be no more than 1 μA. Thus we should remember the following “rules” to check our results:

1. With few exceptions, the dc voltages in our circuits will not exceed the power supply voltages. The peak-to-peak amplitude of an ac signal should not exceed the difference of the power supply voltages.
2. The currents in our circuits will range from microamperes to no more than a hundred milliamperes or so.
3. So if a calculation yields a voltage exceeding that of the power supply, or a current of greater than 1 A, you need to check your work.

⁵ The primary exception is in the area of power electronics, where one encounters much larger voltages and currents than the ones mentioned here.

1.5 IMPORTANT CONCEPTS FROM CIRCUIT THEORY

Analysis and design of electronic circuits make continuous use of a number of important techniques from basic network theory. Circuits are most often analyzed using a combination of **Kirchhoff's voltage law**, abbreviated **KVL**, and **Kirchhoff's current law**, abbreviated **KCL**. Occasionally, the solution relies on systematic application of **nodal** or **mesh analysis**. **Thévenin** and **Norton circuit transformations** are often used to help simplify circuits, and the notions of voltage and current division are extremely useful. Models of active devices invariably involve dependent sources, as mentioned in the last section, and we need to be familiar with dependent sources in all forms. Amplifier analysis also uses two-port network theory. A review of two-port networks is deferred until the introductory discussion of amplifiers in Chapter 6. If the reader feels uncomfortable with any of the concepts just mentioned, this is a good time for review. To help, a brief review of these important circuit techniques follows.

1.5.1 VOLTAGE AND CURRENT DIVISION

Voltage and current division are highly useful circuit analysis techniques that can be derived directly from basic circuit theory. They are both used routinely throughout this text, and it is very important to be sure to understand the conditions for which each technique is valid! Examples of both methods are provided next.

Voltage division is demonstrated by the circuit in Fig. 1.11(a) in which the voltages v_1 and v_2 can be expressed as

$$v_1 = i_i R_1 \quad \text{and} \quad v_2 = i_i R_2 \quad (1.7)$$

Applying KVL to the single loop,

$$v_i = v_1 + v_2 = i_i(R_1 + R_2) \quad \text{and} \quad i_i = \frac{v_i}{R_1 + R_2} \quad (1.8)$$

Combining Eqs. (1.7) and (1.8) yields the basic voltage division formula:

$$v_1 = v_i \frac{R_1}{R_1 + R_2} \quad \text{and} \quad v_2 = v_i \frac{R_2}{R_1 + R_2} \quad (1.9)$$

For the resistor values in Fig. 1.11(a),

$$v_1 = 10 \text{ V} \frac{8 \text{ k}\Omega}{8 \text{ k}\Omega + 2 \text{ k}\Omega} = 8.00 \text{ V} \quad \text{and} \quad v_2 = 10 \text{ V} \frac{2 \text{ k}\Omega}{8 \text{ k}\Omega + 2 \text{ k}\Omega} = 2.00 \text{ V} \quad (1.10)$$

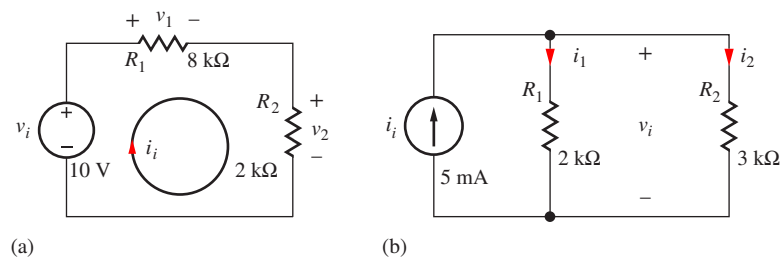
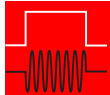


Figure 1.11 (a) A resistive voltage divider; (b) current division in a simple network.



VOLTAGE DIVIDER RESTRICTIONS

Note that the voltage divider relationships in Eq. (1.9) can be applied only when the current through the two resistor branches is the same. Also, note that the formulas are correct if the resistances are replaced by complex impedances and the voltages are represented as **phasors**.

$$\mathbf{V}_1 = \mathbf{V}_i \frac{Z_1}{Z_1 + Z_2} \quad \text{and} \quad \mathbf{V}_2 = \mathbf{V}_i \frac{Z_2}{Z_1 + Z_2}$$

Current division is also very useful. Let us find the currents i_1 and i_2 in the circuit in Fig. 1.11(b). Using KCL at the single node,

$$i_i = i_1 + i_2 \quad \text{where } i_1 = \frac{v_i}{R_1} \text{ and } i_2 = \frac{v_i}{R_2} \quad (1.11)$$

and solving for v_i yields

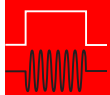
$$v_i = i_i \frac{1}{\frac{1}{R_1} + \frac{1}{R_2}} = i_i \frac{R_1 R_2}{R_1 + R_2} = i_i (R_1 \parallel R_2) \quad (1.12)$$

in which the notation $R_1 \parallel R_2$ represents the parallel combination of resistors R_1 and R_2 . Combining Eqs. (1.11) and (1.12) yields the current division formulas:

$$i_1 = i_i \frac{R_2}{R_1 + R_2} \quad \text{and} \quad i_2 = i_i \frac{R_1}{R_1 + R_2} \quad (1.13)$$

For the values in Fig. 1.11(b),

$$i_1 = 5 \text{ mA} \frac{3 \text{ k}\Omega}{2 \text{ k}\Omega + 3 \text{ k}\Omega} = 3.00 \text{ mA} \quad i_2 = 5 \text{ mA} \frac{2 \text{ k}\Omega}{2 \text{ k}\Omega + 3 \text{ k}\Omega} = 2.00 \text{ mA}$$



CURRENT DIVIDER RESTRICTIONS

It is important to note that the same voltage must appear across both resistors in order for the current division expressions in Eq. (1.13) to be valid. Here again, the formulas are correct if the resistances are replaced by complex impedances and the currents are represented as **phasors**.

$$\mathbf{I}_1 = \mathbf{I}_i \frac{Z_2}{Z_1 + Z_2} \quad \text{and} \quad \mathbf{I}_2 = \mathbf{I}_i \frac{Z_1}{Z_1 + Z_2}$$

1.5.2 THÉVENIN AND NORTON CIRCUIT REPRESENTATIONS

Let us now review the method for finding **Thévenin and Norton equivalent circuits**, including a dependent source; the circuit in Fig. 1.12(a) serves as our illustration. Because the linear network in the dashed box has only two terminals, it can be represented by either the Thévenin or Norton equivalent circuits in Fig. 1.12(b) and (c). The work of Thévenin and Norton permits us to reduce complex circuits to a single source and equivalent resistance. We illustrate these two important techniques with the next four examples.

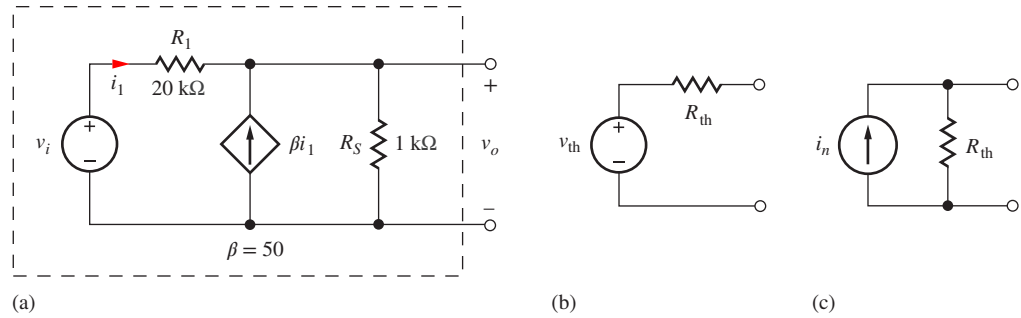


Figure 1.12 (a) Two-terminal circuit and its (b) Thévenin and (c) Norton equivalents.

THÉVENIN EQUIVALENT CIRCUIT

Let's practice finding the Thévenin and Norton equivalent circuits for the network in Fig. 1.12(a).

Find the Thévenin and Norton equivalent representations for the circuit in Fig. 1.12(a).

Known Information and Given Data: Circuit topology and values appear in Fig. 1.12(a).

Unknowns: Thévenin equivalent voltage v_{th} , Thévenin equivalent resistance R_{th} , and Norton equivalent current i_n .

Approach: Voltage source v_{th} is defined as the open-circuit voltage at the terminals of the circuit. R_{th} is the equivalent resistance at the terminals of the circuit with all **independent** sources set to zero. Source i_n represents the short-circuit current available at the output terminals and is equal to v_{th}/R_{th} .

Assumptions: None

Analysis: We will first find the value of v_{th} , then R_{th} and finally i_n . Open-circuit voltage v_{th} can be found by applying KCL at the output terminals where the notational convention for conductance from Sec. 1.3 ($G_S = 1/R_S$) has been applied.

$$\beta i_1 = \frac{v_o - v_i}{R_1} + \frac{v_o}{R_S} = G_1(v_o - v_i) + G_S v_o \quad (1.14)$$

Current i_1 is given by

$$i_1 = G_1(v_i - v_o) \quad (1.15)$$

Substituting Eq. (1.15) into Eq. (1.14) and combining terms yields

$$G_1(\beta + 1)v_i = [G_1(\beta + 1) + G_S]v_o \quad (1.16)$$

The Thévenin equivalent output voltage is then found to be

$$v_{th} = \frac{G_1(\beta + 1)}{[G_1(\beta + 1) + G_S]} v_i = \frac{(\beta + 1)R_S}{[(\beta + 1)R_S + R_1]} v_i \quad (1.17)$$

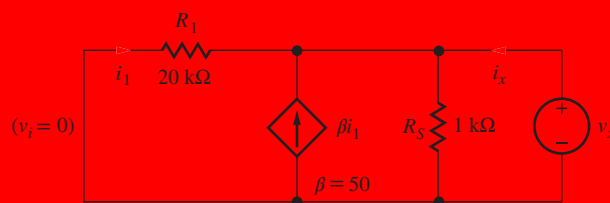
where the second relationship was found by multiplying numerator and denominator by $(R_1 R_S)$. For the values in this problem,

$$v_o = \frac{(50 + 1)1 \text{ k}\Omega}{[(50 + 1)1 \text{ k}\Omega + 20 \text{ k}\Omega]} v_i = 0.718 v_i \quad \text{and} \quad v_{th} = 0.718 v_i \quad (1.18)$$

R_{th} represents the equivalent resistance present at the output terminals with all independent sources set to zero. To find the **Thévenin equivalent resistance** R_{th} , we first set the independent sources in the network to zero. Remember, however, that **dependent sources must remain active**. A test voltage or current source is then applied to the network terminals and the corresponding current or voltage calculated. In Fig. 1.13 v_i is set to zero (i.e., replaced by a short circuit), voltage source v_x is applied to the network, and the current i_x must be determined so that

$$R_{th} = \frac{v_x}{i_x} \quad (1.19)$$

can be calculated.



A test source v_x is applied to the network to find R_{th} .

$$i_x = -i_1 - \beta i_1 + G_S v_x \quad \text{in which } i_1 = -G_1 v_x \quad (1.20)$$

Combining and simplifying these two expressions yield

$$i_x = [(\beta + 1)G_1 + G_S]v_x \quad \text{and} \quad R_{th} = \frac{v_x}{i_x} = \frac{1}{(\beta + 1)G_1 + G_S} \quad (1.21)$$

The denominator of Eq. (1.21) represents the sum of two conductances, which corresponds to the parallel combination of two resistances. Therefore, Eq. (1.21) can be rewritten as

$$R_{th} = \frac{1}{(\beta + 1)G_1 + G_S} = \frac{R_S \frac{R_1}{(\beta + 1)}}{R_S + \frac{R_1}{(\beta + 1)}} = R_S \parallel \frac{R_1}{(\beta + 1)} \quad (1.22)$$

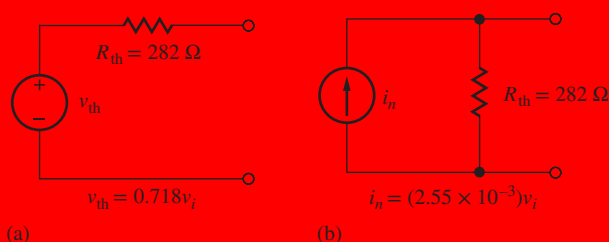
For the values in this example,

$$R_{th} = R_S \parallel \frac{R_1}{(\beta + 1)} = 1 \text{ k}\Omega \parallel \frac{20 \text{ k}\Omega}{(50 + 1)} = 1 \text{ k}\Omega \parallel 392 \text{ }\Omega = 282 \text{ }\Omega \quad (1.23)$$

Norton source i_n represents the short-circuit current available from the original network. Since we already have the Thévenin equivalent circuit, we can use it to easily find the value of i_n .

$$i_n = \frac{v_{th}}{R_{th}} = \frac{0.718 v_i}{282 \text{ }\Omega} = 2.55 \times 10^{-3} v_i$$

The Thévenin and Norton equivalent circuits for Fig. 1.12 calculated in the previous example appear for comparison in Fig. 1.14.



Completed (a) Thévenin and (b) Norton equivalent circuits for the two-terminal network in Fig. 1.12(a).

Check of Results: We have found the three unknowns required. A recheck of the calculations indicates they are done correctly. The value of v_{th} is the same order of magnitude as v_i , so its value should not be unusually large or small. The value of R_{th} is less than 1 kΩ, which seems reasonable, since we should not expect the resistance to exceed the value of R_S that appears in parallel with the output terminals. We can double-check everything by directly calculating i_n from the original circuit. If we short the output terminals in Fig. 1.12, we find the short-circuit current (see Ex. 1.2) to be $i_n = (\beta + 1)v_i/R_1 = 2.55 \times 10^{-3}v_i$ and in agreement with the other method.

NORTON EQUIVALENT CIRCUIT

Practice finding the Norton equivalent circuit for a network containing a dependent source.

Find the Norton equivalent (Fig. 1.12(c)) for the circuit in Fig. 1.12(a).

Known Information and Given Data: Circuit topology and circuit values appear in Fig. 1.12(a). The value of R_{th} was calculated in the previous example.

Unknowns: Norton equivalent current i_n .

Approach: The Norton equivalent current is found by determining the current coming out of the network when a short circuit is applied to the terminals.

Assumptions: None.

Analysis: For the circuit in Fig. 1.15, the output current will be

$$i_n = i_1 + \beta i_1 \quad \text{and} \quad i_1 = v_i/R_1 \quad (1.24)$$

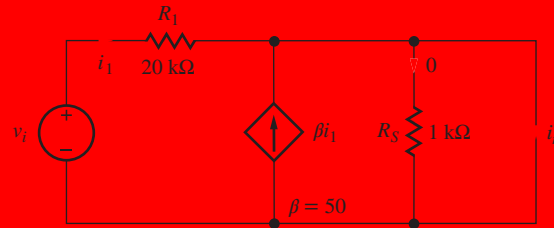
since the short circuit across the output forces the current through R_S to be 0. Combining the two expressions in Eq. (1.24) yields

$$i_n = (\beta + 1)G_1 v_i = \frac{(\beta + 1)}{R_1} v_i \quad (1.25)$$

or

$$i_n = \frac{(50 + 1)}{20 \text{ k}\Omega} v_i = \frac{v_i}{392 \Omega} = (2.55 \text{ mS}) v_i \quad (1.26)$$

The resistance in the Norton equivalent circuit also equals R_{th} found in Eq. (1.23).



Circuit for determining short-circuit output current.

Check of Results: We have found the Norton equivalent current. Note that $v_{th} = i_n R_{th}$ and this result can be used to check the calculations: $i_n R_{th} = (2.55 \text{ mS}) v_s (282 \Omega) = 0.719 v_s$, which agrees within round-off error with the previous example.

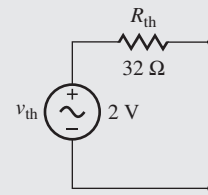
ELECTRONICS IN ACTION

Player Characteristics

The headphone amplifier in a personal music player represents an everyday example of a basic audio amplifier. The traditional audio band spans the frequencies from 20 Hz to 20 kHz, a range that extends beyond the hearing capability of most individuals at both the upper and lower ends.

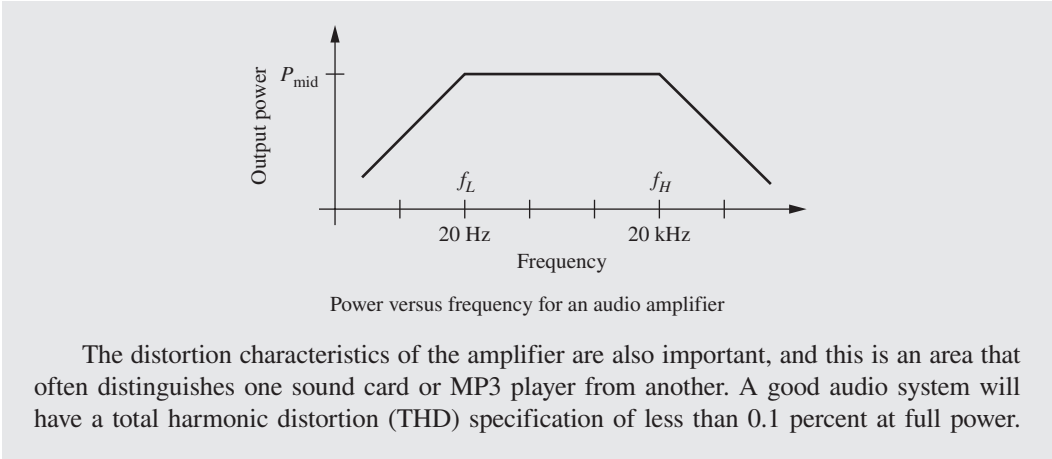


iPad
Framesira/Shutterstock



Thévenin equivalent
circuit for output stage

The characteristics of the Apple iPad in the accompanying figure are representative of a high-quality audio output stage in an MP3 player or a computer sound card. The output can be represented by a Thévenin equivalent circuit with $v_{th} = 2 \text{ V}$ and $R_{th} = 32 \text{ ohms}$, and the output stage is designed to deliver a power of approximately 15 mW into each channel of a headphone with a matched impedance of 32 ohms. The output power is approximately constant over the 20 Hz–20 kHz frequency range. At the lower and upper cutoff frequencies, f_L and f_H , the output power will be reduced by 3 dB, a factor of 2.



1.6 FREQUENCY SPECTRUM OF ELECTRONIC SIGNALS

Fourier analysis and the **Fourier series** represent extremely powerful tools in electrical engineering. Results from Fourier theory show that complicated signals are actually composed of a continuum of sinusoidal components, each having a distinct amplitude, frequency, and phase. The **frequency spectrum** of a signal presents the amplitude and phase of the components of the signal versus frequency.

Nonrepetitive signals have continuous spectra with signals that may occupy a broad range of frequencies. For example, the amplitude spectrum of a television signal measured during a small time interval is depicted in Fig. 1.16. The TV video signal is designed to occupy the frequency range from 0 to 4.5 MHz.⁶ Other types of signals occupy different regions of the frequency spectrum. Table 1.2 identifies the frequency ranges associated with various categories of common signals.

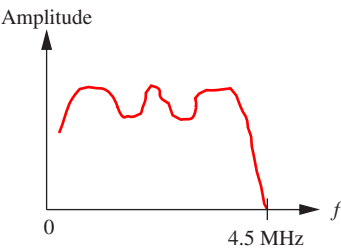


Figure 1.16 Spectrum of a TV signal.

Frequencies Associated with Common Signals

CATEGORY	FREQUENCY RANGE
Audible sounds	20 Hz–20 kHz
Baseband video (TV) signal	0–4.5 MHz
AM radio broadcasting	0.54–1.6 MHz
High-frequency radio communications	1.6–54 MHz
VHF television (Channels 2–6)	54–88 MHz
FM radio broadcasting	88–108 MHz
VHF radio communication	108–174 MHz
VHF television (Channels 7–13)	174–216 MHz
Maritime and government communications	216–450 MHz
Business communications	450–470 MHz
UHF television (Channels 14–69)	470–806 MHz
Fixed and mobile communications including allocations for analog and digital cellular telephones, personal communications, and other wireless devices, 3G/4G/LTE/5G	0.60–5.0 GHz
Satellite television	3.7–4.2 GHz
Wireless devices	5.0–5.5 GHz
Industrial, scientific, and medical (ISM) bands	6 MHz–246 MHz
Automotive radarband	76–81 GHz
Radio Astronomy	73 GHz

⁶ This signal is combined with a much higher carrier frequency prior to transmission.

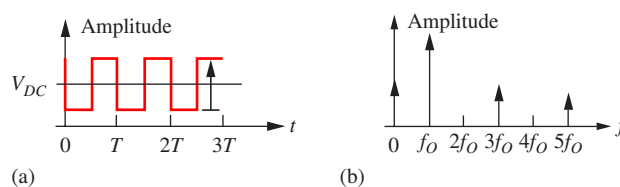


Figure 1.17 A periodic signal (a) and its amplitude spectrum (b).

In contrast to the continuous spectrum in Fig. 1.16, Fourier series analysis shows that *any* periodic signal, such as the square wave of Fig. 1.17, contains spectral components only at discrete frequencies⁷ that are related directly to the period of the signal. For example, the square wave of Fig. 1.17 having an amplitude V_O and period T can be represented by the Fourier series

$$v(t) = V_{DC} + \frac{2V_O}{\pi} \left(\sin \omega_o t + \frac{1}{3} \sin 3\omega_o t + \frac{1}{5} \sin 5\omega_o t + \dots \right) \quad (1.27)$$

in which $\omega_o = 2\pi/T$ (rad/s) is the **fundamental radian frequency** of the square wave. We refer to $f_o = 1/T$ (Hz) as the **fundamental frequency** of the signal, and the frequency components at $2f_o$, $3f_o$, $4f_o$, ... are called the second, third, fourth, and so on **harmonic frequencies**.

1.7 AMPLIFIERS

The characteristics of analog signals are most often manipulated using linear amplifiers that affect the amplitude and/or phase of the signal without changing its frequency. Although a complex signal may have many individual components, as just described in Sec. 1.6, linearity permits us to use the **superposition principle** to treat each component individually.

For example, suppose the amplifier with voltage gain A in Fig. 1.18(a) is fed a sinusoidal input signal component v_i with amplitude V_i , frequency ω_i , and phase ϕ :

$$v_i = V_i \sin(\omega_i t + \phi) \quad (1.28)$$

Then, if the amplifier is linear, the output corresponding to this signal component will also be a sinusoidal signal at the same frequency but with a different amplitude and phase:

$$v_o = V_o \sin(\omega_i t + \phi + \theta) \quad (1.29)$$

Using phasor notation, the input and output signals would be represented as

$$\mathbf{V}_i = V_i \angle \phi \quad \text{and} \quad \mathbf{V}_o = V_o \angle (\phi + \theta) \quad (1.30)$$

The **voltage gain** of the amplifier is defined in terms of these phasors:

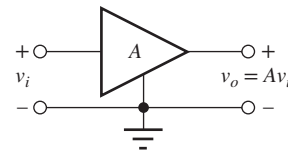
$$A = \frac{\mathbf{V}_o}{\mathbf{V}_i} = \frac{V_o \angle (\phi + \theta)}{V_i \angle \phi} = \frac{V_o}{V_i} \angle \theta \quad (1.31)$$

This amplifier has a voltage gain with magnitude equal to V_o/V_i and a phase shift of θ . In general, both the magnitude and phase of the voltage gain will be a function of frequency. Note that amplifiers also often provide current gain and power gain as well as voltage gain, but these concepts will not be explored further until Chapter 6.

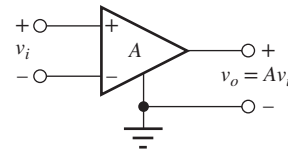
The curves in Fig. 1.19 represent the input and output voltage waveforms for an inverting amplifier with $A_v = -5$ and $v_i = 1 \sin 2000\pi t$ V. Both the factor of five increase in signal amplitude and the 180° phase shift (multiplication by -1) are apparent in the graph.

At this point, a note regarding the phase angle is needed. In Eqs. (1.28) and (1.29), ωt , ϕ , and θ must have the same units. With ωt normally expressed in radians, ϕ should also be in radians. However, in electrical engineering texts, ϕ is often expressed in degrees. We must be

⁷ There are an infinite number of components, however.



(a)



(b)

Figure 1.18 (a) Symbol for amplifier with single input and voltage gain A ; (b) differential amplifier having two inputs and gain A .

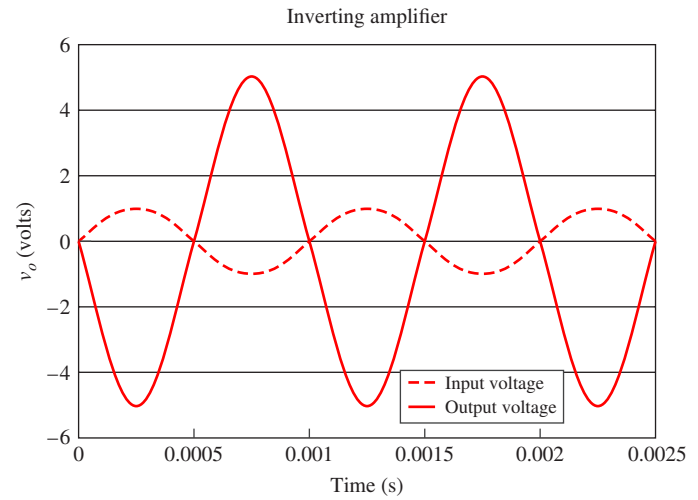


Figure 1.19 Voltage waveforms of input v_i and output v_o for an amplifier with gain $A_v = -5$ and $v_i = 1 \sin 2000\pi t$ V.

aware of this mixed system of units and remember to convert degrees to radians before making any numeric calculations.

The input and output voltages of an amplifier are expressed as

$$v_i = 0.001 \sin(2000\pi t) \text{ V} \quad \text{and} \quad v_o = -5 \cos(2000\pi t + 25^\circ) \text{ V}$$

in which v_i and v_o are specified in volts when t is in seconds. What are \mathbf{V}_i , \mathbf{V}_o , and the voltage gain of the amplifier?

$$0.001\angle 0^\circ; 5\angle -65^\circ; 5000\angle -65^\circ$$

1.7.1 IDEAL OPERATIONAL AMPLIFIERS

The **operational amplifier**, “**op amp**” for short, is a fundamental building block in electronic design and is discussed in most introductory circuit courses. A brief review of the ideal op amp is provided here; an in-depth study of the properties of ideal and nonideal op amps and the circuits used to build the op amp itself are the subjects of Chapters 10–14. Although it is impossible to realize the **ideal operational amplifier**, its use allows us to quickly understand the basic behavior to be expected from a given circuit and serves as a fundamental building block in circuit design.

From our basic circuit courses, we may recall that op amps are differential (or difference) amplifiers that respond to the signal voltage that appears between the $+$ and $-$ input terminals of the amplifier depicted in Fig. 1.18(b). Ideal op amps are assumed to have infinite **voltage gain** and infinite **input resistance**, and these properties lead to two special assumptions that are used to analyze circuits containing ideal op amps:

1. The voltage difference across the input terminals is zero; that is, $v_- = v_+$.
2. Both input currents are zero.

Applying the Assumptions—The Inverting Amplifier

The classic **inverting amplifier** circuit will be used to refresh our memory of the analysis of circuits employing op amps. The inverting amplifier is built by grounding the positive input of the operational amplifier and connecting resistors R_1 and R_2 , called the **feedback network**, between the inverting input and the signal source and amplifier output node, respectively, as in

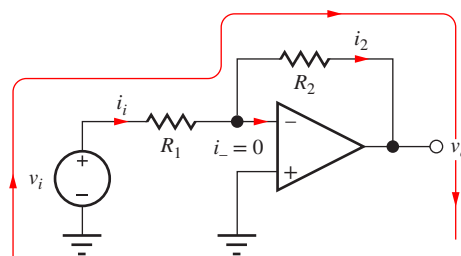


Figure 1.20 Inverting amplifier using op amp.

Fig. 1.20. Note that the ideal op amp is represented by a triangular amplifier symbol without a gain A indicated.

Our goal is to determine the voltage gain A_v of the overall amplifier, and to find A_v , we must find a relationship between v_i and v_o . One approach is to write an equation for the single loop shown in Fig. 1.20:

$$v_i - i_i R_1 - i_2 R_2 - v_o = 0 \quad (1.32)$$

Now we need to express i_i and i_2 in terms of v_i and v_o . By applying KCL at the inverting input to the amplifier, we see that i_2 must equal i_i because Assumption 2 states that i_- must be zero:

$$i_i = i_2 \quad (1.33)$$

Current i_i can be written in terms of v_i as

$$i_i = \frac{v_i - v_-}{R_1} \quad (1.34)$$

where v_- is the voltage at the inverting input (negative input) of the op amp. But Assumption 1 states that the input voltage between the op amp terminals must be zero, so v_- must be zero because the positive input is grounded. Therefore

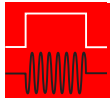
$$i_i = \frac{v_i}{R_1} \quad (1.35)$$

Combining Eqs. (1.32)–(1.35), the voltage gain is given by

$$A_v = \frac{v_o}{v_i} = -\frac{R_2}{R_1} \quad (1.36)$$

Referring to Eq. (1.36), we should note several things. The voltage gain is negative, indicative of an inverting amplifier with a 180° phase shift between its input and output signals. In addition, the magnitude of the gain can be greater than or equal to 1 if $R_2 \geq R_1$ (the most common case), but it can also be less than 1 for $R_2 < R_1$.

In the amplifier circuit in Fig. 1.20, the inverting-input terminal of the operational amplifier is at ground potential, 0 V, and is referred to as a **virtual ground**. The ideal operational amplifier adjusts its output to whatever voltage is necessary to force v_- to be zero. (In practical circuits, the output levels will be limited by the power supply voltages.)



VIRTUAL GROUND IN OP AMP CIRCUITS

Although the inverting input represents a virtual ground, it is *not* connected directly to ground (there is no direct dc path for current to reach ground). Shorting this terminal to ground for analysis purposes is a common mistake that must be avoided.

Suppose $R_2 = 100\text{ k}\Omega$. What value of R_1 gives a gain of -5 ?
 20 k Ω

ELECTRONICS IN ACTION

A Familiar Electronic System—The Cellular Phone

A cellular phone contains many radios operating over a multitude of frequencies including those of Bluetooth, cellular, GPS, ISM, and Wifi, to name a few (see Table 1.2). In addition, a “world” phone must operate on cellular frequency allocations that are different in North America, Europe, and Asia, for example.

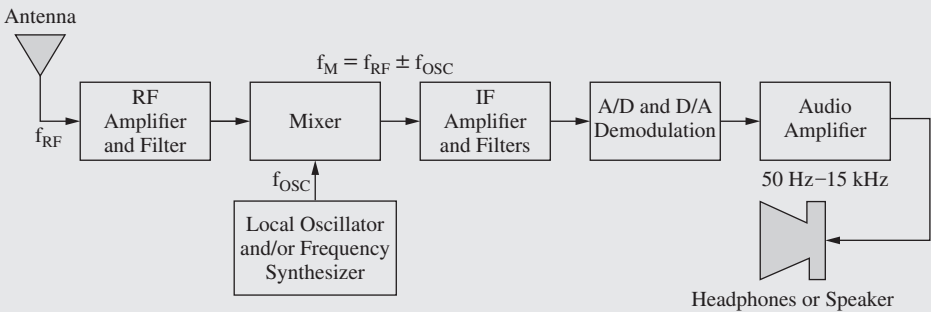
The accompanying block diagram represents a typical mixed-signal (analog and digital) radio-frequency (RF) receiver that uses a number of amplifiers as well as digital technology. The signal from the antenna can be very small, often in the microvolt range. The signal’s amplitude and power level are increased sequentially by the RF, intermediate frequency (IF), and audio amplifiers. The power available from the antenna may amount to only picowatts, whereas at the output, the amplifier may be providing a few tens of milliwatts to a headset or pair of earpods, or a high-power audio system could be delivering a 100-W or larger audio signal to its speaker system.

The local oscillator and frequency synthesizer tune the receiver to the desired frequency or channel and represent special applications of amplifiers and digital hardware. The mixer circuit actually changes the frequency of the incoming signal ($f_M = f_{RF} \pm f_{OSC}$) and is an application of a multiplier whose design draws heavily on linear amplifier and/or digital multiplexing circuit concepts. Finally, the demodulator/detector may be implemented using a combination of analog and digital circuitry employing A/D and D/A converters.

Chapters throughout this text provide in-depth exploration of the design techniques used in linear amplifiers and oscillators as well as the foundation needed to understand more complex circuits such as mixers, modulators, detectors, and A/D and D/A converters.

Examples of Several Radios in a Cellular Phone

TYPE	CAPABILITY
Bluetooth	Receive Rx/Transmit Tx
Cellular	Rx/Tx
GPS	Rx
Wifi (ISM)	Rx/Tx



Block diagram for a basic radio receiver.

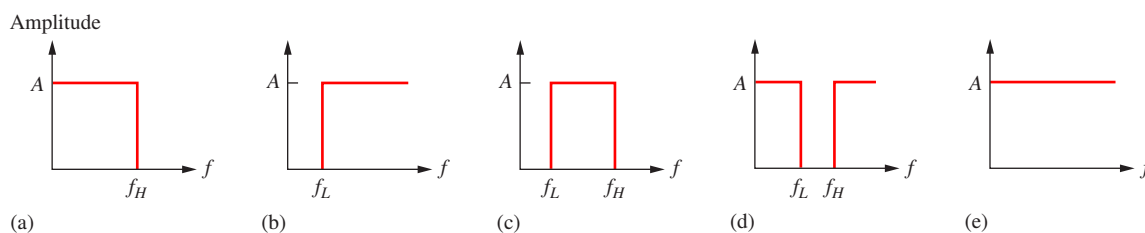


Figure 1.21 Ideal amplifier frequency responses: (a) low-pass, (b) high-pass, (c) band-pass, (d) band-reject, and (e) all-pass characteristics.

1.7.2 AMPLIFIER FREQUENCY RESPONSE

In addition to modifying the voltage, current, and/or power level of a given signal, amplifiers are often designed to selectively process signals of different frequency ranges. Amplifiers are classified into a number of categories based on their frequency response; five possible categories are shown in Fig. 1.21. The **low-pass amplifier**, Fig. 1.21(a), passes all signals below some upper cutoff frequency f_H , whereas the **high-pass amplifier**, Fig. 1.21(b), amplifies all signals above the lower cutoff frequency f_L . The **band-pass amplifier** passes all signals between the two cutoff frequencies f_L and f_H , as in Fig. 1.21(c). The **band-reject amplifier** in Fig. 1.21(d) rejects all signals having frequencies lying between f_L and f_H . Finally, the **all-pass amplifier** in Fig. 1.21(e) amplifies signals at any frequency. The all-pass amplifier is actually used to tailor the phase of the signal rather than its amplitude. Circuits that are designed to amplify specific ranges of signal frequencies are often referred to as **filters**.

(a) The band-pass amplifier in Fig. 1.21(c) has $f_L = 1.5$ kHz, $f_H = 2.5$ kHz, and $A = 10$. If the input voltage is given by

$$v_i = [0.5 \sin(2000\pi t) + \sin(4000\pi t) + 1.5 \sin(6000\pi t)] \text{ V}$$

what is the output voltage of the amplifier? (b) Suppose the same input signal is applied to the low-pass amplifier in Fig. 1.21(a), which has $A = 5$ and $f_H = 1.75$ kHz. What is the output voltage?

$$10.0 \sin 4000\pi t \text{ V}; 2.50 \sin 2000\pi t \text{ V}$$

1.8 ELEMENT VARIATIONS IN CIRCUIT DESIGN

Whether a circuit is built in discrete form or fabricated as an integrated circuit, the passive components and semiconductor device parameters will all have **tolerances** associated with their values. Discrete resistors can be purchased with a number of different tolerances including ± 10 percent, ± 5 percent, ± 1 percent, or better, whereas resistors in ICs can exhibit wide variations (± 30 percent). Capacitors often exhibit asymmetrical tolerance specifications such as $+20$ percent/ -50 percent, and power supply voltage tolerances are often specified in the range of 1 to 10 percent. For the semiconductor devices that we shall study in Chapters 3–5, device parameters may vary by 30 percent or more.

In addition to this initial value uncertainty due to tolerances, the values of the circuit components and parameters will vary with temperature and circuit age. It is important to understand the effect of these element changes on our circuits and to be able to design circuits that will continue to operate correctly in the face of such element variations. We will explore two analysis approaches, worst-case analysis and Monte Carlo analysis, that can help quantify the effects of tolerances on circuit performance.

1.8.1 MATHEMATICAL MODELING OF TOLERANCES

A mathematical model for symmetrical parameter variations is

$$P_{\text{nom}}(1 - \epsilon) \leq P \leq P_{\text{nom}}(1 + \epsilon) \quad (1.37)$$

in which P_{nom} is the nominal specification for the parameter such as the resistor value or independent source value, and ϵ is the fractional tolerance for the component. For example, a resistor R with **nominal value** of 10 k Ω and a 5 percent tolerance could exhibit a resistance anywhere in the following range:

$$10,000 \Omega(1 - 0.05) \leq R \leq 10,000 \Omega(1 + 0.05)$$

or

$$9500 \Omega \leq R \leq 10,500 \Omega$$

A 39-k Ω resistor has a 10 percent tolerance. What is the range of resistor values corresponding to this resistor? Repeat for a 3.6-k Ω resistor with a 1 percent tolerance.

$$35.1 \leq R \leq 42.9 \text{ k}\Omega; 3.56 \leq R \leq 3.64 \text{ k}\Omega$$

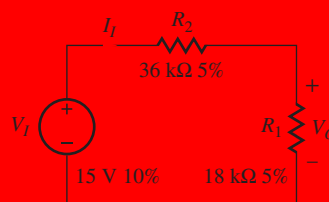
1.8.2 WORST-CASE ANALYSIS

Worst-case analysis is often used to ensure that a design will function under a given set of component variations. Worst-case analysis is performed by choosing values of the various components that make a desired variable (such as voltage, current, power, gain, or bandwidth) as large and as small as possible. These two limits are usually found by analyzing a circuit with the values of the various circuit elements pushed to their extremes. Although a worst-case design is often too conservative and represents “overdesign,” it is important to understand the technique and its limitations. An easy way to explore worst-case analysis is with an example.

WORST-CASE ANALYSIS

Here we apply worst-case analysis to a simple voltage divider circuit.

Find the nominal and worst-case values (highest and lowest) of output voltage V_O and source current I_I for the voltage divider circuit of Fig. 1.22.



Resistor voltage divider circuit with tolerances.

Known Information and Given Data: We have been given the voltage divider circuit in Fig. 1.22; the 15-V source V_I has a 10 percent tolerance; resistor R_1 has a nominal value of 18 k Ω with a 5 percent tolerance; resistor R_2 has a nominal value of 36 k Ω with a 5 percent tolerance. Expressions for V_O and I_I are

$$V_O = V_I \frac{R_1}{R_1 + R_2} \quad \text{and} \quad I_I = \frac{V_I}{R_1 + R_2} \quad (1.38)$$

Unknowns: V_O^{nom} , V_O^{max} , V_O^{min} , I_I^{nom} , I_I^{max} , I_I^{min}

Approach: Find the nominal values of V_O and I_I with all circuit elements set to their nominal (ideal) values. Find the worst-case values by selecting the individual voltage and resistance values that force V_O and I_I to their extremes. Note that the values selected for the various circuit elements to produce V_O^{max} will most likely differ from those that produce I_I^{max} , and so on.

Assumptions: None.

Analysis:

(a) *Nominal Values*

The nominal value of voltage V_O is found using the nominal values for all the parameters:

$$V_O^{\text{nom}} = V_I^{\text{nom}} \frac{R_1^{\text{nom}}}{R_1^{\text{nom}} + R_2^{\text{nom}}} = 15 \text{ V} \frac{18 \text{ k}\Omega}{18 \text{ k}\Omega + 36 \text{ k}\Omega} = 5 \text{ V} \quad (1.39)$$

Similarly, the nominal value of source current I_I is

$$I_I^{\text{nom}} = \frac{V_S^{\text{nom}}}{R_1^{\text{nom}} + R_2^{\text{nom}}} = \frac{15 \text{ V}}{18 \text{ k}\Omega + 36 \text{ k}\Omega} = 278 \text{ }\mu\text{A} \quad (1.40)$$

(b) *Worst-Case Limits*

Now let us find the **worst-case values (the largest and smallest possible values)** of voltage V_O and current I_I that can occur for the given set of element tolerances. First, the values of the components will be selected to make V_O as large as possible. However, it may not always be obvious at first to which extreme to adjust the individual component values. Rewriting Eq. (1.38) for voltage V_O will help:

$$V_O = V_I \frac{R_1}{R_1 + R_2} = \frac{V_I}{1 + R_2/R_1} \quad (1.41)$$

In order to make V_O as large as possible, the numerator of Eq. (1.41) should be large and the denominator small. Therefore, V_I and R_1 should be chosen to be as large as possible and R_2 as small as possible. Conversely, in order to make V_O as small as possible, V_I and R_1 must be small and R_2 must be large. Using this approach, the maximum and minimum values of V_O are

$$V_O^{\text{max}} = \frac{15 \text{ V}(1.1)}{1 + \frac{36 \text{ k}\Omega(0.95)}{18 \text{ k}\Omega(1.05)}} = 5.87 \text{ V} \quad \text{and} \quad V_O^{\text{min}} = \frac{15 \text{ V}(0.9)}{1 + \frac{36 \text{ k}\Omega(1.05)}{18 \text{ k}\Omega(0.95)}} = 4.20 \text{ V} \quad (1.42)$$

The maximum value of V_O is 17 percent greater than the nominal value of 5 V, and the minimum value is 16 percent below the nominal value.

The worst-case values of I_I are found in a similar manner but require different choices for the values of the resistors:

$$I_I^{\text{max}} = \frac{V_I^{\text{max}}}{R_1^{\text{min}} + R_2^{\text{min}}} = \frac{15 \text{ V}(1.1)}{18 \text{ k}\Omega(0.95) + 36 \text{ k}\Omega(0.95)} = 322 \text{ }\mu\text{A}$$

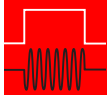
$$I_I^{\text{min}} = \frac{V_I^{\text{min}}}{R_1^{\text{max}} + R_2^{\text{max}}} = \frac{15 \text{ V}(0.9)}{18 \text{ k}\Omega(1.05) + 36 \text{ k}\Omega(1.05)} = 238 \text{ }\mu\text{A} \quad (1.43)$$

The maximum of I_I is 16 percent greater than the nominal value, and the minimum value is 14 percent less than nominal.

Check of Results: The nominal and worst-case values have been determined and range 14 to 17 percent above and below the nominal values. We have three circuit elements that are varying, and the sum of the three tolerances is 20 percent. Our worst-case values differ from the nominal case by somewhat less than this amount, so the results appear reasonable.

Find the nominal and worst-case values of the power delivered by source V_I in Fig. 1.22.

4.17 mW, 3.21 mW, 5.31 mW



BE WARY OF WORST-CASE DESIGN

In a real circuit, the parameters will be randomly distributed between the limits, and it is unlikely that the various components will all reach their extremes at the same time. Thus the worst-case analysis technique will overestimate (often badly) the extremes of circuit behavior, and a design based on worst-case analysis usually represents an unnecessary overdesign that is more costly than necessary to achieve the specifications with satisfactory yield. A better, although more complex, approach is to attack the problem statistically using Monte Carlo analysis. It is also difficult to see how to adjust values to achieve the worst-case situation in complex circuits. However, if every circuit must work no matter what, worst-case analysis may be appropriate.

1.8.3 MONTE CARLO ANALYSIS

Monte Carlo analysis uses randomly selected versions of a given circuit to predict its behavior from a statistical basis. For Monte Carlo analysis, a value for each of the elements in the circuit is selected at random from the possible distributions of parameters, and the circuit is then analyzed using the randomly selected element values. Many such randomly selected realizations (“cases” or “instances”) of the circuit are generated, and the statistical behavior of the circuit is built up from analysis of the many test cases. Obviously, this is a good use of the computer. Before proceeding, we need to refresh our memory concerning a few results from probability and random variables.

Uniformly Distributed Parameters

In this section, the variable parameters will be assumed to be uniformly distributed between the two extremes. In other words, the probability that any given value of the parameter will occur is the same. In fact, when the parameter tolerance expression in Eq. (1.37) was first encountered, most of us probably visualized it in terms of a uniform distribution as depicted by the probability density function $p(r)$ for a uniformly distributed resistor r represented graphically in Fig. 1.23(a). The probability that a resistor value lies between r and $(r + dr)$ is equal to $p(r) dr$. The total probability P must equal unity, so

$$P = \int_{-\infty}^{+\infty} p(r) dr = 1 \quad (1.44)$$

Using this equation with the uniform probability density of Fig. 1.23(a) yields $p(r) = \frac{1}{2r_{\text{nom}}}$ as indicated in the figure.