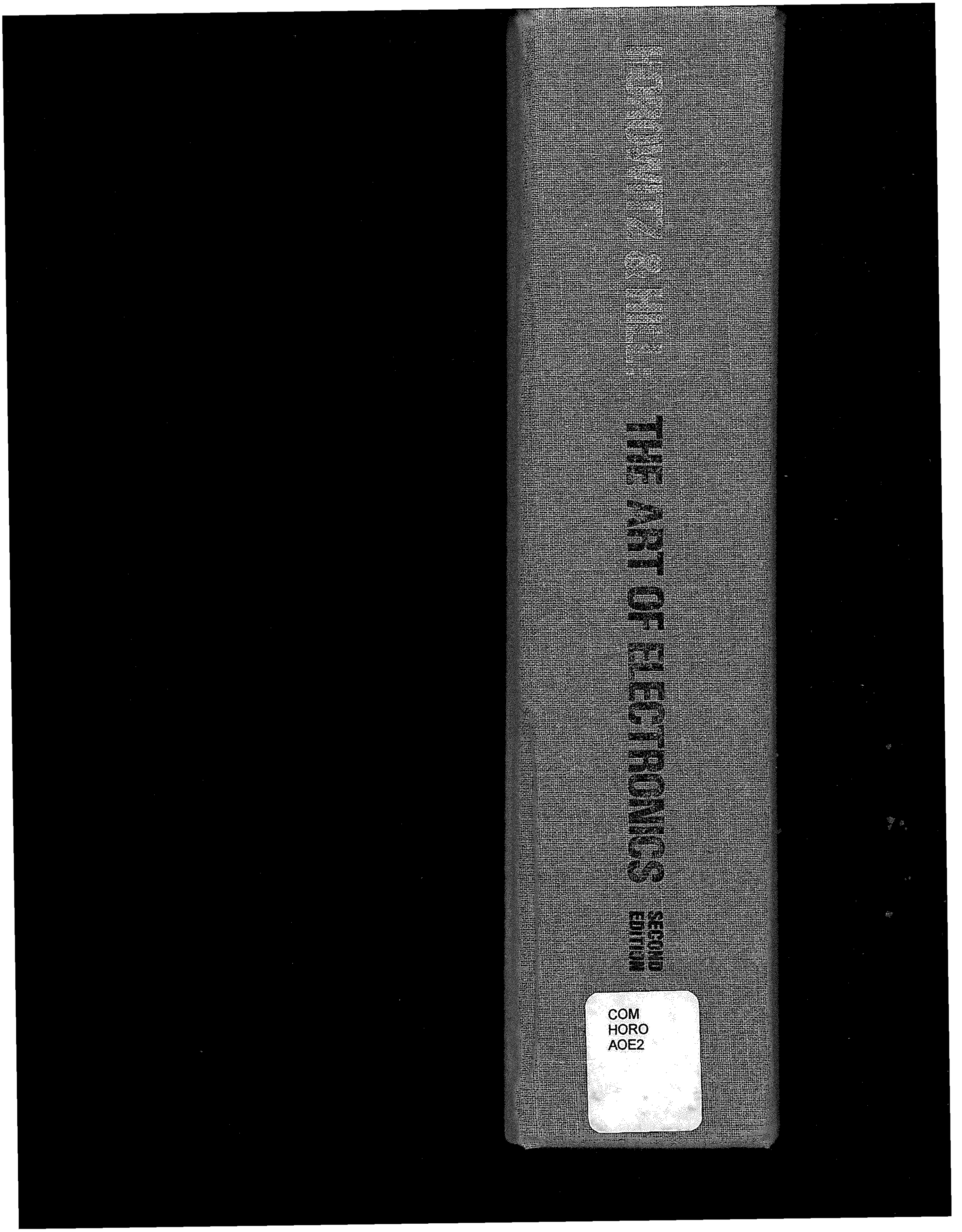
THE ART OF ELECTRONICS
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THE ART OF ELECTRONICS

Second Edition

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THE ART OF ELECTRONICS

Second Edition

Paul Horowitz HARVARD UNIVERSITY

Winfield Hill ROWLAND INSTITUTE FOR SCIENCE, CAMBRIDGE, MASSACHUSETTS

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PREFACE

Electronics, perhaps more than any other field of technology, has enjoyed an explosive development in the last four decades. Thus it was with some trepidation that we attempted, in 1980, to bring out a definitive volume teaching the art of the subject. By "art" we meant the kind of mastery that comes from an intimate familiarity with real circuits, actual devices, and the like, rather than the more abstract approach often favored in textbooks on electronics. Of course, in a rapidly evolving field, such a nuts-and-bolts approach has its hazards – most notably a frighteningly quick obsolescence.

The pace of electronics technology did not disappoint us! Hardly was the ink dry on the first edition before we felt foolish reading our words about "the classic [2Kbyte] 2716 EPROM ... with a price tag of about $25." They're so classic you can't even get them anymore, having been replaced by EPROMs 64 times as large, and costing less than half the price! Thus a major element of this revision responds to improved devices and methods – completely rewritten chapters on microcomputers and microprocessors (using the IBM PC and the 68008) and substantially revised chapters on digital electronics (including PLDs, and the new HC and AC logic families), on op-amps and precision design (reflecting the availability of excellent FET-input op-amps), and on construction techniques (including CAD/CAM). Every table has been revised, some substantially; for example, in Table 4.1 (operational amplifiers) only 65% of the original 120 entries survived, with 135 new op-amps added.

We have used this opportunity to respond to readers' suggestions and to our own experiences using and teaching from the first edition. Thus we have rewritten the chapter on FETs (it was too complicated) and repositioned it before the chapter on op-amps (which are increasingly of FET construction). We have added a new chapter on low-power and micropower design (both analog and digital), a field both important and neglected. Most of the remaining chapters have been extensively revised. We have added many new tables, including A/D and D/A converters, digital logic components, and low-power devices, and throughout the book we have expanded the number of figures. The book now contains 78 tables (available separately as *The Horowitz and Hill Component Selection Tables*) and over 1000 figures.

Throughout the revision we have strived to retain the feeling of informality and easy access that made the first edition so successful and popular, both as reference and text. We are aware of the difficulty students often experience when approaching electronics for the first time: The field is densely interwoven, and there is no path of learning that takes you, by logical steps, from neophyte to broadly competent designer. Thus we have added extensive cross-referencing throughout the text; in addition, we have expanded the separate *Laboratory Manual* into a *Student Manual* (*Student Manual for The Art of Electronics*, by Thomas C. Hayes and Paul Horowitz), complete with additional worked examples of circuit designs, explanatory material, reading assignments, laboratory exercises, and solutions to selected problems. By offering a student supplement, we have been able to keep this volume concise and rich with detail, as requested by our many readers who use the volume primarily as a reference work.

We hope this new edition responds to all our readers' needs – both students and practicing engineers. We welcome suggestions and corrections, which should be addressed directly to Paul Horowitz, Physics Department, Harvard University, Cambridge, MA 02138.

In preparing this new edition, we are appreciative of the help we received from Mike Aronson and Brian Matthews (AOX, Inc.), John Greene (University of Cape Town), Jeremy Avigad and Tom Hayes (Harvard University), Peter Horowitz (EVI, Inc.), Don Stern, and Owen Walker. We thank Jim Mobley for his excellent copyediting, Sophia Prybylski and David Tranah of Cambridge University Press for their encouragement and professional dedication, and the never-sleeping typesetters at Rosenlaui Publishing Services, Inc. for their masterful composition in TEX.

Finally, in the spirit of modern jurisprudence, we remind you to read the legal notice here appended.

Paul Horowitz
Winfield Hill

March 1989

LEGAL NOTICE

In this book we have attempted to teach the techniques of electronic design, using circuit examples and data that we believe to be accurate. However, the examples, data, and other information are intended solely as teaching aids and should not be used in any particular application without independent testing and verification by the person making the application. Independent testing and verification are especially important in any application in which incorrect functioning could result in personal injury or damage to property.

For these reasons, we make no warranties, express or implied, that the examples, data, or other information in this volume are free of error, that they are consistent with industry standards, or that they will meet the requirements for any particular application. THE AUTHORS AND PUBLISHER EXPRESSLY DISCLAIM THE IMPLIED WARRANTIES OF MERCHANTABILITY AND OF FITNESS FOR ANY PARTICULAR PURPOSE, even if the authors have been advised of a particular purpose, and even if a particular purpose is indicated in the book. The authors and publisher also disclaim all liability for direct, indirect, incidental, or consequential damages that result from any use of the examples, data, or other information in this book.

PREFACE TO FIRST EDITION

This volume is intended as an electronic circuit design textbook and reference book; it begins at a level suitable for those with no previous exposure to electronics and carries the reader through to a reasonable degree of proficiency in electronic circuit design. We have used a straightforward approach to the essential ideas of circuit design, coupled with an in-depth selection of topics. We have attempted to combine the pragmatic approach of the practicing physicist with the quantitative approach of the engineer, who wants a thoroughly evaluated circuit design.

This book evolved from a set of notes written to accompany a one-semester course in laboratory electronics at Harvard. That course has a varied enrollment – undergraduates picking up skills for their eventual work in science or industry, graduate students with a field of research clearly in mind, and advanced graduate students and postdoctoral researchers who suddenly find themselves hampered by their inability to "do electronics."

It soon became clear that existing textbooks were inadequate for such a course. Although there are excellent treatments of each electronics specialty, written for the planned sequence of a four-year engineering curriculum or for the practicing engineer, those books that attempt to address the whole field of electronics seem to suffer from excessive detail (the handbook syndrome), from oversimplification (the cookbook syndrome), or from poor balance of material. Much of the favorite pedagogy of beginning textbooks is quite unnecessary and, in fact, is not used by practicing engineers, while useful circuitry and methods of analysis in daily use by circuit designers lie hidden in application notes, engineering journals, and hard-to-get data books. In other words, there is a tendency among textbook writers to represent the theory, rather that the art, of electronics.

We collaborated in writing this book with the specific intention of combining the discipline of a circuit design engineer with the perspective of a practicing experimental physicist and teacher of electronics. Thus, the treatment in this book reflects our philosophy that electronics, as currently practiced, is basically a simple art, a combination of some basic laws, rules of thumb, and a large bag of tricks. For these reasons we have omitted entirely the usual discussions of solid-state physics, the *h*-parameter model of transistors, and complicated network theory, and reduced to a bare minimum the mention of load lines and the *s*-plane. The treatment is largely nonmathematical, with strong encouragement of circuit brainstorming and mental (or, at most, back-of-the-envelope) calculation of circuit values and performance.

In addition to the subjects usually treated in electronics books, we have included the following:

- an easy-to-use transistor model
- extensive discussion of useful subcircuits, such as current sources and current mirrors
- single-supply op-amp design
- easy-to-understand discussions of topics on which practical design information is often difficult to find: op-amp frequency

compensation, low-noise circuits, phase-locked loops, and precision linear design
- simplified design of active filters, with tables and graphs
- a section on noise, shielding, and grounding
- a unique graphical method for streamlined low-noise amplifier analysis
- a chapter on voltage references and regulators, including constant current supplies
- a discussion of monostable multivibrators and their idiosyncrasies
- a collection of digital logic pathology, and what to do about it
- an extensive discussion of interfacing to logic, with emphasis on the new NMOS and PMOS LSI
- a detailed discussion of A/D and D/A conversion techniques
- a section on digital noise generation
- a discussion of minicomputers and interfacing to data buses, with an introduction to assembly language
- a chapter on microprocessors, with actual design examples and discussion – how to design them into instruments, and how to make them do what you want
- a chapter on construction techniques: prototyping, printed circuit boards, instrument design
- a simplified way to evaluate high-speed switching circuits
- a chapter on scientific measurement and data processing: what you can measure and how accurately, and what to do with the data
- bandwidth narrowing methods made clear: signal averaging, multichannel scaling, lock-in amplifiers, and pulse-height analysis
- amusing collections of "bad circuits," and collections of "circuit ideas"
- useful appendixes on how to draw schematic diagrams, IC generic types, *LC* filter design, resistor values, oscilloscopes, mathematics review, and others
- tables of diodes, transistors, FETs, op-amps, comparators, regulators, voltage references, microprocessors, and other devices, generally listing the characteristics of both the most popular and the best types

Throughout we have adopted a philosophy of naming names, often comparing the characteristics of competing devices for use in any circuit, and the advantages of alternative circuit configurations. Example circuits are drawn with real device types, not black boxes. The overall intent is to bring the reader to the point of understanding clearly the choices one makes in designing a circuit – how to choose circuit configurations, device types, and parts values. The use of largely nonmathematical circuit design techniques does not result in circuits that cut corners or compromise performance or reliability. On the contrary, such techniques enhance one's understanding of the real choices and compromises faced in engineering a circuit and represent the best approach to good circuit design.

This book can be used for a full-year electronic circuit design course at the college level, with only a minimum mathematical prerequisite, namely, some acquaintance with trigonometric and exponential functions, and preferably a bit of differential calculus. (A short review of complex numbers and derivatives is included as an appendix.) If the less essential sections are omitted, it can serve as the text for a one-semester course (as it does at Harvard).

A separately available laboratory manual, *Laboratory Manual for the Art of Electronics* (Horowitz and Robinson, 1981), contains twenty-three lab exercises, together with reading and problem assignments keyed to the text.

To assist the reader in navigation, we have designated with open boxes in the margin those sections within each chapter that we feel can be safely passed over in an abbreviated reading. For a one-semester course it would probably be wise to omit, in addition, the materials of Chapter 5 (first half), 7, 12, 13, 14, and possibly 15, as explained in the introductory paragraphs of those chapters.

We would like to thank our colleagues for their thoughtful comments and assistance in the preparation of the manuscript, particularly Mike Aronson, Howard Berg, Dennis Crouse, Carol Davis, David Griesinger, John Hagen, Tom Hayes, Peter Horowitz, Bob Kline, Costas Papaliolios, Jay Sage, and Bill Vetterling. We are indebted to Eric Hieber and Jim Mobley, and to Rhona Johnson and Ken Werner of Cambridge University Press, for their imaginative and highly professional work.

Paul Horowitz
Winfield Hill

April 1980

FOUNDATIONS

CHAPTER 1

INTRODUCTION

Developments in the field of electronics have constituted one of the great success stories of this century. Beginning with crude spark-gap transmitters and "cat's-whisker" detectors at the turn of the century, we have passed through a vacuum-tube era of considerable sophistication to a solid-state era in which the flood of stunning advances shows no signs of abating. Calculators, computers, and even talking machines with vocabularies of several hundred words are routinely manufactured on single chips of silicon as part of the technology of large-scale integration (LSI), and current developments in very large scale integration (VLSI) promise even more remarkable devices.

Perhaps as noteworthy is the pleasant trend toward increased performance per dollar. The cost of an electronic microcircuit routinely decreases to a fraction of its initial cost as the manufacturing process is perfected (see Fig. 8.87 for an example). In fact, it is often the case that the panel controls and cabinet hardware of an instrument cost more than the electronics inside.

On reading of these exciting new developments in electronics, you may get the impression that you should be able to construct powerful, elegant, yet inexpensive, little gadgets to do almost any conceivable task – all you need to know is how all these miracle devices work. If you've had that feeling, this book is for you. In it we have attempted to convey the excitement and know-how of the subject of electronics.

In this chapter we begin the study of the laws, rules of thumb, and tricks that constitute the art of electronics as we see it. It is necessary to begin at the beginning – with talk of voltage, current, power, and the components that make up electronic circuits. Because you can't touch, see, smell, or hear electricity, there will be a certain amount of abstraction (particularly in the first chapter), as well as some dependence on such visualizing instruments as oscilloscopes and voltmeters. In many ways the first chapter is also the most mathematical, in spite of our efforts to keep mathematics to a minimum in order to foster a good intuitive understanding of circuit design and behavior.

Once we have considered the foundations of electronics, we will quickly get into the "active" circuits (amplifiers, oscillators, logic circuits, etc.) that make electronics the exciting field it is. The reader with some background in electronics may wish to skip over this chapter, since it assumes no prior knowledge of electronics. Further generalizations at this time would be pointless, so let's just dive right in.

VOLTAGE, CURRENT, AND RESISTANCE

1.01 Voltage and current

There are two quantities that we like to keep track of in electronic circuits: voltage and current. These are usually changing with time; otherwise nothing interesting is happening.

Voltage (symbol: V, or sometimes E). The voltage between two points is the cost in energy (work done) required to move a unit of positive charge from the more negative point (lower potential) to the more positive point (higher potential). Equivalently, it is the energy released when a unit charge moves "downhill" from the higher potential to the lower. Voltage is also called *potential difference* or *electromotive force* (EMF). The unit of measure is the *volt*, with voltages usually expressed in volts (V), kilovolts ($1\text{kV} = 10^3\text{V}$), millivolts ($1\text{mV} = 10^{-3}\text{V}$), or microvolts ($1\mu\text{V} = 10^{-6}\text{V}$) (see the box on prefixes). A joule of work is needed to move a coulomb of charge through a potential difference of one volt. (The coulomb is the unit of electric charge, and it equals the charge of 6×10^{18} electrons, approximately.) For reasons that will become clear later, the opportunities to talk about nanovolts ($1\text{nV} = 10^{-9}\text{V}$) and megavolts ($1\text{MV} = 10^6\text{V}$) are rare.

Current (symbol: I). Current is the rate of flow of electric charge past a point. The unit of measure is the ampere, or amp, with currents usually expressed in amperes (A), milliamperes ($1\text{mA} = 10^{-3}\text{A}$), microamperes ($1\mu\text{A} = 10^{-6}\text{A}$), nanoamperes ($1\text{nA} = 10^{-9}\text{A}$), or occasionally picoamperes ($1\text{pA} = 10^{-12}\text{A}$). A current of one ampere equals a flow of one coulomb of charge per second. By convention, current in a circuit is considered to flow from a more positive point to a more negative point, even though the actual electron flow is in the opposite direction.

Important: Always refer to voltage *between* two points or *across* two points in a circuit. Always refer to current *through* a device or connection in a circuit.

To say something like "the voltage through a resistor ... " is nonsense, or worse. However, we do frequently speak of the voltage *at a point* in a circuit. This is always understood to mean voltage between that point and "ground," a common point in the circuit that everyone seems to know about. Soon you will, too.

We *generate* voltages by doing work on charges in devices such as batteries (electrochemical), generators (magnetic forces), solar cells (photovoltaic conversion of the energy of photons), etc. We *get* currents by placing voltages across things.

At this point you may well wonder how to "see" voltages and currents. The single most useful electronic instrument is the oscilloscope, which allows you to look at voltages (or occasionally currents) in a circuit as a function of time. We will deal with oscilloscopes, and also voltmeters, when we discuss signals shortly; for a preview, see the oscilloscope appendix (Appendix A) and the multimeter box later in this chapter.

In real circuits we connect things together with wires, metallic conductors, each of which has the same voltage on it everywhere (with respect to ground, say). (In the domain of high frequencies or low impedances, that isn't strictly true, and we will have more to say about this later. For now, it's a good approximation.) We mention this now so that you will realize that an actual circuit doesn't have to look like its schematic diagram, because wires can be rearranged.

Here are some simple rules about voltage and current:

1. The sum of the currents into a point in a circuit equals the sum of the currents out (conservation of charge). This is sometimes called Kirchhoff's current law. Engineers like to refer to such a point as a *node*. From this, we get the following: For a series circuit (a bunch of two-terminal things all connected end-to-end) the current is the same everywhere.

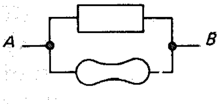


Figure 1.1

2. Things hooked in parallel (Fig. 1.1) have the same voltage across them. Restated, the sum of the "voltage drops" from A to B via one path through a circuit equals the sum by any other route equals the voltage between A and B. Sometimes this is stated as follows: The sum of the voltage drops around any closed circuit is zero. This is Kirchhoff's voltage law.

3. The power (work per unit time) consumed by a circuit device is

$$P = VI$$

This is simply (work/charge) × (charge/time). For V in volts and I in amps, P comes out in watts. Watts are joules per second (1W = 1J/s).

Power goes into heat (usually), or sometimes mechanical work (motors), radiated energy (lamps, transmitters), or stored energy (batteries, capacitors). Managing the heat load in a complicated system (e.g., a computer, in which many kilowatts of electrical energy are converted to heat, with the energetically insignificant by-product of a few pages of computational results) can be a crucial part of the system design.

PREFIXES

These prefixes are universally used to scale units in science and engineering.

Multiple	*Prefix*	*Symbol*
10^{12}	tera	T
10^{9}	giga	G
10^{6}	mega	M
10^{3}	kilo	k
10^{-3}	milli	m
10^{-6}	micro	μ
10^{-9}	nano	n
10^{-12}	pico	p
10^{-15}	femto	f

When abbreviating a unit with a prefix, the symbol for the unit follows the prefix without space. Be careful about upper-case and lower-case letters (especially m and M) in both prefix and unit: 1mW is a milliwatt, or one-thousandth of a watt; 1MHz is 1 million hertz. In general, units are spelled with lower-case letters, even when they are derived from proper names. The unit name is not capitalized when it is spelled out and used with a prefix, only when abbreviated. Thus: hertz and kilohertz, but Hz and kHz; watt, milliwatt, and megawatt, but W, mW, and MW.

Soon, when we deal with periodically varying voltages and currents, we will have to generalize the simple equation $P = VI$ to deal with *average* power, but it's correct as a statement of *instantaneous* power just as it stands.

Incidentally, don't call current "amperage"; that's strictly bush-league. The same caution will apply to the term "ohmage" when we get to resistance in the next section.

1.02 Relationship between voltage and current: resistors

This is a long and interesting story. It is the heart of electronics. Crudely speaking, the name of the game is to make and use gadgets that have interesting and useful I-versus-V characteristics. Resistors (I simply proportional to V), capacitors (I proportional to rate of change of V), diodes (I flows in only one direction), thermistors (temperature-dependent resistor), photoresistors (light-dependent resistor), strain gauges (strain-dependent resistor), etc., are examples. We will gradually get into some of these exotic devices; for now, we will start with the most mundane (and most widely used) circuit element, the resistor (Fig. 1.2).

Figure 1.2

Resistance and resistors

It is an interesting fact that the current through a metallic conductor (or other partially conducting material) is proportional to the voltage across it. (In the case of wire conductors used in circuits, we usually choose a thick enough gauge of wire so that these "voltage drops" will be negligible.) This is by no means a universal law for all objects. For instance, the current through a neon bulb is a highly nonlinear function of the applied voltage (it is zero up to a critical voltage, at which point it rises dramatically). The same goes for a variety of interesting special devices – diodes, transistors, light bulbs, etc. (If you are interested in understanding why metallic conductors behave this way, read sections 4.4–4.5 in the *Berkeley Physics Course*, Vol. II, see Bibliography). A resistor is made out of some conducting stuff (carbon, or a thin metal or carbon film, or wire of poor conductivity), with a wire coming out each end. It is characterized by its resistance:

$$R = V/I$$

R is in ohms for V in volts and I in amps. This is known as Ohm's law. Typical resistors of the most frequently used type (carbon composition) come in values from 1 ohm (1Ω) to about 22 megohms ($22\text{M}\Omega$). Resistors are also characterized by how

RESISTORS

Resistors are truly ubiquitous. There are almost as many types as there are applications. Resistors are used in amplifiers as loads for active devices, in bias networks, and as feedback elements. In combination with capacitors they establish time constants and act as filters. They are used to set operating currents and signal levels. Resistors are used in power circuits to reduce voltages by dissipating power, to measure currents, and to discharge capacitors after power is removed. They are used in precision circuits to establish currents, to provide accurate voltage ratios, and to set precise gain values. In logic circuits they act as bus and line terminators and as "pull-up" and "pull-down" resistors. In high-voltage circuits they are used to measure voltages and to equalize leakage currents among diodes or capacitors connected in series. In radiofrequency circuits they are even used as coil forms for inductors.

Resistors are available with resistances from 0.01 ohm through 10^{12} ohms, standard power ratings from 1/8 watt through 250 watts, and accuracies from 0.005% through 20%. Resistors can be made from carbon-composition moldings, from metal films, from wire wound on a form, or from semiconductor elements similar to field-effect transistors (FETs). But by far the most familiar resistor is the 1/4 or 1/2 watt carbon-composition resistor. These are available in a standard set of values ranging from 1 ohm to 100 megohms with twice as many values available for the 5% tolerance as for the 10% types (see Appendix C). We prefer the Allen-Bradley type AB (1/4 watt, 5%) resistor for general use because of its clear marking, secure lead seating, and stable properties.

Resistors are so easy to use that they're often taken for granted. They're not perfect, though, and it is worthwhile to look at some of their defects. The popular 5% composition type, in particular, although fine for nearly all noncritical circuit applications, is not stable enough for precision applications. You should know about its limitations so that you won't be surprised someday. Its principal defects are variations in resistance with temperature, voltage, time, and humidity. Other defects may relate to inductance (which may be serious at high frequencies), the development of thermal hot spots in power applications, or electrical noise generation in low-noise amplifiers. The following specifications are worst-case values; typically you'll do better, but don't count on it!

SPECIFICATIONS FOR ALLEN-BRADLEY AB SERIES TYPE CB

Standard tolerance is ±5% under nominal conditions. Maximum power for 70°C ambient temperature is 0.25 watt, which will raise the internal temperature to 150°C. The maximum applied voltage specification is $(0.25R)^{1/2}$ or 250 volts, whichever is less. They mean it! (See Fig. 6.53.) A single 5 second overvoltage to 400 volts can cause a permanent change in resistance by 2%.

	Resistance change		*Permanent?*
	(R = 1k)	(R = 10M)	
Soldering (350°C at 1/8 inch)	±2%	±2%	yes
Load cycling (500 ON/OFF cycles in 1000 hours)	+4%–6%	+4%–6%	yes
Vibration (20*g*) and shock (100*g*)	±2%	±2%	yes
Humidity (95% relative humidity at 40°C)	+6%	+10%	no
Voltage coefficient (10V change)	−0.15%	−0.3%	no
Temperature (25°C to −15°C)	+2.5%	+4.5%	no
Temperature (25°C to 85°C)	+3.3%	+5.9%	no

For applications that require any real accuracy or stability a 1% metal-film resistor (see Appendix D) should be used. They can be expected to have stability of better than 0.1% under normal conditions and better than 1% under worst-case treatment. Precision wire-wound resistors are available for the most demanding applications. For power dissipation above about 0.1 watt, a resistor of higher power rating should be used. Carbon-composition resistors are available with ratings up to 2 watts, and wire-wound power resistors are available for higher power. For demanding power applications, the conduction-cooled type of power resistor delivers better performance. These carefully designed resistors are available at 1% tolerance and can be operated at core temperatures up to 250°C with dependable long life. Allowable resistor power dissipation depends on air flow, thermal conduction via the resistor leads, and circuit density; thus, a resistor's power rating should be considered a rough guideline. Note also that resistor power ratings refer to *average* power dissipation and may be substantially exceeded for short periods of time (a few seconds or more, depending on the resistor's "thermal mass").

much power they can safely dissipate (the most commonly used ones are rated at 1/4 watt) and by other parameters such as tolerance (accuracy), temperature coefficient, noise, voltage coefficient (the extent to which R depends on applied V), stability with time, inductance, etc. See the box on resistors and Appendixes C and D for further details.

Roughly speaking, resistors are used to convert a voltage to a current, and vice versa. This may sound awfully trite, but you will soon see what we mean.

Resistors in series and parallel

From the definition of R, some simple results follow:

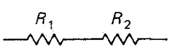


Figure 1.3

1. The resistance of two resistors in series (Fig. 1.3) is

$$R = R_1 + R_2$$

By putting resistors in series, you always get a *larger* resistor.

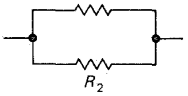


Figure 1.4

2. The resistance of two resistors in parallel (Fig. 1.4) is

$$R = \frac{R_1 R_2}{R_1 + R_2} \quad \text{or} \quad R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2}}$$

By putting resistors in parallel, you always get a *smaller* resistor. Resistance is measured in ohms (Ω), but in practice we frequently omit the Ω symbol when referring to resistors that are more than 1000Ω ($1\text{k}\Omega$). Thus, a $10\text{k}\Omega$ resistor is often referred to as a 10k resistor, and a $1\text{M}\Omega$ resistor as a 1M resistor (or 1 meg). On schematic diagrams the symbol Ω is often omitted altogether. If this bores you, please have patience – we'll soon get to numerous amusing applications.

EXERCISE 1.1
You have a 5k resistor and a 10k resistor. What is their combined resistance (a) in series and (b) in parallel?

EXERCISE 1.2
If you place a 1 ohm resistor across a 12 volt car battery, how much power will it dissipate?

EXERCISE 1.3
Prove the formulas for series and parallel resistors.

EXERCISE 1.4
Show that several resistors in parallel have resistance

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \cdots}$$

A trick for parallel resistors: Beginners tend to get carried away with complicated algebra in designing or trying to understand electronics. Now is the time to begin learning intuition and shortcuts.
Shortcut no. 1 A large resistor in series (parallel) with a small resistor has the resistance of the larger (smaller) one, roughly.
Shortcut no. 2 Suppose you want the resistance of 5k in parallel with 10k. If you think of the 5k as two 10k's in parallel, then the whole circuit is like three 10k's in parallel. Because the resistance of n equal resistors in parallel is $1/n$th the resistance of the individual resistors, the answer in this case is 10k/3, or 3.33k. This trick is handy because it allows you to analyze circuits quickly in your head, without distractions. We want to encourage mental designing, or at least "back of the envelope" designing, for idea brainstorming.

Some more home-grown philosophy: There is a tendency among beginners to want to compute resistor values and other circuit component values to many significant places, and the availability of inexpensive calculators has only made matters worse. There are two reasons you should try to avoid falling into this habit: (a) the components themselves are of finite precision (typical resistors are ±5%; the parameters that characterize transistors, say, frequently are known only to a factor of two); (b) one mark of a good circuit design is insensitivity of the finished circuit to precise values of the components (there are exceptions, of course). You'll also learn circuit intuition more quickly if you get into the habit of doing approximate calculations in your head, rather than watching meaningless numbers pop up on a calculator display.

In trying to develop intuition about resistance, some people find it helpful to think about *conductance*, $G = 1/R$. The current through a device of conductance G bridging a voltage V is then given by $I = GV$ (Ohm's law). A small resistance is a large conductance, with correspondingly large current under the influence of an applied voltage.

Viewed in this light, the formula for parallel resistors is obvious: When several resistors or conducting paths are connected across the same voltage, the total current is the sum of the individual currents. Therefore the net conductance is simply the sum of the individual conductances, $G = G_1 + G_2 + G_3 + \cdots$, which is the same as the formula for parallel resistors derived earlier.

Engineers are fond of defining reciprocal units, and they have designated the unit of conductance the siemens (S $= 1/\Omega$), also known as the mho (that's ohm spelled backward, given the symbol ℧). Although the concept of conductance is helpful in developing intuition, it is not used widely; most people prefer to talk about resistance instead.

Power in resistors

The power dissipated by a resistor (or any other device) is $P = IV$. Using Ohm's law, you can get the equivalent forms $P = I^2R$ and $P = V^2/R$.

EXERCISE 1.5

Show that it is not possible to exceed the power rating of a 1/4 watt resistor of resistance greater than 1k, no matter how you connect it, in a circuit operating from a 15 volt battery.

EXERCISE 1.6

Optional exercise: New York City requires about 10^{10} watts of electrical power, at 110 volts (this is plausible: 10 million people averaging 1 kilowatt each). A heavy power cable might be an inch in diameter. Let's calculate what will happen if we try to supply the power through a cable 1 foot in diameter made of pure copper. Its resistance is 0.05$\mu\Omega$ (5×10^{-8} ohms) per foot. Calculate (a) the power lost per foot from "I^2R losses," (b) the length of cable over which you will lose all 10^{10} watts, and (c) how hot the cable will get, if you know the physics involved ($\sigma = 6 \times 10^{-12}$W/°K^{4}cm^2).

If you have done your computations correctly, the result should seem preposterous. What is the solution to this puzzle?

Input and output

Nearly all electronic circuits accept some sort of applied *input* (usually a voltage) and produce some sort of corresponding *output* (which again is often a voltage). For example, an audio amplifier might produce a (varying) output voltage that is 100 times as large as a (similarly varying) input voltage. When describing such an amplifier, we imagine measuring the output voltage for a given applied input voltage. Engineers speak of the *transfer function* **H**, the ratio of (measured) output divided by (applied) input; for the audio amplifier above, **H** is simply a constant (**H** = 100). We'll get to amplifiers soon enough, in the next chapter. However, with just resistors we can already look at a very important circuit fragment, the *voltage divider* (which you might call a "de-amplifier").

1.03 Voltage dividers

We now come to the subject of the voltage divider, one of the most widespread electronic circuit fragments. Show us any real-life circuit and we'll show you half a dozen voltage dividers. To put it very simply, a voltage divider is a circuit that, given a certain voltage input, produces a predictable fraction of the input voltage as the output voltage. The simplest voltage divider is shown in Figure 1.5.

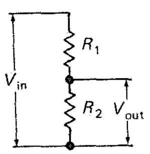


Figure 1.5. Voltage divider. An applied voltage V_{in} results in a (smaller) output voltage V_{out}.

What is V_{out}? Well, the current (same everywhere, assuming no "load" on the output) is

$$I = \frac{V_{\text{in}}}{R_1 + R_2}$$

(We've used the definition of resistance and the series law.) Then, for R_2,

$$V_{\text{out}} = IR_2 = \frac{R_2}{R_1 + R_2} V_{\text{in}}$$

Note that the output voltage is always less than (or equal to) the input voltage; that's why it's called a divider. You could get amplification (more output than input) if one of the resistances were negative. This isn't as crazy as it sounds; it is possible to make devices with negative "incremental" resistances (e.g., the tunnel diode) or even true negative resistances (e.g., the negative-impedance converter that we will talk about later in the book). However, these applications are rather specialized and need not concern you now.

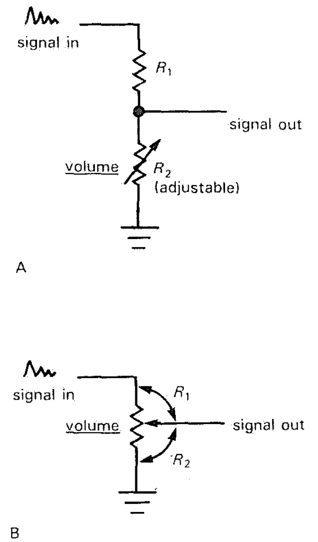


Figure 1.6. An adjustable voltage divider can be made from a fixed and variable resistor, or from a potentiometer.

Voltage dividers are often used in circuits to generate a particular voltage from a larger fixed (or varying) voltage. For instance, if V_{in} is a varying voltage and R_2 is an adjustable resistor (Fig. 1.6A), you have a "volume control"; more simply, the combination R_1R_2 can be made from a single variable resistor, or *potentiometer* (Fig. 1.6B). The humble voltage divider is even more useful, though, as a way of *thinking* about a circuit: the input voltage and upper resistance might represent the output of an amplifier, say, and the lower resistance might represent the input of the following stage. In this case the voltage-divider equation tells you how much signal gets to the input of that last stage. This will all become clearer after you know about a remarkable fact (Thévenin's theorem) that will be discussed later. First, though, a short aside on voltage sources and current sources.

1.04 Voltage and current sources

A perfect voltage source is a two-terminal *black box* that maintains a fixed voltage drop across its terminals, regardless of load resistance. For instance, this means that it must supply a current $I = V/R$ when a resistance R is attached to its terminals. A real voltage source can supply only a finite maximum current, and in addition it generally behaves like a perfect voltage source with a small resistance in series. Obviously, the smaller this series resistance, the better. For example, a standard 9 volt alkaline battery behaves like a perfect 9 volt voltage source in series with a 3 ohm resistor and can provide a maximum current (when shorted) of 3 amps (which, however, will kill the battery in a few minutes). A voltage source "likes" an open-circuit load and "hates" a short-circuit load, for obvious reasons. (The terms "open circuit" and "short circuit" mean the obvious: An open circuit has nothing connected to it, whereas a short circuit is a piece of wire bridging the output.) The symbols used to indicate a voltage source are shown in Figure 1.7.

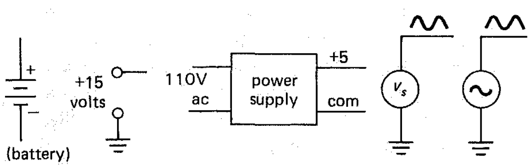


Figure 1.7. Voltage sources can be either steady (dc) or varying (ac).

A perfect current source is a two-terminal black box that maintains a constant current through the external circuit, regardless of load resistance or applied voltage. In order to do this it must be capable of supplying any necessary voltage across its terminals. Real current sources (a much-neglected subject in most textbooks) have a limit to the voltage they can provide (called the *output voltage compliance,* or just *compliance*), and in addition they do not provide absolutely constant output current. A current source "likes" a short-circuit load and "hates" an open-circuit load. The symbols used to indicate a current source are shown in Figure 1.8.

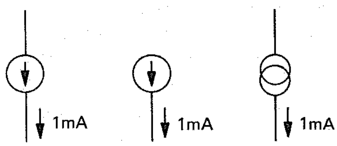


Figure 1.8. Current-source symbols.

A battery is a real-life approximation of a voltage source (there is no analog for a current source). A standard D-size flashlight cell, for instance, has a terminal voltage of 1.5 volts, an equivalent series resistance of about 1/4 ohm, and total energy capacity of about 10,000 watt-seconds (its characteristics gradually deteriorate with use; at the end of its life, the voltage may be about 1.0 volt, with an internal series resistance of several ohms). It is easy to construct voltage sources with far better characteristics, as you will learn when we come to the subject of feedback. Except in devices intended for portability, the use of batteries in electronic devices is rare. We will treat the interesting subject of low-power (battery-operated) design in Chapter 14.

MULTIMETERS

There are numerous instruments that let you measure voltages and currents in a circuit. The oscilloscope (see Appendix A) is the most versatile; it lets you "see" voltages versus time at one or more points in a circuit. Logic probes and logic analyzers are special-purpose instruments for troubleshooting digital circuits. The simple multimeter provides a good way to measure voltage, current, and resistance, often with good precision; however, it responds slowly, and thus it cannot replace the oscilloscope where changing voltages are of interest. Multimeters are of two varieties: those that indicate measurements on a conventional scale with a moving pointer, and those that use a digital display.

The standard VOM (volt-ohm-milliammeter) multimeter uses a meter movement that measures current (typically 50μA full scale). (See a less-design-oriented electronics book for pretty pictures of the innards of meter movements; for our purposes, it suffices to say that it uses coils and magnets.) To measure voltage, the VOM puts a resistor in series with the basic movement. For instance, one kind of VOM will generate a 1 volt (full-scale) range by putting a 20k resistor in series with the standard 50μA movement; higher voltage ranges use correspondingly larger resistors. Such a VOM is specified as 20,000 ohms/volt, meaning that it looks like a resistor whose value is 20k multiplied by the full-scale voltage of the particular range selected. Full scale on any voltage range is 1/20,000, or 50μA. It should be clear that one of these voltmeters disturbs a circuit less on a higher range, since it looks like a higher resistance (think of the voltmeter as the lower leg of a voltage divider, with the Thévenin resistance of the circuit you are measuring as the upper resistor). Ideally, a voltmeter should have infinite input resistance.

Nowadays there are various meters with some electronic amplification whose input resistance may be as large as 10^9 ohms. Most digital meters, and even a number of analog-reading meters that use FETs (field-effect transistors, see Chapter 3), are of this type. Warning: Sometimes the input resistance of FET-input meters is very high on the most sensitive ranges, dropping to a lower resistance for the higher ranges. For instance, an input resistance of 10^9 ohms on the 0.2 volt and 2 volt ranges, and 10^7 ohms on all higher ranges, is typical. Read the specifications carefully! For measurements on most transistor circuits, 20,000 ohms/volt is fine, and there will be little loading effect on the circuit by the meter. In any case, it is easy to calculate how serious the effect is by using the voltage-divider equation. Typically, multimeters provide voltage ranges from a volt (or less) to a kilovolt (or more), full scale.

A VOM can be used to measure current by simply using the bare meter movement (for our preceding example, this would give a range of 50μA full scale) or by shunting (paralleling) the movement with a small resistor. Because the meter movement itself requires a small voltage drop, typically 0.25 volt, to produce a full-scale deflection, the shunt is chosen by the meter manufacturer (all you do is set the range switch to the range you want) so that the full-scale current will produce that voltage drop through the parallel combination of the meter resistance and the shunt resistance. Ideally, a current-measuring meter should have zero resistance in order not to disturb the circuit under test, since it must be put in series with the circuit. In practice, you tolerate a few tenths of a volt drop (sometimes called "voltage burden") with both VOMs and digital multimeters. Typically, multimeters provide current ranges from 50μA (or less) to an amp (or more), full scale.

Multimeters also have one or more batteries in them to power the resistance measurement. By supplying a small current and measuring the voltage drop, they measure resistance, with several ranges to cover values from an ohm (or less) to 10 megohms (or more).

Important: Don't try to measure "the current of a voltage source," for instance by sticking the meter across the wall plug; the same applies for ohms. This is the leading cause of blown-out meters.

EXERCISE 1.7

What will a 20,000 ohms/volt meter read, on its 1 volt scale, when attached to a 1 volt source with an internal resistance of 10k? What will it read when attached to a 10k–10k voltage divider driven by a "stiff" (zero source resistance) 1 volt source?

EXERCISE 1.8

A 50μA meter movement has an internal resistance of 5k. What shunt resistance is needed to convert it to a 0–1 amp meter? What series resistance will convert it to a 0–10 volt meter?

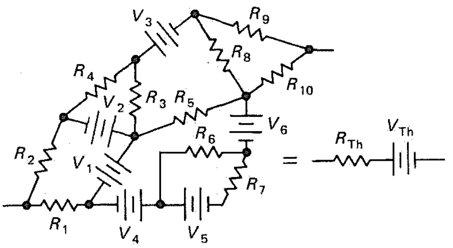


Figure 1.9

1.05 Thévenin's equivalent circuit

Thévenin's theorem states that any two-terminal network of resistors and voltage sources is equivalent to a single resistor R in series with a single voltage source V. This is remarkable. Any mess of batteries and resistors can be mimicked with one battery and one resistor (Fig. 1.9). (Incidentally, there's another theorem, Norton's theorem, that says you can do the same thing with a current source in parallel with a resistor.)

How do you figure out the Thévenin equivalent R_{Th} and V_{Th} for a given circuit? Easy! V_{Th} is the open-circuit voltage of the Thévenin equivalent circuit; so if the two circuits behave identically, it must also be the open-circuit voltage of the given circuit (which you get by calculation, if you know what the circuit is, or by measurement, if you don't). Then you find R_{Th} by noting that the short-circuit current of the equivalent circuit is $V_{\mathrm{Th}}/R_{\mathrm{Th}}$. In other words,

$$V_{\mathrm{Th}} = V \text{ (open circuit)}$$

$$R_{\mathrm{Th}} = \frac{V \text{ (open circuit)}}{I \text{ (short circuit)}}$$

Let's apply this method to the voltage divider, which must have a Thévenin equivalent:

1. The open-circuit voltage is

$$V = V_{\mathrm{in}}\frac{R_2}{R_1 + R_2}$$

2. The short-circuit current is

$$V_{\mathrm{in}}/R_1$$

So the Thévenin equivalent circuit is a voltage source

$$V_{\mathrm{Th}} = V_{\mathrm{in}}\frac{R_2}{R_1 + R_2}$$

in series with a resistor

$$R_{\mathrm{Th}} = \frac{R_1 R_2}{R_1 + R_2}$$

(It is not a coincidence that this happens to be the parallel resistance of R_1 and R_2. The reason will become clear later.)

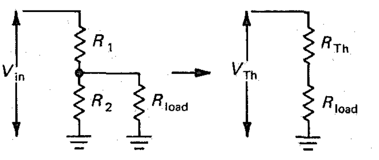


Figure 1.10

From this example it is easy to see that a voltage divider is not a very good battery, in the sense that its output voltage drops severely when a load is attached. As an example, consider Exercise 1.9. You now know everything you need to know to calculate exactly how much the output will drop for a given load resistance: Use the Thévenin equivalent circuit, attach a load, and calculate the new output, noting that the new circuit is nothing but a voltage divider (Fig. 1.10).

EXERCISE 1.9

For the circuit shown in Figure 1.10, with $V_{\mathrm{in}} =$ 30V and $R_1 = R_2 =$ 10k, find (a) the output voltage with no load attached (the open-circuit voltage); (b) the output voltage with a 10k load (treat as voltage divider, with R_2 and R_{load} combined into a single resistor); (c) the Thévenin equivalent circuit; (d) the same as in part b, but using the Thévenin equivalent circuit (again, you wind up with a voltage divider; the answer should agree with the result in part b); (e) the power dissipated in each of the resistors.

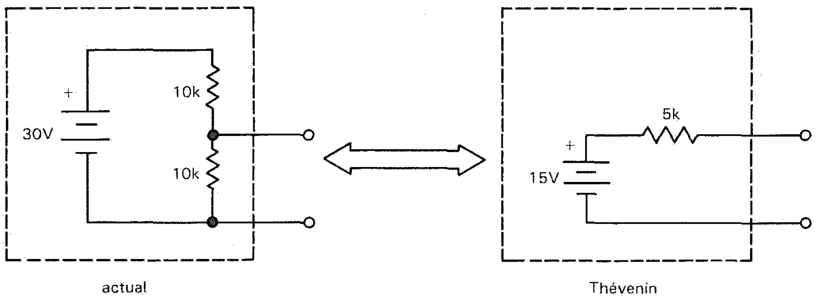


Figure 1.11

Equivalent source resistance and circuit loading

As you have just seen, a voltage divider powered from some fixed voltage is equivalent to some smaller voltage source in series with a resistor; for example, the output terminals of a 10k–10k voltage divider driven by a perfect 30 volt battery are precisely equivalent to a perfect 15 volt battery in series with a 5k resistor (Fig. 1.11). Attaching a load resistor causes the voltage divider's output to drop, owing to the finite *source resistance* (Thévenin equivalent resistance of the voltage divider output, viewed as a source of voltage). This is often undesirable. One solution to the problem of making a stiff voltage source ("stiff" is used in this context to describe something that doesn't bend under load) might be to use much smaller resistors in a voltage divider. Occasionally this brute-force approach is useful. However, it is usually best to construct a voltage source, or power supply, as it's commonly called, using active components like transistors or operational amplifiers, which we will treat in Chapters 2–4. In this way you can easily make a voltage source with internal (Thévenin equivalent) resistance measured in milliohms (thousandths of an ohm), without the large currents and dissipation of power characteristic of a low-resistance voltage divider delivering the same performance. In addition, with an active power supply it is easy to make the output voltage adjustable.

The concept of equivalent internal resistance applies to all sorts of sources, not just batteries and voltage dividers. Signal sources (e.g., oscillators, amplifiers, and sensing devices) all have an equivalent internal resistance. Attaching a load whose resistance is less than or even comparable to the internal resistance will reduce the output considerably. This undesirable reduction of the open-circuit voltage (or signal) by the load is called "circuit loading." Therefore, you should strive to make $R_{\text{load}} \gg R_{\text{internal}}$, because a high-resistance load has little attenuating effect on the source (Fig. 1.12). You will see numerous circuit examples in the chapters ahead. This high-resistance condition ideally

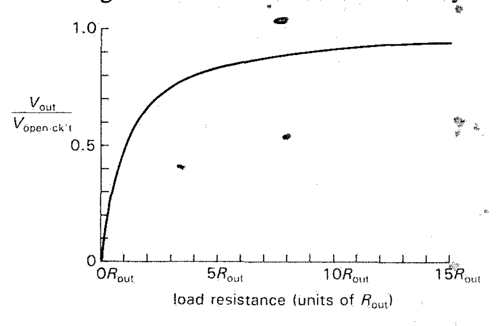


Figure 1.12. To avoid attenuating a signal source below its open-circuit voltage, keep the load resistance large compared with the output resistance.

characterizes measuring instruments such as voltmeters and oscilloscopes. (There are exceptions to this general principle; for example, we will talk about transmission lines and radiofrequency techniques, where you must "match impedances" in order to prevent the reflection and loss of power.)

A word on language: You frequently hear things like "the resistance looking into the voltage divider," or "the output sees a load of so-and-so many ohms," as if circuits had eyes. It's OK (in fact, it's a rather good way to keep straight which resistance you're talking about) to say what part of the circuit is doing the "looking."

Power transfer

Here is an interesting problem: What load resistance will result in maximum power being transferred to the load for a given source resistance? (The terms *source resistance, internal resistance,* and *Thévenin equivalent resistance* all mean the same thing.) It is easy to see that both $R_{\text{load}} = 0$ and $R_{\text{load}} = \infty$ result in zero power transferred, because $R_{\text{load}} = 0$ means that $V_{\text{load}} = 0$ and $I_{\text{load}} = V_{\text{source}}/R_{\text{source}}$, so that $P_{\text{load}} = V_{\text{load}}I_{\text{load}} = 0$. But $R_{\text{load}} = \infty$ means that $V_{\text{load}} = V_{\text{source}}$ and $I_{\text{load}} = 0$, so that $P_{\text{load}} = 0$. There has to be a maximum in between.

EXERCISE 1.10

Show that $R_{\text{load}} = R_{\text{source}}$ maximizes the power in the load for a given source resistance. Note: Skip this exercise if you don't know calculus, and take it on faith that the answer is true.

Lest this example leave the wrong impression, we would like to emphasize again that circuits are ordinarily designed so that the load resistance is much greater than the source resistance of the signal that drives the load.

1.06 Small-signal resistance

We often deal with electronic devices for which I is not proportional to V; in such cases there's not much point in talking about resistance, since the ratio V/I will depend on V, rather than being a nice constant, independent of V. For these devices it is useful to know the slope of the V–I curve, in other words, the ratio of a small change in applied voltage to the resulting change in current through the device, $\Delta V/\Delta I$ (or dV/dI). This quantity has the units of resistance (ohms) and substitutes for resistance in many calculations. It is called the small-signal resistance, incremental resistance, or dynamic resistance.

Zener diodes

As an example, consider the *zener diode,* which has the V–I curve shown in Figure 1.13. Zeners are used to create a constant voltage inside a circuit somewhere, simply by providing them with a (roughly constant) current derived from a higher voltage within the circuit. For example, the zener diode in Figure 1.13 will convert an applied current in the range shown to a corresponding (but narrower) range of voltages. It is important to know how the resulting zener voltage will change with applied current; this is a measure of its "regulation" against changes in the driving current provided to it. Included in the specifications of a zener will be its dynamic resistance, given at a certain current. (Useful fact: the dynamic resistance of a zener diode varies roughly in inverse proportion to current.) For example, a zener might have a dynamic resistance of 10 ohms at 10mA, at its zener voltage of 5 volts. Using the definition of dynamic resistance, we find that a 10% change in applied current will therefore result in a change in voltage of

$$\Delta V = R_{\text{dyn}}\Delta I = 10 \times 0.1 \times 0.01 = 10\text{mV}$$

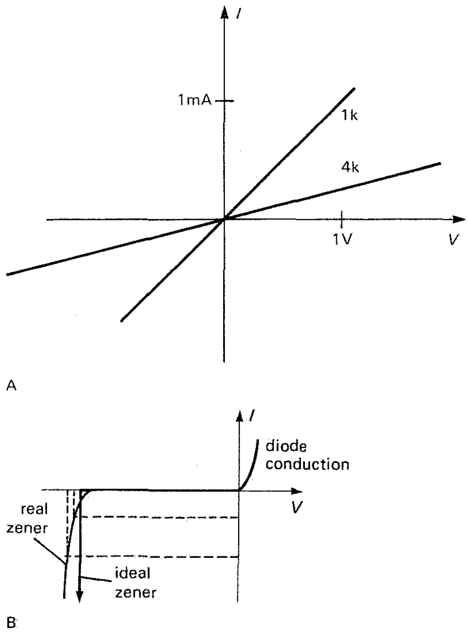


Figure 1.13. *V–I* curves.
A. Resistor (linear).
B. Zener diode (nonlinear).

or

$$\Delta V/V = 0.002 = 0.2\%$$

thus demonstrating good voltage-regulating ability. In this sort of application you frequently get the zener current through a resistor from a higher voltage available somewhere in the circuit, as in Figure 1.14.

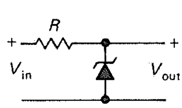


Figure 1.14. Zener regulator.

Then,

$$I = \frac{V_{in} - V_{out}}{R}$$

and

$$\Delta I = \frac{\Delta V_{in} - \Delta V_{out}}{R}$$

so

$$\Delta V_{out} = R_{dyn}\Delta I = \frac{R_{dyn}}{R}(\Delta V_{in} - \Delta V_{out})$$

and finally

$$\Delta V_{out} = \frac{R_{dyn}}{R + R_{dyn}}\Delta V_{in}$$

Thus, for *changes* in voltage, the circuit behaves like a voltage divider, with the zener replaced by a resistor equal to its dynamic resistance at the operating current. This is the utility of incremental resistance. For instance, suppose in the preceding circuit we have an input voltage ranging between 15 and 20 volts and use a 1N4733 (5.1V 1W zener diode) in order to generate a stable 5.1 volt power supply. We choose $R = 300$ ohms, for a maximum zener current of 50mA: (20 − 5.1)/300. We can now estimate the output voltage regulation (variation in output voltage), knowing that this particular zener has a specified maximum dynamic impedance of 7.0 ohms at 50mA. The zener current varies from 50mA to 33mA over the input voltage range; this 17mA change in current then produces a voltage change at the output of $\Delta V = R_{dyn}\Delta I$, or 0.12 volt. You will see more of zeners in Sections 2.04 and 6.14.

In real life, a zener will provide better regulation if driven by a current source, which has, by definition, $R_{incr} = \infty$ (same current regardless of voltage). But current sources are more complex, and therefore in practice we often resort to the humble resistor.

Tunnel diodes

Another interesting application of incremental resistance is the *tunnel diode*, sometimes called the Esaki diode. Its *V–I* curve is shown in Figure 1.15. In the region from A to B it has *negative* incremental resistance. This has a remarkable consequence: A voltage *divider* made with a resistor and

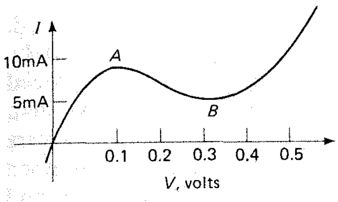


Figure 1.15

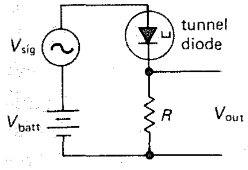


Figure 1.16

a tunnel diode can actually be an *amplifier* (Fig. 1.16). For a wiggly voltage v_{sig}, the voltage divider equation gives us

$$v_{\text{out}} = \frac{R}{R + r_t} v_{\text{sig}}$$

where r_t is the incremental resistance of the tunnel diode at the operating current, and the lower-case symbol v_{sig} stands for a small-signal variation, which we have been calling ΔV_{sig} up to now (we will adopt this widely used convention from now on). The tunnel diode has $r_{t(\text{incr})} < 0$. That is,

$$\Delta V/\Delta I \quad (\text{or } v/i) \quad < \quad 0$$

from A to B on the characteristic curve. If $r_{t(\text{incr})} \approx R$, the denominator is nearly zero, and the circuit amplifies. V_{batt} provides the steady current, or *bias*, to bring the operating point into the region of negative resistance. (Of course, it is always necessary to have a source of power in any device that amplifies.)

A postmortem on these fascinating devices: When tunnel diodes first appeared, late in the 1950s, they were hailed as the solution to a great variety of circuit problems. Because they were fast, they were supposed to revolutionize computers, for instance. Unfortunately, they are difficult devices to use; this fact, combined with stunning improvements in transistors, has made tunnel diodes almost obsolete.

The subject of negative resistance will come up again later, in connection with active filters. There you will see a circuit called a negative-impedance converter that can produce (among other things) a pure negative resistance (not just incremental). It is made with an operational amplifier and has very useful properties.

SIGNALS

A later section in this chapter will deal with capacitors, devices whose properties depend on the way the voltages and currents in a circuit are *changing*. Our analysis of dc circuits so far (Ohm's law, Thévenin equivalent circuits, etc.) still holds, even if the voltages and currents are changing in time. But for a proper understanding of alternating-current (ac) circuits, it is useful to have in mind certain common types of *signals*, voltages that change in time in a particular way.

1.07 Sinusoidal signals

Sinusoidal signals are the most popular signals around; they're what you get out of the wall plug. If someone says something like "take a 10 microvolt signal at 1 megahertz," he means a sine wave. Mathematically, what you have is a voltage described by

$$V = A \sin 2\pi f t$$

where A is called the amplitude, and f is the frequency in cycles per second, or hertz. A sine wave looks like the wave shown in Figure 1.17. Sometimes it is important to know the value of the signal at some arbitrary time $t = 0$, in which case you may see a *phase* ϕ in the expression:

$$V = A \sin(2\pi f t + \phi)$$

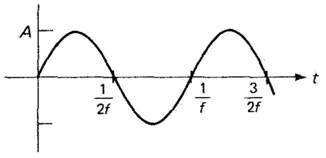


Figure 1.17. Sine wave of amplitude A and frequency f.

The other variation on this simple theme is the use of *angular frequency*, which looks like this:

$V = A \sin \omega t$

Here, ω is the angular frequency in radians per second. Just remember the important relation $\omega = 2\pi f$ and you won't go wrong.

The great merit of sine waves (and the cause of their perennial popularity) is the fact that they are the solutions to certain linear differential equations that happen to describe many phenomena in nature as well as the properties of linear circuits. A linear circuit has the property that its output, when driven by the sum of two input signals, equals the sum of its individual outputs when driven by each input signal in turn; i.e., if $O(A)$ represents the output when driven by signal A, then a circuit is linear if $O(A + B) = O(A) + O(B)$. A linear circuit driven by a sine wave always responds with a sine wave, although in general the phase and amplitude are changed. No other signal can make this statement. It is standard practice, in fact, to describe the behavior of a circuit by its *frequency response*, the way it alters the amplitude of an applied sine wave as a function of frequency. A high-fidelity amplifier, for instance, should be characterized by a "flat" frequency response over the range 20Hz to 20kHz, at least.

The sine-wave frequencies you will usually deal with range from a few hertz to a few megahertz. Lower frequencies, down to 0.0001Hz or lower, can be generated with carefully built circuits, if needed. Higher frequencies, e.g., up to 2000MHz, can be generated, but they require special transmission-line techniques. Above that, you're dealing with microwaves, where conventional wired circuits with lumped circuit elements become impractical, and exotic waveguides or "striplines" are used instead.

1.08 Signal amplitudes and decibels

In addition to its amplitude, there are several other ways to characterize the magnitude of a sine wave or any other signal. You sometimes see it specified by *peak-to-peak amplitude* (pp amplitude), which is just what you would guess, namely, twice the amplitude. The other method is to give the *root-mean-square amplitude* (rms amplitude), which is $V_{\text{rms}} = (1/\sqrt{2})A = 0.707A$ (this is for sine waves only; the ratio of pp to rms will be different for other waveforms). Odd as it may seem, this is the usual method, because rms voltage is what's used to compute power. The voltage across the terminals of a wall socket (in the United States) is 117 volts rms, 60Hz. The *amplitude* is 165 volts (330 volts pp).

Decibels

How do you compare the relative amplitudes of two signals? You could say, for instance, that signal X is twice as large as signal Y. That's fine, and useful for many purposes. But because we often deal with ratios as large as a million, it is easier to use a logarithmic measure, and for this we present the decibel (it's one-tenth as large as something called a bel, which no one ever uses). By definition, the ratio of two signals, in decibels, is

$$\text{dB} = 20 \log_{10} \frac{A_2}{A_1}$$

where A_1 and A_2 are the two signal amplitudes. So, for instance, one signal of twice the amplitude of another is +6dB relative to it, since $\log_{10}2 = 0.3010$. A signal 10 times as large is +20dB; a signal one-tenth as large is −20dB. It is also useful to express the ratio of two signals in terms of power levels:

$$\mathrm{dB} = 10\log_{10}\frac{P_2}{P_1}$$

where P_1 and P_2 represent the power in the two signals. As long as the two signals have the same kind of waveform, e.g., sine waves, the two definitions give the same result. When comparing unlike waveforms, e.g., a sine wave versus "noise," the definition in terms of power (or the amplitude definition, with rms amplitudes substituted) must be used.

Although decibels are ordinarily used to specify the ratio of two signals, they are sometimes used as an absolute measure of amplitude. What is happening is that you are assuming some reference signal amplitude and expressing any other amplitude in decibels relative to it. There are several standard amplitudes (which are unstated, but understood) that are used in this way; the most common references are (a) dBV; 1 volt rms; (b) dBm; the voltage corresponding to 1mW into some assumed load impedance, which for radiofrequencies is usually 50 ohms, but for audio is often 600 ohms (the corresponding 0dBm amplitudes, when loaded by those impedances, are then 0.22V rms and 0.78V rms); and (c) the small noise voltage generated by a resistor at room temperature (this surprising fact is discussed in Section 7.11). In addition to these, there are reference amplitudes used for measurements in other fields. For instance, in acoustics, 0dB SPL is a wave whose rms pressure is 0.0002μbar (a bar is 10^6 dynes per square centimeter, approximately 1 atmosphere); in communications, levels can be stated in dBrnC (relative noise reference weighted in frequency by "curve C"). When stating amplitudes this way, it is best to be specific about the 0dB reference amplitude; say something like "an amplitude of 27 decibels relative to 1 volt rms," or abbreviate "27 dB re 1V rms," or define a term like "dBV."

EXERCISE 1.11

Determine the voltage and power ratios for a pair of signals with the following decibel ratios: (a) 3dB, (b) 6dB, (c) 10dB, (d) 20dB.

1.09 Other signals

The ramp is a signal that looks like the signal shown in Figure 1.18. It is simply a voltage rising (or falling) at a constant rate. That can't go on forever, of course, even in science fiction movies. It is sometimes approximated by a finite ramp (Fig. 1.19) or by a periodic ramp, or sawtooth (Fig. 1.20).

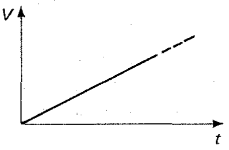


Figure 1.18. Voltage ramp waveform.

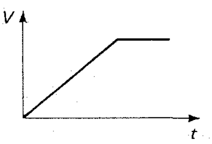


Figure 1.19. Ramp with limit.

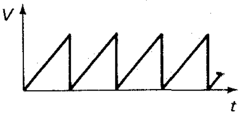


Figure 1.20. Sawtooth wave.

Triangle

The triangle wave is a close cousin of the ramp; it is simply a symmetrical ramp (Fig. 1.21).

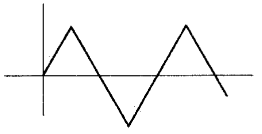


Figure 1.21. Triangle wave.

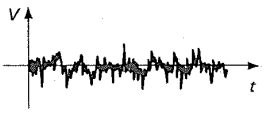


Figure 1.22. Noise.

Noise

Signals of interest are often mixed with *noise*; this is a catchall phrase that usually applies to random noise of thermal origin. Noise voltages can be specified by their frequency spectrum (power per hertz) or by their amplitude distribution. One of the most common kinds of noise is *band-limited white Gaussian noise*, which means a signal with equal power per hertz in some band of frequencies and a Gaussian (bell-shaped) distribution of amplitudes if large numbers of instantaneous measurements of its amplitude are made. This kind of noise is generated by a resistor (Johnson noise), and it plagues sensitive measurements of all kinds. On an oscilloscope it appears as shown in Figure 1.22. We will study noise and low-noise techniques in some detail in Chapter 7. Sections 9.32–9.37 deal with noise-generation techniques.

Square waves

A square wave is a signal that varies in time as shown in Figure 1.23. Like the sine wave, it is characterized by amplitude and frequency. A linear circuit driven by a square wave rarely responds with a square wave. For a square wave, the rms amplitude equals the amplitude.

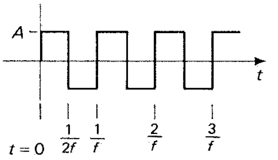


Figure 1.23. Square wave.

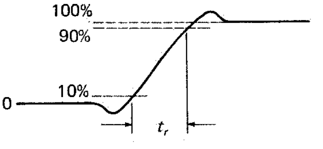


Figure 1.24. Rise time of a step waveform.

The edges of a square wave are not perfectly square; in typical electronic circuits the rise time t_r ranges from a few nanoseconds to a few microseconds. Figure 1.24 shows the sort of thing usually seen. The rise time is defined as the time required for the signal to go from 10% to 90% of its total transition.

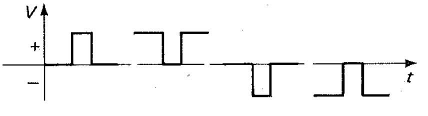


Figure 1.25. Positive- and negative-going pulses of both polarities.

Pulses

A pulse is a signal that looks as shown in Figure 1.25. It is defined by amplitude and pulse width. You can generate a train of periodic (equally spaced) pulses, in which case you can talk about the frequency, or pulse repetition rate, and the "duty cycle," the ratio of pulse width to repetition period (duty cycle ranges from zero to 100%). Pulses can have positive or negative polarity; in addition, they can be "positive-going" or "negative-going." For instance, the second pulse in Figure 1.25

is a negative-going pulse of positive polarity.

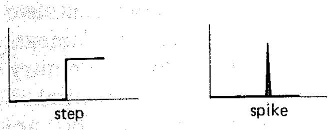


Figure 1.26

Steps and spikes

Steps and spikes are signals that are talked about a lot but are not often used. They provide a nice way of describing what happens in a circuit. If you could draw them, they would look something like the example in Figure 1.26. The step function is part of a square wave; the spike is simply a jump of vanishingly short duration.

1.10 Logic levels

Pulses and square waves are used extensively in digital electronics, where predefined voltage levels represent one of two possible states present at any point in the circuit. These states are called simply HIGH and LOW and correspond to the 0 (false) and 1 (true) states of Boolean logic (the algebra that describes such two-state systems).

Precise voltages are not necessary in digital electronics. You need only to distinguish which of the two possible states is present. Each digital logic family therefore specifies legal HIGH and LOW states. For example, the "74HC" digital logic family runs from a single +5 volt supply, with output levels that are typically 0 volts (LOW) and 5 volts (HIGH), and an input decision threshold of 2.5 volts. Actual outputs can be as much as a volt from ground or +5 volts without malfunction, however. We'll have much more to say about logic levels in Chapters 8 and 9.

1.11 Signal sources

Often the source of a signal is some part of the circuit you are working on. But for test purposes a flexible signal source is invaluable. They come in three flavors: signal generators, pulse generators, and function generators.

Signal generators

Signal generators are sine-wave oscillators, usually equipped to give a wide range of frequency coverage (50kHz to 50MHz is typical), with provision for precise control of amplitude (using a resistive divider network called an *attenuator*). Some units let you *modulate* the output (see Chapter 13). A variation on this theme is the *sweep generator*, a signal generator that can sweep its output frequency repeatedly over some range. These are handy for testing circuits whose properties vary with frequency in a particular way, e.g., "tuned circuits" or filters. Nowadays these devices, as well as many test instruments, are available in configurations that allow you to program the frequency, amplitude, etc., from a computer or other digital instrument.

A variation on the signal generator is the *frequency synthesizer*, a device that generates sine waves whose frequencies can be set precisely. The frequency is set digitally, often to eight significant figures or more, and is internally synthesized from a precise standard (a quartz-crystal oscillator) by digital methods we will discuss later (Sections 9.27–9.31). If your requirement is for no-nonsense accurate frequency generation, you can't beat a synthesizer.

Pulse generators

Pulse generators only make pulses, but what pulses! Pulse width, repetition rate, amplitude, polarity, rise time, etc., may all be adjustable. In addition, many units allow you to generate pulse pairs, with settable spacing and repetition rate, or even coded pulse trains. Most modern pulse generators are provided with logic-level outputs for easy connection to digital circuitry. Like signal generators, these come in the programmable variety.

Function generators

In many ways function generators are the most flexible signal sources of all. You can make sine, triangle, and square waves over an enormous frequency range (0.01 Hz to 10 MHz is typical), with control of amplitude and dc offset (a constant dc voltage added to the signal). Many of them have provision for frequency sweeping, often in several modes (linear or logarithmic frequency variation versus time). They are available with pulse outputs (although not with the flexibility you get with a pulse generator), and some of them have provision for modulation.

Like the other signal sources, function generators come in programmable versions and versions with digital readout of frequency (and sometimes amplitude). The most recent addition to the function-generator family is the synthesized function generator, a device that combines all the flexibility of a function generator with the stability and accuracy of a frequency synthesizer. An example is the HP 8116A, with sine, square, and triangle waves (as well as pulses, ramps, haversines, etc.) from 0.001Hz to 50MHz. Frequency and amplitude (10mV to 16V pp) are programmable, as are linear and logarithmic frequency sweeps. This unit also provides trigger, gate, burst, FM, AM, pulse-width modulation, voltage-controlled frequency, and single cycles. For general use, if you can have only one signal source, the function generator is for you.

CAPACITORS AND AC CIRCUITS

Once we enter the world of changing voltages and currents, or signals, we encounter two very interesting circuit elements that are useless in dc circuits: capacitors and inductors. As you will see, these humble devices, combined with resistors, complete the triad of passive linear circuit elements that form the basis of nearly all circuitry. Capacitors, in particular, are essential in nearly every circuit application. They are used for waveform generation, filtering, and blocking and bypass applications. They are used in integrators and differentiators. In combination with inductors, they make possible sharp filters for separating desired signals from background. You will see some of these applications as we continue with this chapter, and there will be numerous interesting examples in later chapters.

Let's proceed, then, to look at capacitors in detail. Portions of the treatment that follows are necessarily mathematical in nature; the reader with little mathematical preparation may find Appendix B helpful. In any case, an understanding of the details is less important in the long run than an understanding of the results.

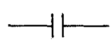


Figure 1.27. Capacitor.

1.12 Capacitors

A capacitor (Fig. 1.27) (the old-fashioned name was *condenser*) is a device that has two wires sticking out of it and has the property

$$Q = CV$$

A capacitor of C farads with V volts across its terminals has Q coulombs of stored charge on one plate, and $-Q$ on the other.

To a first approximation, capacitors are devices that might be considered simply frequency-dependent resistors. They allow you to make frequency-dependent voltage dividers, for instance. For some applications (bypass, coupling) this is almost all you need to know, but for other applications (filtering, energy storage, resonant circuits) a deeper understanding is needed. For example, capacitors cannot dissipate power, even though current can flow through them, because the voltage and current are 90° out of phase.

Taking the derivative of the defining equation above (see Appendix B), you get

$$I = C\frac{dV}{dt}$$

So a capacitor is more complicated than a resistor; the current is not simply proportional to the voltage, but rather to the rate of change of voltage. If you change the voltage across a farad by 1 volt per second, you are supplying an amp. Conversely, if you supply an amp, its voltage changes by 1 volt per second. A farad is very large, and you usually deal in microfarads (μF) or picofarads (pF). (To make matters confusing to the uninitiated, the units are often omitted on capacitor values specified in schematic diagrams. You have to figure it out from the context.) For instance, if you supply a current of 1mA to 1μF, the voltage will rise at 1000 volts per second. A 10ms pulse of this current will increase the voltage across the capacitor by 10 volts (Fig. 1.28).

Capacitors come in an amazing variety of shapes and sizes; with time, you will come to recognize their more common incarnations. The basic construction is simply two conductors near each other (but not touching); in fact, the simplest capacitors are just that. For greater capacitance, you need more area and closer spacing; the usual approach is to plate some conductor onto a thin insulating material (called a dielectric), for instance, aluminized Mylar film rolled up into a small cylindrical configuration. Other popular types are thin ceramic wafers (disc ceramics), metal foils with oxide insulators (electrolytics), and metallized mica. Each of these types has unique properties; for a brief rundown, see the box on capacitors. In general, ceramic and Mylar types are used for most noncritical circuit applications; tantalum capacitors are used where greater capacitance is needed, and electrolytics are used for power-supply filtering.

Capacitors in parallel and series

The capacitance of several capacitors in parallel is the sum of their individual capacitances. This is easy to see: Put voltage V across the parallel combination; then

$$\begin{aligned} C_{\text{total}}V &= Q_{\text{total}} = Q_1 + Q_2 + Q_3 + \ldots \\ &= C_1V + C_2V + C_3V + \ldots \\ &= (C_1 + C_2 + C_3 + \ldots)V \end{aligned}$$

or

$$C_{\text{total}} = C_1 + C_2 + C_3 + \ldots$$

For capacitors in series, the formula is like that for resistors in parallel:

$$C_{\text{total}} = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} + \ldots}$$

or (two capacitors only)

$$C_{\text{total}} = \frac{C_1C_2}{C_1 + C_2}$$

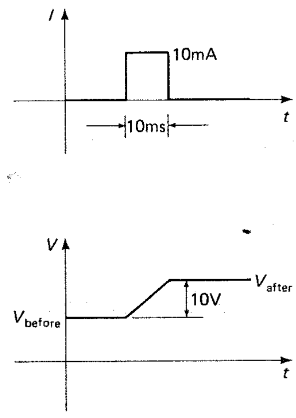


Figure 1.28. The voltage across a capacitor changes when a current flows through it.

CAPACITORS

There is wide variety among the capacitor types available. This is a quickie guide to point out their major advantages and disadvantages. Our judgments should be considered somewhat subjective:

Type	Capacitance range	Maximum voltage	Accuracy	Temperature stability	Leakage	Comments
Mica	1pF–0.01μF	100–600	Good		Good	Excellent; good at RF
Tubular ceramic	0.5pF–100pF	100–600		Selectable		Several tempcos (including zero)
Ceramic	10pF–1μF	50–30,000	Poor	Poor	Moderate	Small, inexpensive, very popular
Polyester (Mylar)	0.001μF–50μF	50–600	Good	Poor	Good	Inexpensive, good, popular
Polystyrene	10pF–2.7μF	100–600	Excellent	Good	Excellent	High quality, large; signal filters
Polycarbonate	100pF–30μF	50–800	Excellent	Excellent	Good	High quality, small
Polypropylene	100pF–50μF	100–800	Excellent	Good	Excellent	High quality, low dielectric absorption
Teflon	1000pF–2μF	50–200	Excellent	Best	Best	High quality, lowest dielectric absorption
Glass	10pF–1000pF	100–600	Good		Excellent	Long-term stability
Porcelain	100pF–0.1μF	50–400	Good	Good	Good	Good long-term stability
Tantalum	0.1μF–500μF	6–100	Poor	Poor		High capacitance; polarized, small; low inductance
Electrolytic	0.1μF–1.6F	3–600	Terrible	Ghastly	Awful	Power-supply filters; polarized; short life
Double layer	0.1F–10F	1.5–6	Poor	Poor	Good	Memory backup; high series resistance
Oil	0.1μF–20μF	200–10,000			Good	High-voltage filters; large, long life
Vacuum	1pF–5000pF	2000–36,000			Excellent	Transmitters

EXERCISE 1.12
Derive the formula for the capacitance of two capacitors in series. Hint: Because there is no external connection to the point where the two capacitors are connected together, they must have equal stored charges.

The current that flows in a capacitor during charging ($I = CdV/dt$) has some unusual features. Unlike resistive current, it's not proportional to voltage, but rather to the rate of change (the "time derivative") of voltage. Furthermore, unlike the situation in a resistor, the power (V times I) associated with capacitive current is not turned into heat, but is stored as energy in the capacitor's internal electric field. You get all that energy back when you discharge the capacitor. We'll see another way to look at these curious properties when we talk about *reactance*, beginning in Section 1.18.

1.13 *RC* circuits: *V* and *I* versus time

When dealing with ac circuits (or, in general, any circuits that have changing voltages and currents), there are two possible approaches. You can talk about V and I versus time, or you can talk about amplitude versus signal frequency. Both approaches have their merits, and you find yourself switching back and forth according to which description is most convenient in each situation. We will begin our study of ac circuits in the time domain. Beginning with Section 1.18, we will tackle the frequency domain.

What are some of the features of circuits with capacitors? To answer this question, let's begin with the simple *RC* circuit (Fig. 1.29). Application of the capacitor rules gives

$$C\frac{dV}{dt} = I = -\frac{V}{R}$$

This is a differential equation, and its solution is

$$V = Ae^{-t/RC}$$

So a charged capacitor placed across a resistor will discharge as in Figure 1.30.

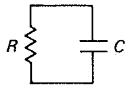


Figure 1.29

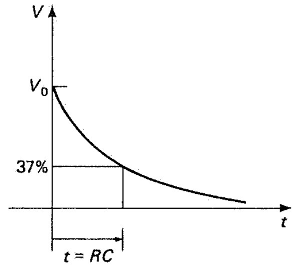


Figure 1.30. *RC* discharge waveform.

Time constant

The product RC is called the *time constant* of the circuit. For R in ohms and C in farads, the product RC is in seconds. A microfarad across 1.0k has a time constant of 1ms; if the capacitor is initially charged to 1.0 volt, the initial current is 1.0mA.

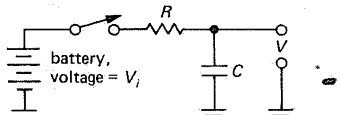


Figure 1.31

Figure 1.31 shows a slightly different circuit. At time $t = 0$, someone connects the battery. The equation for the circuit is then

$$I = C\frac{dV}{dt} = \frac{V_i - V}{R}$$

with the solution

$$V = V_i + Ae^{-t/RC}$$

(Please don't worry if you can't follow the mathematics. What we are doing is getting some important results, which you should remember. Later we will use the results often, with no further need for the mathematics used to derive them.) The constant A is determined by initial conditions (Fig. 1.32): $V = 0$ at $t = 0$; therefore, $A = -V_i$, and

$$V = V_i(1 - e^{-t/RC})$$

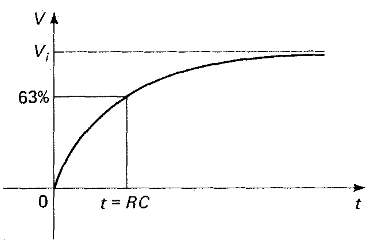


Figure 1.32

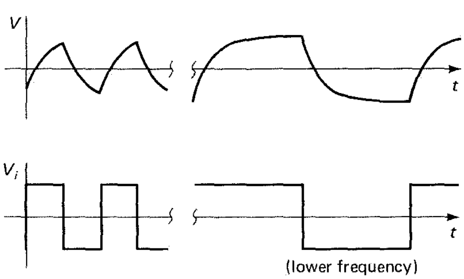


Figure 1.33. Output (top waveform) across a capacitor, when driven by square waves through a resistor.

Decay to equilibrium

Eventually (when $t \gg RC$), V reaches V_i. (Presenting the "5RC rule of thumb": a capacitor charges or decays to within 1% of its final value in 5 time constants.) If we then change V_0 to some other value (say, 0), V will decay toward that new value with an exponential $e^{-t/RC}$. For example, a square-wave input for V_0 will produce the output shown in Figure 1.33.

EXERCISE 1.13

Show that the rise time (the time required to go from 10% to 90% of its final value) of this signal is 2.2RC.

You might ask the obvious next question: What about $V(t)$ for arbitrary $V_i(t)$? The solution involves an inhomogeneous differential equation and can be solved by standard methods (which are, however, beyond the scope of this book). You would find

$$V(t) = \frac{1}{RC}\int_{-\infty}^{t} V_i(\tau)e^{-(t-\tau)/RC}d\tau$$

That is, the RC circuit averages past history at the input with a weighting factor $e^{-\Delta t/RC}$

In practice, you seldom ask this question. Instead, you deal in the *frequency domain* and ask how much of each frequency component present in the input gets through. We will get to this important topic soon (Section 1.18). Before we do, though, there are a few other interesting circuits we can analyze simply with this time-domain approach.

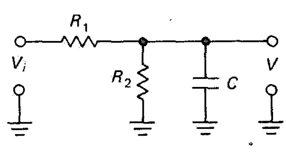


Figure 1.34

Simplification by Thévenin equivalents

We could go ahead and analyze more complicated circuits by similar methods, writing down the differential equations and trying to find solutions. For most purposes it simply isn't worth it. This is as complicated an RC circuit as we will need. Many other circuits can be reduced to it (e.g., Fig. 1.34). By just using the Thévenin equivalent of the voltage divider formed by R_1 and R_2, you can find the output

$V(t)$ produced by a step input for V_0.

EXERCISE 1.14

$R_1 = R_2 = 10\text{k}$, and $C = 0.1\mu\text{F}$ in the circuit shown in Figure 1.34. Find $V(t)$ and sketch it.

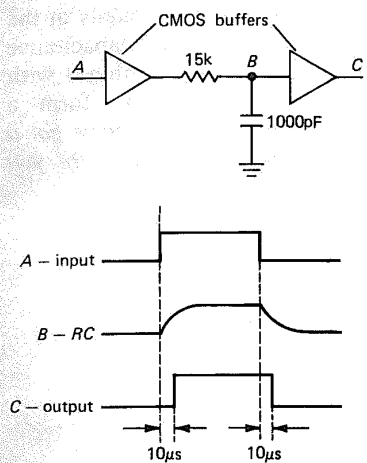


Figure 1.35. Producing a delayed digital waveform with the help of an RC.

Example: time-delay circuit

We have already mentioned logic levels, the voltages that digital circuits live on. Figure 1.35 shows an application of capacitors to produce a delayed pulse. The triangular symbols are "CMOS buffers." They give a HIGH output if the input is HIGH (more than one-half the dc power-supply voltage used to power them), and vice versa. The first buffer provides a replica of the input signal, but with low source resistance, and prevents input loading by the RC (recall our earlier discussion of circuit loading in Section 1.05). The RC output has the characteristic decays and causes the output buffer to switch 10μs after the input transitions (an RC reaches 50% output in $0.7RC$). In an actual application you would have to consider the effect of the buffer input threshold deviating from one-half the supply voltage, which would alter the delay and change the output pulse width. Such a circuit is sometimes used to delay a pulse so that something else can happen first. In designing circuits you try not to rely on tricks like this, but they're occasionally handy.

1.14 Differentiators

Look at the circuit in Figure 1.36. The voltage across C is $V_{\text{in}} - V$, so

$$I = C\frac{d}{dt}(V_{\text{in}} - V) = \frac{V}{R}$$

If we choose R and C small enough so that $dV/dt \ll dV_{\text{in}}/dt$, then

$$C\frac{dV_{\text{in}}}{dt} \approx \frac{V}{R}$$

or

$$V(t) = RC\frac{d}{dt}V_{\text{in}}(t)$$

That is, we get an output proportional to the rate of change of the input waveform.

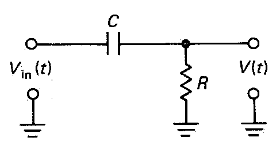


Figure 1.36

To keep $dV/dt \ll dV_{\text{in}}/dt$, we make the product RC small, taking care not to "load" the input by making R too small (at the transition the change in voltage across the capacitor is zero, so R is the load seen by the input). We will have a better criterion for this when we look at things in the frequency domain. If you drive this circuit with a square wave, the output will be as shown in Figure 1.37.

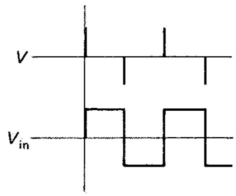


Figure 1.37. Output waveform (top) from differentiator driven by a square wave.

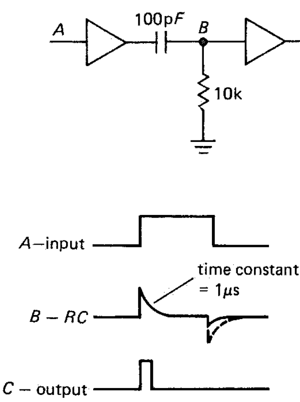


Figure 1.38. Leading-edge detector.

Differentiators are handy for detecting *leading edges* and *trailing edges* in pulse signals, and in digital circuitry you sometimes see things like those depicted in Figure 1.38. The RC differentiator generates spikes at the transitions of the input signal, and the output buffer converts the spikes to short square-topped pulses. In practice, the negative spike will be small because of a diode (a handy device discussed in Section 1.25) built into the buffer.

Unintentional capacitive coupling

Differentiators sometimes crop up unexpectedly, in situations where they're not welcome. You may see signals like those shown in Figure 1.39. The first case is caused by a square wave somewhere in the circuit coupling capacitively to the signal line you're looking at; that might indicate a missing resistor termination on your signal line. If not, you must either reduce the source resistance of the signal line or find a way to reduce capacitive coupling from the offending square wave. The second case is typical of what you might see when you look at a square wave, but have a broken connection somewhere, usually at the scope probe. The very small capacitance of the broken connection combines with the scope input resistance to form a differentiator. *Knowing that you've got a differentiated "something" can help you find the trouble and eliminate it.*

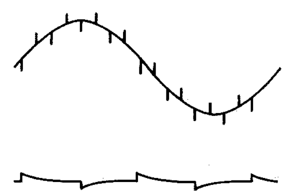


Figure 1.39

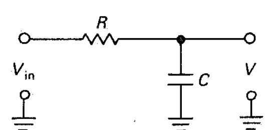


Figure 1.40

1.15 Integrators

Take a look at the circuit in Figure 1.40. The voltage across R is $V_{\text{in}} - V$, so

$$I = C\frac{dV}{dt} = \frac{V_{\text{in}} - V}{R}$$

If we manage to keep $V \ll V_{\text{in}}$, by keeping the product RC large, then

$$C\frac{dV}{dt} \approx \frac{V_{\text{in}}}{R}$$

or

$$V(t) = \frac{1}{RC}\int^{t} V_{\text{in}}(t)\,dt + \text{constant}$$

We have a circuit that performs the integral over time of an input signal! You can see how the approximation works for a square-wave input: $V(t)$ is then the exponential charging curve we saw earlier (Fig. 1.41). The first part of the exponential is a ramp, the integral of a constant; as we increase the time constant RC, we pick off a smaller part of the exponential, i.e., a better approximation to a perfect ramp.

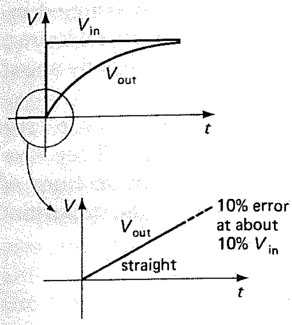


Figure 1.41

Note that the condition $V \ll V_{in}$ is just the same as saying that I is proportional to V_{in}. If we had as input a *current* $I(t)$, rather than a voltage, we would have an exact integrator. A large voltage across a large resistance approximates a current source and, in fact, is frequently used as one.

Later, when we get to operational amplifiers and feedback, we will be able to build integrators without the restriction $V_{out} \ll V_{in}$. They will work over large frequency and voltage ranges with negligible error.

The integrator is used extensively in analog computation. It is a useful subcircuit that finds application in control systems, feedback, analog/digital conversion, and waveform generation.

Ramp generators

At this point it is easy to understand how a ramp generator works. This nice circuit is extremely useful, for example in timing circuits, waveform and function generators, oscilloscope sweep circuits, and analog/digital conversion circuitry. The circuit uses a constant current to charge a capacitor (Fig. 1.42). From the capacitor equation $I = C(dV/dt)$, you get $V(t) = (I/C)t$. The output waveform is as shown in Figure 1.43. The ramp stops when the current source "runs out of voltage," i.e., reaches the limit of its compliance. The curve for a simple RC, with the resistor tied to a voltage source equal to the compliance of the current source, and with R chosen so that the current at zero output voltage is the same as that of the current source, is also drawn for comparison. (Real current sources generally have output compliances limited by the power-supply voltages used in making them, so the comparison is realistic.) In the next chapter, which deals with transistors, we will design some current sources, with some refinements to follow in the chapters on operational amplifiers (op-amps) and field-effect transistors (FETs). Exciting things to look forward to!

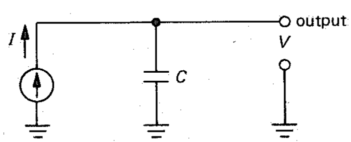


Figure 1.42. A constant current source charging a capacitor generates a ramp voltage waveform.

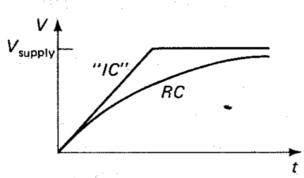


Figure 1.43

EXERCISE 1.15

A current of 1 mA charges a 1 μF capacitor. How long does it take the ramp to reach 10 volts?

INDUCTORS AND TRANSFORMERS

1.16 Inductors

If you understand capacitors, you won't have any trouble with inductors (Fig. 1.44). They're closely related to capacitors; the rate of current change in an inductor depends on the voltage applied across it, whereas the rate of voltage change in a capacitor depends on the current through it. The defining equation for an inductor is

$$V = L\frac{dI}{dt}$$

where L is called the *inductance* and is measured in henrys (or mH, μH, etc.). Putting a voltage across an inductor causes the current to rise as a ramp (for a capacitor, supplying a constant current causes the voltage to rise as a ramp); 1 volt across 1 henry produces a current that increases at 1 amp per second.

Figure 1.44. Inductor.

As with capacitive current, inductive current is not simply proportional to voltage. Furthermore, unlike the situation in a resistor, the power associated with inductive current (V times I) is not turned into heat, but is stored as energy in the inductor's magnetic field. You get all that energy back when you interrupt the inductor's current.

The symbol for an inductor looks like a coil of wire; that's because, in its simplest form, that's all it is. Variations include coils wound on various core materials, the most popular being iron (or iron alloys, laminations, or powder) and ferrite, a black, nonconductive, brittle magnetic material. These are all ploys to multiply the inductance of a given coil by the "permeability" of the core material. The core may be in the shape of a rod, a toroid (doughnut), or even more bizarre shapes, such as a "pot core" (which has to be seen to be understood; the best description we can think of is a doughnut mold split horizontally in half, if doughnuts were made in molds).

Inductors find heavy use in radio-frequency (RF) circuits, serving as RF "chokes" and as parts of tuned circuits (see Chapter 13). A pair of closely coupled inductors forms the interesting object known as a transformer. We will talk briefly about them in the next section.

An inductor is, in a real sense, the opposite of a capacitor. You will see how that works out in the next few sections of this chapter, which deal with the important subject of *impedance*.

1.17 Transformers

A transformer is a device consisting of two closely coupled coils (called primary and secondary). An ac voltage applied to the primary appears across the secondary, with a voltage multiplication proportional to the turns ratio of the transformer and a current multiplication inversely proportional to the turns ratio. Power is conserved. Figure 1.45 shows the circuit symbol for a laminated-core transformer (the kind used for 60Hz ac power conversion).

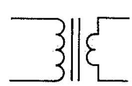


Figure 1.45. Transformer.

Transformers are quite efficient (output power is very nearly equal to input power); thus, a step-up transformer gives higher voltage at lower current. Jumping ahead for a moment, a transformer of turns ratio n increases the impedance by n^2. There is very little primary current if the secondary is unloaded.

Transformers serve two important functions in electronic instruments: They change the ac line voltage to a useful (usually lower) value that can be used by the circuit, and they "isolate" the electronic device from actual connection to the power line, because the windings of a transformer are electrically insulated from each other. *Power transformers* (meant for use from the 110V power line) come in an enormous variety of secondary voltages and currents: outputs as low as 1 volt or so up to several thousand volts, current ratings from a few milliamps to hundreds of amps. Typical transformers for use in electronic instruments might have secondary voltages from 10 to 50 volts, with current ratings of 0.1 to 5 amps or so.

Transformers for use at audiofrequencies and radiofrequencies are also available. At radiofrequencies you sometimes use tuned transformers, if only a narrow range of frequencies is present. There is also an interesting class of transmission-line transformer that we will discuss briefly in Section 13.10. In general, transformers for use at high frequencies must use special core materials or construction to minimize core losses, whereas low-frequency transformers (e.g., power transformers) are burdened instead by large and heavy cores. The two kinds of transformers are in general not interchangeable.

IMPEDANCE AND REACTANCE

Warning: This section is somewhat mathematical; you may wish to skip over the mathematics, but be sure to pay attention to the results and graphs.

Circuits with capacitors and inductors are more complicated than the resistive circuits we talked about earlier, in that their behavior depends on frequency: A "voltage divider" containing a capacitor or inductor will have a frequency-dependent division ratio. In addition, circuits containing these components (known collectively as *reactive* components) "corrupt" input waveforms such as square waves, as we just saw.

However, both capacitors and inductors are *linear* devices, meaning that the amplitude of the output waveform, whatever its shape, increases exactly in proportion to the input waveform's amplitude. This linearity has many consequences, the most important of which is probably the following: *The output of a linear circuit, driven with a sine wave at some frequency* f, *is itself a sine wave at the same frequency (with, at most, changed amplitude and phase).*

Because of this remarkable property of circuits containing resistors, capacitors, and inductors (and, later, linear amplifiers), it is particularly convenient to analyze any such circuit by asking how the output voltage (amplitude and phase) depends on the input voltage, *for sine-wave input at a single frequency*, even though this may not be the intended use. A graph of the resulting *frequency response*, in which the ratio of output to input is plotted for each sine-wave frequency, is useful for thinking about many kinds of waveforms. As an example, a certain "boom-box" loudspeaker might have the frequency response shown in Figure 1.46, where the "output" in this case is of course sound pressure, not voltage. It is desirable for a speaker to have a "flat" response, meaning that the graph of sound pressure versus frequency is constant over the band of audible frequencies. In this case the speaker's deficiencies can be corrected by introducing a passive filter with the inverse response (as shown) into the amplifiers of the radio.

As we will see, it is possible to generalize Ohm's law, replacing the word "resistance" with "impedance," in order to describe any circuit containing these linear passive devices (resistors, capacitors, and inductors). You could think of the subject of impedance and reactance as Ohm's law for circuits that include capacitors and inductors. Some important terminology: Impedance is the "generalized resistance"; inductors

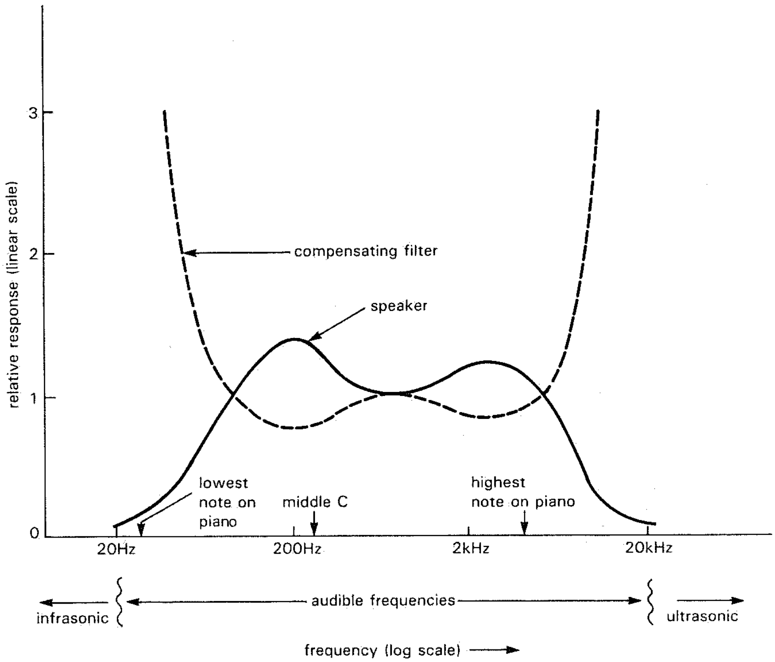


Figure 1.46. Example of frequency analysis: "boom box" loudspeaker equalization.

and capacitors have *reactance* (they are "reactive"); resistors have *resistance* (they are "resistive"). In other words, impedance = resistance + reactance (more about this later). However, you'll see statements like "the impedance of the capacitor at this frequency is ... " The reason you don't have to use the word "reactance" in such a case is that impedance covers everything. In fact, you frequently use the word "impedance" even when you know it's a resistance you're talking about; you say "the source impedance" or "the output impedance" when you mean the Thévenin equivalent resistance of some source. The same holds for "input impedance."

In all that follows, we will be talking about circuits driven by sine waves at a single frequency. Analysis of circuits driven by complicated waveforms is more elaborate, involving the methods we used earlier (differential equations) or decomposition of the waveform into sine waves (Fourier analysis). Fortunately, these methods are seldom necessary.

1.18 Frequency analysis of reactive circuits

Let's start by looking at a capacitor driven by a sine-wave voltage source (Fig. 1.47). The current is

$$I(t) = C\frac{dV}{dt} = C\omega V_0 \cos \omega t$$

i.e., a current of amplitude I, with the phase leading the input voltage by 90°. If we consider amplitudes only, and disregard phases, the current is

$$I = \frac{V}{1/\omega C}$$

(Recall that $\omega = 2\pi f$.) It behaves like a frequency-dependent resistance $R = 1/\omega C$, but in addition the current is 90° out of phase with the voltage (Fig. 1.48).

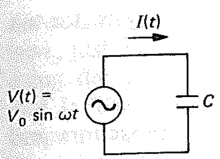


Figure 1.47

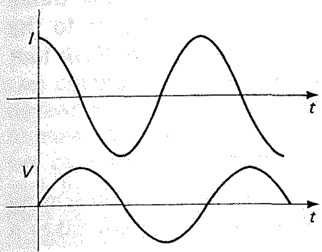


Figure 1.48

For example, a 1μF capacitor put across the 110 volt (rms) 60Hz power line draws a current of rms amplitude

$$I = \frac{110}{1/(2\pi \times 60 \times 10^{-6})} = 41.5\text{mA (rms)}$$

Note: At this point it is necessary to get into some complex algebra; you may wish to skip over the math in some of the following sections, taking note of the results as we derive them. A knowledge of the detailed mathematics is not necessary in order to understand the remainder of the book. Very little mathematics will be used in later chapters. The section ahead is easily the most difficult for the reader with little mathematical preparation. Don't be discouraged!

Voltages and currents as complex numbers

As you have just seen, there can be phase shifts between the voltage and current in an ac circuit being driven by a sine wave at some frequency. Nevertheless, as long as the circuit contains only *linear* elements (resistors, capacitors, inductors), the magnitudes of the currents everywhere in the circuit are still proportional to the magnitude of the driving voltage, so we might hope to find some generalization of voltage, current, and resistance in order to rescue Ohm's law. Obviously a single number won't suffice to specify the current, say, at some point in the circuit, because we must somehow have information about both the magnitude and phase shift.

Although we can imagine specifying the magnitudes and phase shifts of voltages and currents at any point in the circuit by writing them out explicitly, e.g., $V(t) = 23.7\sin(377t + 0.38)$, it turns out that our requirements can be met more simply by using the algebra of complex numbers to *represent* voltages and currents. Then we can simply add or subtract the complex number representations, rather than laboriously having to add or subtract the actual sinusoidal functions of time themselves. Because the actual voltages and currents are real quantities that vary with time, we must develop a rule for converting from actual quantities to their representations, and vice versa. Recalling once again that we are talking about a single sine-wave frequency, ω, we agree to use the following rules:

1. Voltages and currents are *represented* by the complex quantities **V** and **I**. The voltage $V_0 \cos(\omega t + \phi)$ is to be represented by the complex number $V_0 e^{j\phi}$. Recall that $e^{j\theta} = \cos\theta + j\sin\theta$, where $j = \sqrt{-1}$.

2. *Actual* voltages and currents are obtained by multiplying their complex number representations by $e^{j\omega t}$ and then taking the real part: $V(t) = \mathcal{R}e(\mathbf{V}e^{j\omega t})$, $I(t) = \mathcal{R}e(\mathbf{I}e^{j\omega t})$

In other words,

circuit voltage versus time		complex number representation
$V_0\cos(\omega t + \phi)$	$\rightleftarrows$ multiply by $e^{j\omega t}$ and take real part	$V_0 e^{j\phi} = a + jb$

(In electronics, the symbol j is used instead of i in the exponential in order to avoid confusion with the symbol i meaning current.) Thus, in the general case the actual voltages and currents are given by

$$\begin{aligned} V(t) &= \mathcal{R}e(\mathbf{V}e^{j\omega t}) \\ &= \mathcal{R}e(\mathbf{V})\cos\omega t - \mathcal{I}m(\mathbf{V})\sin\omega t \\ I(t) &= \mathcal{R}e(\mathbf{I}e^{j\omega t}) \\ &= \mathcal{R}e(\mathbf{I})\cos\omega t - \mathcal{I}m(\mathbf{I})\sin\omega t \end{aligned}$$

For example, a voltage whose complex representation is

$$\mathbf{V} = 5j$$

corresponds to a (real) voltage versus time of

$$\begin{aligned} V(t) &= \mathcal{R}e[5j\cos\omega t + 5j(j)\sin\omega t] \\ &= -5\sin\omega t \text{ volts} \end{aligned}$$

Reactance of capacitors and inductors

With this convention we can apply complex Ohm's law to circuits containing capacitors and inductors, just as for resistors, once we know the reactance of a capacitor or inductor. Let's find out what these are. We have

$$V(t) = \mathcal{R}e(V_0 e^{j\omega t})$$

For a capacitor, using $I = C(dV/dt)$, we obtain

$$\begin{aligned} I(t) &= -V_0 C\omega \sin\omega t = \mathcal{R}e\left(\frac{V_0 e^{j\omega t}}{-j/\omega C}\right) \\ &= \mathcal{R}e\left(\frac{V_0 e^{j\omega t}}{X_C}\right) \end{aligned}$$

i.e., for a capacitor

$$X_C = -j/\omega C$$

X_C is the reactance of a capacitor at frequency ω. As an example a 1μF capacitor has a reactance of $-2653j$ ohms at 60Hz and a reactance of $-0.16j$ ohms at 1MHz. Its reactance at dc is infinite.

If we did a similar analysis for an inductor, we would find

$$X_L = j\omega L$$

A circuit containing only capacitors and inductors always has a purely imaginary impedance, meaning that the voltage and current are always 90° out of phase – it is purely reactive. When the circuit contains resistors, there is also a real part to the impedance. The term "reactance" in that case means the imaginary part only.

Ohm's law generalized

With these conventions for representing voltages and currents, Ohm's law takes a simple form. It reads simply

$$\mathbf{I} = \mathbf{V}/\mathbf{Z}$$
$$\mathbf{V} = \mathbf{I}\mathbf{Z}$$

where the voltage represented by $\mathbf{V}$ is applied across a circuit of impedance $\mathbf{Z}$, giving a current represented by $\mathbf{I}$. The complex impedance of devices in series or parallel obeys the same rules as resistance:

$$\mathbf{Z} = \mathbf{Z}_1 + \mathbf{Z}_2 + \mathbf{Z}_3 + \cdots \quad \text{(series)}$$
$$\mathbf{Z} = \frac{1}{\frac{1}{\mathbf{Z}_1} + \frac{1}{\mathbf{Z}_2} + \frac{1}{\mathbf{Z}_3} + \cdots} \quad \text{(parallel)}$$

Finally, for completeness we summarize here the formulas for the impedance of resistors, capacitors, and inductors:

$$\mathbf{Z}_R = R \quad \text{(resistor)}$$
$$\mathbf{Z}_C = -j/\omega C = 1/j\omega C \quad \text{(capacitor)}$$
$$\mathbf{Z}_L = j\omega L \quad \text{(inductor)}$$

With these rules we can analyze many ac circuits by the same general methods we used in handling dc circuits, i.e., application of the series and parallel formulas and Ohm's law. Our results for circuits such as voltage dividers will look nearly the same as before. For multiply connected networks we may have to use Kirchhoff's laws, just as with dc circuits, in this case using the complex representations for V and I: The sum of the (complex) voltage drops around a closed loop is zero, and the sum of the (complex) currents into a point is zero. The latter rule implies, as with dc circuits, that the (complex) current in a series circuit is the same everywhere.

EXERCISE 1.16

Use the preceding rules for the impedance of devices in parallel and in series to derive the formulas (Section 1.12) for the capacitance of two capacitors (a) in parallel and (b) in series. Hint: In each case, let the individual capacitors have capacitances C_1 and C_2. Write down the impedance of the parallel or series combination; then equate it to the impedance of a capacitor with capacitance C. Find C.

Let's try out these techniques on the simplest circuit imaginable, an ac voltage applied across a capacitor, which we considered just previously. Then, after a brief look at power in reactive circuits (to finish laying the groundwork), we'll analyze some simple but extremely important and useful RC filter circuits.

Imagine putting a 1μF capacitor across a 110 volt (rms) 60Hz power line. What current flows? Using complex Ohm's law, we have

$$\mathbf{Z} = -j/\omega C$$

Therefore, the current is given by

$$\mathbf{I} = \mathbf{V}/\mathbf{Z}$$

The phase of the voltage is arbitrary, so let us choose $\mathbf{V} = A$, i.e. $V(t) = A\cos\omega t$, where the amplitude $A = 110\sqrt{2} \approx 156$ volts. Then

$$\mathbf{I} = j\omega CA \approx 0.059\sin\omega t$$

The resulting current has an amplitude of 59mA (41.5mA rms) and leads the voltage by 90°. This agrees with our previous calculation. Note that if we just wanted to know the magnitude of the current, and didn't care what the relative phase was, we could have avoided doing any complex algebra: If

$$\mathbf{A} = \mathbf{B}/\mathbf{C}$$

then

$$A = B/C$$

where A, B, and C are the magnitudes of the respective complex numbers; this holds for multiplication, also (see Exercise 1.17). Thus, in this case,

$$I = V/Z = \omega CV$$

This trick is often useful.

Surprisingly, there is no power dissipated by the capacitor in this example. Such activity won't increase your electric bill; you'll see why in the next section. Then we will go on to look at circuits containing resistors and capacitors with our complex Ohm's law.

EXERCISE 1.17

Show that if $\mathbf{A}=\mathbf{BC}$, then $A=BC$, where A, B, and C are magnitudes. Hint: Represent each complex number in polar form, i.e., $\mathbf{A}=Ae^{i\theta}$.

Power in reactive circuits

The instantaneous power delivered to any circuit element is always given by the product $P = VI$. However, in reactive circuits where V and I are not simply proportional, you can't just multiply them together. Funny things can happen; for instance, the sign of the product can reverse over one cycle of the ac signal. Figure 1.49 shows an example. During time intervals A and C, power is being delivered to the capacitor (albeit at a variable rate), causing it to charge up; its stored energy is increasing (power is the rate of change of energy). During intervals B and D, the power delivered to the capacitor is negative; it is discharging. The average power over a whole cycle for this example is in fact exactly zero, a statement that is always true for any purely reactive circuit element (inductors, capacitors, or any combination thereof). If you know your trigonometric integrals, the next exercise will show you how to prove this.

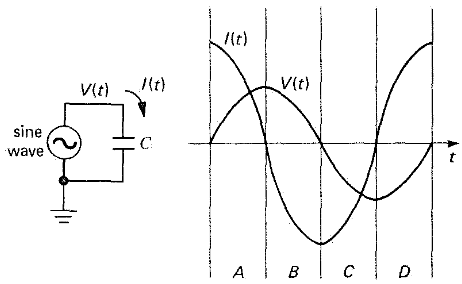


Figure 1.49. When driven by a sine wave, the current through a capacitor leads the voltage by 90°.

EXERCISE 1.18

Optional exercise: Prove that a circuit whose current is 90° out of phase with the driving voltage consumes no power, averaged over an entire cycle.

How do we find the average power consumed by an arbitrary circuit? In general, we can imagine adding up little pieces of VI product, then dividing by the elapsed time. In other words,

$$P = \frac{1}{T}\int_0^T V(t)I(t)\,dt$$

where T is the time for one complete cycle. Luckily, that's almost never necessary. Instead, it is easy to show that the average power is given by

$$P = \mathcal{R}e(\mathbf{V}\mathbf{I}^*) = \mathcal{R}e(\mathbf{V}^*\mathbf{I})$$

where $\mathbf{V}$ and $\mathbf{I}$ are complex rms amplitudes.

Let's take an example. Consider the preceding circuit, with a 1 volt (rms) sine wave driving a capacitor. We'll do everything with rms amplitudes, for simplicity. We have

$$\mathbf{V} = 1$$

$$\mathbf{I} = \frac{\mathbf{V}}{-j/\omega C} = j\omega C$$

$$P = \mathcal{R}e(\mathbf{V}\mathbf{I}^*) = \mathcal{R}e(-j\omega C) = 0$$

That is, the average power is zero, as stated earlier.

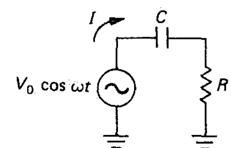


Figure 1.50

As another example, consider the circuit shown in Figure 1.50. Our calculations go like this:

$$\mathbf{Z} = R - \frac{j}{\omega C}$$

$$\mathbf{V} = V_0$$

$$\mathbf{I} = \frac{\mathbf{V}}{\mathbf{Z}} = \frac{V_0}{R-(j/\omega C)} = \frac{V_0[R+(j/\omega C)]}{R^2+(1/\omega^2C^2)}$$

$$P = \mathcal{R}e(\mathbf{V}\mathbf{I}^*) = \frac{V_0^2 R}{R^2+(1/\omega^2C^2)}$$

(In the third line we multiplied numerator and denominator by the complex conjugate of the denominator, in order to make the denominator real.) This is less than the product of the magnitudes of $\mathbf{V}$ and $\mathbf{I}$. In fact, the ratio is called the *power factor*:

$$|\mathbf{V}|\,|\mathbf{I}| = \frac{V_0^2}{[R^2+(1/\omega^2C^2)]^{1/2}}$$

$$\text{power factor} = \frac{\text{power}}{|\mathbf{V}|\,|\mathbf{I}|} = \frac{R}{[R^2+(1/\omega^2C^2)]^{1/2}}$$

in this case. The power factor is the cosine of the phase angle between the voltage and the current, and it ranges from 0 (purely reactive circuit) to 1 (purely resistive). A power factor less than 1 indicates some component of reactive current.

EXERCISE 1.19
Show that all the average power delivered to the preceding circuit winds up in the resistor. Do this by computing the value of V_R^2/R. What is that power, in watts, for a series circuit of a 1μF capacitor and a 1.0k resistor placed across the 110 volt (rms), 60Hz power line?

Power factor is a serious matter in large-scale electrical power distribution, because reactive currents don't result in useful power being delivered to the load, but cost the power company plenty in terms of I^2R heating in the resistance of generators, transformers, and wiring. Although residential users are only billed for "real" power $[\mathcal{R}e(\mathbf{VI}^*)]$, the power company charges industrial users according to the power factor. This explains the capacitor yards that you see behind large factories, built to cancel the inductive reactance of industrial machinery (i.e., motors).

EXERCISE 1.20
Show that adding a series capacitor of value $C = 1/\omega^2 L$ makes the power factor equal 1.0 in a series RL circuit. Now do the same thing, but with the word "series" changed to "parallel."

Voltage dividers generalized

Our original voltage divider (Fig. 1.5) consisted of a pair of resistors in series to ground, input at the top and output at the junction. The generalization of that simple resistive divider is a similar circuit in which either or both resistors are replaced by a capacitor or inductor (or a more complicated network made from R, L, and C), as in Figure 1.51. In general, the division ratio V_{out}/V_{in} of such a divider is not constant, but depends on frequency. The analysis is straightforward:

$$\mathbf{I} = \frac{\mathbf{V}_{in}}{\mathbf{Z}_{total}}$$

$$\mathbf{Z}_{total} = \mathbf{Z}_1 + \mathbf{Z}_2$$

$$\mathbf{V}_{out} = \mathbf{I}\,\mathbf{Z}_2 = \mathbf{V}_{in}\frac{\mathbf{Z}_2}{\mathbf{Z}_1 + \mathbf{Z}_2}$$

Rather than worrying about this result in general, let's look at some simple, but very important, examples.

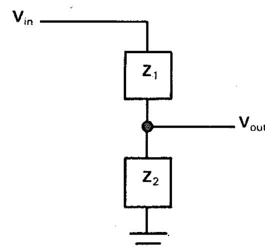


Figure 1.51. Generalized voltage divider: a pair of arbitrary impedances.

1.19 *RC* filters

By combining resistors with capacitors it is possible to make frequency-dependent voltage dividers, owing to the frequency dependence of a capacitor's impedance $\mathbf{Z}_C = -j/\omega C$. Such circuits can have the desirable property of passing signal frequencies of interest while rejecting undesired signal frequencies. In this section you will see examples of the simplest such RC filters, which we will be using frequently throughout the book. Chapter 5 and Appendix H describe filters of greater sophistication.

High-pass filters

Figure 1.52 shows a voltage divider made from a capacitor and a resistor. Complex Ohm's law gives

$$\mathbf{I} = \frac{\mathbf{V}_{in}}{\mathbf{Z}_{total}} = \frac{\mathbf{V}_{in}}{R - (j/\omega C)}$$
$$= \frac{\mathbf{V}_{in}[R + (j/\omega C)]}{R^2 + 1/\omega^2 C^2}$$

(For the last step, multiply top and bottom by the complex conjugate of the denominator.) So the voltage across R is just

$$\mathbf{V}_{out} = \mathbf{I}\,\mathbf{Z}_R = \mathbf{I}\,R = \frac{\mathbf{V}_{in}[R + (j/\omega C)]R}{R^2 + (1/\omega^2 C^2)}$$

Most often we don't care about the phase of V_{out}, just its amplitude:

$$\begin{aligned}\mathbf{V}_{\text{out}} &= (\mathbf{V}_{\text{out}}\mathbf{V}^*_{\text{out}})^{1/2}\\ &= \frac{R}{[R^2 + (1/\omega^2C^2)]^{1/2}}V_{\text{in}}\end{aligned}$$

Note the analogy to a resistive divider, where

$$V_{\text{out}} = \frac{R_1}{R_1 + R_2}V_{\text{in}}$$

Here the impedance of the series RC combination (Fig. 1.53) is as shown in Figure 1.54. So the "response" of this circuit, ignoring phase shifts by taking magnitudes of the complex amplitudes, is given by

$$\begin{aligned}V_{\text{out}} &= \frac{R}{[R^2 + (1/\omega^2C^2)]^{1/2}}V_{\text{in}}\\ &= \frac{2\pi fRC}{[1 + (2\pi fRC)^2]^{1/2}}V_{\text{in}}\end{aligned}$$

and looks as shown in Figure 1.55. We could have gotten this result immediately by taking the ratio of the *magnitudes* of impedances, as in Exercise 1.17 and the example immediately preceding it; the numerator is the magnitude of the impedance of the lower leg of the divider (R), and the denominator is the magnitude of the impedance of the series combination of R and C.

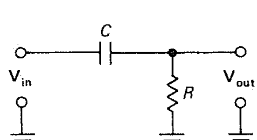


Figure 1.52. High-pass filter.

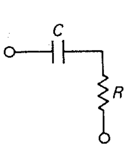
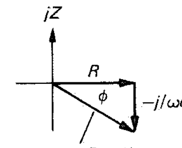


Figure 1.53



$Z_{\text{total}} = R - j/\omega C$

$|Z_{\text{total}}| = \sqrt{R^2 + \frac{1}{\omega^2C^2}}$

$\phi = \tan^{-1}\left(\frac{-1/\omega C}{R}\right)$

Figure 1.54

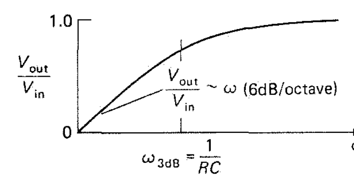


Figure 1.55. Frequency response of high-pass filter.

You can see that the output is approximately equal to the input at high frequencies (how high? $\gtrsim 1/RC$) and goes to zero at low frequencies. This is a very important result. Such a circuit is called a high-pass filter, for obvious reasons. It is very common. For instance, the input to the oscilloscope (Appendix A) can be switched to ac coupling. That's just an RC high-pass filter with the bend at about 10Hz (you would use ac coupling if you wanted to look at a small signal riding on a large dc voltage). Engineers like to refer to the −3dB "breakpoint" of a filter (or of any circuit that behaves like a filter). In the case of the simple RC high-pass filter, the −3dB breakpoint is given by

$$f_{3\text{dB}} = 1/2\pi RC$$

Note that the capacitor lets no steady current through ($f = 0$). This use as a dc *blocking capacitor* is one of its most frequent applications. Whenever you need to couple a signal from one amplifier to another, you almost invariably use a capacitor. For instance, every hi-fi audio amplifier has all its inputs capacitively coupled, because it doesn't know what dc level its input signals might be riding on. In such a coupling application you always pick R and C so that all frequencies of interest (in this case, 20Hz–20kHz) are passed without loss (attenuation).

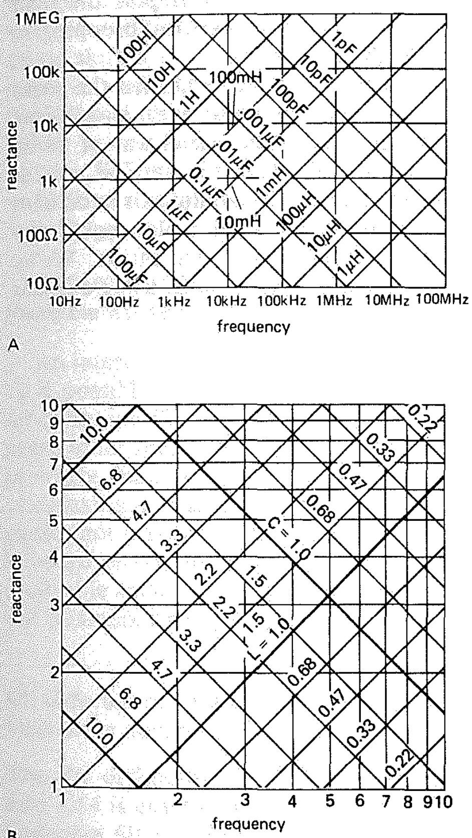


Figure 1.56. A. Reactance of inductors and capacitors versus frequency; all decades are identical, except for scale.
B. A single decade from part A expanded, with standard 20% component values shown.

You often need to know the impedance of a capacitor at a given frequency (e.g., for design of filters). Figure 1.56 provides a very useful graph covering large ranges of capacitance and frequency, giving the value of $|\mathbf{Z}| = 1/2\pi fC$.

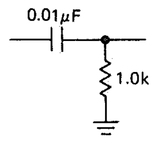


Figure 1.57

As an example, consider the filter shown in Figure 1.57. It is a high-pass filter with the 3dB point at 15.9kHz. The impedance of a load driven by it should be much larger than 1.0k in order to prevent circuit loading effects on the filter's output, and the driving source should be able to drive a 1.0k load without significant attenuation (loss of signal amplitude) in order to prevent circuit loading effects by the filter on the signal source.

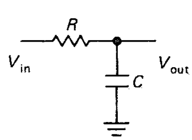


Figure 1.58. Low-pass filter.

Low-pass filters

You can get the opposite frequency behavior in a filter by interchanging R and C (Fig. 1.58). You will find

$$V_{out} = \frac{1}{(1+\omega^2R^2C^2)^{1/2}}V_{in}$$

as seen in Figure 1.59. This is called a low-pass filter. The 3dB point is again at a frequency

$$f = 1/2\pi RC$$

Low-pass filters are quite handy in real life. For instance, a low-pass filter can be used to eliminate interference from nearby radio and television stations (550kHz–800MHz), a problem that plagues audio amplifiers and other sensitive electronic equipment.

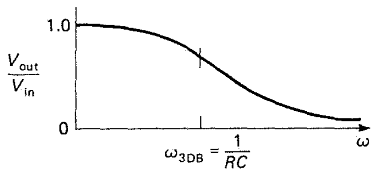


Figure 1.59. Frequency response of low-pass filter.

EXERCISE 1.21

Show that the preceding expression for the response of an RC low-pass filter is correct.

The low-pass filter's output can be viewed as a signal source in its own right. When driven by a perfect ac voltage (zero source impedance), the filter's output looks like R at low frequencies (the perfect signal source can be replaced by a short, i.e., by its small-signal source impedance, for the purpose of impedance calculations). It drops to zero impedance at high frequencies, where the capacitor dominates the output impedance. The signal driving the filter sees a load of R plus the load resistance at low frequencies, dropping to R at high frequencies.

In Figure 1.60, we've plotted the same low-pass filter response with *logarithmic* axes, which is a more usual way of doing it. You can think of the vertical axis as decibels, and the horizontal axis as octaves (or decades). On such a plot, equal distances correspond to equal ratios. We've also plotted the phase shift, using a linear

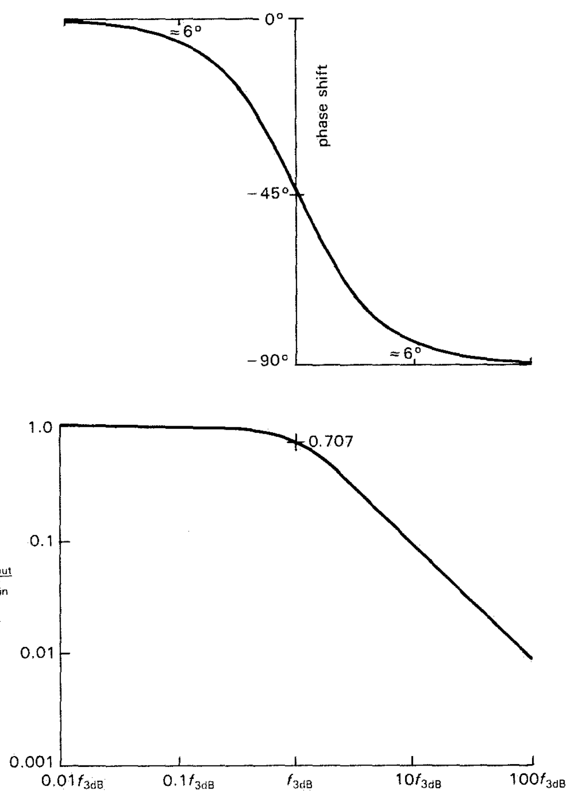


Figure 1.60. Frequency response (phase and amplitude) of low-pass filter, plotted on logarithmic axes. Note that the phase shift is 45° at the 3dB point and is within 6° of its asymptotic value for a decade of frequency change.

vertical axis (degrees) and the same logarithmic frequency axis. This sort of plot is good for seeing the detailed response even when it is greatly attenuated (as at right); we'll see a number of such plots in Chapter 5, when we treat active filters. Note that the filter curve plotted here becomes a straight line at large attenuations, with a slope of −20dB/decade (engineers prefer to say "−6dB/octave"). Note also that the phase shift goes smoothly from 0° (at frequencies well below the breakpoint) to 90° (well above it), with a value of 45° at the −3dB point. A rule of thumb for single-section RC filters is that the phase shift is $\approx 6°$ from its asymptotic value at $0.1 f_{3dB}$ and $10 f_{3dB}$.

EXERCISE 1.22
Prove the last assertion.

An interesting question is the following: Is it possible to make a filter with some arbitrary specified amplitude response and some other specified phase response? Surprisingly, the answer is no: The demands of causality (i.e., that response must follow cause, not precede it) force a relationship between phase and amplitude response of realizable analog filters (known officially as the Kramers-Kronig relation).

RC differentiators and integrators in the frequency domain

The RC differentiator that we saw in Section 1.14 is exactly the same circuit as the high-pass filter in this section. In fact, it can be considered as either, depending on whether you're thinking of waveforms in the time domain or response in the frequency domain. We can restate the earlier time-domain condition for its proper operation ($V_{out} \ll V_{in}$) in terms of the frequency response: For the output to be small compared with the input, the signal frequency (or frequencies) must be well below the 3dB point. This is easy to check. Suppose we have the input signal

$V_{in} = \sin \omega t$

Then, using the equation we obtained earlier for the differentiator output,

$$V_{out} = RC\frac{d}{dt}\sin \omega t = \omega RC \cos \omega t$$

and so $V_{out} \ll V_{in}$ if $\omega RC \ll 1$, i.e., $RC \ll 1/\omega$. If the input signal contains a range of frequencies, this must hold for the highest frequencies present in the input.

The RC integrator (Section 1.15) is the same circuit as the low-pass filter; by similar reasoning, the criterion for a good integrator is that the lowest signal frequencies must be well above the 3dB point.

Inductors versus capacitors

Inductors could be used, instead of capacitors, in combination with resistors to make low-pass (or high-pass) filters. In practice, however, you rarely see RL low- or high-pass filters. The reason is that inductors tend to be more bulky and expensive and perform less well (i.e., they depart further from the ideal) than capacitors. If you have a choice, use a capacitor. One exception to this general statement is the use of ferrite beads and chokes in high-frequency circuits. You just string a few beads here and there in the circuit; they make the wire interconnections slightly inductive, raising the impedance at very high frequencies and preventing "oscillations," without the added resistance you would get with an RC filter. An RF "choke" is an inductor, usually a few turns of wire wound on a ferrite core, used for the same purpose in RF circuits.

□ 1.20 Phasor diagrams

There's a nice graphic method that can be very helpful when trying to understand reactive circuits. Let's take an example, namely the fact that an RC filter attenuates 3dB at a frequency $f = 1/2\pi RC$, which we derived in Section 1.19. This is true for both high-pass and low-pass filters. It is easy to get a bit confused here, because at that frequency the reactance of the capacitor equals the resistance of the resistor; so you might at first expect 6dB attenuation. That is what you would get, for example, if you were to replace the capacitor by a resistor of the same impedance (recall that 6dB means half voltage). The confusion arises because the capacitor is reactive, but the matter is clarified by a phasor diagram (Fig. 1.61). The axes are the real (resistive) and imaginary (reactive) components of the impedance. In a series circuit like this, the axes also represent the (complex) voltage, because the current is the same everywhere. So for this circuit (think of it as an R–C voltage divider) the input voltage (applied across the series R–C pair) is proportional to the length of the hypotenuse, and the output voltage (across R only) is proportional to the length of the R leg of the triangle. The diagram represents the situation at the frequency where the magnitude of the capacitor's reactance equals R, i.e., $f = 1/2\pi RC$, and shows that the ratio of output voltage to input voltage is $1/\sqrt{2}$, i.e., -3dB.

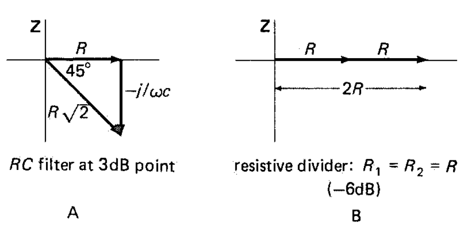


Figure 1.61

The angle between the vectors gives the phase shift from input to output. At the 3dB point, for instance, the output amplitude equals the input amplitude divided by the square root of 2, and it leads by 45° in phase. This graphic method makes it easy to read off amplitude and phase relationships in RLC circuits. For example, you can use it to get the response of the high-pass filter that we previously derived algebraically.

EXERCISE 1.23

Use a phasor diagram to derive the response of an RC high-pass filter:

$$V_{\text{out}} = \frac{R}{[R^2 + (1/\omega^2C^2)]^{1/2}} V_{\text{in}}$$

EXERCISE 1.24

At what frequency does an RC low-pass filter attenuate by 6dB (output voltage equal to half the input voltage)? What is the phase shift at that frequency?

EXERCISE 1.25

Use a phasor diagram to obtain the low-pass filter response previously derived algebraically.

In the next chapter (Section 2.08) you will see a nice example of phasor diagrams in connection with a constant-amplitude phase-shifting circuit.

1.21 "Poles" and decibels per octave

Look again at the response of the RC low-pass filter (Fig. 1.59). Far to the right of the "knee" the output amplitude is dropping proportional to $1/f$. In one octave (as in music, one octave is twice the frequency) the output amplitude will drop to half, or -6dB; so a simple RC filter has a 6dB/octave falloff. You can make filters with several RC sections; then you get 12dB/octave (two RC sections), 18dB/octave (three sections), etc. This is the usual way of describing how a filter behaves beyond the cutoff. Another popular way is to say a "3-pole filter," for instance, meaning a filter with three RC sections (or one that behaves like one). (The word "pole" derives from a method of analysis that is beyond the scope of this book and that involves complex transfer functions in the complex frequency plane, known by engineers as the "s-plane.")

A caution on multistage filters: You can't simply cascade several identical filter sections in order to get a frequency response that is the concatenation of the individual responses. The reason is that each stage will load the previous one significantly (since they're identical), changing the overall response. Remember that the response function we derived for the simple RC filters was based on a zero-impedance driving source and an infinite-impedance load. One solution is to make each successive filter section have much higher impedance than the preceding one. A better solution involves active circuits like transistor or operational amplifier (op-amp) interstage "buffers," or active filters. These subjects will be treated in Chapters 2 through 5.

1.22 Resonant circuits and active filters

When capacitors are combined with inductors or are used in special circuits called active filters, it is possible to make circuits that have very sharp frequency characteristics (e.g., a large peak in the response at a particular frequency), as compared with the gradual characteristics of the RC filters we've seen so far. These circuits find applications in various audiofrequency and radiofrequency devices. Let's now take a quick look at LC circuits (there will be more on them, and active filters, in Chapter 5 and Appendix H).

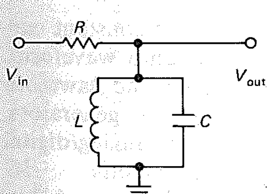


Figure 1.62. LC resonant circuit: bandpass filter.

First, consider the circuit shown in Figure 1.62. The reactance of the LC combination at frequency f is just

$$\frac{1}{\mathbf{Z}_{LC}} = \frac{1}{\mathbf{Z}_L} + \frac{1}{\mathbf{Z}_C} = \frac{1}{j\omega L} - \frac{\omega C}{j}$$
$$= j\left(\omega C - \frac{1}{\omega L}\right)$$

i.e.,

$$\mathbf{Z}_{LC} = \frac{j}{(1/\omega L) - \omega C}$$

In combination with R it forms a voltage divider; because of the opposite behaviors of inductors and capacitors, the impedance of the parallel LC goes to infinity at the *resonant frequency* $f_0 = 1/2\pi\sqrt{LC}$ (i.e., $\omega_0 = 1/\sqrt{LC}$), giving a peak in the response there. The overall response is as shown in Figure 1.63.

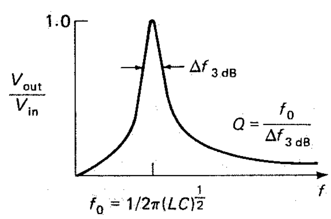


Figure 1.63

In practice, losses in the inductor and capacitor limit the sharpness of the peak, but with good design these losses can be made very small. Conversely, a Q-spoiling resistor is sometimes added intentionally to reduce the sharpness of the resonant peak. This circuit is known simply as a parallel LC resonant circuit or a tuned circuit and is used extensively in radiofrequency circuits to select a particular frequency for amplification (the L or C can be variable, so you can tune the resonant frequency). The higher the driving impedance, the sharper the peak; it is not uncommon to drive them with something approaching a current source, as you will see later. The *quality factor Q* is a measure of the sharpness of the peak. It equals the resonant frequency divided by the width at the -3dB points. For a parallel RLC circuit, $Q = \omega_0 RC$.

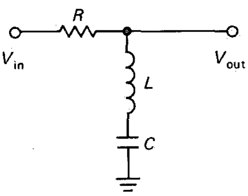


Figure 1.64. LC notch filter ("trap").

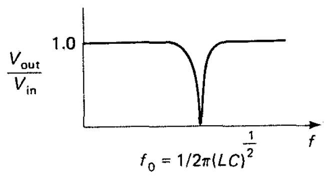


Figure 1.65

Another variety of LC circuit is the series LC (Fig. 1.64). By writing down the impedance formulas involved, you can convince yourself that the impedance of the LC goes to zero at resonance [$f_0 = 1/2\pi(LC)^{1/2}$]; such a circuit is a "trap" for signals at or near the resonant frequency, shorting them to ground. Again, this circuit finds application mainly in radiofrequency circuits. Figure 1.65 shows what the response looks like. The Q of a series RLC circuit is $Q = \omega_0 L/R$.

EXERCISE 1.26

Find the response (V_{out}/V_{in} versus frequency) for the series LC trap circuit in Figure 1.64.

1.23 Other capacitor applications

In addition to their uses in filters, resonant circuits, differentiators, and integrators, capacitors are needed for several other important applications. We will treat these in detail later in the book, mentioning them here only as a preview.

Bypassing

The impedance of a capacitor goes down with increasing frequency. This is the basis of another important application: bypassing. There are places in circuits where you want to allow a dc (or slowly varying) voltage, but don't want signals present. Placing a capacitor across that circuit element (usually a resistor) will help to kill any signals there. You choose the capacitor value so that its impedance at signal frequencies is small compared with what it is bypassing. You will see much more of this in later chapters.

Power-supply filtering

Power-supply filtering is really a form of bypassing, although we usually think of it as energy storage. The dc voltages used in electronics are usually generated from the ac line voltage by a process called *rectification* (which will be treated later in this chapter); some residue of the 60Hz input remains, and this can be reduced as much as desired by means of bypassing with suitably large capacitors. These capacitors really are large – they're the big shiny round things you see inside most electronic instruments. You will see how to design power supplies and filters later in this chapter and again in Chapter 6.

Timing and waveform generation

A capacitor supplied with a constant current charges up with a ramp waveform. This is the basis of ramp and sawtooth generators, used in function generators, oscilloscope sweep circuits, analog/digital converters, and timing circuits. RC circuits are also used for timing, and they form the basis of digital delay circuits (monostable multivibrators). These timing and waveform applications are important in many areas of electronics and will be covered in Chapters 3, 5, 8, and 9.

TABLE 1.1. DIODES

Type	$V_{R(max)}$[a] (V)	$I_{R(max)}$[b] (μA)	Continuous V_F (V)	@ I_F (mA)	Peak V_F (V)	@ I_F (A)	Reverse recovery (ns)	Capacitance (10V) (pF)	Class	Comments
PAD-1	45	1pA@20V	0.8	5	–	–	–	0.8	lowest I_R	Siliconix
FJT1100	30	0.001	–	–	1.1	0.05	–	1.2	very low I_R	1pA@5V, 10pA@15V
ID101	30	10pA@10V	0.8	1	1.1	0.03	–	0.8	very low I_R	Intersil; dual
1N3595	150	3	0.7	10	<1.0	0.2	3000	8.0	low I_R	1nA@125V
1N914	75	5	0.75	10	1.1	0.1	4	1.3	gen purp sig diode	indus std; same as 1N4148
1N6263	60	10	0.4	1	0.7	0.01	0	1.0	Schottky: low V_F	
1N3062	75	50	<1.0	20[b]	–	–	2	0.6	low cap, sig diode	1pF at 0 volts
1N4305	75	50	0.6	1	–	–	4	1.5	controlled V_F	
1N4002 }	100	50	0.9	1000	2.3	25	3500	15	1-amp rect	indus std; 7-member fam
1N4007 }	1000	50	0.9	1000	2.3	25	5000	10		
1N5819	40	10000	0.4	1000	1.1	20	–	50	pwr Schottky	lead mounted
1N5822	40	20000	0.45	3000	1.3	50	–	180	pwr Schottky	lead mounted
1N5625	400	50	1.1	5000	2.0	50	2500	45	5-amp rect	lead mounted
1N1183A	50	1000	1.1	40000	1.3	100	–	–	high curr rect	1N1183RA reverse

(a) $V_{R(max)}$ is repetitive peak reverse voltage, 25°C, 10μA leakage. (b) $I_{R(max)}$ is reverse leakage current at V_R and 100°C ambient temperature.

1.24 Thévenin's theorem generalized

When capacitors and inductors are included, Thévenin's theorem must be restated: Any two-terminal network of resistors, capacitors, inductors, and signal sources is equivalent to a single complex impedance in series with a single signal source. As before, you find the impedance and the signal source from the open-circuit output voltage and the short-circuit current.

DIODES AND DIODE CIRCUITS

1.25 Diodes

The circuit elements we've discussed so far (resistors, capacitors, and inductors) are all *linear*, meaning that a doubling of the applied signal (a voltage, say) produces a doubling of the response (a current, say). This is true even for the reactive devices (capacitors and inductors). These devices are also *passive*, meaning that they don't have a built-in source of power. And they are all two-terminal devices, which is self-explanatory.

Figure 1.66. Diode.

The diode (Fig. 1.66) is a very important and useful two-terminal passive *nonlinear* device. It has the V–I curve shown in Figure 1.67. (In keeping with the general philosophy of this book, we will not attempt to describe the solid-state physics that makes such devices possible.)

The diode's arrow (the anode terminal) points in the direction of forward current flow. For example, if the diode is in a circuit in which a current of 10mA is flowing from anode to cathode, then (from the graph) the anode is approximately 0.5 volt more positive than the cathode; this is called the "forward voltage drop." The reverse current, which is measured in the nanoamp range for a general-purpose diode (note the different scales in the graph for forward and reverse current), is almost never of any consequence until you reach the reverse breakdown voltage (also called the peak inverse voltage, PIV), typically 75 volts for a general-purpose diode like the 1N914. (Normally you never subject a diode to voltages large enough to cause reverse breakdown; the exception is the zener diode we mentioned earlier.) Frequently, also, the forward voltage drop of about 0.5 and 0.8 volt is of little concern, and the diode can be treated as a good approximation to an ideal one-way conductor. There are other important characteristics that distinguish the thousands of diode types available, e.g., maximum forward current, capacitance, leakage current, and reverse recovery time (see Table 1.1 for characteristics of some typical diodes).

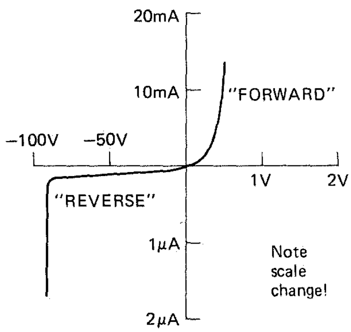


Figure 1.67. Diode V–I curve.

Before jumping into some circuits with diodes, we should point out two things: (a) A diode doesn't actually have a resistance (it doesn't obey Ohm's law). (b) If you put some diodes in a circuit, it won't have a Thévenin equivalent.

1.26 Rectification

A rectifier changes ac to dc; this is one of the simplest and most important applications of diodes (diodes are sometimes called rectifiers). The simplest circuit is shown in Figure 1.68. The "ac" symbol represents a source of ac voltage; in electronic circuits it is usually provided by a transformer, powered from the ac power line. For a sine-wave input that is much larger than the forward drop (about 0.6V for silicon diodes, the usual type), the output will look like that in Figure 1.69. If you think of the diode as a one-way conductor, you won't have any trouble understanding how the circuit works. This circuit is called a *half-wave rectifier*, because only half of the input waveform is used.

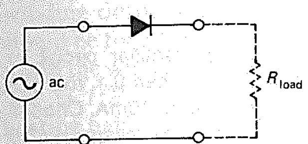


Figure 1.68. Half-wave rectifier.

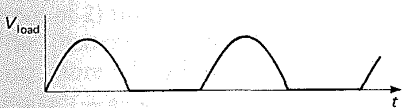


Figure 1.69

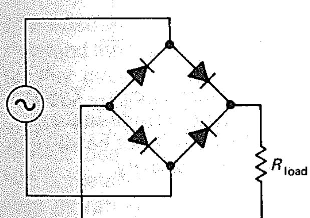


Figure 1.70. Full-wave bridge rectifier.

Figure 1.70 shows another rectifier circuit, a full-wave bridge. Figure 1.71 shows the voltage across the load for which the whole input waveform is used. The gaps at zero voltage occur because of the diodes' forward voltage drop. In this circuit, two diodes are always in series with the input; when you design low-voltage power supplies, you have to remember that.

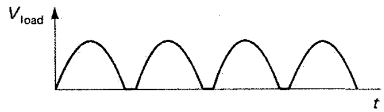


Figure 1.71

1.27 Power-supply filtering

The preceding rectified waveforms aren't good for much as they stand. They're dc only in the sense that they don't change polarity. But they still have a lot of "ripple" (periodic variations in voltage about the steady value) that has to be smoothed out in order to generate genuine dc. This we do by tacking on a low-pass filter (Fig. 1.72). Actually, the series resistor is unnecessary and is always omitted (although you sometimes see a very small resistor used to limit the peak rectifier current). The reason is that the diodes prevent flow of current back out of the capacitors, which are really serving more as energy-storage devices than as part of a classic low-pass filter. The energy stored in a capacitor is $U = \frac{1}{2}CV^2$. For C in farads and V in volts, U comes out in joules (watt-seconds).

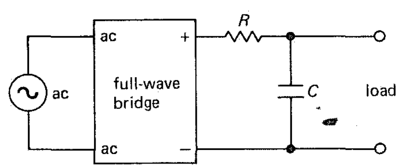


Figure 1.72

The capacitor value is chosen so that

$$R_{\text{load}}C \gg 1/f$$

(where f is the ripple frequency, here 120Hz) in order to ensure small ripple, by making the time constant for discharge much longer than the time between recharging. We will make this vague statement clearer in the next section.

Calculation of ripple voltage

It is easy to calculate the approximate ripple voltage, particularly if it is small compared with the dc (see Fig. 1.73). The load causes the capacitor to discharge somewhat between cycles (or half cycles, for full-wave rectification). If you assume that the load current stays constant (it will, for small ripple), you have

$$\Delta V = \frac{I}{C}\Delta t \qquad \left(\text{from } I = C\frac{dV}{dt}\right)$$

Just use $1/f$ (or $1/2f$ for full-wave rectification) for Δt (this estimate is a bit on the safe side, since the capacitor begins charging again in less than a half cycle). You get

$$\Delta V = \frac{I_{\text{load}}}{fC} \qquad \text{(half wave)}$$

$$\Delta V = \frac{I_{\text{load}}}{2fC} \qquad \text{(full wave)}$$

(While teaching electronics we've noticed that students love to memorize these equations! An informal poll of the authors showed that two out of two engineers don't memorize them. Please don't waste brain cells that way – instead, learn how to derive them.)

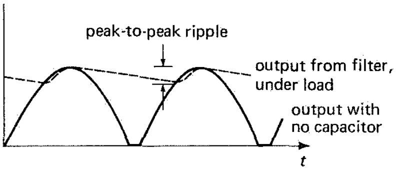


Figure 1.73. Power-supply ripple calculation.

If you wanted to do the calculation without any approximation, you would use the exact exponential discharge formula. You would be misguided in insisting on that kind of accuracy, though, for two reasons:

1. The discharge is an exponential only if the load is a resistance; many loads are not. In fact, the most common load, a *voltage regulator,* looks like a constant-current load.

2. Power supplies are built with capacitors with typical tolerances of 20% or more. Realizing the manufacturing spread, you design conservatively, allowing for the worst-case combination of component values.

In this case, viewing the initial part of the discharge as a ramp is in fact quite accurate, especially if the ripple is small, and in any case it errs in the direction of conservative design – it overestimates the ripple.

EXERCISE 1.27

Design a full-wave bridge rectifier circuit to deliver 10 volts dc with less than 0.1 volt (pp) ripple into a load drawing up to 10mA. Choose the appropriate ac input voltage, assuming 0.6 volt diode drops. Be sure to use the correct ripple frequency in your calculation.

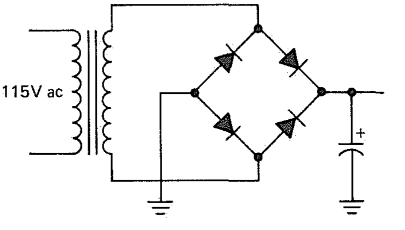


Figure 1.74. Bridge rectifier circuit. The polarity marking and curved electrode indicate a polarized capacitor, which must not be allowed to charge with the opposite polarity.

1.28 Rectifier configurations for power supplies

Full-wave bridge

A dc power supply using the bridge circuit we just discussed looks as shown in Figure 1.74. In practice, you generally buy the bridge as a prepackaged module. The smallest ones come with maximum current ratings of 1 amp average, with breakdown voltages going from 100 volts to 600 volts, or even 1000 volts. Giant bridge rectifiers are available with current ratings of 25 amps or more. Take a look at Table 6.4 for a few types.

Center-tapped full-wave rectifier

The circuit in Figure 1.75 is called a center-tapped full-wave rectifier. The output voltage is half what you get if you use a bridge rectifier. It is not the most efficient circuit in terms of transformer design, because each half of the secondary is used only half the time. Thus the current through the winding during that time is twice what it would be for a true full-wave circuit. Heating in the windings, calculated from Ohm's law, is I^2R, so you have four times the heating half the time, or twice the average heating of an equivalent full-wave bridge circuit. You would have to choose a transformer with a current rating 1.4 (square root of 2) times as large, as compared with the (better) bridge circuit; besides costing more, the resulting supply would be bulkier and heavier.

EXERCISE 1.28

This illustration of I^2R heating may help you understand the disadvantage of the center-tapped rectifier circuit. What fuse rating (minimum) is required to pass the current waveform shown in Figure 1.76, which has 1 amp average current? Hint: A fuse "blows out" by melting (I^2R heating) a metallic link, for steady currents larger than its rating. Assume for this problem that the thermal time constant of the fusible link is much longer than the time scale of the square wave, i.e., that the fuse responds to the value of I^2 averaged over many cycles.

Split supply

A popular variation of the center-tapped full-wave circuit is shown in Figure 1.77. It gives you split supplies (equal plus and minus voltages), which many circuits need. It is an efficient circuit, because both halves of the input waveform are used in each winding section.

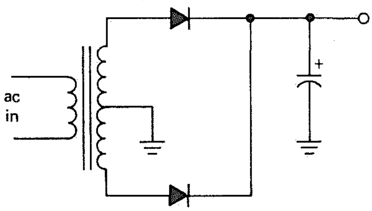


Figure 1.75. Full-wave rectifier using center-tapped transformer.

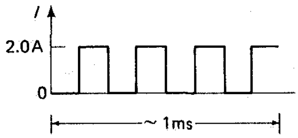


Figure 1.76

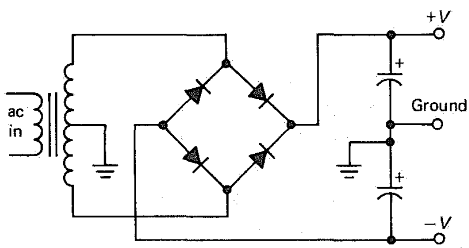


Figure 1.77. Dual-polarity (split) supply.

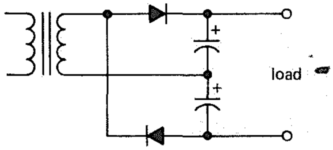


Figure 1.78. Voltage doubler.

□ *Voltage multipliers*

The circuit shown in Figure 1.78 is called a voltage doubler. Think of it as two half-wave rectifier circuits in series. It is officially a full-wave rectifier circuit, since

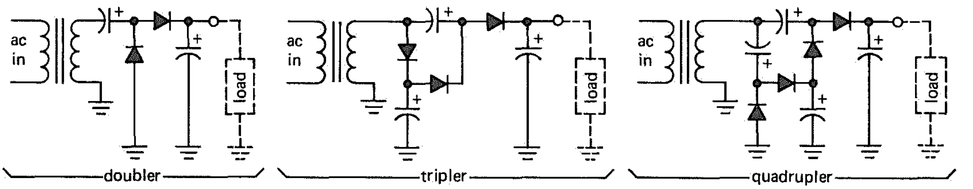


Figure 1.79. Voltage multipliers; these configurations don't require a floating voltage source.

both halves of the input waveform are used – the ripple frequency is twice the ac frequency (120Hz for the 60Hz line voltage in the United States).

Variations of this circuit exist for voltage triplers, quadruplers, etc. Figure 1.79 shows doubler, tripler, and quadrupler circuits that let you ground one side of the transformer.

1.29 Regulators

By choosing capacitors that are sufficiently large, you can reduce the ripple voltage to any desired level. This brute-force approach has two disadvantages:

1. The required capacitors may be prohibitively bulky and expensive.

2. Even with the ripple reduced to negligible levels, you still have variations of output voltage due to other causes, e.g., the dc output voltage will be roughly proportional to the ac input voltage, giving rise to fluctuations caused by input line voltage variations. In addition, changes in load current will cause the output voltage to change because of the finite internal resistances of the transformer, diode, etc. In other words, the Thévenin equivalent circuit of the dc power supply has $R > 0$.

A better approach to power-supply design is to use enough capacitance to reduce ripple to low levels (perhaps 10% of the dc voltage), then use an active *feedback circuit* to eliminate the remaining ripple. Such a feedback circuit "looks at" the output, making changes in a controllable series resistor (a transistor) as necessary to keep the output constant (Fig. 1.80).

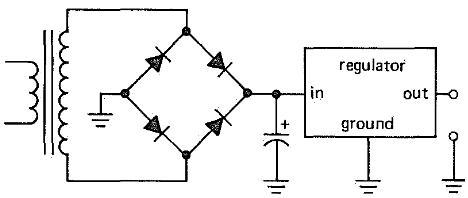


Figure 1.80. Regulated dc power supply.

These voltage regulators are used almost universally as power supplies for electronic circuits. Nowadays complete voltage regulators are available as inexpensive integrated circuits (priced under one dollar). A power supply built with a voltage regulator can be made easily adjustable and self-protecting (against short circuits, overheating, etc.), with excellent properties as a voltage source (e.g., internal resistance measured in milliohms). We will deal with regulated dc power supplies in Chapter 6.

1.30 Circuit applications of diodes

Signal rectifier

There are other occasions when you use a diode to make a waveform of one polarity only. If the input waveform isn't a sine wave, you usually don't think of it as a rectification in the sense of a power supply. For instance, you might want a train of pulses corresponding to the rising edge of a square wave. The easiest way is to rectify the differentiated wave (Fig. 1.81). Always keep in mind the 0.6 volt (approximately) forward drop of the diode. This circuit, for instance, gives no output for square waves smaller than 0.6 volt pp. If this is a problem, there are various tricks to circumvent this limitation. One possibility is to use *hot carrier diodes* (Schottky diodes), with a forward drop of about 0.25 volt (another device called a *back diode* has nearly zero forward drop, but its usefulness is limited by very low reverse breakdown voltage).

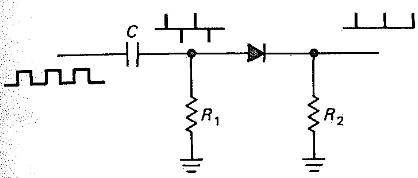


Figure 1.81

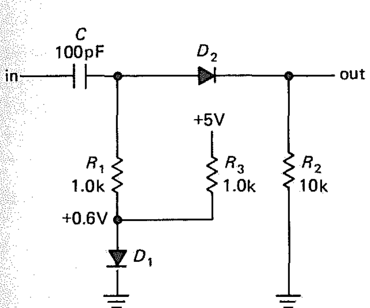


Figure 1.82. Compensating the forward voltage drop of a diode signal rectifier.

A possible circuit solution to this problem of finite diode drop is shown in Figure 1.82. Here D_1 compensates D_2's forward drop by providing 0.6 volt of *bias* to hold D_2 at the threshold of conduction. Using a diode (D_1) to provide the bias (rather than, say, a voltage divider) has several advantages: There is nothing to adjust, the compensation will be nearly perfect, and changes of the forward drop (e.g., with changing temperature) will be compensated properly. Later we will see other instances of matched-pair compensation of forward drops in diodes, transistors, and FETs: it is a simple and powerful trick.

Diode gates

Another application of diodes, which we will recognize later under the general heading of *logic*, is to pass the higher of two voltages without affecting the lower. A good example is *battery backup*, a method of keeping something running (e.g, a precision electronic clock) that must not stop when there is a power failure. Figure 1.83 shows a circuit that does the job. The battery does nothing until the power fails; then it takes over without interruption.

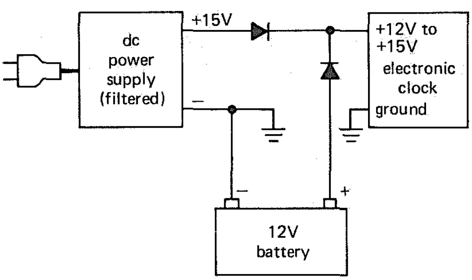


Figure 1.83. Diode OR gate: battery backup.

EXERCISE 1.29
Make a simple modification to the circuit so that the battery is charged by the dc supply (when power is on, of course) at a current of 10mA (such a circuit is necessary to maintain the battery's charge).

Diode clamps

Sometimes it is desirable to limit the range of a signal (i.e., prevent it from exceeding certain voltage limits) somewhere in a circuit. The circuit shown in Figure 1.84 will accomplish this. The diode prevents the output from exceeding about +5.6 volts, with no effect on voltages less than that (including negative voltages); the only limitation is that the input must not go so negative that the reverse breakdown voltage of the diode is exceeded (e.g., −70V for a 1N914). Diode clamps are standard equipment on all inputs in the CMOS family of digital logic. Without them, the delicate input circuits are easily destroyed by static electricity discharges during handling.

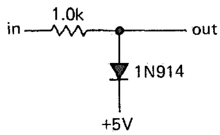


Figure 1.84. Diode voltage clamp.

EXERCISE 1.30

Design a symmetrical clamp, i.e., one that confines a signal to the range −5.6 volts to +5.6 volts.

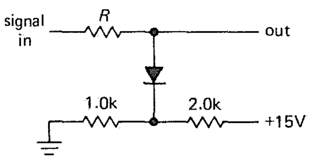


Figure 1.85

A voltage divider can provide the reference voltage for a clamp (Fig. 1.85). In this case you must ensure that the impedance looking into the voltage divider (R_{vd}) is small compared with R, because what you have looks as shown in Figure 1.86 when the voltage divider is replaced by its Thévenin equivalent circuit. When the diode conducts (input voltage exceeds clamp voltage), the output is really just the output of a voltage divider, with the Thévenin equivalent resistance of the voltage reference as the lower resistor (Fig. 1.87). So, for the values shown, the output of the clamp for a triangle-wave input would look as shown in Figure 1.88. The problem is that the voltage divider doesn't provide a stiff reference, in the language of electronics. A stiff voltage source is one that doesn't bend easily, i.e., it has low internal (Thévenin) impedance.

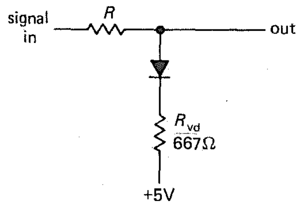


Figure 1.86

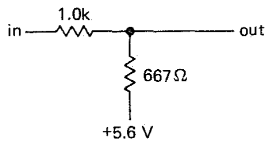


Figure 1.87

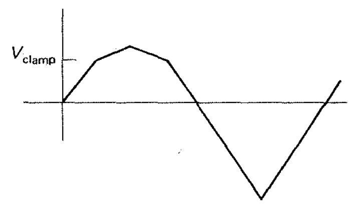


Figure 1.88

A simple way to stiffen the clamp circuit of Figure 1.85, at least for *high-frequency* signals, is to add a bypass capacitor across the 1k resistor. For example, a 15μF capacitor to ground reduces the impedance seen looking into the divider below 10 ohms for frequencies above 1kHz. (You could similarly add a bypass capacitor across D_1 in Fig. 1.82.) Of course, the effectiveness of this trick drops at low frequencies, and it does nothing at dc.

In practice, the problem of finite impedance of the voltage-divider reference can be easily solved using a transistor or operational amplifier (op-amp). This is usually a better solution than using very small resistor values, because it doesn't consume large currents, yet it provides impedances of a few ohms or less. Furthermore, there are other ways to construct a clamp, using an op-amp as part of the clamp circuit. You will see these methods in Chapter 4.

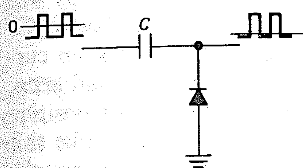


Figure 1.89. dc restoration.

One interesting clamp application is "dc restoration" of a signal that has been ac-coupled (capacitively coupled). Figure 1.89 shows the idea. This is particularly important for circuits whose inputs look like diodes (e.g., a transistor with grounded emitter); otherwise an ac-coupled signal will just fade away.

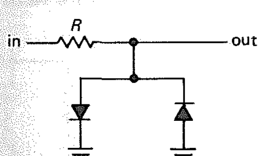


Figure 1.90. Diode limiter.

Limiter

One last clamp circuit is shown in Figure 1.90. This circuit limits the output "swing" (again, a common electronics term) to one diode drop, roughly 0.6 volt. That might seem awfully small, but if the next stage is an amplifier with large voltage amplification, its input will always be near zero volts; otherwise the output is in "saturation" (e.g., if the next stage has a gain of 1000 and operates from ±15V supplies, its input must stay in the range ±15mV in order for its output not to saturate). This clamp circuit is often used as input protection for a high-gain amplifier.

Diodes as nonlinear elements

To a good approximation the forward current through a diode is proportional to an exponential function of the voltage across it at a given temperature (for a discussion of the exact law, see Section 2.10). So you can use a diode to generate an output voltage proportional to the logarithm of a current (Fig. 1.91). Because V hovers in the region of 0.6 volt, with only small voltage changes that reflect input current variations, you can generate the input current with a resistor if the input voltage is much larger than a diode drop (Fig. 1.92).

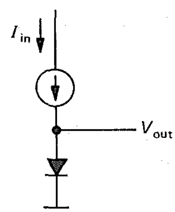


Figure 1.91. Exploiting the diode's nonlinear V–I curve: logarithmic converter.

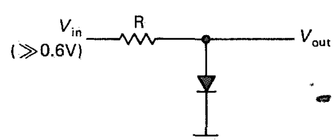


Figure 1.92

In practice, you may want an output voltage that isn't offset by the 0.6 volt diode drop. In addition, it would be nice to have a circuit that is insensitive to changes in temperature. The method of diode drop compensation is helpful here (Fig. 1.93). R_1 makes D_2 conduct, holding point A at about -0.6 volt. Point B is then near ground (making I_{in} accurately proportional to V_{in}, incidentally). As long as the two (identical) diodes are at the same temperature, there is good cancellation of the forward drops, except, of course, for the difference owing to input current through D_1, which produces the desired output. In this circuit, R_1 should be chosen so that the current through D_2 is much larger than the maximum input current, in order to keep D_2 in conduction.

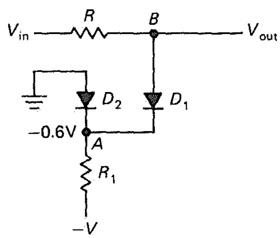


Figure 1.93. Diode drop compensation in the logarithmic converter.

In the chapter on op-amps we will examine better ways of constructing logarithmic converter circuits, along with careful methods of temperature compensation. With such methods it is possible to construct logarithmic converters accurate to a few percent over six decades or more of input current. A better understanding of diode and transistor characteristics, along with an understanding of op-amps, is necessary first. This section is meant to serve only as an introduction for things to come.

1.31 Inductive loads and diode protection

What happens if you open a switch that is providing current to an inductor? Because inductors have the property

$$V = L\frac{dI}{dt}$$

it is not possible to turn off the current suddenly, since that would imply an infinite voltage across the inductor's terminals. What happens instead is that the voltage across the inductor suddenly rises and keeps rising until it forces current to flow. Electronic devices controlling inductive loads can be easily damaged, especially the component that "breaks down" in order to satisfy the inductor's craving for continuity of current. Consider the circuit in Figure 1.94. The switch is initially closed, and current is flowing through the inductor (which might be a relay, as will be described later). When the switch is opened, the inductor "tries" to keep current flowing from A to B, as it had been. That means that terminal B goes positive relative to terminal A. In a case like this it may go 1000 volts positive before the switch contact "blows over." This shortens the life of the switch and also generates impulsive interference that may affect other circuits nearby. If the switch happens to be a transistor, it would be an understatement to say that its life is shortened; its life is ended!

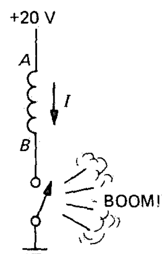


Figure 1.94. Inductive "kick."

The best solution is to put a diode across the inductor, as in Figure 1.95. When the switch is on, the diode is back-biased (from the dc drop across the inductor's winding resistance). At turn-off the diode goes into conduction, putting the switch terminal a diode drop above the positive supply voltage. The diode must be able to handle the initial diode current, which equals the steady current that had been flowing through the inductor; something like a 1N4004 is fine for nearly all cases.

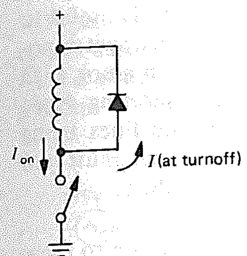


Figure 1.95. Blocking inductive kick.

The only disadvantage of this protection circuit is that it lengthens the decay of current through the inductor, since the rate of change of inductor current is proportional to the voltage across it. For applications where the current must decay quickly (high-speed impact printers, high-speed relays, etc.), it may be better to put a resistor across the inductor, choosing its value so that $V_{supply} + IR$ is less than the maximum allowed voltage across the switch. (For fastest decay with a given maximum voltage, a zener could be used instead, giving a ramp-down of current rather than an exponential decay.)

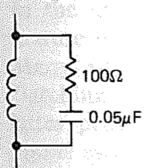


Figure 1.96. RC "snubber" for suppressing inductive kick.

For inductors driven from ac (transformers, ac relays), the diode protection just described will not work, since the diode will conduct on alternate half cycles when the switch is closed. In that case a good solution is an RC "snubber" network (Fig. 1.96). The values shown are typical for small inductive loads driven from the ac power line. Such a snubber should be included in all instruments that run from the ac power line, since a transformer is inductive. An alternative protection device is a metal-oxide varistor, or transient suppressor, an inexpensive device that looks something like a disc ceramic capacitor and behaves electrically like a bi-directional zener diode. They are available at voltage ratings from 10 to 1000 volts and can handle transient currents up to thousands of amperes (see Section 6.11 and Table 6.2). Putting a transient suppressor across the ac power-line terminals makes good sense in a piece of electronic equipment, not only to prevent inductive spike interference to other nearby instruments but also to prevent occasional large power-line spikes from damaging the instrument itself.

OTHER PASSIVE COMPONENTS

In the following sections we would like to introduce briefly an assortment of miscellaneous but essential components. If you are experienced in electronic construction, you may wish to proceed to the next chapter.

1.32 Electromechanical devices

Switches

These mundane but important devices seem to wind up in most electronic equipment. It is worth spending a few paragraphs on the subject. Figure 1.97 shows some common switch types.

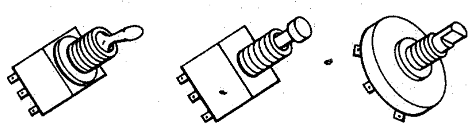


Figure 1.97. Panel switches.

Toggle switches. The simple toggle switch is available in various configurations, depending on the number of poles; Figure 1.98 shows the usual ones (SPDT indicates a single-pole double-throw switch, etc.). Toggle switches are also available with "center OFF" positions and with up to 4 poles switched simultaneously. Toggle switches are always "break before make," e.g., the moving contact never connects to both terminals in an SPDT switch.

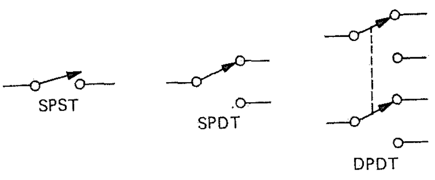


Figure 1.98. Fundamental switch types.

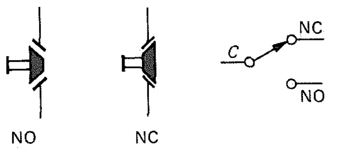


Figure 1.99. Momentary-contact (push-button) switches.

Push-button switches. Push-button switches are useful for momentary-contact applications; they are drawn schematically as shown in Figure 1.99 (NO and NC mean normally open and normally closed). For SPDT momentary-contact switches, the terminals must be labeled NO and NC, whereas for SPST types the symbol is self-explanatory. Momentary-contact switches are always "break before make." In the electrical (as opposed to electronic) industry, the terms form A, form B, and form C are used to mean SPST (NO), SPST (NC), and SPDT, respectively.

Rotary switches. Rotary switches are available with many poles and many positions, often as kits with individual wafers and shaft hardware. Both shorting (make before break) and nonshorting (break before make) types are available, and they can be mixed on the same switch. In many applications the shorting type is useful to prevent an open circuit between switch positions, because circuits can go amok with unconnected inputs. Nonshorting types are necessary if the separate lines being switched to one common line must not ever be connected to each other.

Other switch types. In addition to these basic switch types, there are available various exotic switches such as Hall-effect switches, reed switches, proximity switches, etc. All switches carry maximum current and voltage ratings; a small toggle switch might be rated at 150 volts and 5 amps. Operation with inductive loads drastically reduces switch life because of arcing during turn-off.

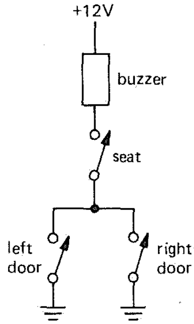


Figure 1.100

Switch examples. As an example of what can be done with simple switches, let's consider the following problem: Suppose you want to sound a warning buzzer if the driver of a car is seated and one of the car doors is open. Both doors and the driver's seat have switches, all normally open. Figure 1.100 shows a circuit that does what you want. If one OR the other door is open (switch closed) AND the seat switch is closed, the buzzer sounds. The words OR and AND are used in a logic sense here, and we will see this example again in Chapters 2 and 8 when we talk about transistors and digital logic.

Figure 1.101 shows a classic switch circuit used to turn a ceiling lamp on or off from a switch at either of two entrances to a room.

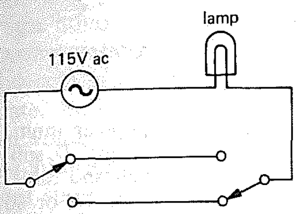


Figure 1.101. Electrician's "three-way" switch wiring.

EXERCISE 1.31
Although few electronic circuit designers know how, every *electrician* can wire up a light fixture so that any of N switches can turn it on or off. See if you can figure out this generalization of Figure 1.101. It requires two SPDT switches and $N-2$ DPDT switches. (Hint: First figure out how to use a DPDT switch to crisscross a pair of wires.)

Relays

Relays are electrically controlled switches. In the usual type, a coil pulls in an armature when sufficient coil current flows. Many varieties are available, including "latching" and "stepping" relays; the latter provided the cornerstone for telephone switching stations, and they're still popular in pinball machines. Relays are available for dc or ac excitation, and coil voltages from 5 volts up to 110 volts are common. "Mercury-wetted" and "reed" relays are intended for high-speed (~1ms) applications, and giant relays intended to switch thousands of amps are used by power companies. Many previous relay applications are now handled with transistor or FET switches, and devices known as solid-state relays are now available to handle ac switching applications. The primary uses of relays are in remote switching and high-voltage (or high-current) switching. Because it is important to keep electronic circuits electrically isolated from the ac power line, relays are useful to switch ac power while keeping the control signals electrically isolated.

Connectors

Bringing signals in and out of an instrument, routing signal and dc power around between the various parts of an instrument, providing flexibility by permitting circuit boards and larger modules of the instrument to be unplugged (and replaced) – these are the functions of the connector, an essential ingredient (and usually the most unreliable part) of any piece of electronic equipment. Connectors come in a bewildering variety of sizes and shapes.

Single-wire connectors. The simplest kind of connector is the simple pin jack or banana jack used on multimeters, power supplies, etc. It is handy and inexpensive, but not as useful as the shielded-cable or multiwire connectors you often need. The humble binding post is another form of single-wire connector, notable for the clumsiness it inspires in those who try to use it.

Shielded-cable connectors. In order to prevent capacitive pickup, and for other reasons we'll go into in Chapter 13, it is usually desirable to pipe signals around from one instrument to another in shielded coaxial cable. The most popular connector is the BNC ("baby N" connector) type that adorns most instrument front panels. It connects with a quarter-turn twist and completes both the shield (ground) circuit and inner conductor (signal) circuit simultaneously. Like all connectors used to mate a cable to an instrument, it comes in both panel-mounting and cable-terminating varieties (Fig. 1.102).

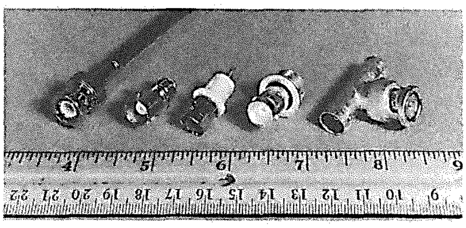


Figure 1.102. BNC connectors are the most popular type for use with shielded (coaxial) cable. From left to right: A male connector on a length of cable, a standard panel-mounted female connector, two varieties of insulated panel-mounted female connectors, and a BNC "T," a handy device to have in the laboratory.

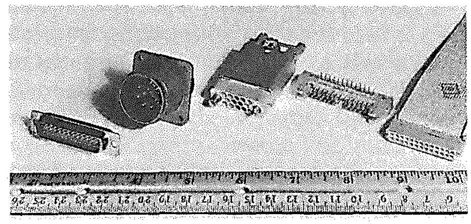


Figure 1.103. A selection of popular multipin connectors. From left to right: *D* subminiature type, available in panel- and cable-mounting versions, with 9, 15, 25, 37, or 50 pins; the venerable MS-type connector, available in many (too many!) pin and mounting configurations, including types suitable for shielded cables; a miniature rectangular connector (Winchester MRA type) with integral securing jackscrews, available in several sizes; a circuit-board-mounting mass-termination connector with its mating female ribbon connector.

Among the other connectors for use with coaxial cable are the TNC (a close cousin of the BNC, but with threaded outer shell), the high-performance but bulky type N, the miniature SMA, the subminiature LEMO and SMC, and the MHV, a high-voltage version of the standard BNC connector. The so-called phono jack used in audio equipment is a nice lesson in bad design, because the inner conductor mates before the shield when you plug it in; furthermore, the design of the connector is such that both shield and center conductor tend to make poor contact. You've undoubtedly *heard* the results! Not to be outdone, the television industry has responded with its own bad standard, the type F coax "connector," which uses the unsupported inner wire of the coax as the pin of the male plug, and a shoddy arrangement to mate the shield.

Multipin connectors. Very frequently electronic instruments demand multiwire cables and connectors. There are literally dozens of different kinds. The simplest example is a 3-wire line cord connector. Among the more popular are the excellent type D subminiature, the Winchester MRA series, the venerable MS type, and the flat ribbon-cable mass-termination connectors (Fig. 1.103).

Beware of connectors that can't tolerate being dropped on the floor (the miniature hexagon connectors are classic) or that don't provide a secure locking mechanism (e.g., the Jones 300 series).

Card-edge connectors. The most common method used to make connection to printed-circuit cards is the card-edge connector, which mates to a row of gold-plated contacts at the edge of the card. Card-edge connectors may have from 15 to 100 pins, and they come with different lug styles according to the method of connection. You can solder them to a "motherboard" or "backplane," which is itself just another printed-circuit board containing the interconnecting wiring between the individual circuit cards. Alternatively, you may want to use edge connectors with standard solder-lug terminations, particularly in a system with only a few cards (see Chapter 12 for some photographs).

1.33 Indicators

Meters

To read out the value of some voltage or current, you have a choice between the time-honored moving-pointer type of meter and digital-readout meters. The latter are more expensive and more accurate. Both types are available in a variety of voltage and current ranges. There are, in addition, exotic panel meters that read out such things as VU (volume units, an audio dB scale), expanded-scale ac volts (e.g., 105 to 130 volts), temperature (from a thermocouple), percentage motor load, frequency, etc. Digital panel meters often provide the option of logic-level outputs, in addition to the visible display, for internal use by the instrument.

Lamps and LEDs

Flashing lights, screens full of numbers and letters, eerie sounds – these are the stuff of science fiction movies, and except for the latter, they form the subject of lamps and displays (see Section 9.10). Small incandescent lamps used to be standard for front-panel indicators, but they have been replaced by light-emitting diodes (LEDs). The latter behave electrically like ordinary diodes, but with a forward voltage drop in the range of 1.5 to 2.5 volts. When current flows in the forward direction, they light up. Typically, 5mA to 20mA produces adequate brightness. LEDs are cheaper than incandescent lamps, they last forever, and they are even available in three colors (red, yellow, and green). They come in convenient panel-mounting packages; some even provide built-in current limiting.

LEDs are also used for digital displays, most often the familiar 7-segment numeric display you see in calculators. For displaying letters as well as numbers (alphanumeric display), you can get 16-segment displays or dot-matrix displays. For low power or outdoor use, liquid-crystal displays are superior.

1.34 Variable components

Resistors

Variable resistors (also called volume controls, potentiometers, pots, or trimmers) are useful as panel controls or internal adjustments in circuits. The most common panel type is known as a 2 watt type AB potentiometer; it uses the same basic material as the fixed carbon-composition resistor, with a rotatable "wiper" contact. Other panel types are available with ceramic or plastic resistance elements, with improved characteristics. Multiturn types (3, 5, or 10 turns) are available, with counting dials, for improved resolution and linearity. "Ganged" pots (several independent sections on one shaft) are also manufactured, although in limited variety, for applications that demand them.

For use inside an instrument, rather than on the front panel, *trimmer pots* come in single-turn and multiturn styles, most intended for printed-circuit mounting. These are handy for calibration adjustments of the "set-and-forget" type. Good advice: Resist the temptation to use lots of trimmers in your circuits. Use good design instead.

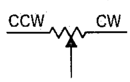


Figure 1.104. Potentiometer (3-terminal variable resistor).

The symbol for a variable resistor, or pot, is shown in Figure 1.104. Sometimes the symbols CW and CCW are used to indicate clockwise and counterclockwise.

One important point about variable resistors: Don't attempt to use a potentiometer as a substitute for a precise resistor value somewhere within a circuit. This is tempting, because you can trim the resistance to the value you want. The trouble is that potentiometers are not as stable as good (1%) resistors, and in addition they may not have good resolution (i.e., they can't be set to a precise value). If you must have a precise and settable resistor value somewhere, use a combination of a 1% (or better) precision resistor and a potentiometer, with the fixed resistor contributing most of the resistance. For example, if you need a 23.4k resistor, use a 22.6k (a 1% value) 1% fixed resistor in series with a 2k trimmer pot. Another possibility is to use a series combination of several precision resistors, choosing the last (and smallest) resistor to give the desired series resistance.

As you will see later, it is possible to use FETs as voltage-controlled variable resistors in some applications. Transistors can be used as variable-gain amplifiers, again controlled by a voltage. Keep an open mind when design brainstorming.

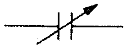


Figure 1.105. Variable capacitor.

Capacitors

Variable capacitors are primarily confined to the smaller capacitance values (up to about 1000pF) and are commonly used in radiofrequency (RF) circuits. Trimmers are available for in-circuit adjustments, in addition to the panel type for user tuning. Figure 1.105 shows the symbol for a variable capacitor.

Diodes operated with applied reverse voltage can be used as voltage-variable capacitors; in this application they're called varactors, or sometimes varicaps or epicaps. They're very important in RF applications, especially automatic frequency control (AFC), modulators, and parametric amplifiers.

Inductors

Variable inductors are usually made by arranging to move a piece of core material in a fixed coil. In this form they're available with inductances ranging from microhenrys to henrys, typically with a 2:1 tuning range for any given inductor. Also available are rotary inductors (coreless coils with a rolling contact).

Transformers

Variable transformers are very handy devices, especially the ones operated from the 115 volt ac line. They're usually "autoformers," which means that they have only one winding, with a sliding contact. They're also commonly called Variacs, and they are made by Technipower, Superior Electric, and others. Typically they provide 0 to 135 volts ac output when operated from 115 volts, and they come in current ratings from 1 amp to 20 amps or more. They're good for testing instruments that seem to be affected by powerline variations, and in any case to verify worst-case performance. Warning: Don't forget that the output is not electrically isolated from the power line, as it would be with a transformer!

ADDITIONAL EXERCISES

(1) Find the Norton equivalent circuit (a current source in parallel with a resistor) for the voltage divider in Figure 1.106. Show that the Norton equivalent gives the same output voltage as the actual circuit when loaded by a 5k resistor.

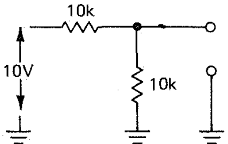


Figure 1.106

(2) Find the Thévenin equivalent for the circuit shown in Figure 1.107. Is it the

same as the Thévenin equivalent for exercise 1?

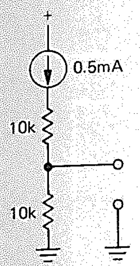


Figure 1.107

(3) Design a "rumble filter" for audio. It should pass frequencies greater than 20Hz (set the −3dB point at 10Hz). Assume zero source impedance (perfect voltage source) and 10k (minimum) load impedance (that's important so that you can choose R and C such that the load doesn't affect the filter operation significantly).

(4) Design a "scratch filter" for audio signals (3dB down at 10kHz). Use the same source and load impedances as in exercise 3.

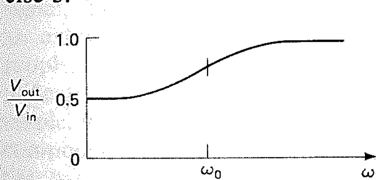


Figure 1.108

(5) How would you make a filter with Rs and Cs to give the response shown in Figure 1.108?

(6) Design a bandpass RC filter (as in Fig. 1.109); f_1 and f_2 are the 3dB points. Choose impedances so that the first stage isn't much affected by the loading of the second stage.

(7) Sketch the output for the circuit shown in Figure 1.110.

(8) Design an oscilloscope "×10 probe"

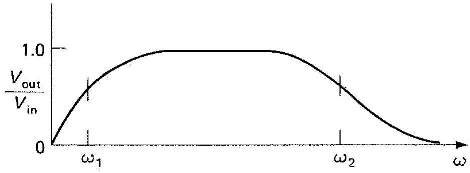


Figure 1.109

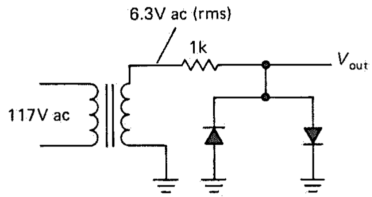


Figure 1.110

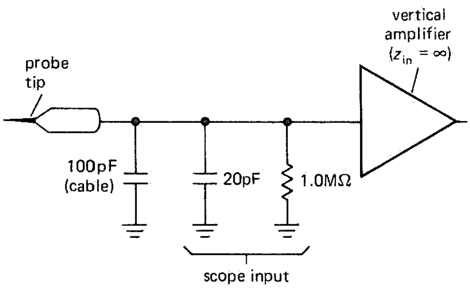


Figure 1.111

(see Appendix A) to use with a scope whose input impedance is 1MΩ in parallel with 20pF. Assume that the probe cable adds an additional 100pF and that the probe components are placed at the tip end (rather than at the scope end) of the cable (Fig. 1.111). The resultant network should have 20dB (×10) attenuation at all frequencies, including dc. The reason for using a ×10 probe is to increase the load impedance seen by the circuit under test, which reduces loading effects. What input impedance (R in parallel with C) does your ×10 probe present to the circuit under test, when used with the scope?

VOLTAGE REGULATORS AND POWER CIRCUITS

CHAPTER 6

Nearly all electronic circuits, from simple transistor and op-amp circuits up to elaborate digital and microprocessor systems, require one or more sources of stable dc voltage. The simple transformer-bridge-capacitor unregulated power supplies we discussed in Chapter 1 are not generally adequate because their output voltages change with load current and line voltage and because they have significant amounts of 120Hz ripple. Fortunately, it is easy to construct stable power supplies using negative feedback to compare the dc output voltage with a stable voltage reference. Such regulated supplies are in universal use and can be simply constructed with integrated circuit voltage regulator chips, requiring only a source of unregulated dc input (from a transformer-rectifier-capacitor combination, a battery, or some other source of dc input) and a few other components.

In this chapter you will see how to construct voltage regulators using special-purpose integrated circuits. The same circuit techniques can be used to make regulators with discrete components (transistors, resistors, etc.), but because of the availability of inexpensive high-performance regulator chips, there is no advantage to using discrete components in new designs. Voltage regulators get us into the domain of high power dissipation, so we will be talking about heat sinking and techniques like "foldback limiting" to limit transistor operating temperatures and prevent circuit damage. These techniques can be used for all sorts of power circuits, including power amplifiers. With the knowledge of regulators you will have at that point, we will be able to go back and discuss the design of the unregulated supply in some detail. In this chapter we will also look at voltage references and voltage-reference ICs, devices with uses outside of power-supply design.

BASIC REGULATOR CIRCUITS WITH THE CLASSIC 723

6.01 The 723 regulator

The μA723 voltage regulator is a classic. Designed by Bob Widlar and first introduced in 1967, it is a flexible, easy-to-use regulator with excellent performance.

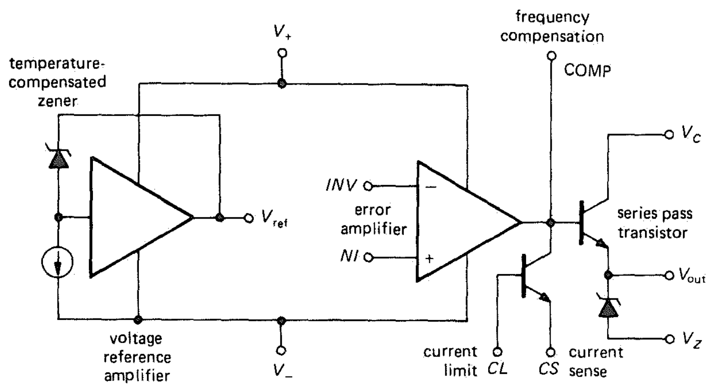


Figure 6.1. Simplified circuit of the 723 regulator. (Courtesy of Fairchild Camera and Instrument Corp.)

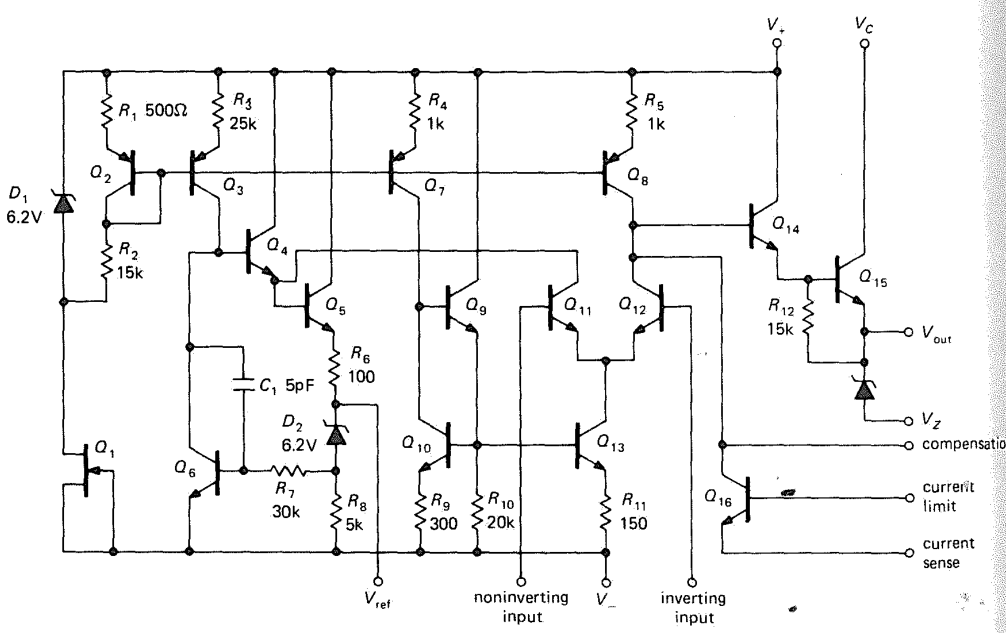


Figure 6.2. Schematic of the 723 regulator. (Courtesy of Fairchild Camera and Instrument Corp.)

Although you would not choose it for a new design nowadays, it is worth looking at in some detail, since more recent regulators work on the same principles. Its circuit is shown in Figures 6.1 and 6.2. As you can see, it is really a power-supply kit, containing a temperature-compensated voltage reference, differential amplifier, series pass transistor, and current-limiting protective circuit. As it comes, the 723 doesn't regulate anything. You have to hook up an external circuit to make it do what you want. Before going on to design regulators with it, let's look briefly at its internal circuit. It is straightforward and easy to understand (the innards of many ICs aren't).

The heart of the regulator is the temperature-compensated zener reference. Zener D_2 has a positive temperature coefficient, so its voltage is added to Q_6's base-emitter drop (remember, V_{BE} has a negative temperature coefficient of roughly $-2\text{mV}/^\circ\text{C}$) to form a voltage reference (nominally 7.15V) of nearly zero temperature coefficient (typically 0.003%/°C). Q_4 through Q_6 are arranged to bias D_2 at $I = V_{BE}/R_8$ via negative feedback at dc, as indicated on the block diagram. Q_2 and Q_3 form an unsymmetrical current mirror to bias the reference; current to the mirror is set by D_1 and R_2 (their junction is fixed at 6.2V below V_+), which in turn is biased by Q_1 (the FET behaves roughly like a current source).

Q_{11} and Q_{12} form the differential amplifier (sometimes called the "error amplifier," thinking of the whole thing as an exercise in negative feedback), a classic long-tailed pair with emitter current source Q_{13}. The latter is half of a current mirror (Q_9, Q_{10}, and Q_{13}), driven in turn from current mirror Q_7 (Q_3, Q_7, and Q_8 all mirror the current generated by the D_1 reference, as we mentioned in Section 2.14). Q_{11}'s collector is tied to the fixed positive voltage at Q_4's emitter, and the error amplifier's output is taken from Q_{12}'s collector. Current mirror Q_8 supplies the latter's collector load. Q_{14} drives the pass transistor Q_{15}, in a not-quite-Darlington connection. Note that Q_{15}'s collector is brought out separately, to allow for separate positive supplies. By turning on Q_{16} you cut off drive to the pass transistors; this is used to limit output currents to nondestructive levels. Unlike many of the newer regulators, the 723 does not incorporate internal shut-down circuitry to protect against excessive load current or chip dissipation. The SG3532 and LAS1000 are improved 723-type regulators, with low-voltage bandgap reference (Section 6.15), internal current limiting, and thermal-overload shutdown circuitry.

6.02 Positive regulator

Figure 6.3 shows how to make a positive voltage regulator with the 723. All the components except the four resistors and the two capacitors are contained on the 723. Voltage divider R_1R_2 compares a fraction of the output with the voltage reference, and the 723 components do the rest; this circuit is identical with the op-amp noninverting amplifier with emitter follower, with V_{ref} as the "input." R_4 is chosen for about 0.5 volt drop at maximum desired output current, since a V_{BE} drop applied across the CL-CS inputs will turn on the current-limiting transistor (Q_{16} in Fig. 6.2), shutting off base drive to the output pass transistor. The 100pF capacitor stabilizes the loop. R_3 (sometimes omitted) is chosen so that the differential amplifier sees equal impedances at its inputs. This makes the output insensitive to changes in bias current (with changes in temperature, say), in the same way as we saw with op-amps (Section 4.12).

With this circuit, a regulated supply with output voltage ranging from V_{ref} to the maximum allowable output voltage (37V) can be made. Of course, the input voltage must stay a few volts more positive than the output at all times, including the effects of ripple on the unregulated supply. The "dropout voltage" (the amount by which the input voltage must exceed the regulated output voltage) is specified as 3 volts (minimum) for the 723, a value typical of most regulators. R_1 or R_2 is usually made adjustable, or trimmable, so the output voltage can be set precisely. The production spread in V_{ref} is 6.8 to 7.5 volts.

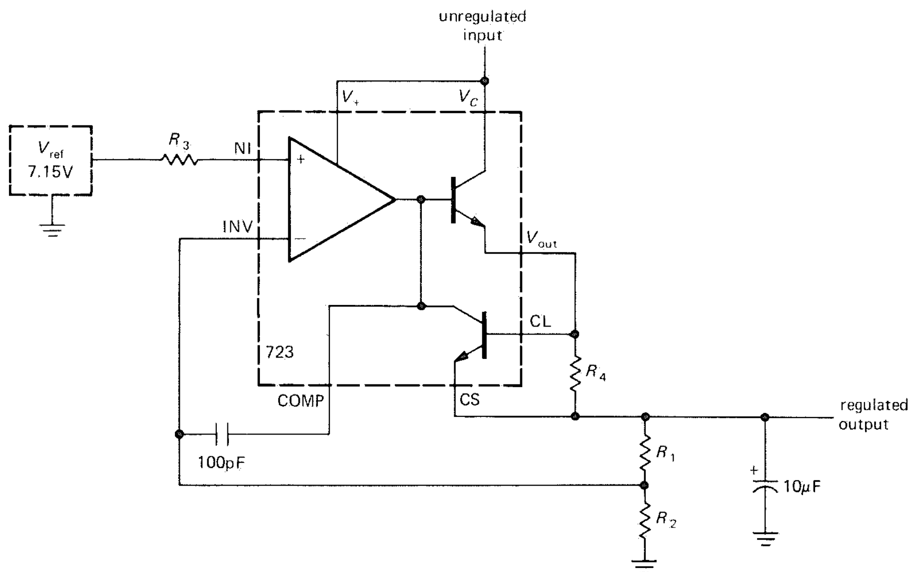


Figure 6.3. 723 regulator: $V_{out} > V_{ref}$.

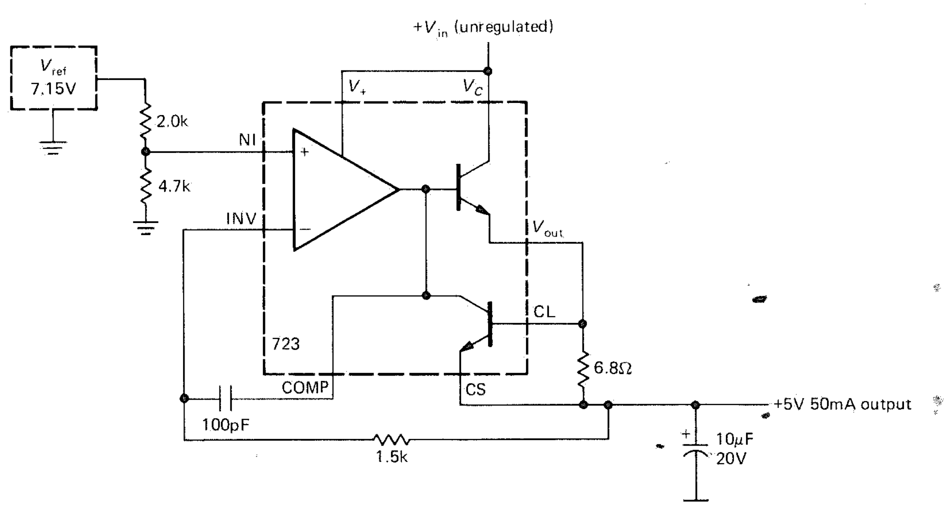


Figure 6.4. 723 regulator: $V_{out} < V_{ref}$.

It is usually a good idea to put a capacitor of a few microfarads across the output, as shown. This keeps the output impedance low even at high frequencies, where the feedback becomes less effective. It is best to use the output capacitor value recommended on the specification sheet, since oscillations can occur otherwise. In general, it is a good idea to bypass power-supply leads to ground liberally throughout a circuit, using a combination of ceramic types (0.01–0.1μF) and electrolytic or tantalum types (1–10μF).

For output voltages less than V_{ref}, you just put the voltage divider on the reference (Fig. 6.4). Now the full output voltage is compared with a fraction of the reference. The values shown are for +5 volts 50mA max. With this circuit configuration, output voltages from +2 volts to V_{ref} can be produced. The output cannot be adjusted down to zero volts because the differential amplifier will not operate below 2 volts input. This is given as a manufacturer's specification (see Table 6.9). With this circuit the unregulated input voltage must never drop below +9.5 volts, the voltage necessary to power the reference.

A third variation of this circuit is necessary if you want a regulator that is continuously adjustable through a range of output voltages around V_{ref}. In such cases, just compare a divided fraction of the output with a fraction of V_{ref} chosen to be less than the minimum output voltage desired.

EXERCISE 6.1

Design a regulator to deliver up to 50mA load current over an output voltage range of +5 to +10 volts, using a 723. Hint: Compare a fraction of the output voltage with $0.5V_{\text{ref}}$.

6.03 High-current regulator

The internal pass transistor in the 723 is rated at 150mA maximum; in addition, the power dissipation must not exceed 1 watt at 25°C (less at higher ambient temperatures; the 723 must be "derated" at 8.3mW/°C above 25°C in order to keep the junction temperature within safe limits). Thus, for instance, a 5 volt regulator with +15 volts input cannot deliver more than about 80mA to the load. To construct a higher-current supply, an external pass transistor must be used. It is easy to add one as a Darlington pair with the internal transistor (Fig. 6.5). Q_1 is the external pass transistor; it must be mounted on a *heat sink*, most often a finned metal plate designed to carry off heat (alternatively, the transistor can be mounted to one wall of the metal chassis housing the power supply). We will deal with thermal problems like these in the next section. A trimmer potentiometer has been used so that the output can be set accurately to +5 volts; its range of adjustment should be sufficient to allow for resistor tolerances as well as the maximum specified spread in V_{ref} (this is an example of worst-case design), and in this case it allows about ±1 volt adjustment from the nominal output voltage. Note the low-resistance high-power current-limiting resistor necessary for a 2 amp supply.

Pass transistor dropout voltage

One problem with this circuit is the high power dissipation in the pass transistor (at least 10W at full load current). This is unavoidable if the regulator chip is powered by the unregulated input, since it needs a few volts of "headroom" to operate (specified by the dropout voltage). With the use of a separate low-current supply for the 723 (e.g., +12V), the minimum unregulated input to the external pass transistor can be only a volt or so above the regulated output voltage (although you will always have to allow at least a few volts, since worst-case design dictates proper operation even at 105V ac line input).

Overvoltage protection

Also shown in this circuit is an *overvoltage crowbar* protection circuit consisting of

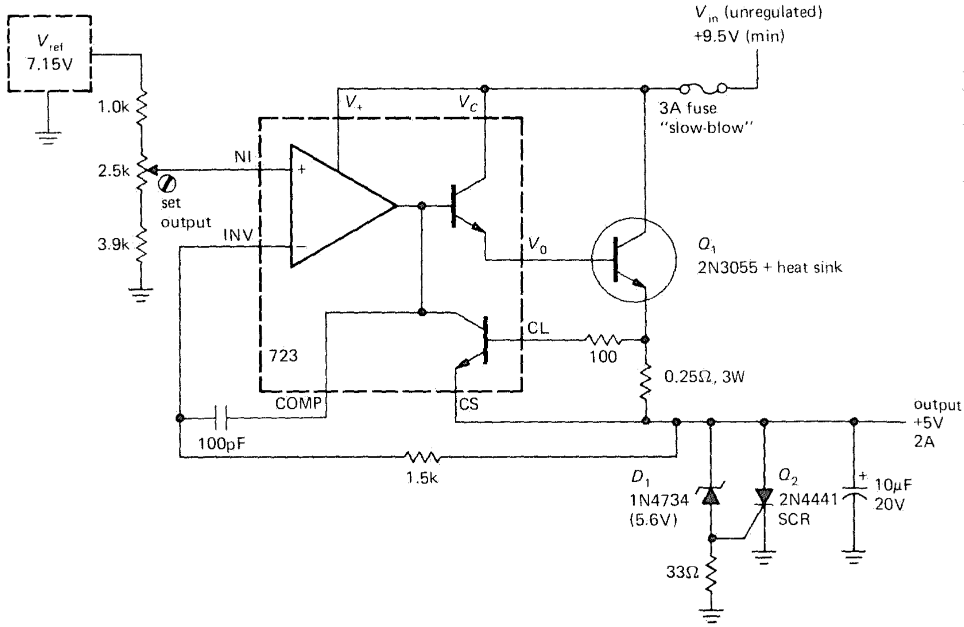


Figure 6.5. Five volt regulator with outboard pass transistor and crowbar.

D_1, Q_2, and the 33 ohm resistor. Its function is to short the output if some circuit fault causes the output voltage to exceed about 6.2 volts (this could happen if one of the resistors in the divider were to open up, for instance, or if some component in the 723 were to fail). Q_2 is an SCR (silicon-controlled rectifier), a device that is normally nonconducting but that goes into saturation when the gate-cathode junction is forward-biased. Once turned on, it will not turn off again until anode current is removed externally. In this case, gate current flows when the output exceeds D_1's zener voltage plus a diode drop. When that happens, the regulator will go into a current-limiting condition, with the output held near ground by the SCR. If the failure that produces the abnormally high output also disables the current-limiting circuit (e.g., a collector-to-emitter short in Q_1), then the crowbar will sink a very large current. For this reason it is a good idea to include a fuse somewhere in the power supply, as shown. We will treat overvoltage crowbar circuits in more detail in Section 6.06.

HEAT AND POWER DESIGN

6.04 Power transistors and heat sinking

As in the preceding circuit, it is often necessary to use power transistors or other high-current devices like SCRs or power rectifiers that can dissipate many watts. The 2N3055, an inexpensive power transistor of great popularity, can dissipate as much as 115 watts if properly mounted. All power devices are packaged in cases that permit contact between a metal surface and an external heat sink. In most cases the metal surface of the device is electrically connected to one terminal (e.g., for power transistors the case is always connected to the collector).

The whole point of heat sinking is to keep the transistor junction (or the junction of some other device) below some maximum specified operating temperature. For silicon transistors in metal packages the maximum junction temperature is usually 200°C, whereas for transistors in plastic packages it is usually 150°C. Table 6.1 lists some useful power transistors, along with their thermal properties. Heat sink design is then simple: Knowing the maximum power the device will dissipate in a given circuit, you calculate the junction temperature, allowing for the effects of heat conductivity in the transistor, heat sink, etc., and the maximum ambient temperature in which the circuit is expected to operate. You then choose a heat sink large enough to keep the junction temperature well below the maximum specified by the manufacturer. It is wise to be conservative in heat sink design, since transistor life drops rapidly at operating temperatures near or above maximum.

Thermal resistance

To carry out heat sink calculations, you use *thermal resistance*, θ, defined as heat rise (in degrees) divided by power transferred. For heat transferred entirely by conduction, the thermal resistance is a constant, independent of temperature, that depends only on the mechanical properties of the joint. For a succession of thermal joints in "series," the total thermal resistance is the sum of the thermal resistances of the individual joints. Thus, for a transistor mounted on a heat sink, the total thermal resistance from transistor junction to the outside (ambient) world is the sum of the thermal resistance from junction to case θ_{JC}, the thermal resistance from case to heat sink, θ_{CS}, and the thermal resistance from heat sink to ambient θ_{SA}. The temperature of the junction is therefore

$$T_J = T_A + (\theta_{JC} + \theta_{CS} + \theta_{SA})P$$

where P is the power being dissipated.

Let's take an example. The preceding power-supply circuit, with external pass transistor, has a maximum transistor dissipation of 20 watts for an unregulated input of +15 volts (10V drop, 2A). Let's assume that the power supply is to operate at ambient temperatures up to 50°C, not unreasonable for electronic equipment packaged together in close quarters. And let's try to keep the junction temperature below 150°C, well below its specified maximum of 200°C. The thermal resistance from junction to case is 1.5°C per watt. A TO-3 power transistor package mounted with an insulating washer and heat-conducting compound has a thermal resistance from case to heat sink of about 0.3°C per watt. Finally, a Wakefield model 641 heat sink (Fig. 6.6) has a thermal resistance from sink to ambient of about 2.3°C per watt. So the total thermal resistance from junction to ambient is about 4.1°C per watt. At 20 watts dissipation the junction will be 84°C above ambient, or 134°C (at maximum ambient temperature) in this example. The chosen heat sink will be adequate; in fact, a smaller one could be used if necessary to save space.

Comments on heat sinks

1. Where very high power dissipation (several hundred watts, say) is involved, forced air cooling may be necessary. Large heat sinks designed to be used with a blower are available with thermal resistances (sink to ambient) as small as 0.05°C to 0.2°C per watt.

2. When the transistor must be insulated from the heat sink, as is usually necessary (especially if several transistors are mounted on the same sink), a thin insulating washer is used between the transistor and sink, and insulating bushings are used around the mounting screws. Washers are available in standard

TABLE 6.1. SELECTED BIPOLAR POWER TRANSISTORS

npn	pnp	Pkg[a]	V_{CEO} max (V)	I_C max (A)	h_{FE} typ	@ I_C (A)	f_T min (MHz)	C_{cb}[b] typ (pF)	P_{diss} (T_C=25°C) (W)	Θ_{JC} (°C/W)	T_J max (°C)	Comments
Regular power: V_{CE}(sat) = 0.4V (typ); V_{BE}(on) = 0.8V (typ)												
2N5191	**2N5194**	A	60	4	100	0.2	2	80	40	3.1	150	low cost, gen purp
2N5979	**2N5976**	B	80	5	50	0.5	2	60	70	1.8	150	
2N3055	**MJ2955**	TO-3	60	15	50	2	2.5	125	115	1.5	200	metal, indus std
MJE3055	**MJE2955**	B	60	10	50	2	2.5	125	90	1.4	150	plastic, indus std
2N5886	**2N5884**	TO-3	80	25	50	10	4	400	200	0.9	200	
2N5686	**2N5684**	TO-3	80	50	30	25	2	700	300	0.6	200	for real power jobs
2N6338	**2N6437**	TO-3	100	25	50	8	40	200	200	0.9	200	premium audio
2N6275	**2N6379**	TO-3	120	50	50	20	30	400	250	0.7	200	premium audio
Darlington power: V_{CE}(sat) = 0.8V (typ); V_{BE}(on) = 1.4V (typ)												
2N6038	**2N6035**	A	60	4	2000	2	–	30	40	3.1	150	low cost
2N6044	**2N6041**	B	80	8	2500	4	4	80	75	1.7	150	
2N6059	**2N6052**	TO-3	100	12	3500	5	4	100	150	1.2	200	
2N6284	**2N6287**	TO-3	100	20	3000	10	4	150	160	1.1	200	high current

(a) A: small plastic pwr pkg (TO-126). B: large plastic pwr pkg (TO-127). (b) C_{cb} (npn) at V_{CB}=10V; C_{cb} (pnp) ≈ $2C_{cb}$ (npn).

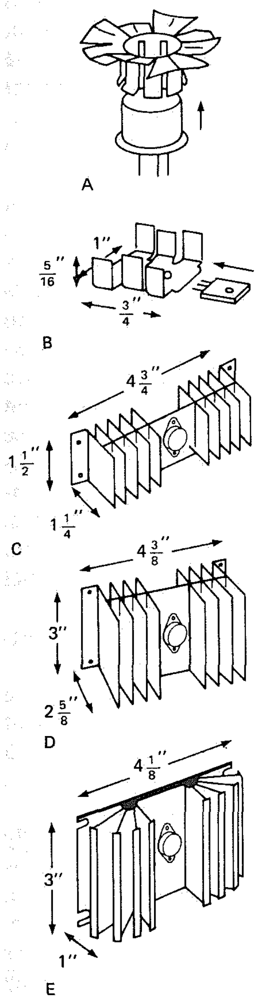
transistor-shape cutouts made from mica, insulated aluminum, or beryllia (BeO). Used with heat-conducting grease, these add from 0.14°C per watt (beryllia) to about 0.5°C per watt.

An attractive alternative to the classic mica-washer-plus-grease is provided by greaseless silicone-based insulators that are loaded with a dispersion of thermally conductive compound, usually boron nitride or aluminum oxide. They're clean and dry, and easy to use; you don't get white slimy stuff all over your hands, your electronic device, and your clothes. You save lots of time. They have thermal resistances of about 0.2–0.4°C per watt, comparable to values with the messy method. Bergquist calls its product "Sil-Pad," Chomerics calls its "Cho-Therm," SPC calls it "Koolex," and Thermalloy calls its "Thermasil." We've been using these insulators, and we like them.

3. Small heat sinks are available that simply clip over the small transistor packages (like the standard TO-5). In situations of relatively low power dissipation (a watt or two) this often suffices, avoiding the nuisance of mounting the transistor remotely on a heat sink with its leads brought back to the circuit. An example is shown in Figure 6.6. In addition, there are various small heat sinks intended for use with the plastic power packages (many regulators, as well as power transistors, come in this package) that mount right on a printed-circuit board underneath the package. These are very handy in situations of a few watts dissipation; a typical unit is illustrated in Figure 6.6.

4. Sometimes it may be convenient to mount power transistors directly to the chassis or case of the instrument. In such cases it is wise to use conservative design (keep it cool), especially since a hot case will subject the other circuit components to high temperatures and shorten component life.

5. If a transistor is mounted to a heat sink without insulating hardware, the heat sink must be insulated from the chassis.



style	part number	thermal resistance °C/W @ ΔT ($T_{sink} - T_{ambient}$) $\Delta T = 25°C$	$\Delta T = 50°C$	$\Delta T = 75°C$
A	I TXBF-032-025B	70	70	70
B	I PA2	30	27	
	T 6107	22	18	16
C	I E1000-03 T 6401 W 401	3.5	3.1	2.8
D	I E2000-06 T 6421 W 421	1.3	1.1	1.0
E	T 6169 W 641	2.6	2.2	1.9

Figure 6.6. Power transistor heat sinks. I, IERC; T, Thermalloy; W, Wakefield.

The use of insulating washers (e.g., Wakefield model 103) is recommended (unless, of course, the transistor case happens to be at ground). When the transistor is insulated from the sink, the heat sink may be attached directly to the chassis. But if the transistor is accessible from outside the instrument (e.g., if the heat sink is mounted externally on the rear wall of the box), it is a good idea to use an insulating cover over the transistor (e.g., Thermalloy 8903N) to prevent someone from accidentally coming in contact with it, or shorting it to ground.

6. The thermal resistance from heat sink to ambient is usually specified for the sink mounted with the fins vertical and with unobstructed flow of air. If the sink is mounted differently, or if the air flow is obstructed, the efficiency will be reduced (higher thermal resistance); usually it is best to mount it on the rear of the instrument with fins vertical.

EXERCISE 6.2

A 2N5320, with a thermal resistance from junction to case of 17.5°C per watt, is fitted with an IERC TXBF slip-on heat sink of the type shown in Figure 6.6. The maximum permissible junction temperature is 200°C. How much power can you dissipate with this combination at 25°C ambient temperature? How much must the dissipation be decreased per degree rise in ambient temperature?

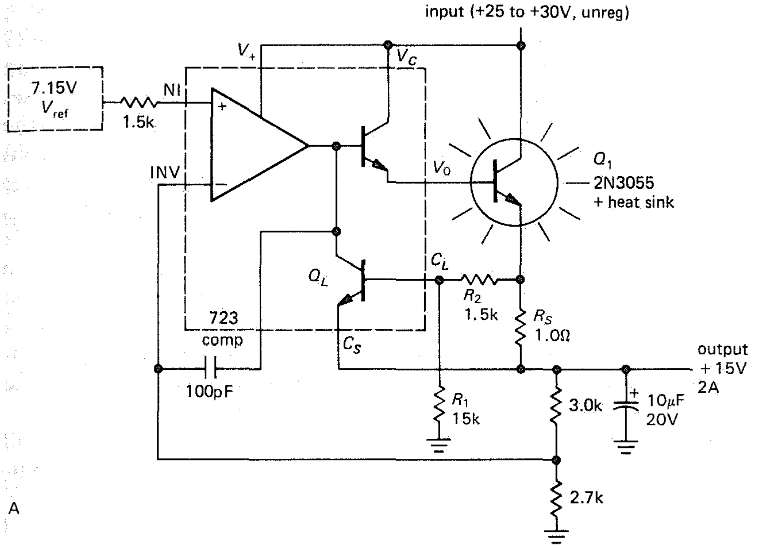
□ 6.05 Foldback current limiting

For a regulator with simple current limiting, transistor dissipation is maximum when the output is shorted to ground (either accidentally or through some circuit malfunction), and it usually exceeds the maximum value of dissipation that would otherwise occur under normal load conditions. For instance, the pass transistor in the preceding +5 volt 2 amp regulator circuit will dissipate 30 watts with the output shorted (+15V input, current limit at 2A), whereas the worst-case dissipation under normal load conditions is 20 watts (10V drop at 2A). The situation is even worse in circuits in which the voltage normally dropped by the pass transistor is a smaller fraction of the output voltage. For instance, in a +15 volt 2 amp regulated supply with +25 volt unregulated input, the transistor dissipation rises from 20 watts (full load) to 50 watts (short circuit).

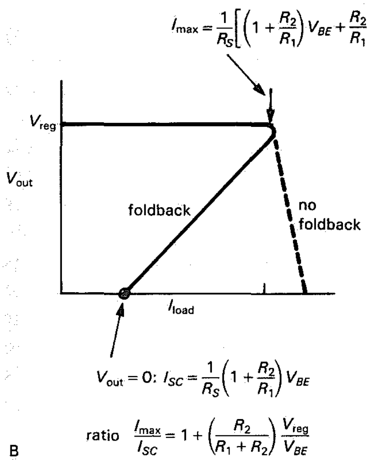
You get into a similar problem with push-pull power amplifiers. Under normal conditions you have maximum load current when the voltage across the transistors is minimum (near the extremes of output swing), and you have maximum voltage across the transistors when the current is nearly zero (zero output voltage). With a short-circuit load, on the other hand, you have maximum load current at the worst possible time, namely, with full supply voltage across the transistor. This results in much higher transistor dissipation than normal.

The brute-force solution to this problem is to use massive heat sinks and transistors of higher power rating (and safe operating area, see Section 6.07) than necessary. Even so, it isn't a good idea to have large currents flowing into the powered circuit under fault conditions, since other components in the circuit may then be damaged. The best solution is to use *foldback* current limiting, a circuit technique that reduces the output current under short-circuit or overload conditions. Figure 6.7 shows the basic configuration, again illustrated with a 723 with external pass transistor.

The divider at the base of the current-limiting transistor Q_L provides the foldback. At +15 volts output (the normal value) the circuit will limit at about 2 amps, since Q_L's base is then at +15.5 volts while its emitter is at +15 (V_{BE} is about 0.5V at the elevated temperatures at which regulator chips are normally run). But the short-circuit current is less; with the output shorted to ground, the output current is about 0.5 amp, holding Q_1's dissipation down to less than in the full-load case. This is highly desirable, since excessive heat sinking is not now required, and the thermal design need only satisfy the full-load requirements. The choice of the three resistors in the current-limiting circuit sets the short-circuit current, for a given full-load current limit. Warning: Use care in choosing the short-circuit current, since it is possible to be overzealous and design a supply that will not "start up" into a normal load. The short-circuit current should not be too small; as a



A



B

Figure 6.7. A. Power regulator with foldback current limiting. B. Output voltage versus load current.

rough guide, the short-circuit current limit should be set at about one-third the maximum load current at full output voltage.

EXERCISE 6.3

Design a 723 regulator with outboard pass transistor and foldback current limiting to provide up to 1.0 amp when the output is at its regulated value of +5.0 volts, but only 0.4 amp into a short-circuit load.

6.06 Overvoltage crowbars

As we remarked in Section 6.03, it is often a good idea to include some sort of overvoltage protection at the output of a regulated supply. Take, for instance, a +5 volt supply used to power a large digital system (you'll see examples beginning in Chapter 8). The input to the regulator is probably in the range of +10 to +15 volts. If the series pass transistor fails by shorting its collector to emitter (a common failure mode), the full unregulated voltage will be applied to the circuit, with devastating results. Although a fuse probably will blow, what's involved is a race between the fuse and the "silicon fuse" that is constituted by the rest of the circuit; the rest of the circuit will probably respond first! This problem is most serious with TTL logic, which operates from a +5 volt supply, but cannot tolerate more than +7 volts without damage. Another situation with considerable disaster potential arises when you operate something from a wide-range "bench" supply, where the unregulated input may be 40 volts or more, regardless of the output voltage.

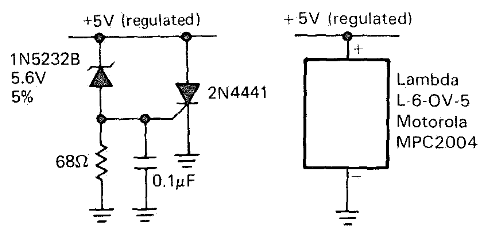


Figure 6.8. Overvoltage crowbars.

□ *Zener sensing*

Figure 6.8 shows a popular crowbar circuit and a crowbar module. You hook the circuit between the regulated output terminal and ground. If the voltage exceeds the zener voltage plus a diode drop (about 6.2V for the zener shown), the SCR is turned on, and it remains in a conducting state until its anode current drops below a few milliamps. An inexpensive SCR like the 2N4441 can sink 5 amps continuously and withstand 80 amp surge currents; its voltage drop in the conducting state is typically 1.0 volt at 5 amps. The 68 ohm resistor is provided to generate a reasonable zener current (10mA) at SCR turn-on, and the capacitor is added to prevent crowbar triggering on harmless short spikes.

The preceding circuit, like all crowbars, puts an unrelenting 1 volt "short circuit" across the supply when triggered by an overvoltage condition, and it can be reset only by turning off the supply. Since the SCR maintains a low voltage while conducting, there isn't much problem with the crowbar itself failing from overheating. As a result, it is a reliable crowbar circuit. It is essential that the regulated supply have some sort of current limiting, or at least fusing, to handle the short. There may be overheating problems with the supply after the crowbar fires. In particular, if the supply includes internal current limiting, the fuse won't blow, and the supply will sit in the "crowbarred" state, with the output at low voltage, until someone notices. Foldback current limiting of the regulated supply would be a good solution here.

There are several problems with this simple crowbar circuit, mostly involving the choice of zener voltage. Zeners are available in discrete values only, with generally poor tolerances and (often) soft knees in the *VI* characteristic. The desired crowbar trigger voltage may involve rather tight tolerances. Consider a 5 volt supply used to power digital logic. There is typically a 5% or 10% tolerance on the supply voltage, meaning that the crowbar cannot be set less than 5.5 volts. The minimum permissible crowbar voltage is raised by the problem of transient response of a regulated supply: When the load current is changed quickly, the voltage can jump, creating a spike followed by some "ringing." This problem is exacerbated by remote sensing via long (inductive) sense leads. The resultant ringing puts glitches on the supply that we don't want to trigger the crowbar. The result is that the crowbar voltage should not be set less than about 6.0 volts, but it cannot exceed 7.0 volts without risk of damage to the logic circuits. When you fold in zener tolerance, the discrete voltages actually available, and SCR trigger voltage tolerances, you've got a tricky problem. In the example shown earlier, the crowbar threshold could lie between 5.9 volts and 6.6 volts, even using the relatively precise 5% zener indicated.

□ *IC sensing*

A nice solution to the problems of predictability and lack of adjustability in the simple zener/SCR crowbar circuit is to use a special crowbar trigger IC such as the MC3423-5, the TL431, or the MC34061-2. These inexpensive chips come in convenient packages (8-pin mini-DIP or 3-pin TO-92), they drive the SCR directly, and they're very easy to use. For example, the MC3425 has adjustable threshold and response time for its crowbar output, and in addition an *under*voltage sensor to signal your circuit that the supply voltage is low (very handy for circuits with microprocessors). It includes an internal reference and several comparators and drivers, and it requires only two external resistors, an optional capacitor, and an SCR to form a complete crowbar. These crowbar chips belong to a class of "power-supply supervisory circuits," which includes complex chips like the MAX691 that not only sense undervoltage but even switch over to battery backup when ac power fails, generate a power-on reset signal on return of normal power, and continually check for lockup conditions in microprocessor circuitry.

Modular crowbars

Why build it when you can buy it! From the designer's point of view the simplest crowbar of all is a 2-terminal gadget that says "crowbar" on top. You can buy just such a device from Lambda or Motorola, who offer a series of overvoltage protection modules in several current ranges. You just pick the voltage and current rating you need, and connect the crowbar across the regulated dc output. For example, the smallest units from Lambda are rated at 2 amps maximum, with the following set of fixed voltages (5V, 6V, 12V, 15V, 18V, 20V, and 24V). They're monolithic, come in a TO-66 package (small metal power transistor case), and cost \$2.50 in small quantities. The Lambda monolithic 6 amp series comes in TO-3 packages (large metal power transistor case) and costs \$5. They also make hybrid 12, 20, and 35 amp crowbars. Motorola's MPC2000 series are all monolithic (5V, 12V, and 15V only, rated at 7.5A, 15A, or 35A). The first two come in TO-220 (plastic power) packages, the last (available in 5V only) in TO-3 (metal power). The good news from Motorola is the incredibly low price: \$1.96, \$2.36, and \$6.08 in small quantities for the three current ratings. One nice feature of these crowbars is the good accuracy; for example, the 5 volt units from Lambda have a specified trip point of 6.6 ± 0.2 volts.

□ *Clamps*

Another possible solution to overvoltage protection is to put a power zener, or its equivalent, across the supply terminals. This avoids the problems of false triggering on spikes, since the zener will stop drawing current when the overvoltage condition disappears (unlike an SCR, which has the memory of an elephant). Figure 6.9 shows

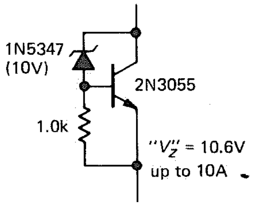


Figure 6.9. Active power zener.

the circuit of an "active zener." Unfortunately, a crowbar constructed from a power zener clamp has its own problems. If the regulator fails, the crowbar has to contend with high power dissipation ($V_{zener}I_{limit}$)

and may itself fail. We witnessed just such a failure in a commercial 15 volt 4 amp magnetic disc supply. When the pass transistor failed, the 16 volt 50 watt zener found itself dissipating more than rated power, and it proceeded to fail too.

□ 6.07 Further considerations in high-current power-supply design

□ *Separate high-current unregulated supply*

As we mentioned in Section 6.03, it is usually a good idea to use a separate supply to power the regulator in very high current supplies. In that way the dissipation in the pass transistors can be minimized, since the unregulated input to the pass transistor can then be chosen just high enough to allow sufficient "headroom" (regulators like the 723 have separate V_+ terminals for this purpose). For instance, a +5 volt 10 amp regulator might use a 10 volt unregulated input with a volt or two of ripple, with a separate low-current +15 volt supply for the regulator components (reference, error amplifier, etc.). As mentioned earlier, the unregulated input voltages must be chosen large enough to allow for worst-case ac power-line voltage (105V) as well as transformer and capacitor tolerances.

□ *Connection paths*

With high-current supplies, or supplies of highly precise output voltage, careful thought must be given to the connection paths, both within the regulator and between the regulator and its load. If several loads are run from the same supply, they should connect to the supply at the place where the output voltage is sensed; otherwise, fluctuations in the current of one load will affect the voltage seen by the other loads (Fig. 6.10).

In fact, it is a good idea to have one common ground point (a "mecca"), as shown, to which the unregulated supply, reference, etc., are all returned. The problem of unregulated voltage drops in the connecting leads from power supply to high-current load is sometimes solved by remote sensing: The connections back to the error amplifier and reference are brought out to the rear of the supply separately and may either be connected to the output terminals right there (the normal method) or brought out and connected to the load at a remote location along with the output voltage leads (this requires four wires, two of which must be able to handle

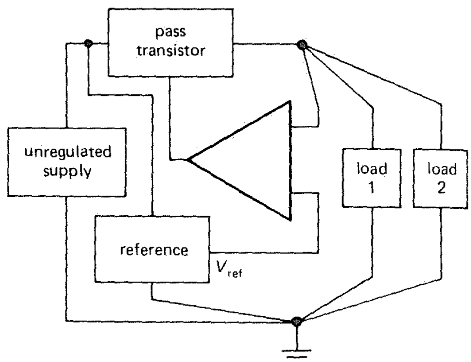


Figure 6.10. A power-supply ground "mecca."

the high load currents). Most commercially available power supplies come with jumpers at the rear that connect the sensing circuitry to the output and that may be removed for remote sensing. Four-wire resistors are used in an analogous manner to sense load currents accurately when constructing precision constant-current supplies. This will be discussed in greater detail in Section 6.24.

□ *Parallel pass transistors*

When very high output currents are needed, it may be necessary to use several pass transistors in parallel. Since there will be a spread of V_{BE}s, it is necessary to add a small resistance in series with each emitter, as in Figure 6.11. The Rs ensure that the current is shared approximately equally among the pass transistors. R should be chosen for about 0.2 volt drop at maximum output current. Power FETs can be connected in parallel without any external components, owing to their negative temperature coefficient of drain current (Fig. 3.13).

Safe operating area

One last point about bipolar power transistors: A phenomenon known as "second breakdown" restricts the simultaneous

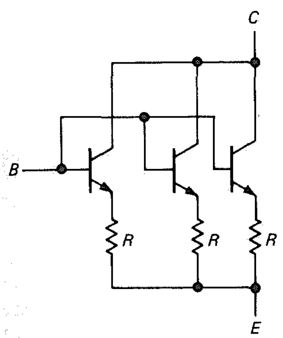


Figure 6.11. Use emitter ballasting resistors when paralleling bipolar power transistors.

voltage and current that may be applied for any given transistor, and it is specified on the data sheet as the safe operating area (SOA) (it's a family of safe voltage-versus-current regions, as a function of time duration). Second breakdown involves the formation of "hot spots" in the transistor junctions, with consequent uneven sharing of the total load. Except at low collector-to-emitter voltages, it sets a limit that is more restrictive than the maximum power dissipation specification. As an example, Figure 6.12 shows the SOA for the ever-popular 2N3055. For $V_{CE} > 40$ volts, second breakdown limits the dc collector current to values corresponding to less

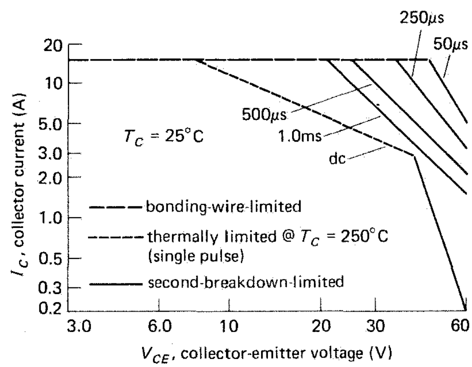


Figure 6.12. Safe operating area for 2N3055 bipolar power transistor. (Courtesy of Motorola, Inc.)

than the maximum allowable dissipation of 115 watts. Figure 6.13 shows the SOA for two similar high-performance power transistors: the 2N6274 *npn* bipolar transistor and the comparable Siliconix VNE003A *n*-channel MOSFET. For $V_{CE} > 10$ volts, second breakdown limits the *npn* transistor dc collector current to values corresponding to less than the maximum allowable dissipation of 250 watts. The problem is less severe for short pulses, and it effectively disappears for pulses of 1ms duration or less. Note that the MOSFET has no second breakdown; its SOA is bounded by maximum current (bonding-wire limited, therefore higher for short pulses), maximum dissipation, and maximum allowable drain-source voltage. See Chapter 3 for more details on power MOSFETs.

□ 6.08 Programmable supplies

There is frequently the need for power supplies that can be adjusted right down to zero volts, especially in bench applications where a flexible source of power is essential. In addition, it is often desirable to be able to "program" the output voltage with another voltage or with a digital input (via digital thumbwheel switches, for instance).

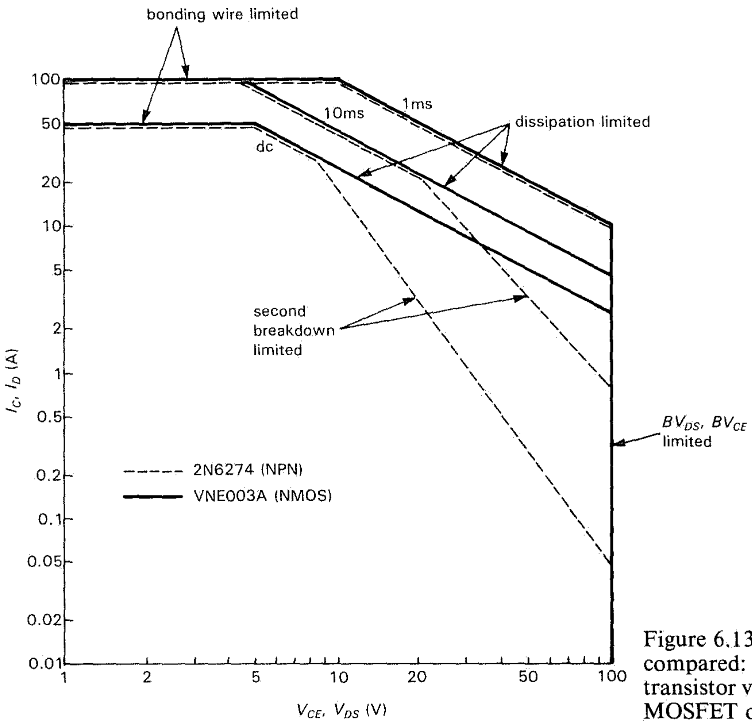


Figure 6.13. Safe operating areas compared: bipolar *npn* power transistor versus *n*-channel power MOSFET of same ratings.

Figure 6.14 shows the classic scheme for a supply that is adjustable down to zero output voltage (as our 723 circuits so far are not). A separate split supply provides power for the regulator and also generates an accurate negative reference voltage (more on references in Sections 6.14 and 6.15). R_1 sets the output voltage (since the inverting input will be at ground), which can be adjusted all the way down to zero (at zero resistance). When the regulator circuitry (which can be an integrated circuit or discrete components) is run from a split supply, no problems are encountered at low output voltages.

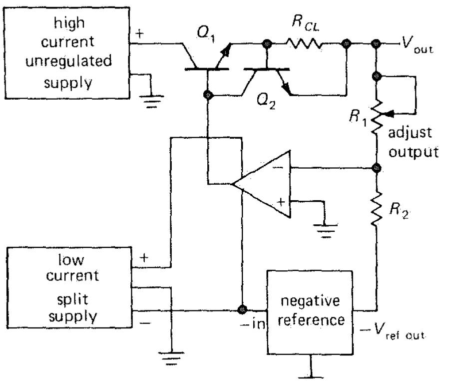


Figure 6.14. Regulator adjustable down to zero volts.

To make the supply programmable with an external voltage, just replace V_{ref} with an externally controlled voltage (Fig. 6.15). The rest of the circuit is unchanged. R_1 now sets the scale of $V_{control}$.

Digital programmability can be added by replacing V_{ref} with a device called a DAC (digital-to-analog converter) with current-sinking output. These devices, which we will discuss later, convert a binary input code to a proportional current (or voltage) output. A good choice here is the AD7548, a monolithic 12-bit DAC with current-sinking output and a price tag of about \$9. By replacing R_2 with the DAC, you get a digitally programmed supply, with step size of 1/4096 (2^{-12}) of the full-scale output voltage. Since the inverting input is a virtual ground, the DAC doesn't even have to have any output compliance. In practice, R_1 would be adjusted to set a convenient scale for the output, say 1mV per input digit.

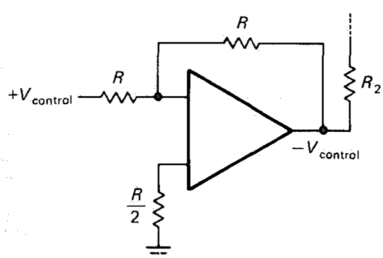


Figure 6.15

□ 6.09 Power-supply circuit example

The "laboratory" bench supply shown in Figure 6.16 should help pull all these design ideas together. It is important to be able to adjust the regulated output voltage right down to zero volts in a general-purpose bench supply, so an additional split supply is used to power the regulator. IC_1 is a high-voltage op-amp, which can operate with 80 volts total supply voltage. We used paralleled power MOSFETs as the output pass transistor, both because of its easy gate drive requirements and its excellent safe operating area (characteristic of all power MOSFETs). The combination can dissipate plenty of power (60W per transistor at 100°C case temperature), which is necessary even for moderate output current when such a wide range of output voltage is provided. This is because the unregulated input voltage has to be high enough for the maximum regulated output voltage, resulting in a large voltage drop across the pass transistors when the regulated output voltage is low. Some supplies solve this problem by having several ranges of output voltage, switching the unregulated input voltage accordingly. There are even supplies with the unregulated supply driven from a variable-voltage transformer ganged to the same control as the output voltage. In both cases you lose the capability of remote programmability.

EXERCISE 6.4

What is the maximum power dissipation in the pass transistors for this circuit?

R_1 is a precision multidecade potentiometer for precise and linear adjustment of the output voltage. The output voltage is referenced to the 1N829 precision zener (5ppm/°C tempco at 7.5mA zener current). The current-limiting circuitry is considerably better than the simple protective current limiters we have been discussing, since it is sometimes desirable to be able to set a precise and stable current limit when using a bench supply. Note the unusual (but convenient) method of current limiting by sinking current from the compensation pin of IC_1, which has unity gain to the output while operating at low current. By providing both precision-regulated voltage (all the way down to 0V) and current, the device becomes a flexible laboratory power supply. With this current-limit method, the supply becomes a flexible constant-current source. Q_4 provides a constant 100mA load, maintaining good performance near zero output voltage (or current) by keeping the pass transistors well into the active region. This current sink also allows the load to source some current into the supply without its output voltage rising. This is useful with the bizarre loads you sometimes encounter, e.g., an instrument that contains some additional supplies of its own capable of sourcing some current into the power-supply output terminal.

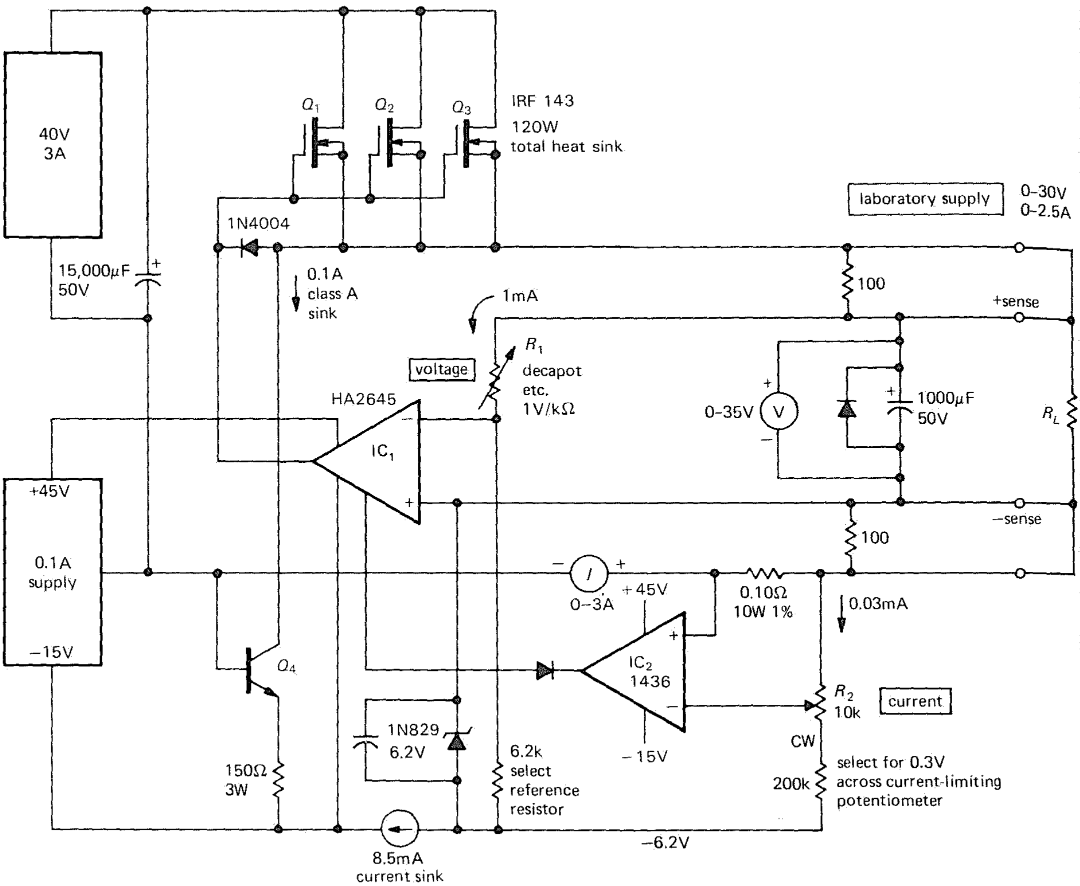


Figure 6.16. Laboratory bench supply.

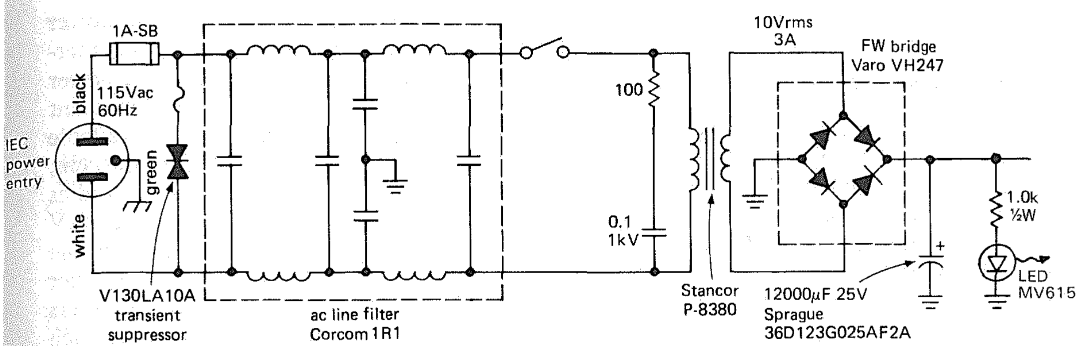


Figure 6.17. Unregulated supply with ac line connections. Note color convention of ac line cord.

Note the external sense leads, with default connection to the power-supply output terminals. For precise regulation of output voltage at the load, you would bring external sense leads to the load itself, eliminating (through feedback) voltage drops in the connecting leads.

6.10 Other regulator ICs

The 723 was the original voltage regulator IC, and it is still a useful chip. There are a few improved versions that work much the same way, however, and you should consider them when you design a regulated power supply. The LAS1000 and LAS1100 from Lambda and the SG3532 from Silicon General can operate down to 4.5 volts input voltage, because they use an internal 2.5 volt "bandgap reference" (see Section 6.15) rather than the 7.15 volt zener of the 723. They also have internal circuitry that shuts off the chip if it overheats; compare the 723's solution (burnout!). Although these regulators have the same pin names, you can't just plug these regulators into a socket intended for a 723, because (among other things) they assume a lower reference voltage. Another 723-like regulator is the MC1469 (and its negative twin, the MC1463) from Motorola.

If you look at modern power-supply circuits, you won't see many 723s, or even the improved versions we just mentioned. Instead, you'll see mostly ICs like the 7805 or 317, with a remarkable absence of external components (the 7805 requires none!). Most of the time you can get all the performance you need from these highly integrated and easy-to-use "three-terminal" regulators, including high output current (up to 10A) without external pass transistors, adjustable output voltage, excellent regulation, and internal current limiting and thermal shut-down. We'll talk about these shortly, but first an interlude on (a) the design of the unregulated supply and (b) voltage references.

THE UNREGULATED SUPPLY

All regulated supplies require a source of "unregulated" dc, a subject we introduced in Section 1.27 in connection with rectifiers and ripple calculations. Let's look at this subject in more detail, beginning with the circuit shown in Figure 6.17. This is an unregulated +13 volt (nominal) supply for use with a +5 volt 2 amp regulator. Let's go through it from left to right, pointing out some of the things to keep in mind when you do this sort of design.

6.11 ac line components

Three-wire connection

Always use a 3-wire line cord with neutral (green) connected to the instrument case. Instruments with ungrounded cases can become lethal devices in the event of transformer insulation failure or accidental connection of one side of the power line to the case. With a grounded case, such a failure simply blows a fuse. You often see instruments with the line cord attached to the chassis (permanently) using a plastic "strain relief," made by Heyco or Richco. A better way is to use an IEC three-prong male chassis-mounted connector, to mate with those popular line cords that have the three-prong IEC female molded onto the end. That way the line cord is conveniently removable. Better yet, you can get a combined "power entry module," containing IEC connector, fuse holder, line filter, and switch (as described later). Note that ac wiring uses a nonintuitive color convention: black = "hot," white = neutral, and green = ground.

Line filter and transient suppressor

In this supply we have used a simple *LC* line filter. Although they are often omitted, such filters are a good idea, since they serve the purpose of preventing possible radiation of radiofrequency interference (RFI) from the instrument via the power line, as well as filtering out incoming interference that may be present on the power line. Power-line filters with excellent performance characteristics are available from several manufacturers, e.g., Corcom, Cornell-Dubilier, and Sprague. Studies have shown that spikes as large as 1kV to 5kV are occasionally present on the power lines at most locations, and smaller spikes occur quite frequently. Line filters are reasonably effective in reducing such interference.

In many situations it is desirable to use a "transient suppressor," as shown, a device that conducts when its terminal voltage exceeds certain limits (it's like a bidirectional high-power zener). These are inexpensive and small and can short out hundreds of amperes of potentially harmful current in the form of spikes. Transient suppressors are made by a number of companies, e.g., GE and Siemens. Tables 6.2 and 6.3 list some useful RFI filters and transient suppressors.

Fuse

A fuse is essential in every piece of electronic equipment. The large wall fuses or circuit breakers (typically 15–20A) in house or lab won't protect electronic equipment, since they are chosen to blow only when the current rating of the wiring in the wall is exceeded. For instance, a

TABLE 6.2. 130 VOLT AC TRANSIENT SUPPRESSORS

Type	Manuf.	Diameter (in)	Energy (W-s)	Peak curr (A)	Capacitance (pF)
V130LA1	GE	0.34	4	500	180
S07K130	Siemens	0.35	6	500	130
V130LA10A	GE	0.65	30	4000	1000
S14K130	Siemens	0.67	22	2000	1000
V130LA20B	GE	0.89	50	6000	1900
S20K130	Siemens	0.91	44	4000	2300

TABLE 6.3. 115 VOLT AC POWER FILTERS (IEC CONNECTOR[a])

Manuf.	Part No.	Circuit	Current (A)	Attenuation[b] (line-to-gnd, 50Ω/50Ω) 150kHz (dB)	500kHz (dB)	1MHz (dB)	Comments
Corcom	3EF1	π	3	15	25	30	general purpose
	3EC1	π	3	20	30	37	higher attenuation
	3EDSC2-2	π	3	32	37	44	with fuse
	2EDL1S	π	2	14	–	24	with fuse and switch
Curtis	F2100CA03	π	3	15	25	30	general purpose
	F2400CA03	π	3	22	35	40	higher attenuation
	F2600FA03	π	3	21	35	41	with fuse
	PE810103	π	3	18	24	30	with fuse and switch
Delta	03GEEG3H	π	3	24	30	38	general purpose
	03SEEG3H	dual-π	3	42	65	70	higher attenuation
	04BEEG3H	π	4	26	35	40	with fuse
	03CK2	π	3	35	40	40	with fuse and switch
	03CR2	dual-π	3	50	60	55	same, higher attenuation
Schaffner	FN323-3	π	3	22	32	36	general purpose
	FN321-3	π	3	35	43	46	higher attenuation
	FN361-2	π	2	25	40	46	with fuse
	FN291-2.5	π	2.5	25	40	46	with fuse and switch
	FN1393-2.5	π	2.5	40	45	42	same, higher attenuation
Sprague	3JX5421A	π	3	15	25	30	general purpose
	3JX5425C	π	3	20	30	37	higher attenuation
	200JM6-2	π	6	12	25	–	with fuse

(a) these units are representative of a large selection, many of which do not include an IEC input connector. (b) rf attenuation figures are measured in a 50Ω system, and should not be relied upon to predict performance in an ac line circuit.

house wired with 14 gauge wire will have 15 amp breakers. Now, if the filter capacitor in the preceding supply becomes short-circuited someday (a typical failure mode), the transformer might then draw 5 amps primary current (instead of its usual 0.25A). The house breaker won't open, but your instrument becomes an incendiary device, with its transformer dissipating over 500 watts!

Some notes on fuses: (a) It is best to use a "slow-blow" type in the power-line circuit, because there is invariably a large current transient at turn-on (caused mostly by rapid charging of the power-supply filter capacitors). (b) You may think you know how to calculate the fuse current rating, but you're probably wrong. A dc power supply has a high ratio of rms current to average current, because of the small conduction angle (fraction of the cycle over which the diodes are conducting). The problem is worse if overly large filter capacitors are used. The result is an rms current considerably higher than you would estimate. The best procedure is to use a "true rms" ac current meter to measure the actual rms line current, then choose a fuse of at least 50% higher current rating (to allow for high line voltage, the effects of fuse "fatigue," etc.). (c) When wiring cartridge-type fuse holders (used with the popular 3AG fuse, which is almost universal in electronic equipment), be sure to connect the leads so that anyone changing the fuse cannot come in contact with the power line. This means connecting the "hot" lead to the rear terminal of the fuse holder (the authors learned this the hard way!). Commercial power-entry modules with integral fuse holders are usually arranged so that the fuse cannot be reached without removing the line cord.

Shock hazard

Incidentally, it is a good idea to insulate all exposed 110 volt power connections inside any instrument, using Teflon heat-shrink tubing, for instance (the use of "friction tape" or electrical tape inside electronic instruments is strictly bush-league). Since most transistorized circuits operate on relatively low dc voltages (±15V to ±30V or so), from which it is not possible to receive a shock, the power line wiring is the only place where any shock hazard exists in most electronic devices (there are exceptions, of course). The front-panel ON/OFF switch is particularly insidious in this respect, since it is close to other low-voltage wiring. Your test instruments (or, worse, your fingers) can easily come in contact with it when you go to pick up the instrument while testing it.

Miscellany

We favor "power-entry modules," combining a 3-prong IEC connector (use a removable line cord) and some combination of line filter, fuse holder, and power switch. For example, the Schaffner FN380 series (or Corcom L series) has all these features, and they are available with maximum currents from 2 to 6 amps. They give you options for fusing or switching either one or both sides of the line, and they offer several filter configurations. Some other manufacturers offering similar products are Curtis, Delta, and Power Dynamics (Table 6.3).

Our circuit shows an LED pilot light (with current-limiting resistor) running from the unregulated dc voltage. It is generally better practice to power the LED from the regulated dc, so that it doesn't flicker with load or power-line variations.

The series combination of 100 ohms and 0.1μF capacitor across the transformer primary prevents the large inductive transient that would otherwise occur at turn-off. This is often omitted, but it is highly desirable, particularly in equipment intended for use near computers or other digital devices. Sometimes this *RC* "snubber" network is wired across the switch, which is equivalent.

6.12 Transformer

Now for the transformer. Never build an instrument to run off the power line without a transformer! To do so is to flirt with disaster. Transformerless power supplies, which are popular in some consumer electronics (radios and televisions, particularly) because they're cheap, put the circuit at high voltage with respect to external ground (water pipes, etc.). This has no place in instruments intended to interconnect with any other equipment and should always be avoided. And use extreme caution when servicing any such equipment; just connecting your oscilloscope probe to the chassis can be a shocking experience.

The choice of transformer is more involved than you might at first expect. One problem is that manufacturers have been slow to introduce transformers with voltages and currents appropriate for transistorized circuitry (the catalogs are still cluttered with transformers designed for vacuum tubes), and you wind up making compromises you'd rather avoid. We have found the Signal Transformer Company unusual, with their nice selection of transformers and quick delivery. Don't overlook the possibility of having transformers custom-made if your application requires more than a few.

Even assuming that you can get the transformer you want, you still have to decide what voltage and current are best. The lower the input voltage to the regulator, the lower the dissipation in the pass transistors. But you must be absolutely certain the input to the regulator will never drop below the minimum necessary for regulation, typically 2 to 3 volts above the regulated output voltage, or you may encounter 120Hz dips in the regulated output. The amount of ripple in the unregulated output is involved here, since it is the minimum input to the regulator that must stay above some critical voltage, but it is the average input to the regulator that determines the transistor dissipation.

As an example, for a +5 volt regulator you might use an unregulated input of +10 volts at the minimum of the ripple, which itself might be a volt or two. From the secondary voltage rating you can make a pretty good guess of the dc output from the bridge, since the peak voltage (at the top of the ripple) is approximately 1.4 times the rms secondary voltage, less two diode drops. But it is essential to make actual measurements if you are designing a power supply with near-minimum drop across the regulator, because the actual output voltage of the unregulated supply depends on poorly specified parameters of the transformer, such as winding resistance and magnetic coupling, both of which contribute to voltage drop under load. Be sure to make measurements under worst-case conditions: full load and low power-line voltage (105V). Remember that large filter capacitors typically have loose tolerances: −30% to +100% about the nominal value is not unusual. It is a good idea to use transformers with multiple taps on the primary, when available, for final adjustment of output voltage. The Triad F-90X series and the Stancor TP series are very flexible this way.

One further note on transformers: Current ratings are sometimes given as rms secondary current, particularly for transformers intended for use into a resistive load (filament transformers, for instance). Since a rectifier circuit draws current only over a small part of the cycle (during the time the capacitor is actually charging), the rms current, and therefore the I^2R heating, is likely to exceed specifications for a load current approaching the rated rms current of the transformer. The situation gets worse as you increase capacitor size to reduce preregulator ripple; this simply requires a transformer of larger rating. Full-wave rectification is better in this respect, since a greater portion of the transformer waveform is used.

6.13 dc components

Filter capacitor

The filter capacitor is chosen large enough to provide acceptably low ripple voltage, with voltage rating sufficient to handle the worst-case combination of no load and high line voltage (125–130V rms). For the circuit shown in Figure 6.17, the ripple is about 1.5 volts pp at full load. Good design practice calls for the use of computer-type electrolytics (they come in a cylindrical package with screw terminals at one end), e.g., the Sprague 36D type. In smaller capacitance values most manufacturers provide capacitors of equivalent quality in an axial-lead package (one wire sticking out each end), e.g., the Sprague 39D type. Watch out for the loose capacitance tolerance!

At this point it may be helpful to look back at Section 1.27, where we first discussed the subject of ripple. With the exception of switching regulators (see Section 6.19 and following), you can always calculate ripple voltage by assuming a constant-current load equal to the maximum output load current. In fact, the input to a series regulator looks just like a constant-current sink. This simplifies your arithmetic, since the capacitor discharges with a ramp, and you don't have to worry about time constants or exponentials (Fig. 6.18).

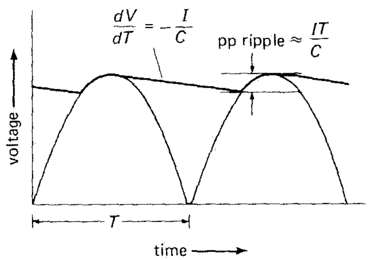


Figure 6.18

For example, suppose you want to choose a filter capacitor for the unregulated portion of a +5 volt 1 amp regulated supply, and suppose you have already chosen a transformer with a 10 volt rms secondary, to give an unregulated dc output of 12 volts (at the peak of the ripple) at full load current. With a typical regulator dropout voltage of 2 volts, the input to the regulator should never dip below +7 volts (the 723 will require +9.5V, but the convenient 3-terminal regulators discussed in Section 6.16 are more friendly). Since you have to contend with a ±10% worst-case line-voltage variation, you should keep ripple to less than 2 volts pp. Therefore,

$$2 = T(dV/dT) = TI/C = 0.008 \times 1.0/C$$

from which C=4000μF. A 5000μF 25 volt electrolytic would be a minimum choice, with allowance for a 20% tolerance in capacitor value. When choosing filter capacitors, don't get carried away: An oversize capacitor not only wastes space but also increases transformer heating (by reducing the conduction angle, hence increasing the ratio $I_{\mathrm{rms}}/I_{\mathrm{avg}}$). It also increases stress on the rectifiers.

The LED shown across the output in Figure 6.17 acts as a "bleeder" to discharge the capacitor in a few seconds under no-load conditions. This is a good feature, because power supplies that stay charged after things have been shut off can easily lead you to damage some circuit components if you mistakenly think that no voltage is present.

Rectifier

The first point to be made is that the diodes used in power supplies are quite different from the small 1N914-type signal diodes used in circuitry. Signal diodes are generally designed for high speed (a few nanoseconds), low leakage (a few nanoamps), and low capacitance (a few picofarads), and they can generally handle currents up to about 100mA, with breakdown voltages rarely exceeding 100 volts. By contrast, rectifier diodes and bridges for use in power supplies are hefty objects with current ratings going from 1 amp to 25 amps or more and breakdown voltages going from 100 volts to 1000 volts. They have relatively high leakage currents (in the range of microamps to milliamps) and plenty of junction capacitance. They are not intended for high speed. Table 6.4 lists a selection of popular types.

Typical of rectifiers is the popular 1N4001–1N4007 series, rated at 1 amp, with reverse-breakdown voltages ranging from 50 to 1000 volts. The 1N5625 series is rated at 3 amps, which is about the highest current available in a lead-mounted (cooled by conduction through the leads) package. The popular 1N1183A series typifies high-current stud-mounted rectifiers, with a current rating of 40 amps and breakdown voltages to 600 volts. Plastic-encapsulated bridge rectifiers are quite popular also, with lead-mounted 1 and 2 amp types and chassis-mounted packages in ratings up to 25 amps or more. For rectifier applications where high speed is important (e.g., dc-to-dc converters, see Section 6.19), fast-recovery diodes are

TABLE 6.4. RECTIFIERS

Type	Breakdown voltage V_{BR} (V)	Forward drop V_F typ (V)	@	Average current I_o (A)	Package	Comments
General purpose						
1N4001-07	50-1000	0.9		1	lead-mounted	popular
1N5059-62	200-800	1.0		2	lead-mounted	
1N5624-27	200-800	1.0		5	lead-mounted	
1N1183A-90A	50-600	0.9		40	stud-mounted	popular; -R for rev. pol.
Fast recovery (t_{rr} = 0.1μs typ)						
1N4933-37	50-600	1.0		1	lead-mounted	
1N5415-19	50-500	1.0		3	lead-mounted	
1N3879-83	50-400	1.2		6	stud-mounted	-R for reverse polarity
1N5832-34	50-400	1.0		20	stud-mounted	-R for reverse polarity
Schottky (low V_F, very fast)						
1N5817-19	20-40	0.6^m		1	lead-mounted	
1N5820-22	20-40	0.5^m		3	lead-mounted	
1N5826-28	20-40	0.5^m		15	stud-mounted	
1N5832-34	20-40	0.6^m		40	stud-mounted	
Full-wave bridge						
3N246-52	50-1000	0.9		1	plastic SIP	MDA100A
3N253-59	50-1000			2	plastic SIP	MDA200
MDA970A1-A5	50-400	0.85		8	chassis mtd	
MDA3500-10	50-1000			35	chassis mtd	
Exotic						
GE A570A-A640L	100-2000	1.0^m		1500	giant button	high current!
Semtech SCH5000-25000	5kV-25kV	$7\text{-}33^m$		0.5	lead-mounted	HV, curr; fast (0.2μs)
Varo VF25-5 to -40	5kV-40kV	$12\text{-}50^m$		0.025	lead-mounted	high voltage
Semtech SCKV100K3-200K3	100kV-200kV	150-300		0.1	plastic rod	very high voltage

$^{(m)}$ maximum.

available, e.g., the 1N4933 series of 1 amp diodes. For low-voltage applications it may be desirable to use Schottky barrier rectifiers, e.g., the 1N5823 series, with forward drops of less than 0.4 volt at 5 amps.

VOLTAGE REFERENCES

There is frequently the need for good voltage references within a circuit. For instance, you might wish to construct a precision regulated supply with characteristics better than those you can obtain using complete regulators like the 723 (since integrated voltage regulator chips usually dissipate considerable power because of the built-in pass transistor, they tend to heat up, with consequent drift). Or you might want to construct a precision constant-current supply. Another application that requires a precision reference, but not a precision power supply, is design of an accurate voltmeter, ohmmeter, or ammeter.

There are two kinds of voltage references – *zener diodes* and *bandgap* *references*; each can be used alone or as an internal part of an integrated circuit voltage reference.

6.14 Zener diodes

The simplest form of voltage reference is the zener diode, a device we discussed in Section 1.06. Basically, it is a diode operated in the reverse-bias region, where current begins to flow at some voltage and increases dramatically with further increases in voltage. To use it as a reference, you simply provide a roughly constant current; this is often done with a resistor from a higher supply voltage, forming the most primitive kind of regulated supply.

Zeners are available in selected voltages from 2 to 200 volts (they come in the same series of values as standard 5% resistors), with power ratings from a fraction of a watt to 50 watts and tolerances of 1% to 20%. As attractive as they might seem for use as general-purpose voltage references, zeners are actually somewhat difficult to use, for a variety of reasons. It is necessary to stock a selection of values, the voltage tolerance is poor except in high-priced precision zeners, they are noisy, and the zener voltage depends on current and temperature. As an example of the last two effects, a 27 volt zener in the popular 1N5221 series of 500mW zeners has a temperature coefficient of +0.1%/°C, and it will change voltage by 1% when its current varies from 10% to 50% of maximum.

There is an exception to this generally poor performance of zeners. It turns out that in the neighborhood of 6 volts, zener diodes become very stiff against changes in current and simultaneously achieve a nearly zero temperature coefficient. The graphs in Figure 6.19, plotted from measurements on zeners with different voltages, illustrate the effects. This peculiar behavior comes about because "zener" diodes actually employ two different mechanisms: zener breakdown (low voltage) and avalanche breakdown (high voltage). If you need a zener for use as a stable voltage reference only, and you don't care what voltage it is, the best thing to use is one of the compensated zener references constructed from a 5.6 volts zener (approximately) in series with a forward-biased diode. The zener voltage is chosen to give a positive coefficient to cancel the diode's temperature coefficient of −2.1mV/°C.

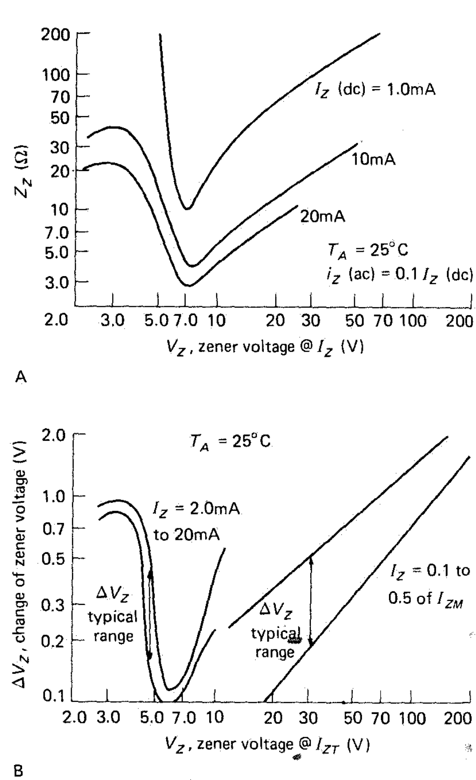


Figure 6.19. Zener diode impedance and regulation for zener diodes of various voltages. (Courtesy of Motorola, Inc.)

As you can see from the graph in Figure 6.20, the temperature coefficient depends on operating current and also on the zener voltage. Therefore, by choosing the zener current properly, you can "tune" the temperature coefficient somewhat. Such zeners with built-in series diodes make particularly good references. As an example, the 1N821 series of inexpensive 6.2 volt references offers temperature coefficients going from 100ppm/°C (1N821) down to 5ppm/°C (1N829); the 1N940 and 1N946 are 9 volt and 11.7 volt references with tempcos of 2ppm/°C.

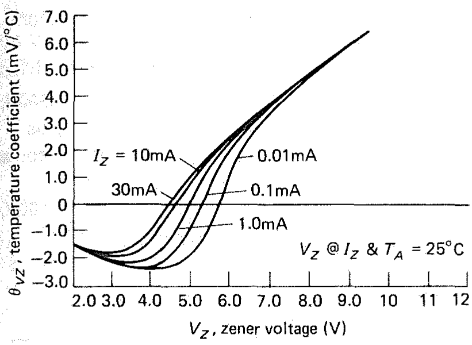


Figure 6.20. Temperature coefficient of zener diode breakdown voltage versus the voltage of the zener diode. (Courtesy of Motorola, Inc.)

Providing operating current

These compensated zeners can be used as stable voltage references within a circuit, but they must be provided with constant current. The 1N821 series is specified as 6.2 volts ±5% at 7.5mA, with an incremental resistance of about 15 ohms; thus, a change in current of 1mA changes the reference voltage three times as much as a change in temperature from −55°C to +100°C for the 1N829. Figure 6.21 shows a simple way to provide constant bias current for a precision zener. The op-amp is wired as a noninverting amplifier in order to generate an output of exactly +10.0 volts. That stable output is itself used to provide a precision 7.5mA bias current. This circuit is self-starting, but it can turn on with either polarity of output! For the "wrong" polarity, the zener operates as an ordinary forward-biased diode. Running the op-amp from a single supply, as shown, overcomes this bizarre problem. Be sure to use an op-amp that has common-mode input range to the negative rail ("single-supply" op-amps).

There are special compensated zeners available with guaranteed stability of zener voltage with *time*, a specification that normally tends to get left out. Examples are the 1N3501 and 1N4890 series. Zeners of this type are available with guaranteed stability of better than 5ppm/1000h. They're not cheap. Table 6.5 lists the characteristics of some useful zeners and reference diodes, and Table 6.6 shows part numbers for two popular 500mW general-purpose zener families.

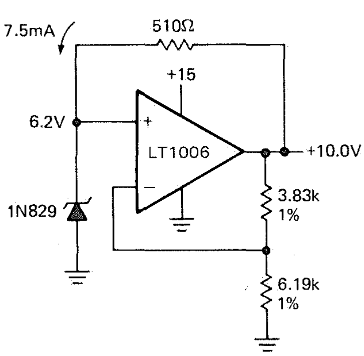


Figure 6.21

IC zeners

The 723 regulator uses a compensated zener reference to achieve its excellent performance (30ppm/°C stability of V_{ref}). The 723, in fact, is quite respectable as a voltage reference all by itself, and you can use the other components of the IC to generate a stable reference output at any desired voltage.

The 723 used as a voltage reference is an example of a *3-terminal reference*, meaning that it requires a power supply to

TABLE 6.5. ZENER AND REFERENCE DIODES[a]

Type	Zener voltage V_Z (V)	@	Test current I_{ZT} (mA)	Tolerance (±%)	Tempco max (ppm/°C)	Regulation ΔV for ±10% I_{ZT} max (mV)	P_{diss} max (W)	Comments
Reference zeners								
1N821A-	6.2		7.5	5	±100	7.5	0.4	5 member family, graded by
1N829A	6.2		7.5	5	±5	7.5	0.4	tempco; best and worst shown
1N4890-	6.35		7.5	5	±20		0.4	long-term stab < 100ppm/1000h
1N4895	6.35		7.5	5	±5		0.4	long-term stab < 10ppm/1000h
Regulator zeners								
1N5221A	2.4		20	10	-850	60	0.5	60 member family, 2.4V to 200V,
1N5231A	5.1		20	10	±300	34	0.5	in "5% resistor values," plus
1N5281A	200		0.65	10	+1100	160	0.5	some extras. -B = ±5%; popular[b]
1N4728A	3.3		76	10	-750	76	1.0	37 member family, 3.3V to 100V,
1N4735A	6.2		41	10	+500	8	1.0	in "5% resistor values."
1N4764A	100		2.5	10	+1100	88	1.0	-B = ±5%; popular

(a) see also Table 6.7 (IC Voltage References). (b) see Table 6.6 (500mW Zeners).

operate, and includes internal circuitry to bias the zener and buffer the output voltage. Improved 3-terminal IC zeners include the excellent LM369 from National (1.5ppm/°C typ), and the REF10KM from Burr-Brown (1ppm/°C max tempco); we've often used the inexpensive Motorola MC1404 (which is actually a bandgap reference, see below) in our circuits. We'll treat 3-terminal precision references in more detail shortly, after discussing the simpler 2-terminal types.

Precision temperature-compensated zener ICs are available as *2-terminal references* also; electrically they look just like zeners, although they actually include a number of active devices to give improved performance (most notably, constancy of "zener" voltage with applied current). An example is the inexpensive LM329, with a zener voltage of 6.9 volts. Its best version has a temperature coefficient of 6ppm/°C (typ), 10ppm/°C (max), when provided with a constant current of 1mA. Some unusual IC zeners include the temperature-stabilized LM399 (0.3ppm/°C

TABLE 6.6. 500mW ZENER DIODES

1N5221 series	1N746 series	V_Z (V)	@	I_{ZT} (mA)
1N5230	1N750	4.7		20
1N5231	1N751	5.1		20
1N5232	1N752	5.6		20
1N5233	–	6.0		20
1N5235	1N754	6.8		20
1N5236	1N755	7.5		20
1N5237	1N756	8.2		20
1N5240	1N758	10		20
1N5242	1N759	12		20
1N5245	1N965	15		8.5
1N5248	1N967	18		7.0
1N5250	1N968	20		6.2
1N5253	–	25		5.0
1N5256	1N972	30		4.2
1N5259	1N975	39		3.2
1N5261	1N977	47		2.7
1N5267	1N982	75		1.7
1N5271	1N985	100		1.3
1N5276	1N989	150		0.85
1N5281	1N992	200		0.65

typ), the micropower LM385 (which operates down to 10μA), and the astounding LTZ1000 from Linear Technology, with its 0.05ppm/°C typical tempco, 0.3ppm per square-root-month drift, and 1.2μV low-frequency noise.

Zener diodes can be very noisy, and some IC zeners suffer from the same disease. The noise is related to surface effects, however, and *buried* (or *subsurface*) zener diodes are considerably quieter. In fact, the LTZ1000 buried zener just mentioned is the quietest reference of any kind. The LM369 and REF10KM also have very low noise.

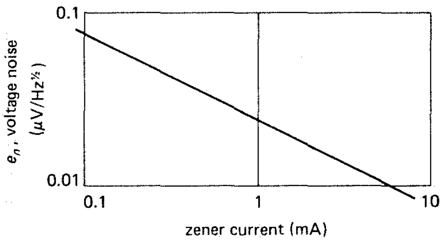


Figure 6.22. Voltage noise for a low-noise zener reference diode similar to the type used in the 723 regulator.

Table 6.7 lists the characteristics of nearly all available IC references, both zener and bandgap.

□ 6.15 Bandgap (V_{BE}) reference

More recently, a circuit known as a "bandgap" reference has become popular. It should properly be called a V_{BE} reference, and it is easily understandable using the Ebers-Moll diode equation. Basically, it involves the generation of a voltage with a positive temperature coefficient the same as V_{BE}'s negative coefficient; when added to a V_{BE}, the resultant voltage has zero tempco.

We start with a current mirror with two transistors operating at different emitter

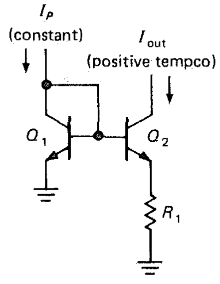


Figure 6.23

current densities (typically a ratio of 10:1) (see Fig. 6.23). Using the Ebers-Moll equation, it is easy to show that I_{out} has a positive temperature coefficient, since the difference in V_{BE}s is just $(kT/q)\log_e r$, where r is the ratio of current densities (see the graph in Fig. 2.53). You may wonder where we get the constant programming current I_P. Don't worry; you'll see the clever method at the end. Now all you do is convert that current to a voltage with a resistor and add a normal V_{BE}.

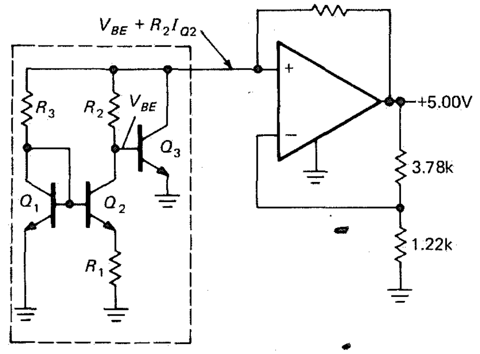


Figure 6.24. Classic V_{BE} bandgap voltage reference.

Figure 6.24 shows the circuit. R_2 sets the amount of positive-coefficient voltage you have added to V_{BE}, and by choosing it appropriately, you get zero overall temperature coefficient. It turns out that zero temperature coefficient occurs when the

TABLE 6.7. IC VOLTAGE REFERENCES

Type	Mfg[a]	B'gap/zener	Terminals	Trim	Voltage (V)	Acc'y (%)	Tempco typ (ppm/°C)	Min supply voltage (V)	Supply curr (mA)	Output curr max (mA)	Noise voltage 0.1-10Hz typ (μV pp)	Long-term stability typ (ppm/1000h)	Regulation: Line typ (%/V)	Regulation: Load 0-10mA typ (%)
Regulator type														
LM10C	NS+	B	8	•	0.20	5	30	1.1	0.3	20	–	–	0.001	0.01[a]
μA723C	FS+	Z	14	•	7.15	3	20	9.5	2.3	65	–	1000	0.003	0.03
SG3532J	SG+	B	10	•	2.50	4	50	4.5	1.6	150	–	300	0.005	0.02
Two-terminal (zener) type														
LM129A	NS	Z	2	–	6.9	5	6	–	1	15[b]	–	20	–	0.1
VR182C	DA	B	2	–	2.455	1.4	23	–	2	120[b]	10[e]	10	–	0.1
LM313	NS	B	2	–	1.22	5	100	–	1[l]	20[b]	5[f]	–	–	0.5[l]
LM329C	NS	Z	2	–	6.9	5	30	–	1	15[b]	–	20	–	0.1
LM336-2.5	NS	B	3	•	2.50	4	10	–	1	10[b]	–	20	–	0.1
LM336B-5	NS	B	3	•	5.0	1	15	–	1[o]	10[b]	–	20	–	0.1[o]
LM385B	NS	B	2	–	1.23	1	20	–	0.1[d]	20[b]	25	–	–	0.02[a]
LM385BX-1.2	NS	B	2	–	1.235	1	30[m]	–	0.1[d]	20[b]	60[f]	20	–	0.8[n]
LM385BX-2.5	NS	B	2	–	2.50	1.5	30[m]	–	0.1[p]	20[b]	120[f]	20	–	0.4[n]
LM299A	NS	Z	4	–	6.95	2	0.2	9	17	10[b]	–	20	–	0.1
LM399	NS	Z	4	–	6.95	5	0.3	9	17	10[b]	–	20	–	0.1
LM3999	NS	Z	3	–	6.95	5	2.0	9	17	10[b]	–	20	–	0.1
TL430	TI	B	3	•	2.75	5	120	–	10	100[b]	50	–	–	0.5
TL431	TI	B	3	•	2.75	2	10	–	10	100[b]	50	–	–	0.5
AD589M	AD	B	2	–	1.235	2	10[m]	–	0.1[h]	5[b]	5[f]	–	–	0.05[a]
LTZ1000	LT	Z	2	–	7.2	4	0.05	–	5		1.2	0.3[s]	–	1[r]
LT1004C-1.2	LT	B	2	–	1.235	0.3	20	–	0.1[d]	20[b]	60[f]	20	–	0.8[mn]
LT1009C	LT	B	3	•	2.50	0.2	15	–	1[o]	10[b]	–	20	–	0.1[o]
LT1029A	LT	B	3	•	5.0	0.2	8	–	1[o]	10[b]	–	20	–	0.04[o]
LT1034B	LT	B	3	–	1.225	1	10	–	0.1[p]	20[b]	4	–	–	0.3[p]
"	LT	Z			7.0	4	40	–	0.1[q]	20[b]	–	–	–	4[q]
HS5010N	HS	B	2	–	1.22	2	3	–	0.1[h]	5[b]	5[f]	–	–	0.05[a]
ICL8069A	IL	B	2	–	1.23	2	10	–	0.5	10[b]	–	–	–	0.2[a]
TSC9491	TS	B	2	–	1.22	2	30	–	0.1[k]	0.5[b]	–	–	–	1.2[k]
Three-terminal type														
REF-01A	PM	B	8	•	10.0	0.3	3	12	1	10	20	–	0.006	0.005
REF-02A	PM	B	8	•	5.0	0.3	3	7	1	10	10	–	0.006	0.005
REF-03E	PM	B	8	•	2.5	0.3	3	4.5	1	10	5	–	0.006	0.05
REF-05	PM	B	8	•	5.0	0.3	3	7	1	10	10	100[m]	0.006	0.05
REF-08G	PM	Z	8	•	-10.0	0.2	10[m]	-11.4	2[m]	10	10	–	0.02[m]	0.2[m]
REF-10	PM	B	8	•	10.0	0.3	3	12	1	10	20	50[m]	0.006	0.05
REF10KM	BB	Z	8	•	10.0	0.05	1[m]	13.5	4.5	10	6	10	0.001	0.01
REF-43E	PM	B	8	•	2.5	0.05	3[m]	4.5	0.2[m]	10	8[gm]	–	0.0002[m]	0.03[m]
LH0070-1	NS	Z	3	–	10.0	0.1	4	12.5	3	10	20	–	0.001	0.01
REF101KM	BB	Z	8	•	10.0	0.05	1[m]	13.5	4.5	10	6	25	0.0003	0.003
LM368Y-2.5	NS	B	8	•	2.5	0.2	11	4.9	0.35	10	12	–	0.0001	0.003
LM368-5	NS	B	4	•	5.0	0.1	15	7.5	0.25	10	16	–	0.0001	0.003
LM368-10	NS	B	4	•	10.0	0.1	15	12.5	0.25	10	30	–	0.0001	0.003
LM369B	NS	Z	3,8	•	10.0	0.05	1.5	13	1.4	10	4	6	0.0002	0.003
AD580M	AD	B	3	–	2.5	1	10	4.5	1	10	60	25	0.04	0.4

Type	Mfg[a]	B'gap/zener	Terminals	Trim	Voltage (V)	Acc'y (%)	Tempco typ (ppm/°C)	Min supply voltage (V)	Supply curr (mA)	Output curr max (mA)	Noise voltage 0.1-10Hz typ (μV pp)	Long-term stability typ (ppm/1000h)	Regulation Line typ (%/V)	Regulation Load 0-10mA typ (%)
Three-terminal type (cont'd)														
AD581L	AD+	B	3	–	10.0	0.05	5	12	0.75	10	50	25	0.005	0.002
AD584L	AD	B	8	•	2.5	0.05	10	5	0.75	18	50	25	0.005	0.002
"	AD				5.0	0.06	5	7.5	0.75	15	50	25	0.005	0.002
"	AD				7.5	0.06	5	10	0.75	13	50	25	0.005	0.002
"	AD				10.0	0.1	5	12.5	0.75	10	50	25	0.005	0.002
AD586L	AD	Z	8	•	5.0	0.05	5[m]	–	5[m]	10	–	15	–	–
AD587L	AD	Z	8	•	10.0	0.05	5[m]	–	5[m]	10	–	15	–	–
AD588B	AD	Z	14	•	±10.0	0.01	1.5[m]	±14	±10	±10	10	25[m]	0.002[m]	0.01[m]
MAX671C	MA	Z	14	•	10.0	0.01	1[m]	13.5	9	10	12	50	0.005[m]	0.01[m]
AD689L	AD	Z	8	•	8.192	0.05	5[m]	10.8	2	±10	2	15	0.002[m]	0.01[m]
R675C-3	HS	Z	14	•	±10.0	0.05	5	±13	+15,-3[m]	10	–	–	0.003[m]	0.02[m]
LT1019A-2.5	LT	B	8	•	2.5	0.002	3	4	0.7	10	6	–	0.00005	0.008
LT1021B-5	LT	Z	8	•	5.0	1	2	7	0.8	10	3	15	0.0004	0.01
LT1031B	LT	Z	3	–	10.0	0.05	3	11	1.2	10	6	15	0.00005	0.01
MC1403A	MO	B	8	–	2.5	1	10	4.5	1.2	10	–	–	0.002	0.06
MC1404AU5	MO	B	8	•	5.0	1	10	7.5	1.2	10	12	25	0.001	0.06
MC1404AU10	MO	B	8	•	10.0	1	10	12.5	1.2	10	12	25	0.0006	0.06
AD2702L[i]	AD+	Z	14	•	±10.0	0.05	5[m]	±13	+12,-2	±10	50	100	0.03[m]	0.05[m]
AD2712L[i]	AD+	Z	14	•	±10.0	0.01	1[m]	±13	+12,-2	±5	30	25	0.013	0.003[j]
LP2950ACZ	NS	B	3	–	5.0	0.5	20	5.4	0.08	100	–	–	0.002	0.004
ICL8212	IL	B	8	•	1.15	3	200	1.8	0.035	20	–	–	0.2	
TSC9495	TS	B	8	•	5.0	1	20	7	1	8	12	–	0.01	0.06
TSC9496	TS	B	8	•	10.0	1	20	12	1	8	25	–	0.01	0.06

(a) 0 to 1mA. (b) max zener curr. (c) on-chip heater/thermostat. (d) specified for 10μA to 20mA operating curr. (e) 1Hz to 10Hz. (f) 10Hz to 10kHz, rms. (g) 10Hz to 1kHz, rms. (h) spec'd for 50μA to 5mA. (i) 2700,2710: +10V; 2701: -10V; 2702,2712: ±10V. (j) 0 to 5mA. (k) spec'd for 50μA to 500μA. (l) spec'd for 0.5 to 20mA. (m) min or max. (n) 1 to 20mA, max. (o) specified for 0.5 to 10mA. (p) specified for 20μA to 20mA. (q) specified for 100μA to 20mA. (r) specified for 1mA to 5mA.

total voltage equals the silicon bandgap voltage (extrapolated to absolute zero), about 1.22 volts. The circuit in the box is the reference. Its own output is used (via R_3) to create the constant current we initially assumed.

Figure 6.25 shows another very popular bandgap reference circuit (it replaces the components in the box in Figure 6.24). Q_1 and Q_2 are a matched pair, forced to operate at a ratio of emitter currents of 10:1 by feedback from the collector voltages. The difference in V_{BE}s is $(kT/q)\log_e 10$, making Q_2's emitter current proportional to T (the preceding voltage applied across R_1). But since Q_1's collector current is larger by a factor of 10, it also is proportional to T. Thus, the total emitter current is proportional to T, and therefore it generates a positive-tempco voltage across R_2. That voltage can be used as a thermometer output, by the way, as will be discussed shortly. R_2's voltage is added to Q_1's V_{BE} to generate a stable reference of zero tempco at the base. Bandgap references appear in many variations, but they all feature the summation of V_{BE} with a voltage generated from a pair of transistors operated with some ratio of current densities.

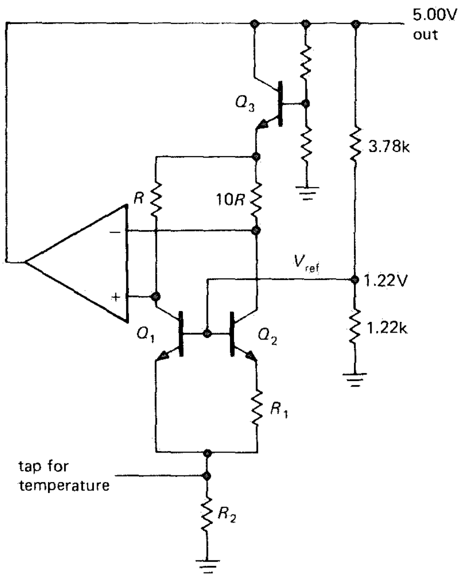


Figure 6.25

□ *IC bandgap references*

An example of an IC bandgap reference is the inexpensive 2-terminal LM385-1.2, with a nominal operating voltage of 1.235 volts, ±1% (the companion LM385-2.5 uses internal circuitry to generate 2.50V), usable down to 10μA. That's much less than you can run any zener at, making these references excellent for micropower equipment (see Chapter 14). The low reference voltage (1.235V) is often much more convenient than the approximately 5 volt minimum usable voltage for zeners (you can get zeners rated at voltages as low as 3.3V, but they are pretty awful, with very soft knees). The best grade of LM385 guarantees 30ppm/°C maximum tempco and has a typical dynamic impedance of 1 ohm at 100μA. Compare this with the equivalent figures for a 1N4370 2.4 volt zener diode: tempco 800ppm/°C (typ), dynamic impedance ≈ 3000 ohms at 100μA, at which the "zener voltage" (specified as 2.4V at 20mA) is about 1.1 volt! When you need a precision stable voltage reference, these excellent bandgap ICs put conventional zener diodes to shame.

If you're willing to spend a bit more money, you can find bandgap references of excellent stability, for example, the 2-terminal LT1029, or the 3-terminal REF-43 (2.50V, 3ppm/°C max). The latter type, like the 3-terminal references based on zener technology, requires a dc supply. Table 6.7 lists most available bandgap (and zener) references, both 2-terminal and 3-terminal.

One other interesting voltage reference is the TL431C. It is an inexpensive "programmable zener" reference, and it is used as shown in Figure 6.26. The "zener" (made from a V_{BE} circuit) turns on when the control voltage reaches 2.75 volts; the device draws only a few microamps from the control terminal and gives a typical tempco of output voltage of 10ppm/°C. The circuit values shown give a zener voltage of 10.0 volts, for example. This device comes in a mini-DIP package and can handle currents to 100mA.

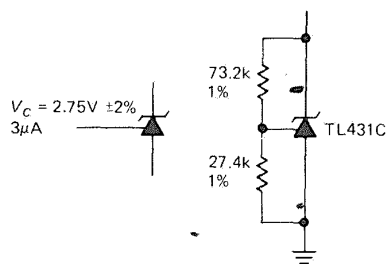


Figure 6.26

□ *Bandgap temperature sensors*

The predictable V_{BE} variation with temperature can be exploited to make a temperature-measuring IC. The REF-02, for instance, generates an additional output voltage that varies linearly with temperature (see preceding discussion). With some simple external circuitry you can generate an output voltage that tells you the chip temperature, accurate to 1% over the full "military" temperature range (−55°C to +125°C). The AD590, intended for temperature measurement only, generates an accurate current of 1μA/°K. It's a 2-terminal device; you just put a voltage across it (4–30V) and measure the current. The LM334 can also be used in this manner. Other sensors, such as the LM35 and LM335, generate accurate voltage outputs with a slope of +10mV/°C. Section 15.01 has a detailed discussion on all these temperature "transducers."

Three-terminal precision references

As we remarked earlier, it is possible to make voltage references of remarkable temperature stability (down to 1ppm/°C or less). This is particularly impressive when you consider that the venerable Weston cell, the traditional voltage reference through the ages, has a temperature coefficient of 40ppm/°C (see Section 15.11). There are two techniques used to make such references.

1. Temperature-stabilized references. A good approach to achieving excellent temperature stability in a voltage reference circuit (or any other circuit, for that matter) is to hold the reference, and perhaps its associated electronics, at a constant elevated temperature. You will see simple techniques for doing this in Chapter 15 (one obvious method is to use a bandgap temperature sensor to control a heater). In this way the circuit can deliver equivalent performance with a greatly relaxed temperature coefficient, since the actual circuit components are isolated from external temperature fluctuations. Of greater interest for precision circuitry is the ability to deliver significantly improved performance by putting an already well-compensated reference circuit into a constant-temperature environment.

This technique of temperature-stabilized or "ovenized" circuits has been used for many years, particularly for ultrastable oscillator circuits. There are commercially available power supplies and precision voltage references that use ovenized reference circuits. This method works well, but it has the drawbacks of bulkiness, relatively large heater power consumption, and sluggish warm-up (typically 10min or more). These problems are effectively eliminated if the thermal stabilization is done at the chip level by integrating a heater circuit (with sensor) onto the integrated circuit itself. This approach was pioneered in the 1960s by Fairchild with the μA726 and μA727 temperature-stabilized differential pair and preamp, respectively.

More recently, temperature-stabilized voltage references such as the National LM199 series have appeared. It offers a temperature coefficient of 0.00002%/°C (typ), which is a mere 0.2ppm/°C. These references are packaged in standard metal transistor cans (TO-46); they consume about 0.25 watt of heater power and come up to temperature in 3 seconds. Users should be aware that the subsequent op-amp circuitry, and even precision wire-wound resistors with their ±2.5ppm/°C tempco, may degrade performance considerably, unless extreme care is used in design. In particular, low-drift precision op-amps such as the OP-07, with 0.2μV/°C (typ) input-stage drift, are essential. These aspects of precision circuit design are discussed in Sections 7.01 to 7.06.

One caution when using the LM399: The chip can be damaged if the heater supply hovers below 7.5 volts for any length of time.

The LT1019 bandgap reference, though normally operated unheated, has an on-chip heater and temperature sensor. So

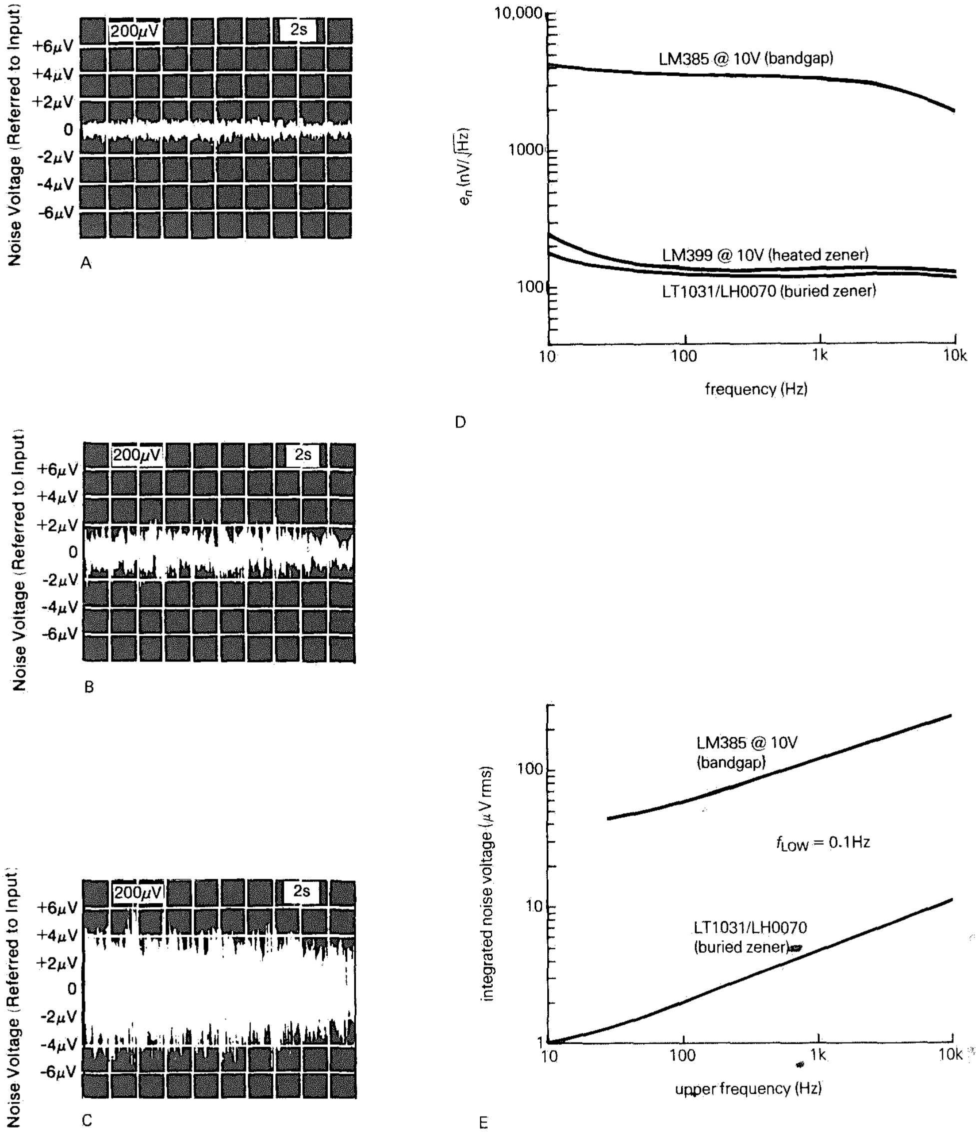


Figure 6.27. Buried zener references (A) have lower noise than either heated zeners (B) or bandgap references (C). (Courtesy of Burr-Brown Corporation.) (D) Noise density (e_n) comparison; (E) integrated noise voltage comparison.

you can use it like the LM399, to get tempcos less than 2ppm/°C. However, unlike the LM399, the LT1019 requires some external circuitry to implement the thermostat (an op-amp and a half dozen components).

☐ 2. *Precision unheated references.* The thermostated LM399 has excellent tempco, but it does not exhibit extraordinary noise or long-term drift specs (see Table 6.7). The chip also takes a few seconds to heat up, and it uses plenty of power (4W at start-up, 250mW stabilized).

Clever chip design has made possible unheated references of equivalent stability. The REF10KM and REF101KM from Burr-Brown have tempcos of 1ppm/°C (max), with no heater power or warm-up delays. Furthermore, they exhibit lower long-term drift and noise than the LM399-style references. Other 3-terminal references with 1ppm/°C maximum tempco are the MAX671 from Maxim and the AD2710/2712 references from Analog Devices. In 2-terminal configurations the only contender is the magnificent LTZ1000 from Linear Technology, with its claimed 0.05ppm/°C tempco. It also claims long-term drift and noise specs that are a factor of 10 better than any other reference of any kind. The LTZ1000 does require a good external biasing circuit, which you can make with an op-amp and a few parts. All of these high stability references (including the heated LM399) use buried zeners, which additionally provide much lower noise than ordinary zener or bandgap references (Figure 6.27).

THREE-TERMINAL AND FOUR-TERMINAL REGULATORS

6.16 Three-terminal regulators

For most noncritical applications the best choice for a voltage regulator is the simple 3-terminal type. It has only three connections (input, output, and ground) and is factory-trimmed to provide a fixed output. Typical of this type is the 78*xx*. The voltage is specified by the last two digits of the part number and can be any of the following: 05, 06, 08, 10, 12, 15, 18, or 24. Figure 6.28 shows how easy it is to make a +5 volt regulator, for instance, with one of these regulators. The capacitor across the output improves transient response and keeps the impedance low at high frequencies (an input capacitor of at least 0.33μF should be used in addition if the regulator is located a considerable distance from the filter capacitors). The 7800 series is available in plastic or metal power packages (same as power transistors). A low-power version, the 78L*xx*, comes in the same plastic and metal packages as small-signal transistors (see Table 6.8). The 7900 series of negative regulators works the same way (with negative input voltage, of course). The 7800 series can provide up to 1 amp load current and has on-chip circuitry to prevent damage in the event of overheating or excessive load current; the chip simply shuts down, rather than blowing out. In addition, on-chip circuitry prevents operation outside the transistor safe operating area (see Section 6.07) by reducing available output current for large input–output voltage differential. These regulators are inexpensive and easy to use, and they make it practical to design a system with many printed-circuit boards in which the unregulated dc is brought to each board and regulation is done locally on each circuit card.

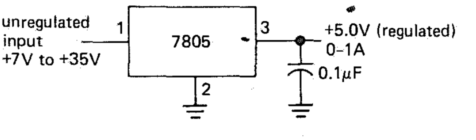


Figure 6.28

Three-terminal fixed regulators come in some highly useful variants. The LP2950 works just like a 7805, but draws only

TABLE 6.8. FIXED VOLTAGE REGULATORS

				Output current (max)[a]							
				@75°C case	No heatsink[b]		Regulation (typ)			Input voltage	
Type	Pkg	V_{out} (V)	Accuracy (%)	I_{out} (A)	I_{out} (A)	P_{diss} (W)	Load[c] (mV)	Line[d] (mV)	Θ_{JC} (°C/W)	min[i] (V)	max (V)
Positive											
LM2950CZ-5.0	TO-92	5	1	0.08	0.1	0.5	2	1.5	160	5.4	30
LM2931Z-5.0	TO-92	5	5	0.1	0.1	0.5	14	3	160	5.3	26
LM78L05ACZ	TO-92	5	4	0.1	0.1	0.6	5	50	160	7	35
LM330T-5.0[g]	TO-220	5	4	0.15	0.15	1.5	14	20	4	5.3	26
TL750L05	TO-92	5	4	0.15	0.15	0.6	20	6	160	5.6	26
LM2984CT	TO-220[h]	5	3	0.5	0.5	2	12	4	3	5.5	26
LM2925T	TO-220	5	5	0.75	0.5	2	10	8	3	5.6	26
LM2935T	TO-220	5	5	0.75	0.5	2	10	8	3	5.5	26
LM309K	TO-3	5	4	1	0.6	2.2	20	4	3	7	35
LT1005CT	TO-220	5	2	1	0.5	2	5	5	3	7	20
LM2940T-5.0	TO-220	5	3	1	0.5	2	35	20	3	5.5	26
LM7805CK	TO-3	5	4	1	0.6	2.2	10	3	3.5	7	35
LM7805CT	TO-220	5	4	1	0.45	1.7	10	3	3	7	35
LM7815CT	TO-220	15	4	1	0.15	1.7	12	4	3	17	35
LT1086-5CT	TO-220	5	1	1.5	0.5	2	5	0.5	3	6.3	30
LAS16A05	TO-3	5	2	2	0.75	2.8	30[m]	100[m]	2.5	7.6	30
LM323K	TO-3	5	4	3	0.6	2	25	5	2	7	20
LT1035CK	TO-3	5	2	3	0.8	3	10	5	1.5	7.3	20
LT1085-5CT	TO-220	5	1	3	0.5	2	5	0.5	3	6.3	30
LAS14A05	TO-3	5	2	3	0.8	3	30[m]	50[m]	2.3	7.5	35
LT1003CK	TO-3	5	2	5	0.8	3	25	5	1	7.3	20
LT1084-5CK	TO-3	5	1	5	0.8	3	5	0.5	1.6	6.3	30
LAS19A05	TO-3	5	2	5	0.8	3	30[m]	50[m]	0.9	7.6	30
LT1083-5CK	TO-3	5	1	7.5	0.8	3	5	0.5	1.6	6.3	30
LAS3905	TO-3	5	5	8	0.8	3	20[m]	100[m]	0.7	7.6	25
Negative											
LM79L15ACZ	TO-92	-15	4	0.1	0.05	0.6	75[m]	45[m]	160	-17	-35
LM7915CK	TO-3	-15	4	1	0.2	2.2	4	3	3.5	-16.5	-35
LM7915CT	TO-220	-15	4	1	0.15	1.7	4	3	3	-16.5	-35
LM345K-5.0	TO-3	-5	4	3	0.2	2.1	10	5	2	-7.5	-20

(a) with V_{in}=1.75V_{out}. (b) 50°C ambient. (c) 0 to I_{max}. (d) ΔV_{in}=15V. (e) ΔV_{out} for 0°C to 100°C junc temp. (f) 1000 hours. (g) similar to LM2930T-5.0, LM2931T-5.0. (h) wide TO-220. (i) at I_{max}. (m) min or max. (t) typical. All include internal thermal shutdown and current-limiting circuitry. Most are available in ±5, 6, 8, 10, 12, 15, 18, and 24V units; a few are available in -2, -3, -4, -5.2, -9, +2.6, +9, and +17V units.

Type	120Hz ripple reject typ (dB)	Temp stab[e] typ (mV)	Long-term stab[f] max (%)	Output impedance 10Hz (Ω)	Output impedance 10kHz (Ω)	Comments
LM2950CZ-5.0	70	10	–	0.01	0.5	micropower, 1%
LM2931Z-5.0	80	–	0.4[t]	0.1	0.2	low dropout, low power
LM78L05ACZ	50	–	0.25	0.2	0.2	small; LM240LAZ-5.0
LM330T-5.0[g]	56	25	0.4[t]	0.1	0.2	low dropout; 2930
TL750L05	65	50	–	–	–	TL751 has enable
LM2984CT	70	3	0.4[t]	0.01	0.02	dual outputs (μP); reset, on/off
LM2925T	66	–	0.4[t]	0.2	0.2	microprocessor; reset
LM2935T	66	–	0.4[t]	0.02	0.02	dual outputs (μP); reset, on/off
LM309K	80	50	0.4	0.04	0.05	original +5V regulator
LT1005CT	70	25	–	0.003	0.01	dual outputs (μP)
LM2940T-5.0	72	20	0.4[t]	0.03	0.03	
LM7805CK	80	30	0.4	0.01	0.03	LM340K-5
LM7805CT	80	30	0.4	0.01	0.03	popular; LM340T-5
LM7815CT	70	100	0.4	0.02	0.05	LM340T-15
LT1086-5CT	63	25	1	–	–	low dropout
LAS16A05	75	–	–	0.002	0.02	Lambda, monolithic
LM323K	70	30	0.7	0.01	0.02	
LT1035CK	70	25	–	0.003	0.01	dual +5; 1036 is +12/+5
LT1085-5CT	63	25	1	–	–	low dropout
LAS14A05	70	100[m]	–	0.001	0.003	Lambda, monolithic
LT1003CK	66	25	0.7	0.003	0.02	
LT1084-5CK	63	25	1	–	–	low dropout
LAS19A05	70	150[m]	–	0.01	0.2	lambda, monolithic
LT1083-5CK	63	25	1	–	–	low dropout
LAS3905	60[m]	100	–	0.004	0.01	Lambda, monolithic
LM79L15ACZ	40	–	0.4[t]	0.05	0.05	small; LM320LZ-15
LM7915CK	60	60	0.4	0.06	0.07	LM320KC-15
LM7915CT	60	60	0.4	0.06	0.07	LM320T-15
LM345K-5.0	65	25	1.0	0.02	0.04	

75μA of quiescent current (compared with the 7805's 5mA, or the 78L05's 3mA); it also regulates with as little as a 0.4 volt drop from unregulated input to regulated output (called the "dropout voltage"), compared with 2 volts dropout for the classic 7805. The LM2931 is also low-dropout, but you might call it *milli*power (0.4mA quiescent current), compared with the "micropower" LP2950. Low-dropout regulators also come in high-current versions – for example, the LT1085/4/3 series from

LTC (3A, 5A, and 7.5A, respectively, with both +5V and +12V available in each type). Regulators like the LM2984 are basically 3-terminal fixed regulators, but with extra outputs to signal a microprocessor that power has failed, or resumed. Finally, regulators like the 4195 contain a pair of 3-terminal 15 volt regulators, one positive and one negative. We'll say a bit more about these special regulators shortly.

6.17 Three-terminal adjustable regulators

Sometimes you want a nonstandard regulated voltage (say +9V, to emulate a battery) and can't use a 78*xx*-type fixed regulator. Or perhaps you want a standard voltage, but set more accurately than the ±3% accuracy typical of fixed regulators. By now you're spoiled by the simplicity of 3-terminal fixed regulators, and therefore you can't imagine using a 723-type regulator circuit, with all its required external components. What to do? Get an "adjustable 3-terminal regulator"! Table 6.8 lists the characteristics of a representative selection of 3-terminal fixed regulators.

These wonderful ICs are typified by the classic LM317 from National. This regulator has no ground terminal; instead, it adjusts V_{out} to maintain a constant 1.25 volts (bandgap) from the output terminal to the "adjustment" terminal. Figure 6.29 shows the easiest way to use it. The regulator puts 1.25 volts across R_1, so 5mA flows through it. The adjustment terminal draws very little current (50–100μA), so the output voltage is just

$$V_{out} = 1.25(1 + R_2/R_1) \text{ volts}$$

In this case the output voltage is adjustable from 1.25 volts to 25 volts. For a fixed-output-voltage application, R_2 will normally be adjustable only over a narrow range, to improve settability (use a fixed resistor in series with a trimmer). Choose your resistive divider values low enough to allow for a 50μA change in adjustment current with temperature. Because the loop compensation for the regulator is the output capacitor, larger values must be used compared with other designs. At least a 1μF tantalum is required, but we recommend something more like 6.8μF.

The 317 is available in several packages, including the plastic power package (TO-220), the metal power package (TO-3), and the small transistor packages (metal, TO-5; plastic, TO-92). In the power packages it can deliver up to 1.5 amps, with proper heat sinking. Because it doesn't "see" ground, it can be used for high-voltage regulators, as long as the input–output voltage differential doesn't exceed the rated maximum of 40 volts (60V for the LM317HV high-voltage variant).

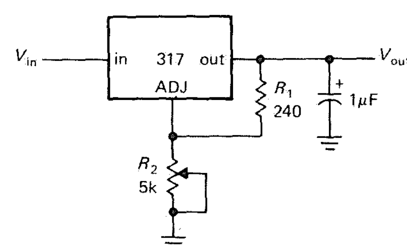


Figure 6.29. Three-terminal adjustable regulator.

EXERCISE 6.5

Design a +5 volt regulator with the 317. Provide ±20% voltage adjustment range with a trimmer pot.

Three-terminal adjustable regulators are available with higher current ratings, e.g., the LM350 (3A), the LM338 (5A), and the LM396 (10A), and also with higher voltage ratings, e.g., the LM317H (60V) and the TL783 (125V). Read the data sheets carefully before using these parts, noting bypass capacitor requirements and safety diode suggestions. As with the fixed

3-terminal regulators, you can get low-dropout versions (e.g., the LT1085, with 1.3V dropout at 3.5A), and you can get micropower versions (e.g., the LP2951, the adjustable variant of the fixed 5V LP2950; both have $I_Q = 75\mu A$). You can also get *negative* versions, although there's less variety: The LM337 is the negative cousin of the LM317 (1.5A), and the LM333 is a negative LM350 (3A).

Four-terminal regulators

Three-terminal adjustable regulators are the favorite for noncritical requirements. Historically they were preceded by four-terminal adjustable regulators, which you connect as shown in Figure 6.30. You drive the "control" terminal with a sample of the output; the regulator adjusts the output to keep the control terminal at a fixed voltage (+3.8V for the Lambda regulators in Table 6.9, +5V for the μA79G, and −2.2V for the negative regulators). Four-terminal regulators aren't any better than the simpler 3-terminal variety (but they aren't any worse, either), and we mention them here for completeness.

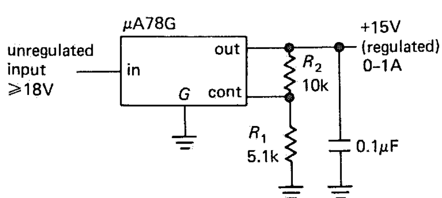


Figure 6.30

6.18 Additional comments about 3-terminal regulators

General characteristics of 3- and 4-terminal regulators

The following specifications are typical for most 3- and 4-terminal regulators, both fixed and adjustable, and they may be useful as a rough guide to the performance you can expect:

Output voltage tolerance:	1–2%
Dropout voltage:	0.5–2 volts
Maximum input voltage:	35 volts (except TL783 to +125V)
Ripple rejection:	0.01–0.1%
Spike rejection:	0.1–0.3%
Load regulation:	0.1–0.5%, full load change
dc input rejection:	0.2%
Temperature stability:	0.5%, over full temp range

Improving ripple rejection

The circuit of Figure 6.29 is the standard 3-terminal regulator, and it works fine. However, the addition of a 10μF bypass capacitor from the adjust (ADJ) terminal to ground (Fig. 6.31) improves the ripple (and spike) rejection by about 15dB (factor of 5 in voltage). For example, the LM317 ripple rejection factor goes from 65dB to 80dB (the latter is 0.1mV output ripple when supplied with 1V input ripple, a typical value). Be sure to include the safety discharge diode; look at the specification sheet of the particular regulator for more details.

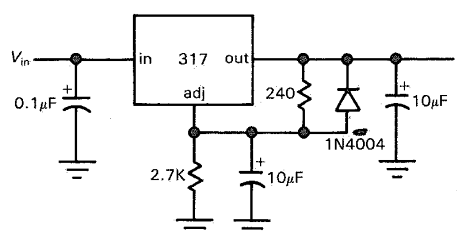


Figure 6.31. The ADJ pin may be bypassed for lower noise and ripple, but a safety discharge diode must be included.

Low-dropout regulators

As we mentioned earlier, most series regulators need at least 2 volts of "headroom" to function; that's because the base of the *npn* pass transistor is a V_{BE} drop above

TABLE 6.9. ADJUSTABLE VOLTAGE REGULATORS

Type	Polarity	Pkg	Output voltage min (V)	Output voltage max (V)	I_{max} (A)	Regulation (typ) Load[a] (%)	Regulation (typ) Line[b] (%)	Θ_{JC} (°C/W)	Input voltage min (V)	Input voltage max (V)	Dropout voltage @I_{max} max (V)	120Hz ripple reject typ (dB)	Temp stab[c] typ (%)	Long-term stab[d] max (%)	Output impedance 10Hz (Ω)	Output impedance 10kHz (Ω)	Therm lim	Curr lim[i]	Comments
Three-terminal																			
LM317L	+	TO-92	1.2	37	0.1	0.1	0.15	160[h]	–	40[e]	2.5[t]	65	0.5	1	0.07	4	•	I	miniature
LM337L	–	TO-92	1.2	37	0.1	0.1	0.15	160[h]	–	-40[e]	2.5[t]	65	0.5	–	–	–	•	I	miniature (neg 317L)
LM317H	+	TO-39	1.2	37	0.5	0.1	0.2	12	–	40[e]	2[t]	80	0.6	0.3	0.01	0.03	•	I	317 in TO-39
LM337H	–	TO-39	-1.2	-37	0.5	0.3	0.2	12	–	-40[e]	2[t]	75	0.5	0.3	0.02	0.02	•	I	negative 317H
TL783C	+	TO-220	1.3	125	0.7	0.2[f]	0.02	4	–	125[e]	10	50	0.3	0.2	0.05	0.3	•	I	MOSFET, high voltage
LM317T	+	TO-220	1.2	37	1.5	0.1	0.2	4	–	40[e]	2.5[t]	80	0.6	0.3	0.01	0.03	•	I	popular
LM317HVK	+	TO-3	1.2	57	1.5	0.1	0.2	2.3	–	60[e]	2.5[t]	80	0.6	0.3	0.01	0.03	•	I	high voltage 317
LM337T	–	TO-220	-1.2	-37	1.5	0.3	0.2	4	–	-40[e]	2.5[t]	75	0.5	0.3	0.02	0.02	•	I	negative 317
LM337HVK	–	TO-3	-1.2	-47	1.5	0.3	0.2	2.3	–	-50[e]	2.5[t]	75	0.5	0.3	0.02	0.02	•	I	high voltage 337
LT1086CP	+	TO-220	1.3	30	1.5	0.1	0.02	–	–	30[e]	1.5	75	0.5	1	–	–	•	I	low dropout
LM350K	+	TO-3	1.2	32	3	0.1	0.1	2	–	35[e]	2.5[t]	80	0.6	0.3	0.005	0.02	•	I	3A monolithic
IP3R07T	+	TO-220	1.2	37	3	0.1	0.08	2.3	–	15[e]	0.8[t]	65	–	–	–	–	•	I	two unreg inputs
LM333T	–	TO-220	-1.2	-32	3	0.2	0.02	50	–	-35[e]	2.5[t]	60	0.5	0.2	–	–	•	I	neg 350; LT1033 is imprvd
LT1085CT	+	TO-220	1.3	30	3	0.1	0.02	3	–	30[e]	1.5	75	0.5	1	–	–	•	I	low dropout
LM338K	+	TO-3	1.2	32	5	0.1	0.1	2	–	35[e]	2.5[t]	80	0.6	0.3	–	–	•	I	5A monolithic
LT1084CP	+	TO-247	1.3	30	5	0.1	0.02	2.3	–	30[e]	1.5	75	0.5	1	–	–	•	I	low dropout
LT1083CP	+	TO-247	1.3	30	7.5	0.1	0.02	1.6	–	30[e]	1.5	75	0.5	1	–	–	•	I	low dropout
LM396K	+	TO-3	1.2	15	10	0.4[m]	0.08	1	–	20[e]	2.1[t]	74	0.3	1	0.01	0.02	•	I	10A monolithic
LT1038CK	+	TO-3	1.2	32	10	0.1	0.08	1	–	35[e]	2.5[t]	60	1	1	0.005	0.1	•	I	10A monolithic, 1% acc'y

Four-terminal																			
µA78GU1C	+	TO-220	5	30	1	1^{m}	1^{m}	7.5	7.5	40	2.5	80	3^{m}	–	–	–	•	I	TO-39 pkg avail
µA79GU1C	–	TO-220	-2.5	-30	1	1^{m}	1^{m}	7.5	-7	-40	2^{t}	60	3^{m}	–	–	–	•	I	TO-39 pkg avail
LAS15U	+	TO-3	4	30	1.5	0.6^{m}	2^{m}	3	6.5	40	2.4	70	3^{m}	–	0.003	0.02	•	I	Lambda
LAS18U	–	TO-3	-2.6	-30	1.5	0.6^{m}	2^{m}	3	-5	-40	2.1	60	3^{m}	–	0.02	0.04	•	I	Lambda
LAS16U	+	TO-3	4	30	2	0.6^{m}	2^{m}	2.5	6.5	35	2.6	70	2^{m}	–	0.002	0.02	•	I	Lambda
LAS14AU	+	TO-3	4	35	3	0.6^{m}	1^{m}	1.5	6.5	40	2.3	70	2^{m}	–	0.001	0.01	•	I	Lambda
LAS19U	+	TO-3	4	30	5	0.6^{m}	1^{m}	0.9	6.5	35	2.6	65	2^{m}	–	0.0005	0.004	•	I	Lambda
LAS39U	+	TO-3	4	16	8	0.6^{m}	2^{m}	0.7	6.6	25	2.6	60^{m}	3^{m}	–	0.002	0.01	•	I	Lambda
Multiterminal																			
LM376N	+	DIP-8	5	37	0.03	0.2^{m}	0.6^{m}	190^{h}	9	40	3	60^{m}	1^{m}	–	–	–	–	E	
LM304H	–	TO-5	0	-40	0.03	1mV	0.2	45	-8	-40	2	65	0.3	0.01	–	–	–	E	orig neg reg
ICL7663S	+	DIP-8	1.3	16	0.04	0.4^{f}	0.5	200^{h}	1.5	16	1^{t}	20	1	–	–	–	–	E	micropower; also MAX663
MAX664	–	DIP-8	-1.3	-16	0.04	0.8^{f}	0.5	120^{h}	-2	-16	0.3^{t}	15	1	–	–	–	–	E	µpwr, impr 7664; low dropout
LM305AH	+	TO-5	4.5	40	0.05	0.03	0.3	45	8.5	50	3	80	0.3	0.1	–	–	–	E	
LM2931CT	+	TO-220	3	24	0.1	0.3	0.06^{f}	3	3.6	26	0.3^{t}	60^{f}	–	0.4	0.1^{f}	0.2^{f}	•	I	low dropout, low power
LP2951CN	+	DIP-8	1.3	29	0.1	0.1^{f}	0.03^{f}	105	1.7	30	0.4^{t}	70^{f}	0.5	–	0.01	0.5	•	I	low dropout, micropower
LT1020CN	+	DIP-14	2.5	35	0.13	0.2	0.15	60?	4.5	36	0.4^{t}	60	1	–	–	–	•	I	micropower
NE550N	+	DIP-14	2	40	0.15	0.03	0.08	150^{h}	8.5	40	3	90	0.2	0.1^{t}	0.1	0.1	–	E	
µA723PC	+	DIP-14	2	37	0.15	0.03	0.1	150^{h}	9.5	40	3	75	0.3	0.1	0.05	0.1	–	E	classic
LAS1000	+	TO-5	3	38	0.15	0.1^{m}	0.2	150^{h}	5	40	2	60^{m}	1.5^{m}	–	0.004	0.05	•	E	Lambda, improved 723
LAS1100	+	TO-5	3	48	0.15	0.1^{m}	0.2	150^{h}	5	50	2	60^{m}	1.5^{m}	–	0.004	0.05	•	E	high voltage LAS1000
SG3532J	+	DIP-14	2	38	0.17	0.1	0.1	125^{h}	4.7	40	2	66	0.5	0.3	–	–	•	E	improved 723
MC1469R	+	TO-66	2.5	32	0.6	0.005	0.05	7	9	35	3	100	0.2	–	0.05	0.1	–	E	precision, may oscillate
MC1463R	–	TO-66	-3.8	-32	0.6	0.005	0.05	17	-9	-35	3	90	0.2	–	0.02	0.03	–	E	neg MC1469
LM2941CT	+	TO-220	1.3	25	1			3		26	1	74	0.4	0.4^{t}	0.04	0.1	–	I	low dropout
LAS2200	+	module	2.5	28	5	0.2^{m}	0.15^{m}	2	9.6	40	2.5	60^{m}	0.7^{m}	–	–	–	•	I	
LAS3000	+	module	2.7	29	10	0.2^{m}	0.15^{m}	1.3	7.9	40	2.5	60^{m}	1.5^{m}	–	–	–	•	I	Lambda hybrid; 2 unreg inputs
LAS5000	+	module	4.8	29	20	0.2^{m}	0.2^{m}	0.7	11.9	40	2.5	60^{m}	1.5^{m}	–	–	–	•	I	Lambda hybrid; 2 unreg inputs
LAS7000	+	module	4.8	29	30	0.2^{m}	0.2^{m}	0.4	12.3	40	2.5	60^{m}	1.5^{m}	–	–	–	•	I	Lambda hybrid; 2 unreg inputs
MC1466L	+	DIP-14	0	1000	–	0.02	0.05	170^{h}	–	–	2^{t}	70	0.4	–	–	–	–	E	lab supply; good curr lim
LAS3700	+	TO-5	0	1000	–	0.003	0.15	220	–	–	–	65	0.5^{m}	–	–	–	–	E	floating reg with on-chip heater

(a) 10% to 50% I_{max}. (b) for ΔV_{in}=15V. (c) ΔV_{out} for 0°C to 100°C T_J. (d) 1000 hrs. (e) max V_{in}-V_{out}. (f) at 5V. (h) Θ_{JA}. (i) E – external; I – internal. (m) min/max. (t) typ.

the output, and it has to be driven by a driver transistor, usually another *npn* whose base is pulled up with a current mirror. That's already two V_{BE} drops. Furthermore, you need to allow another V_{BE} drop across the current-sensing resistor for short-circuit protection; see Figure 6.32A, a simplified schematic of the 78L*xx*. The three V_{BE}s add up to about 2 volts, below which the regulator drops out of regulation at full current.

By using a *pnp* (or *p*-channel MOSFET) pass transistor, the dropout voltage can be reduced from the three V_{BE} drop of the conventional *npn* scheme, down nearly to the transistor saturation voltage. Figure 6.32B shows a simplified schematic of the LM330 low-dropout fixed +5 volt

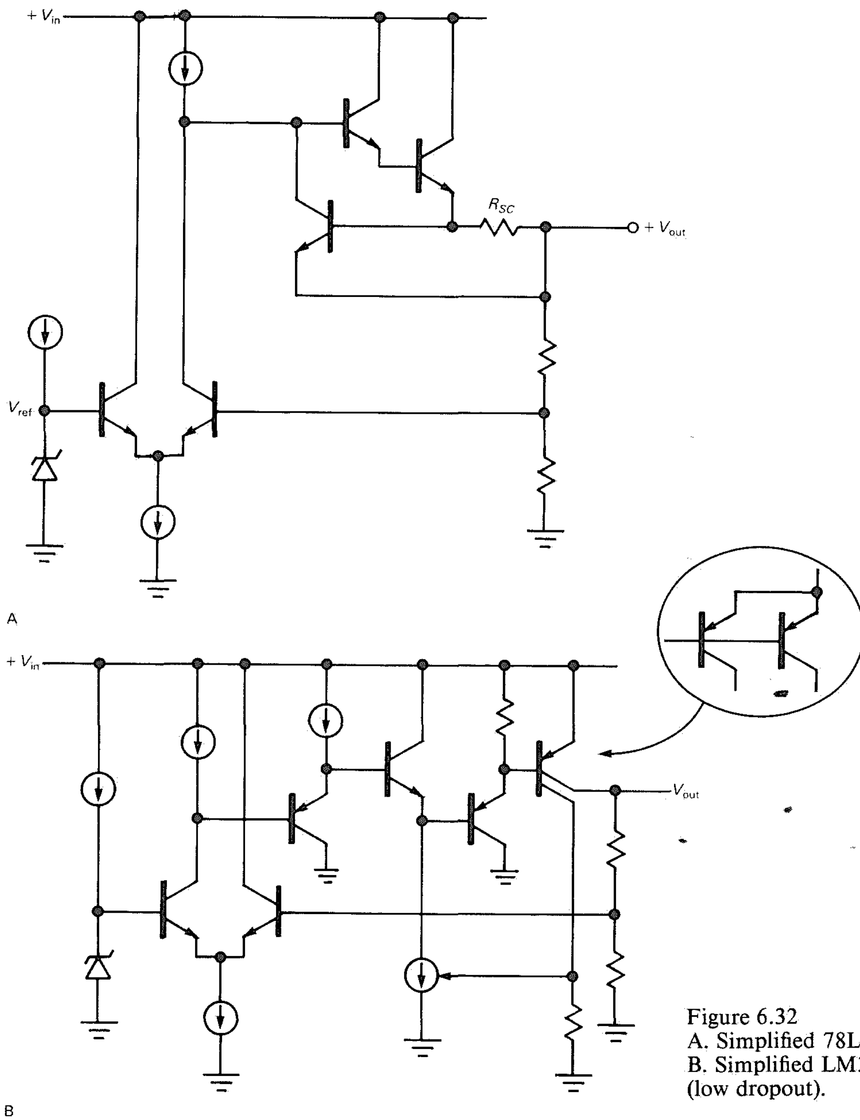


Figure 6.32
A. Simplified 78L*xx*.
B. Simplified LM330 (low dropout).

(150mA) regulator. The output can be brought within a saturation voltage of the unregulated input voltage by the *pnp* pass transistor. Having thus eliminated the Darlington V_{BE} drops of the *npn* regulator circuit, the designers weren't about to waste a diode drop with the usual (series resistor) short-circuit protection scheme. So they used a clever trick, deriving a sample of the output current via a second collector: That current is a fixed fraction of the output current and is used to shut off base drive as shown. This current-limit scheme is not particularly precise (I_L is specified as 150mA min, 700mA max), but it's good enough to protect the regulator, which also has internal thermal limiting.

Low-dropout regulators are available in most of the popular types, for example 3-terminal fixed voltage [LM2931, LM330, LT1083/4/5 (5V and 12V), TL750], 3-terminal adjustable (LT1083/4/5, LM2931), and micropower (LP2950/1, MAX664, LT1020). Tables 6.8 and 6.9 include all low-dropout regulators available at the time of writing.

Processor-oriented regulators

Electronic devices that include microprocessors (the subject of Chapters 10 and 11) require more than a simple regulated voltage. In order to retain the contents of volatile memory (and in order to keep track of elapsed time) they need a separate source of low-current dc when the regular power source is off; this may happen because the device is shut off, or because of a power failure. They also need to know when ordinary power has resumed, so they can "wake up" in a known state. Furthermore, a microprocessor-based device may need to have a few milliseconds warning that normal power is about to fail, so it can put data into safe memory.

Until recently you had to design extra circuitry to do these things. Now life is easy – you can get "(micro)processor-oriented" regulator ICs, with various combinations of these functions built in. These ICs sometimes go by the name of "power supply supervisory chips" or "watchdog" chips. An example is the LM2984, which has two high-current +5 volt outputs (one for the microprocessor, one for other circuitry) and a low-current +5 volt output (for memory), a delayed RESET flag output to initialize your microprocessor when power resumes, and an ON/OFF controlling input for the high-current outputs. It also has an input that monitors microprocessor activity, resetting the processor if it grinds to a halt. An example of a watchdog chip *without* regulator is the MAX691 from Maxim, which monitors the regulated supply voltage and microprocessor activity, and provides reset (and "interrupt") signals to the microprocessor, just like the LM2984. However, it adds both power-fail *warning* and battery switchover circuitry to the other capabilities of the LM2984. Used with an ordinary +5 volt regulator, the MAX691 does everything you need to keep a microprocessor happy. We'll learn much more about the care and feeding of microprocessors in Chapters 10 and 11.

□ Micropower regulators

As we suggested earlier, most regulator chips draw a few milliamps of quiescent current to run the internal voltage reference and error amplifiers. That's no problem for an instrument powered from the ac mains, but it's undesirable in a battery-operated instrument, powered by a 400mA-hour 9 volt alkaline battery, and it's intolerable in a micropower instrument that must run a thousand hours, say, on a battery.

The solution is a micropower regulator. The most miserly of these are the ICL7663/4, positive and negative adjustable regulators with quiescent currents of 4μA. At that current a 9 volt battery would last 100,000 hours (more than 10 years), which exceeds the "shelf life" (self-discharge time) of all batteries except some lithium-based types. Micropower design is challenging and fun, and we'll tell you all about it in Chapter 14.

Dual-polarity regulated supplies

Most of our op-amp circuits in Chapter 4 ran from symmetrical bipolarity supplies, typically ±15 volts. That's a common requirement in analog circuit design, where you often deal with signals near ground, and the simplest way to generate symmetrical split supplies is to use a pair of 3-terminal regulators. For example, to generate regulated ±15 volts, you could use a 7815 and a 7915, as in Figure 6.33A. We tend to favor the use of adjustable 3-terminal regulators, because (a) you only need to stock one type for each polarity and current range, and (b) you can trim the voltage exactly, if needed; Figure 6.33B shows how the circuit looks with a 317 and 337.

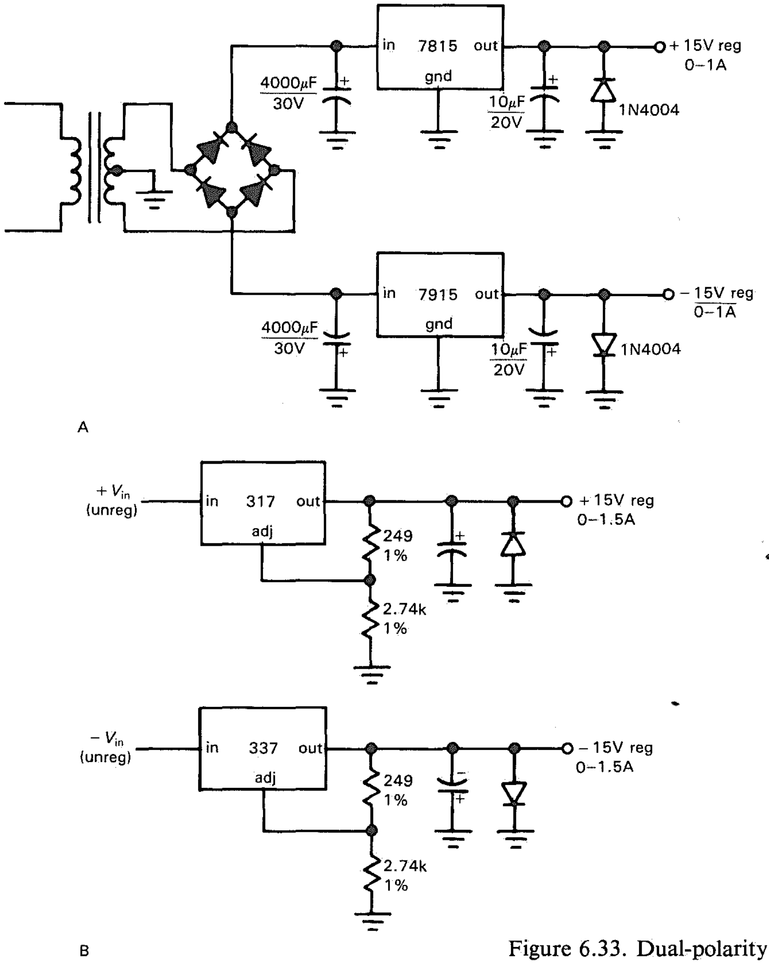


Figure 6.33. Dual-polarity regulated supplies.

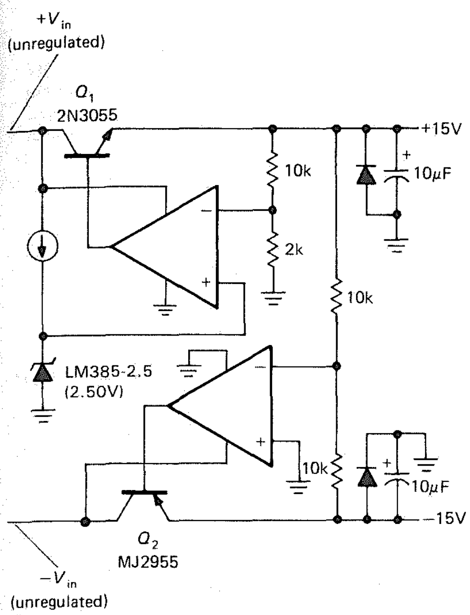


Figure 6.34. Dual-tracking regulator.

□ *Dual tracking regulators.* Given the need for regulated split supplies, you might wonder why there aren't "dual 3-terminal regulators." Wonder no more – they exist and are known as "dual tracking regulators." To understand why they have such a complicated name, take a look at Figure 6.34, which shows the classic dual tracking regulator circuit. Q_1 is the pass transistor for a conventional positive regulated supply. The positive regulated output is then simply used as the reference for a negative supply. The lower error amplifier controls the negative output by comparing the average of the two output voltages with ground, thus giving equal 15 volt positive and negative regulated outputs. The positive supply can be any of the configurations we have already talked about; if it is an adjustable regulator, the negative output follows any changes in the positive regulated output. In practice, it is wise to include current-limiting circuitry, not shown in Figure 6.34 for simplicity.

As with single-polarity regulators, dual-tracking regulators are available as complete integrated circuits in both fixed and adjustable versions, though in considerably less variety. Table 6.10 lists the characteristics of most types now available. Typical are the 4194 and 4195 regulators from Raytheon, which are used as shown in Figure 6.35. The 4195 is factory-trimmed for ±15 volt outputs, whereas the 4194's symmetrical outputs are adjustable via the single resistor R_1. Both regulators are available in power packages as well as the small DIP packages, and both have internal thermal shutdown and current limiting. For higher output currents you can add outboard pass transistors (see below).

Many of the preceding regulators (e.g., the 4-terminal adjustable regulators) can be connected as dual-tracking regulators. The manufacturers' data sheets often give suggested circuit configurations. It is worth keeping in mind that the idea of referencing one supply's output to another supply can be used even if the two supplies are not of equal and opposite voltages. For instance, once you have a stable +15 volt supply, you can use it to generate a regulated +5 volt output, or even a regulated −12 volt output.

EXERCISE 6.6
Design a ±12 volt regulator using the 4194.

Reverse-polarity protection. An additional caution with dual supplies: Almost any electronic circuit will be damaged extensively if the supply voltages are reversed. The only way that can happen with a single supply is if you connect the wires backward; sometimes you see a high-current rectifier connected across the circuit in the reverse direction to protect against this error. With circuits that use several supply voltages (a split supply, for instance), extensive damage can result if there is a component failure that shorts the two supplies together; a common

TABLE 6.10. DUAL-TRACKING REGULATORS

Type	Pkg	V_{out} (V)	Adj. output	Balance trim	Adj. cur. limit	Thermal limit	Max V_+-V_- Input (V)	Maximum output current[a] (each supply): @75°C case I_{out} (mA)	Maximum output current[a] (each supply): No sink[b] I_{out} (mA)	Maximum output current[a] (each supply): No sink[b] P_{diss} (W)	Regulation typ: Load[c] (mV)	Regulation typ: Line[d] (mV)	Θ_{JC} (°C/W)	120Hz ripple reject typ (dB)	Temp stab[e] typ (mV)	Noise[f] (μV rms)
Motorola																
MC1468L	DIP	±15	•	•	•	–	60	55	30	0.5	10[m]	10[m]	50	75	45	100
MC1468R	TO-66	±15	•	•	•	–	60	100	65	1.2	10[m]	10[m]	17	75	45	100
National																
LM325H[g]	TO-5	±15	–	–	•	•	60	100	50	0.5	6	2	12	75	45	150
LM325N[g]	DIP	±15	–	–	•	•	60	–	50	0.5	6	2	90[h]	75	45	150
LM326H[g]	TO-5	adj	–	–	•	•	60	100	70	0.5	6	2	12	75	35	100
LM326N[g]	DIP	adj	–	–	•	•	60	–	70	0.5	6	2	90[h]	75	35	100
Raytheon																
RC4194DB	DIP	adj	•	•	–	•	70	30[i]	25[i]	0.5	0.1%	0.2%	160[h]	70	0.2%	250[j]
RC4194TK	TO-66	adj	•	•	–	•	70	250[i]	90[i]	1.8	0.2%	0.2%	7	70	0.2%	250[j]
RC4195NB	miniDIP	±15	–	•	–	•	60	–	20	0.35	2	2	210[h]	75	75	60
RC4195TK	TO-66	±15	–	•	–	•	60	150	70	1.2	3	2	11	75	75	60
Silicon General																
SG3501AN	DIP	±15	–	•	–	•	60	60	30	0.6	30	20	125[h]	75	150	50
SG3502N	DIP	adj	•	•	•	•	50	50[i]	30	0.6	0.3%	0.2%	125[h]	75	1%	50

(a) V_{in}=1.6V_{out} (each supply). (b) for 50°C ambient. (c) 10% to 50% I_{max}. (d) for ΔV_{in}=15V. (e) ΔV_{out} for 0°C to 100°C T_J. (f) 100Hz to 10kHz. (g) intended for use with a pair of external pass transistors. (h) Θ_{JA}. (i) 10V drop (each supply). (j) 10Hz to 100kHz. (m) max.

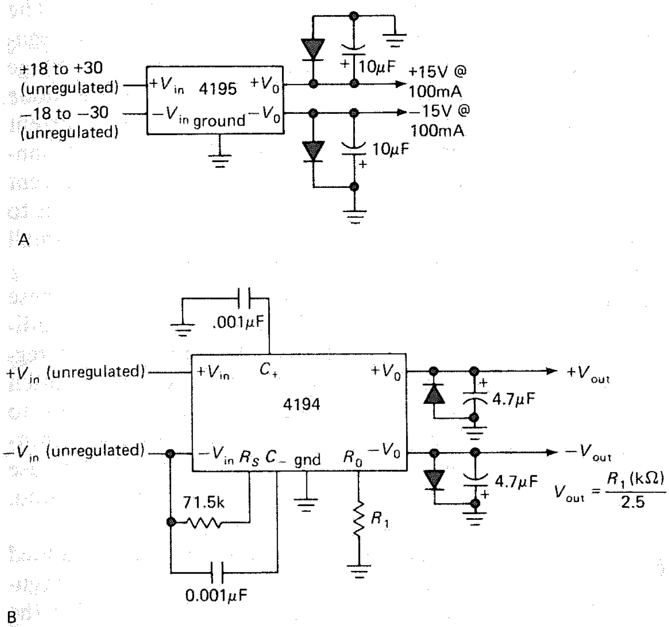


Figure 6.35

situation is a collector-to-emitter short in one transistor of a push-pull pair operating between the supplies. In that case the two supplies find themselves tied together, and one of the regulators will win out. The opposite supply voltage is then reversed in polarity, and the circuit starts to smoke. For this reason it is wise to connect a power rectifier (e.g., a 1N4004) in the reverse direction from each regulated output to ground, as we drew in Figure 6.33.

Outboard pass transistors

Three-terminal fixed regulators are available with 5 amps or more of output current, for example the adjustable 10 amp LM396. However, such high current operation may be undesirable, since the maximum chip operating temperature for these regulators is lower than for power transistors, mandating oversize heat sinks. Also, they are expensive. An alternative solution is the use of external pass transistors, which can be added to the 3- and 4-terminal regulators (and dual-tracking regulators) just as with the classic 723. Figure 6.36 shows the basic circuit.

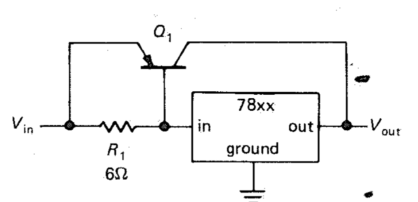


Figure 6.36. Three-terminal regulator with current-boosting outboard transistor.

The circuit works normally for load currents less than 100mA. For greater load currents, the drop across R_1 turns on Q_1, limiting the actual current through the 3-terminal regulator to about 100mA. The 3-terminal regulator maintains the output at the correct voltage, as usual, by reducing input current and hence drive to Q_1 if the output voltage rises, and vice versa. It never even realizes the load is drawing more than 100mA! With this circuit the input voltage must exceed the output voltage by the dropout voltage of the 78xx (2V) plus a V_{BE} drop.

In practice, the circuit must be modified to provide current limiting for Q_1, which could otherwise supply an output current equal to h_{FE} times the regulator's internal current limit, i.e., 20 amps or more! That's enough to destroy Q_1, as well as the unfortunate load that happens to be connected at the time. Figure 6.37 shows two methods of current limiting.

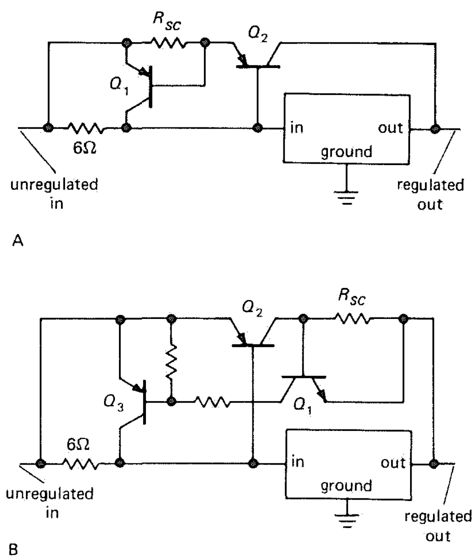


Figure 6.37. Current-limit circuits for outboard transistor booster.

In both circuits, Q_2 is the high-current pass transistor, and its emitter-to-base resistor has been chosen to turn it on at 100mA load current. In the first circuit, Q_1 senses the load current via the drop across R_{SC}, cutting off Q_2's drive when the drop exceeds a diode drop. There are a couple of drawbacks to this circuit: The input voltage must now exceed the regulated output voltage by the dropout voltage of the 3-terminal regulator plus two diode drops, for load currents near the current limit. Also, Q_1 must be capable of handling high currents (equal to the current limit of the regulator), and it is difficult to add foldback limiting because of the small resistor values required in Q_1's base.

The second circuit helps solve these problems, at the expense of some additional complexity. With high-current regulators, a low dropout voltage is often important to reduce power dissipation to acceptable levels. To add foldback limiting to the latter circuit, just tie Q_1's base to a divider from Q_2's collector to ground, rather than directly to Q_2's collector.

External pass transistors can be added to the adjustable 3- and 4-terminal regulators in exactly the same way. See the manufacturers' data sheets for further details.

Current source

A 3-terminal adjustable regulator makes a handy high-power constant-current source. Figure 6.38 shows one to source 1 amp. The addition of an op-amp follower, as in the second circuit, is necessary if the circuit is used to source small currents, since the "ADJ" (adjust) input contributes a current error of about 50μA. As with the previous regulators, there is on-chip current limit, thermal-overload protection, and safe operating area protection.

EXERCISE 6.7

Design an adjustable current source for output currents from 10μA to 1mA using a 317. If $V_{in} = +15$V, what is the output compliance? Assume a dropout voltage of 2 volts.

Note that the current source in Figure 6.38A is a 2-terminal device. Thus, the load can be connected on either side. The figure shows how you might connect things

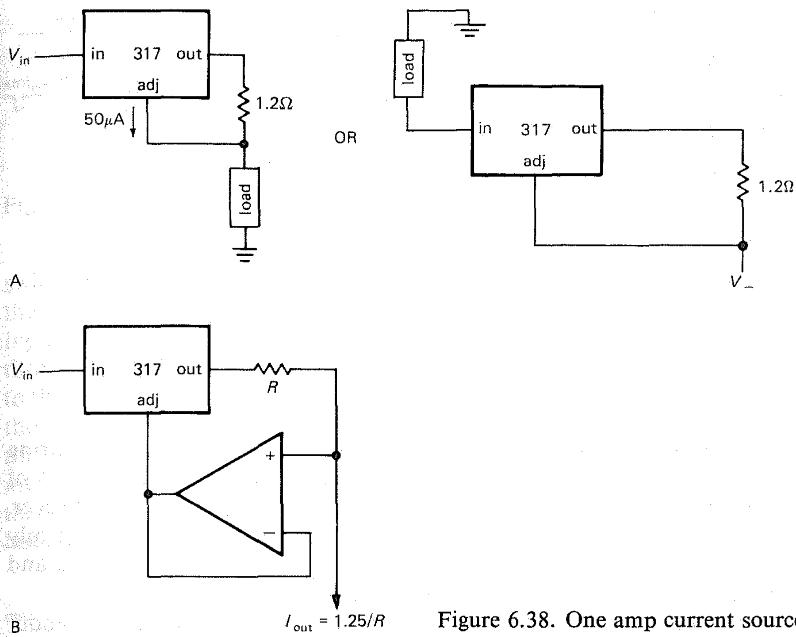


Figure 6.38. One amp current sources.

to *sink* current from a load returned to ground (of course, you could always use the negative-polarity 337, in the configuration analogous to Fig. 6.38A).

National makes a special 3-terminal device, the LM334, optimized for use as a low-power current source. It comes in the small plastic transistor package (TO-92), as well as the standard DIP IC package. You can use it all the way down to 1μA, because the "adj" current is a small fraction of the total current. It has one peculiarity, however: The output current is temperature-dependent, in fact, precisely proportional to absolute temperature. So although it is not the world's most stable current source, you can use it (Section 15.01) as a temperature sensor!

6.19 Switching regulators and dc-dc converters

All the voltage regulator circuits we have discussed so far work the same way: A linear control element (the "pass transistor") in series with the unregulated dc is used, with feedback, to maintain constant output voltage (or perhaps constant current). The output voltage is always lower in voltage than the unregulated input voltage, and some power is dissipated in the control element [the average value of $I_{out}(V_{in} - V_{out})$, to be precise]. A minor variation on this theme is the *shunt regulator*, in which the control element is tied from the output to ground, rather than in series with the load; the simple resistor-plus-zener is an example.

There is another way to generate a regulated dc voltage, fundamentally different from what we've seen so far; look at Figure 6.39. In such a *switching regulator* a transistor operated as a saturated switch periodically applies the full unregulated voltage across an inductor for short intervals. The inductor's current builds up during each pulse, storing $\frac{1}{2}LI^2$ of energy in its magnetic field; the stored energy is transferred to a filter capacitor at the output,

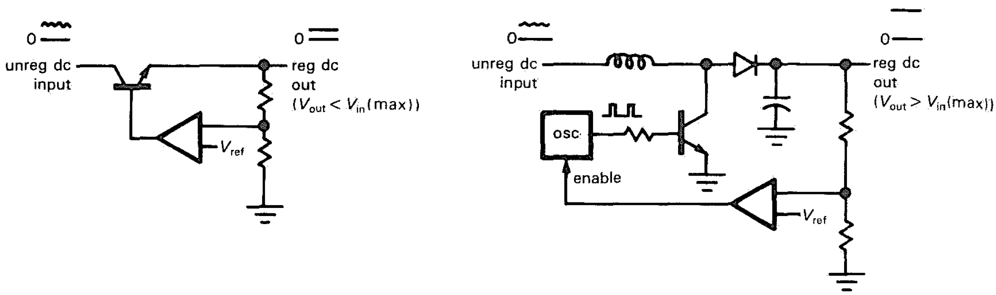


Figure 6.39. Two kinds of regulators.
A. Linear (series).
B. Step-up switcher.

which also smooths the output (to carry the output load between charging pulses). As with a linear regulator, feedback compares the output with a voltage reference – but in a switching regulator it controls the output by changing the oscillator's pulse width or switching frequency, rather than by linearly controlling the base or gate drive.

Switching regulators have unusual properties that have made them very popular: Since the control element is either off or saturated, there is very little power dissipation; switching supplies are thus very efficient, even if there is a large drop from input to output. Switchers (slang for "switching power supplies") can generate output voltages *higher* than the unregulated input, as in Figure 6.39B; they can just as easily generate outputs *opposite in polarity* to the input! Finally, switchers can be designed with no dc path from input to output; that means they can run directly from the rectified power line, with no ac power transformer! The result is a very small, lightweight, and efficient dc supply. For these reasons, switching supplies are used almost universally in computers.

Switching supplies have their problems, too. The dc output has some switching "noise," and they can put hash back onto the power line. They used to have a bad reputation for reliability, with occasional spectacular pyrotechnic displays during episodes of catastrophic failure. Most of these problems have been solved, however, and the switching supply is now firmly entrenched in electronic instruments and computers.

In this section we'll tell you all about switching supplies, in two steps: First, we'll describe the basic switching regulator, operating from a conventional unregulated dc supply. There are three circuits, used for (a) step-down (output voltage less than input), (b) step-up (output voltage greater than input), and (c) inverting (output polarity opposite to input). Then we'll take a radical step, describing the heretical (and most widely used!) designs that run straight from the rectified ac power line, without an isolation transformer. Both kinds of power supplies are in wide use, so our treatment is practical (not just pedagogically pleasing). Finally, we'll give you plenty of advice on the subject: When to use switchers, when to avoid them; when to design your own, when to buy them. With characteristic overstatement, we won't leave you in any doubt!

□ *Step-down regulator*

Figure 6.40 shows the basic step-down (or "bucking") switching circuit, with feedback omitted for simplicity. When the MOS

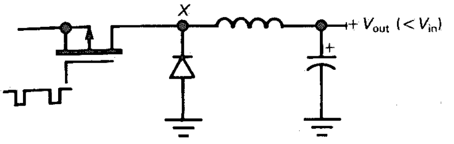


Figure 6.40. Step-down switcher.

switch is closed, $V_{out} - V_{in}$ is applied across the inductor, causing a linearly increasing current (recall $dI/dt = V/L$) to flow through the inductor. (This current flows to the load and capacitor, of course.) When the switch opens, inductor current continues to flow in the same direction (remember that inductors don't like to change their current suddenly, according to the last equation), with the "catch diode" now

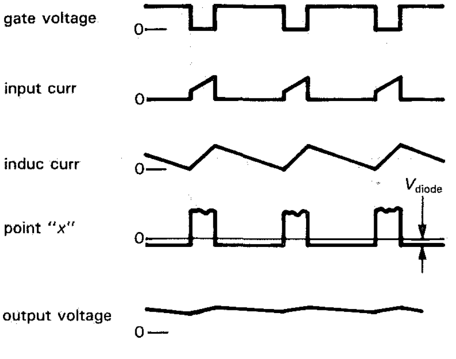


Figure 6.41

conducting to complete the circuit. The output capacitor acts as an energy "flywheel," smoothing the inevitable sawtooth ripple (the larger the capacitor, the less the ripple). The inductor current now finds fixed voltage V_{out} − 0.6V across it, causing its current to decrease linearly. Figure 6.41 shows the corresponding voltage and current waveforms. To complete the circuit as a *regulator*, you would of course add feedback, controlling either the pulse width (at constant pulse repetition rate) or the repetition rate (with constant pulse width) from an error amplifier that compares the output voltage with a reference.

Figure 6.42 shows a low-current +5 volt regulator using the MAX638 from Maxim. This nice chip gives you a choice of fixed +5 volt output (no external divider needed) or adjustable positive output, with external resistive divider. It includes nearly all components in a convenient mini-DIP package. In the MAX638 the oscillator runs at a constant 65kHz, with the error amplifier either permitting or cutting off gate drive pulses, according to the output voltage. The circuit shown gives about 85% efficiency, pretty much independent of the input voltage. Compare that with a linear regulator by doing the next problem:

EXERCISE 6.8

What is the maximum theoretical efficiency of a linear (series pass) regulator, when used to generate regulated +5 volt from a +12 volt unregulated input?

EXERCISE 6.9

What does a step-down regulator's high efficiency imply about the ratio of output current to input current? What is the corresponding ratio of currents, for a linear regulator?

□ *Step-up regulator; inverting regulator*

Apart from its high efficiency, the step-down switching regulator of the previous paragraph has no significant advantage (and some significant disadvantages – component count, switching noise) over a linear regulator. However, when there is a need for an output voltage greater than the unregulated input, or for an output voltage of opposite polarity to the unregulated input, switching supplies become very attractive indeed. Figure 6.43 shows the basic step-up (or "boosting") and inverting (sometimes called "flyback") circuits.

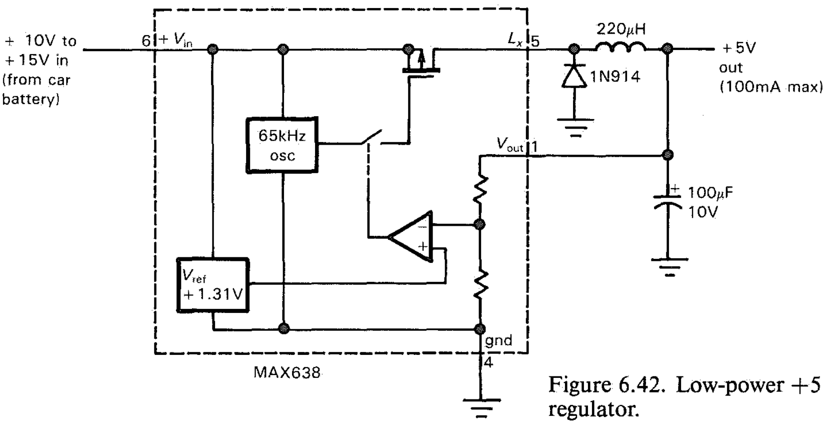


Figure 6.42. Low-power +5 volt switching regulator.

We showed the step-up circuit of Figure 6.39A earlier, in comparison with the linear regulator. The inductor current ramps up during switch conduction (point X near ground); when the switch is turned off, the voltage at point X rises rapidly as the inductor attempts to maintain constant current. The diode turns on, and the inductor dumps current into the capacitor. The output voltage can be much larger than the input voltage.

EXERCISE 6.10
Draw waveforms for the step-up switcher, showing voltage at point X, inductor current, and output voltage.

EXERCISE 6.11
Why can't the step-up circuit be used as a step-down regulator?

The inverting circuit is shown in Figure 6.43B. During switch conduction, a linearly increasing current flows from point X to ground. To maintain the current when the switch is open, the inductor pulls point X negative, as much as needed to maintain current flow. Now, however, the current is flowing *into* the inductor from the filter capacitor. The output is thus negative, and its average value can be larger or smaller in magnitude than the input (as determined by feedback); in other words, the inverting regulator can be either step-up or step-down.

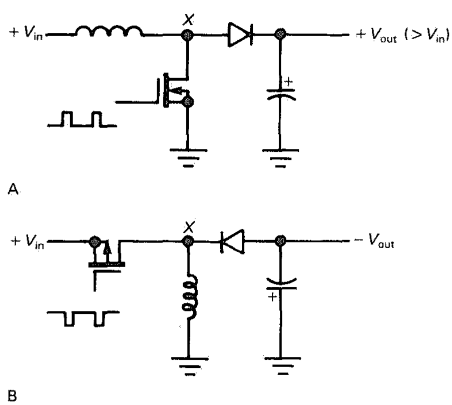


Figure 6.43. Two switching-element configurations.
A. Step-up ("boost").
B. Inverting.

EXERCISE 6.12
Draw waveforms for the inverting switcher, showing voltage at point X, inductor current, and output voltage.

Figure 6.44 shows how you might use low-power switching regulators to generate ±15 volt op-amp supply voltages from a single +12 volt automotive battery, a trick that is impossible with linear regulators. Here we've again used low-power

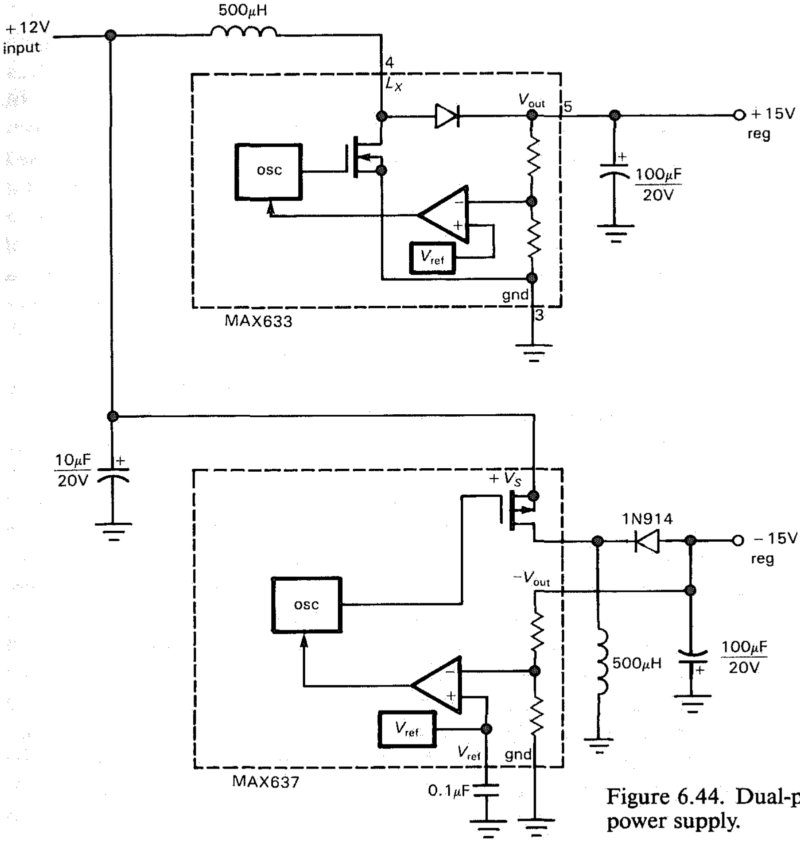


Figure 6.44. Dual-polarity switching power supply.

fixed-output ICs from Maxim, in this case the step-up MAX633 and the inverting MAX637. The external components shown were chosen according to the manufacturer's data sheets. They're not critical, but, as always in electronic design, there are trade-offs. For example, a larger value of inductor lowers the peak currents and increases the efficiency, at the expense of maximum available output current. This circuit is rather insensitive to input voltage, as long as it doesn't exceed the output voltage, and will work all the way down to +2 volts input, with greatly reduced maximum output current.

Before leaving the subject of inverting and step-up regulators, we should mention that there is one other way to accomplish the same goal, namely with "flying capacitors." The basic idea is to use MOS switches to (a) charge a capacitor from the dc input, and then (b) change the switches to connect the now-charged capacitor in series with another (step-up), or with reversed polarity across the output (inverting). Flying-capacitor voltage converters (e.g., the popular 7662) have some advantages (no inductors) and some disadvantages (low power, poor regulation, limited voltage). We'll discuss them later in the chapter.

General comments on switching regulators

As we've seen, the ability of switchers to generate stepped-up or inverted outputs makes them quite handy for making, say, low-current ± 12 volt supplies on an otherwise all-digital $+5$ volt circuit board. You'll often need such bipolarity supplies to power "serial ports" (more in Chapters 10 and 11) or linear circuitry using op-amps or A/D (analog-to-digital) and D/A (digital-to-analog) converters. Another good use for step-up switchers is to power displays that require relatively high voltage, for example using fluorescent or plasma technology. In these applications, where the dc input (typically +5V) is already regulated, you often use the phrase "dc-to-dc converter," rather than "switching regulator," although it's really the same circuit. Finally, in battery-operated equipment you often want high efficiency over a wide range of battery voltage; for example, a 9 volt alkaline "transistor" battery begins life at about 9.5 volts, dropping steadily to about 6 volts at the end of its useful life. A +5 volt low-power step-down regulator maintains high efficiency, with current step-up over most of the battery's life.

Note that the inductor and capacitor in a switching regulator are not functioning as an LC filter. In the simple step-down regulator, that might seem to be true, but obviously a circuit that *inverts* a dc level is hardly a filter! The inductor is a lossless energy-storage device (stored energy = $\frac{1}{2}LI^2$), able to transform impedance in order to conserve energy. This is an accurate statement from a physicist's point of view, in which the magnetic field contains stored energy. We're more used to thinking of capacitors as energy storage devices (stored energy = $\frac{1}{2}CV^2$), which is their role in switching supplies, as in conventional series regulators.

A bit of nomenclature: You sometimes see the phrases "PWM switch-mode regulator" and "current-mode regulator." They refer to the particular way in which the switching waveform is modified according to the feedback (error) signal. In particular, PWM means pulse-width modulation, in which the error signal is used to control the conduction pulse width (at fixed frequency), whereas in current-mode control the error voltage is used to control the peak inductor current (as sensed by a resistor) via width on a pulse-by-pulse basis. Current-mode regulators have some significant advantages and are becoming more popular now that good current-mode controller ICs have become available.

Keep in mind, when considering any switching supply, the noise generated by the switching process. This takes three forms, namely (a) output ripple, at the switching frequency, typically of order 10mV–100mV peak-to-peak, (b) ripple, again at the switching frequency, impressed onto the input supply, and (c) radiated noise, at the switching frequency and its harmonics, from switching currents in the inductor and leads. You can get into plenty of trouble with switching supplies in a circuit that has low-level signals (say 100μV or less). Although an aggressive job of shielding and filtering may solve such problems, you're probably better off with linear regulators from the outset.

Line-powered switching supplies

As we have seen, switching supplies have high efficiency even when the output voltage is nowhere near the input voltage. It may help our understanding to think of the inductor as an "impedance converter," since the average dc output current can be larger (step-down) or smaller (step-up) than the average dc input current. This is in stark contrast to linear series regulators, where the average values of input and output currents are always equal (ignoring the quiescent current of the regulator circuitry, of course).

This leads to a radical idea: We could eliminate the heavy 60Hz step-down transformer if we ran the regulator directly from rectified (unregulated) and filtered ac power. Two immediate comments: (a) The dc input voltage will be approximately 160 volts (for 115V ac power) – this is a dangerous circuit to tinker with! (b) The absence of a transformer means that the dc input is not isolated from the power line. Thus, the switching circuit itself must be modified to provide isolation.

The usual way to isolate the switching circuit is to wind a secondary onto the energy-storage inductor and use an isolation device (either transformer or opto-isolator) to couple the feedback to the switching oscillator; see the simplified block diagram in Figure 6.45. Note that the oscillator circuitry is powered from the high-voltage unregulated dc, whereas the feedback control circuitry (error amplifier, reference) is powered from the regulated dc output. Sometimes an auxiliary low-current unregulated supply (with its own 60Hz low-voltage transformer) is used to power the control elements. The box labeled "isolation" is often a small pulse transformer, although optical isolation can also be used (more on this later).

It might seem as if the advantage of a transformerless unregulated supply is more than overcome by the need for at least *two* other transformers! Not so. The size of a transformer is determined by the core size, which decreases dramatically at high frequencies. As a result, line-powered switching supplies are much smaller and lighter than the equivalent linear supply; they also run cooler, owing to higher efficiency. For example, Power-One manufactures both kinds of supplies. Comparing their model F5-25 (5V, 25A) linear supply with their comparably priced SPL130-1005 (5V, 26A) switcher, we find that the switcher weighs 2.5 pounds, compared with 19 pounds for the linear, and occupies just one-fourth the volume. Furthermore, the switcher will run cool, while the 19-pound linear will run hot, dissipating up to 75 watts at full load.

□ *Real-world switcher example*

In order to give you a feel for the real complexity of line-powered switching supplies, we've reproduced in Figure 6.46 the complete schematic of a commercial switcher, in fact the power supply used by Tandy

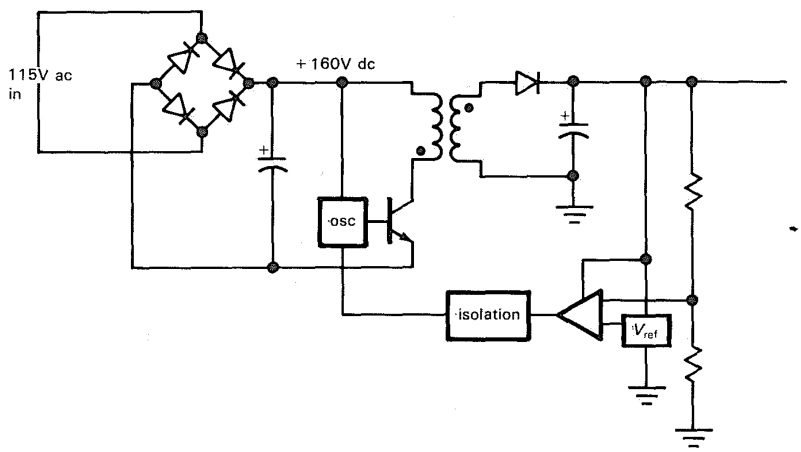


Figure 6.45. Direct ac-line-powered switching supply.

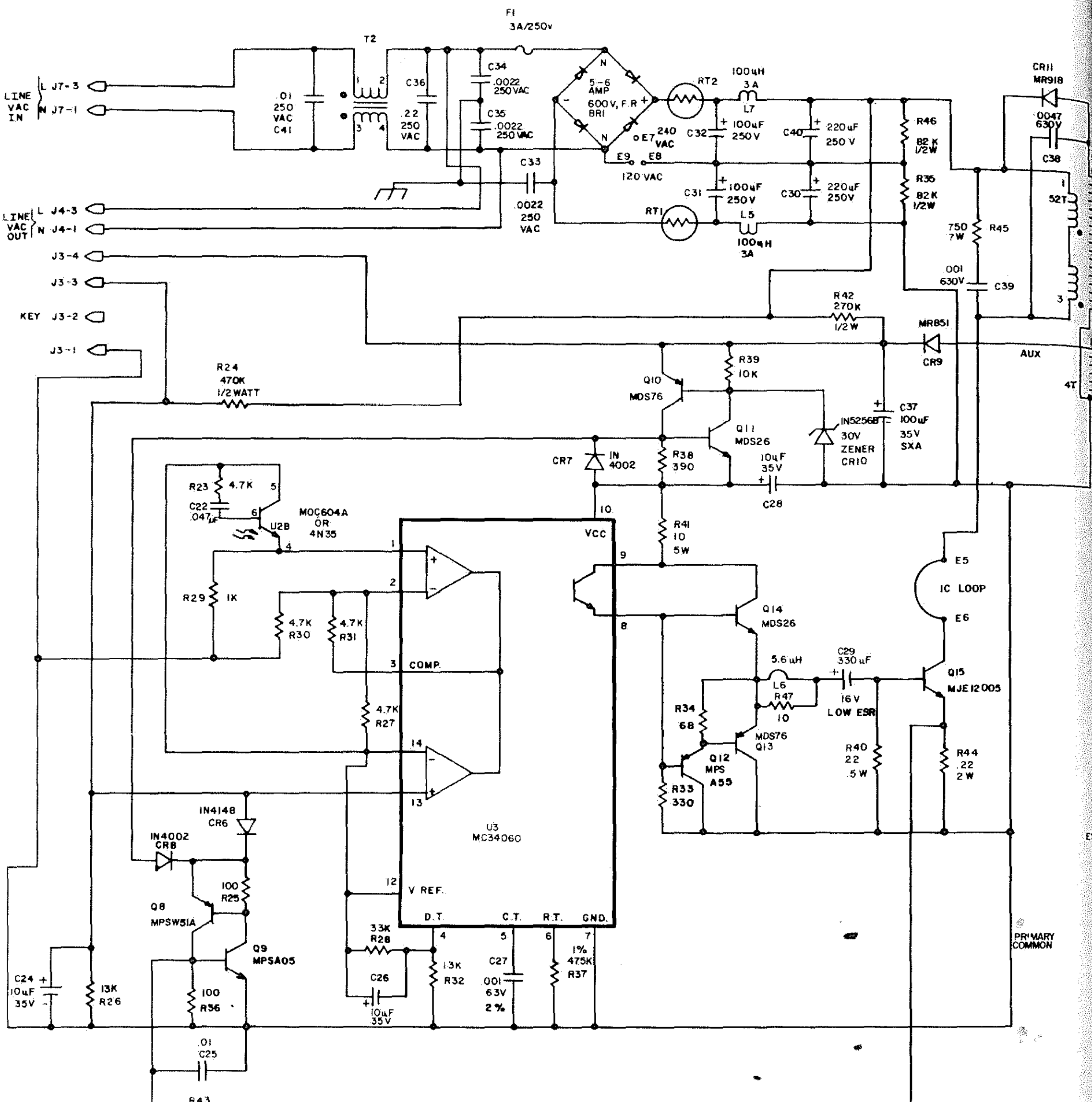
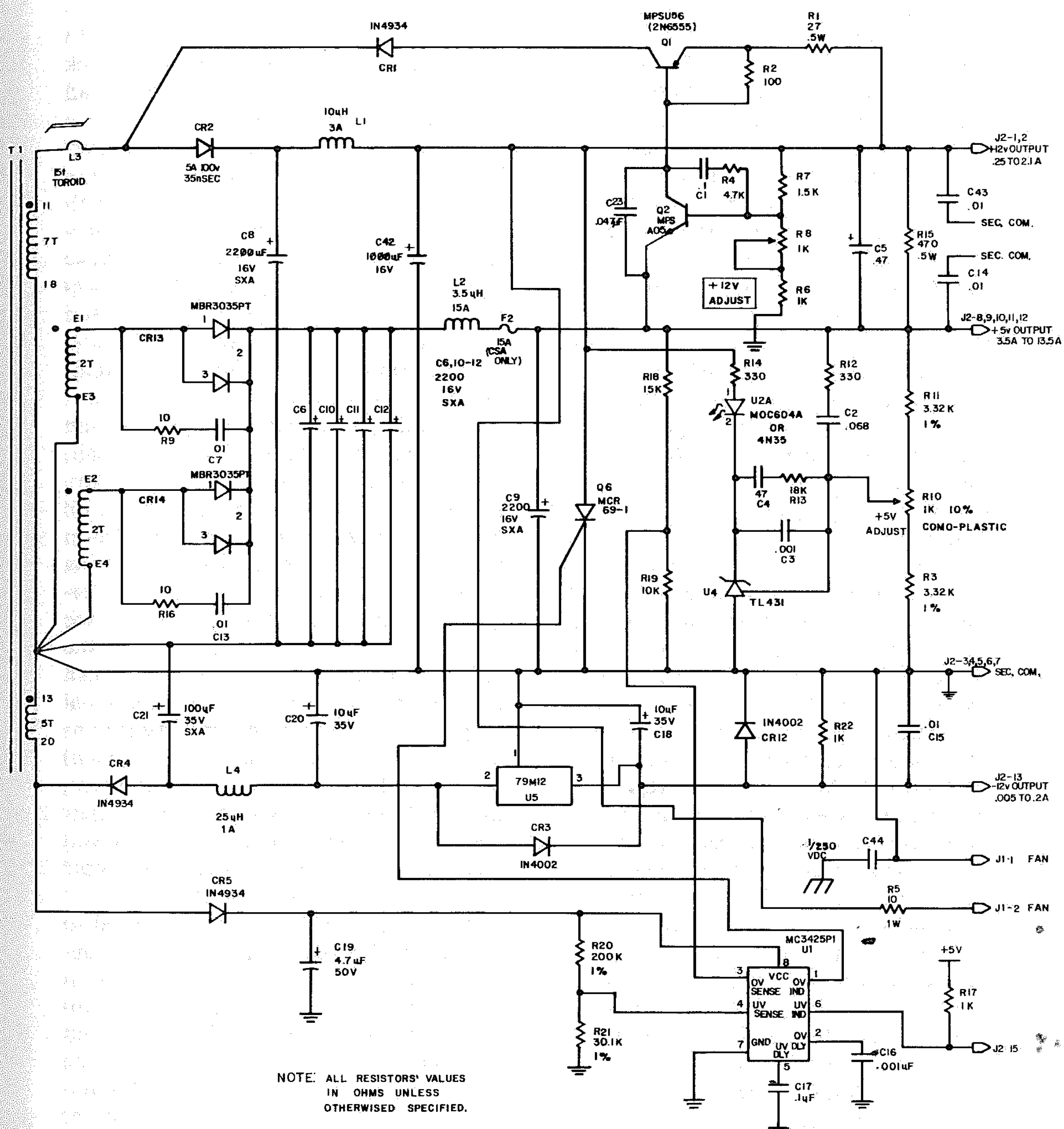


Figure 6.46. Switching power supply used in the Tandy model 2000 personal computer. Feedback from the +5 volt output is provided via opto-isolator U_{2A}-U_{2B}. (Courtesy of Tandy Corporation. Copyright 1984.)



NOTE: ALL RESISTORS' VALUES IN OHMS UNLESS OTHERWISED SPECIFIED.

(Radio Shack) to power their model 2000 personal computer. (We tried to get power-supply schematics from both IBM and Apple, but were ignored or haughtily rebuffed. Tandy, by comparison, publishes excellent documentation, with complete schematics and extensive circuit description.) It provides regulated outputs of +5 volts at 13 amps, +12 volts at 2.5 amps, and −12 volts at 0.2 amp (95W total), which are used to power the logic circuits and floppy-disk drives in the computer.

Let's take a walk through Figure 6.46 to see how a line-powered switcher copes with real-world problems. The circuit topology chosen by Tandy's designers is precisely that shown in Figure 6.45, though there are a few more components! Begin by comparing the figures: The line-powered bridge rectifier (BR_1) charges filter capacitors C_{30}, C_{31}, C_{32}, and C_{40} (T_2 is not a transformer – note the connections – but rather an interference filter). The charged capacitors are switched across the transformer primary (pins 1 and 3) by power transistor Q_{15}, whose switching waveform (a fixed-frequency square wave of variable pulse width) is provided by IC U_3 (a "PWM switch-mode regulator"). The secondary winding (there are actually three windings, one for each output voltage) is half-wave-rectified to generate the dc output: The +12 volts are produced by CR_2 from the 7-turn winding of pins 11 and 18, the −12 volts by CR_4 from the 5-turn winding of pins 13 and 20, and the +5 volts by the paralleled combination of CR_{13} and CR_{14}, each powered from its own (2-turn!) winding.

With multioutput switchers, only one output can be used for voltage-regulating feedback. It is conventional to use the +5 volts logic supply for this purpose, as Tandy has done here: R_{10} selects a fraction (nominally 50%) of the +5 volt output to compare with U_4's internal +2.5 volt reference, turning on photodiode U_{2a} if the output is too high. This photodiode couples optically to phototransistor U_{2b}, which varies the pulse width of U_3 to maintain +5 volts output. Thus the block labeled "isolation" in Figure 6.45 is an opto-coupler (see Section 9.10).

At this point we have accounted for perhaps 25% of the components in Figure 6.46. The rest are needed to cope with problems such as (a) short-circuit protection, (b) overvoltage and undervoltage shut-down, (c) auxiliary power for the regulating circuitry, (d) ac power filtering, and (e) linear post-regulation of the (tracking) ±12 volt supplies. Let's explore the circuit in some more detail.

Beginning at the ac line input, we find four capacitors and a series inductor pair, together forming an RFI filter. It's always a good idea, of course, to clean up the ac power entering an instrument (see Section 6.11); here, however, the careful filtering is additionally needed to keep radiofrequency hash generated *inside* the machine (mostly from the switching action in the power supply) from radiating *out* through the power line. Note next the optional jumper at E_8E_9, which converts the input from full-wave bridge (jumper open) to full-wave doubler (jumper shorted); manufacturers who wish to export their electronic wares must provide 110/220 volt compatibility, which is remarkably simple in the case of switching supplies.

Thermistors RT_1 and RT_2 are used to limit the high inrush current when the supply is first switched on, at which point the power line sees a few hundred microfarads of uncharged capacitor. Without the thermistors (or some other trick) the inrush current can easily exceed 100 amperes! The thermistors provide an ohm or two of series resistance, dropping to near zero when they warm up. Even with thermistors, the inrush current is impressive: The power supply has a specified "Input Surge Current" of 70 amps, maximum.

The 100μH series inductors L_5 and L_7 in the unregulated supply further clean up transmitted switching hash, and the 82k shunt resistors (R_{35} and R_{46}) are "bleeders," to make sure the power-supply filter capacitors discharge fully after power is turned off. Some additional passive "snubber" components (C_{38}, C_{39}, and R_{45}) are used to damp the large voltage spikes that otherwise might destroy the switching transistor Q_{15}. CR_{11}'s function is more subtle – it cleverly returns unused transformer energy to filter capacitors C_{30} and C_{40}.

Moving down the page, we encounter some real trickery, namely the "auxiliary supply." The circuits need some low-voltage, low-current dc to run the PWM controller chip and associated circuits. One possibility is to use a separate little linear supply, with its own line-powered transformer, etc. However, the temptation is overwhelming to hang another small winding (with half-wave rectifier) on T_1, thus saving a separate transformer. That's what the designer has done here, with a 4-turn winding (pins 9 and 10), rectified and filtered by CR_9 and C_{37}. This simple supply generates a nominal 15 volt output.

Sharp-eyed readers will have noticed a flaw in this scheme: The circuit cannot start itself, since the auxiliary power is only present if the supply is already running! This turns out to be an old problem: Designers of television sets love to play the same trick, deriving all their low-voltage supplies from auxiliary windings on the high-frequency horizontal drive transformer. The solution is the so-called kick-start circuit, in which some of the unregulated dc is brought over to start the circuit; once going, the supply keeps itself going from its transformer-derived dc power. In this circuit the kick-start comes via R_{42}, which begins charging up C_{37} at power-on. Nothing happens until the capacitor reaches a diode drop above CR_{10}'s zener voltage, at which point the SCR-like combination of Q_{10} and Q_{11} is switched into conduction (figure out how that works), dumping C_{37}'s charge across C_{28}, thus momentarily powering the control circuitry (U_3 and all components to its left). Once the oscillation starts, CR_9 provides 15 volts with enough current to power the control circuitry continuously (which R_{42} cannot do).

Most of the components surrounding U_3 pander to its needs (C_{27} and R_{37}, for example, set the pulse repetition rate at 25kHz). At the input side, U_{2b} provides overall feedback to maintain the output at +5 volts, as described earlier. Q_8 and Q_9 are another SCR-like latch, this time triggered to shut down the oscillator (and the series latching switch $Q_{10}Q_{11}$) if driver Q_{15}'s emitter current (sensed by R_{44}) is excessive, for example if the power supply sees a short-circuit load. The series combination $R_{43}C_{25}$ provides a $1\,\mu s$ time constant so that the circuit is not triggered by switching spikes. The shutdown circuitry also derives an input from divider $R_{26}R_{24}$, quenching oscillation if the ac input drops below 90 volts ac. At the output side of the controller U_3, $Q_{12}-Q_{14}$ provide high-current push-pull drive to Q_{15}'s base from the single-ended on-chip *npn* driver transistor (figure out how). Note the "I_C loop," an accessible length of wire in Q_{15}'s collector, which lets you observe the current waveform on a 'scope by using a clip-on current probe (see, for example, the Tektronix catalog).

Things are considerably simpler on the output side of T_1. The +5 volt supply uses paralleled power Schottky diodes (CR_{13} and CR_{14}) for fast recovery and low forward drop (the MBR3035PT is rated at 30A average current with 20kHz drive, 35V reverse breakdown, and 0.5V typical forward drop at 10A), with "snubber networks" ($10\Omega/0.01\mu F$) to protect the diodes from high-voltage spikes. The "π-section" filter consists of $8800\mu F$ of input capacitance, a $3.5\mu H$ series inductor, and a $2200\mu F$ output capacitor. (The lower-current $\pm 12V$ outputs also use half-wave

Schottky rectifiers and π-section filters, with smaller-valued components.) This degree of filtering might seem extreme by linear regulator standards, but remember that there is no post-regulation – what comes out of the filter is the "regulated dc" – therefore lots of filtering is needed to reduce ripple, predominantly at the switching frequency, to the requisite $\approx$50mV or so at the output.

The +5 volt output is sensed via divider $R_3R_{10}R_{11}$, driving TI's TL431 "3-terminal zener" (U_4), which, in combination with a few resistors and capacitors for feedback compensation, provides isolated feedback via opto-coupler U_{2ab}. The +5 volt output is also sensed, via $R_{18}R_{19}$, to trigger the overvoltage-sensor IC (U_1: $V_{thresh} = +2.5V$); the latter drives the gate of SCR Q_6, which crowbars the +12 volt supply, shutting things down via current limiting in the primary side, as described earlier. U_1 is also wired to sense an *under*voltage condition, via its dedicated auxiliary power from CR_5 and C_{19}; the undervoltage signal (a saturated *npn* transistor to ground) is sent to the microprocessor, alerting the system to imminent power failure so that the program can be brought to an orderly shut-down during the few remaining milliseconds without loss of data.

The power-supply designers used a bit of trickery to improve regulation in the ±12 volt supplies, which otherwise ride virtually open-loop on what is basically a +5 volt supply. For the +12 volt supply they used the +5 volt output as a reference for error amplifier Q_2, which controls a "magnetic amplifier." The latter consists of series saturable reactor L_3, provided with an opposing "reset current" via Q_1. The reset current determines how many volt-seconds the inductor will block before reaching the state of magnetic saturation, in which it acts as a good conductor. A magnetic amplifier deserves its name, because a small control current modifies a large output current. Mag-amp controllers are available as complete integrated circuits, for example the UC3838 from Unitrode.

For the lower-current −12 volt supply the designers opted for the simpler solution of a linear 7912-type post-regulator, complete with diodes for protection against reverse polarity. Throughout, the designers have used bypass capacitors and bleeders on the dc outputs.

This power-supply circuit illustrates most of the details that seldom get mentioned in textbooks, but are essential in the real world. The extra component count in this circuit pays handsomely in ensuring a power supply that is robust under field conditions. Although this extra care in design might appear to be a display of unnecessary compulsiveness, in fact it is hard-nosed cost-effectiveness – each field failure under warranty costs the manufacturer at least a hundred dollars in real shipping and repair costs, not to mention the tarnished reputation produced by persistent failure.

General comments on line-powered switching power supplies

1. Line-powered switchers (also called "off-line" switchers, though we don't like the term) make excellent *high-power* supplies. Their high efficiency keeps them cool, and the absence of a low-frequency transformer makes them considerably lighter and smaller than the equivalent linear supply. As a result, they are used almost exclusively to power computers, even desktop personal computers. They are finding their way into other portable instruments, too, even such noise-sensitive applications as oscilloscopes.

2. Switchers are noisy! Their outputs have tens of millivolts of switching ripple at their outputs, they put garbage onto the power line, and they can even scream audibly! One cure for output ripple, if that's a problem, is to add an external high-current *LC* low-pass filter; alternatively, you can add a low-dropout linear post-regulator. Some dc-dc converters include this feature, as well as complete shielding and extensive input filtering.

3. Switchers with multiple outputs are available and are popular in computer systems. However, the separate outputs are generated from additional windings on a common transformer. Typically, feedback is taken from the highest current output (usually the +5V output), which means that the other outputs are not particularly well regulated. There is usually a "cross-regulation" specification, which tells, for example, how much the +12 volt output, say, changes when you vary the load on the +5 volt output from 75% of full load to either 50% or 100% of full load; a typical cross-regulation specification is 5%. Some multiple-output switchers achieve excellent regulation by using linear post-regulators on the auxiliary outputs, but this is the exception. Check the specs!

4. Line-powered switchers may have a minimum load current requirement. If your load-current may drop below the minimum, you'll have to add some resistive loading; otherwise the output may soar or oscillate. For example, the +5 volt, 26 amp switching supply above has a minimum load current of 1.3 amps.

5. When working on a line-powered switcher, *watch out!* Many components are at line potential and can be lethal. You can't clip the ground of your scope probe to the circuit without catastrophic consequences.

6. When you first turn on the power, the ac line sees a large discharged electrolytic filter capacitor across it (through a diode bridge, of course). The resulting "inrush" current can be enormous; for our Power One switcher it's specified as 17 amps, maximum (compared with a full-load input current of 1.6A). Commercial switchers use various "soft-start" tricks to keep the inrush current within civilized bounds. One method is to put a negative-tempco resistor (a low-resistance thermistor) in series with the input; another method is to actively switch out a small (10Ω) series resistor a fraction of a second after the supply is turned on.

7. Switchers usually include overvoltage "shut-down" circuitry, analogous to our SCR crowbar circuits, in case something goes wrong. However, this circuit often is simply a zener sensing circuit at the output that shuts off the oscillator if the dc output exceeds the trip point. There are imaginable failure modes in which such a "crowbar" wouldn't crowbar anything. For maximum safety you may want to add an autonomous outboard SCR-type crowbar.

8. Switchers used to have a bad reputation for reliability, but recent designs seem much better. However, when they decide to blow out, they sometimes do it with great panache! We had one blow its guts out in a "catastrophic deconstruction," spewing black crud all over its innards and innocent electronic bystanders as well.

9. Line-powered switchers are definitely complex and tricky to design reliably. You need special inductors and transformers (and lots of them; Fig. 6.46). Our advice is to avoid the design phase entirely, by *buying* what you need! After all, why build what you can buy?

10. A switching supply presents a peculiar load to the power line that drives it. In particular, an increase in line voltage results in a *decrease* in average current, because the switcher operates at roughly constant efficiency: That's a negative resistance load (averaged over the 60Hz wave), and it can cause some crazy effects. If there's a lot of inductance in the power line, the system may oscillate.

Advice

Luckily for you, we're not bashful about giving advice! Here it is:

1. For *digital* systems, you usually need +5 volts, often at high current (10A or more).

Advice: (a) Use a line-powered switcher. (b) Buy it (perhaps adding filtering, if needed).
2. For analog circuits with low-level signals (small-signal amplifiers, signals less than 100μV, etc.). *Advice:* Use a linear regulator; switchers are too noisy – they will ruin your life. *Exception:* For some battery-operated circuits it may be better to use a low-power dc-dc switching converter.
3. For high-power anything. *Advice:* Use a line-powered switcher. It's smaller, lighter, and cooler.
4. For high-voltage, low-power applications (photomultiplier tubes, flash tubes, image intensifiers, plasma displays). *Advice:* Use a low-power step-up converter.

In general, low-power dc-dc converters are easy to design and require few components, thanks to handy chips like the Maxim series we saw earlier. Don't hesitate to build your own. By contrast, high-power switchers (generally line-powered) are complex and tricky and extremely trouble-prone. If you must design your own, be careful, and test your design very thoroughly. Better yet, swallow your pride and buy the best switcher you can find.

SPECIAL-PURPOSE POWER-SUPPLY CIRCUITS

□ 6.20 High-voltage regulators

Some special problems arise when you design linear regulators to deliver high voltages. Since ordinary transistors typically have breakdown voltages of less than 100 volts, supplies to deliver voltages higher than that require some clever circuit trickery. This section will present a collection of such techniques.

□ *Brute force: high-voltage components*

Power transistors, both bipolar and MOSFET, are available with breakdown voltages to 1000 volts and higher, and they're not even very expensive. Motorola's MJ12005, for example, is an 8 amp *npn* power transistor with conventional (V_{CEO}) collector-to-emitter breakdown of 750 volts, and base back-biased breakdown (V_{CEX}) of 1500 volts; it costs less than 5 dollars in single quantities. Their MTP1N100 (similar to the European BUZ-50) is a 1 amp n-channel power MOSFET with 1000 volt breakdown and a price tag of a few dollars. Power MOSFETs in particular are often excellent choices for high-voltage regulators, owing to their excellent safe operating area (absence of thermally induced second breakdown).

By running the error amplifier near ground (the output-voltage-sensing divider gives a low-voltage sample of the output), you can build a high-voltage regulator with only the pass transistor and its driver seeing high voltage. Figure 6.47 shows the idea, in this case a +100 to +500 volt regulated supply using NMOS pass transistor and driver. Q_2 is the series pass transistor, driven by inverting amplifier Q_1. The op-amp serves as error amplifier, comparing an adjustable fraction of the output with a precision +5 volt reference. Q_3 provides current limiting by shutting off drive to Q_2 when the drop across the 33 ohm resistor equals a V_{BE} drop. The remaining components serve more subtle, but necessary, functions: The diode protects Q_2 from reverse gate breakdown if Q_1 decides to pull its drain down rapidly (while the output capacitor holds up Q_2's source). The various small capacitors in the circuit provide compensation, which is needed because Q_1 is operated as an inverting amplifier with voltage gain, thus making the op-amp loop unstable (especially considering the circuit's capacitive load). This circuit is an exception to the general rule that transistor circuits do not present a shock hazard!

We can't resist an aside here: In slightly modified form (reference replaced by

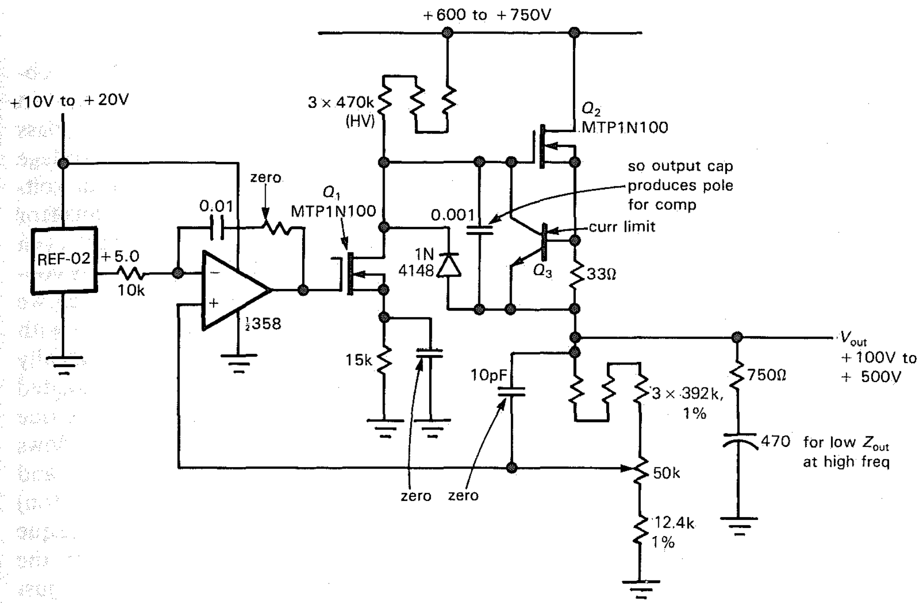


Figure 6.47. High-voltage regulated supply.

signal input) this circuit makes a very nice high-voltage amplifier, useful for driving crazy loads such as piezoelectric transducers. For that particular application the circuit must be able both to sink and to source current into the capacitive load. Oddly enough, the circuit acts like a "pseudo-push-pull" output, with Q_2 sourcing current and Q_1 sinking current (via the diode), as needed; see Section 3.14.

If a high-voltage regulator is designed to provide a fixed output only, the pass transistor may have a breakdown voltage less than the output voltage. In the preceding circuit, replacing the voltage-adjustment resistors with a fixed 12.4k resistor results in a fixed +500 volt regulator. A 300 volt pass transistor will then be fine, provided that the circuit ensures that the voltage across it never exceeds 300 volts, even during turn-on, turn-off, and output short-circuit conditions. The latter condition presents a challenge, but bridging Q_2 with a 300 volt zener may solve the problem. If the zener can handle high current, it can also protect the pass transistor against short-circuit loads, if suitable fusing is provided ahead of the regulator. The active zener circuit mentioned in Section 6.06 would be a good choice here.

□ *Regulating the ground return*

Figure 6.48 shows another way to regulate high voltages with low-voltage components. Q_1 is a series pass transistor, but it is connected in the low side of the supply; its "output" goes to ground. It has only a fraction of the output voltage across it, and it sits near ground, simplifying the driver circuitry. As before, protection must be provided during on/off transients and overloads. The simple zener protection shown is adequate, but remember that the zener must be able to handle the full short-circuit current.

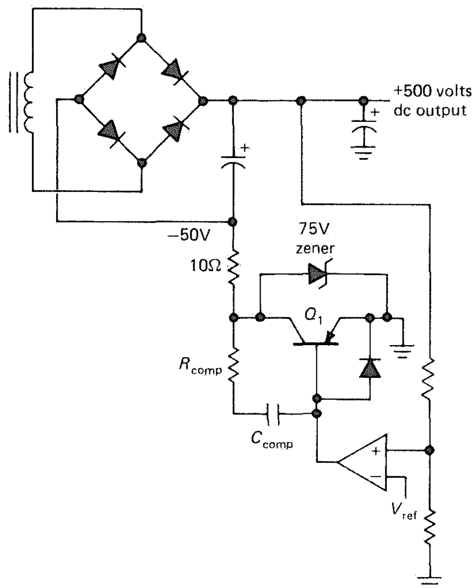


Figure 6.48. Regulating the ground return.

□ *Lifting the regulator above ground*

Another method sometimes used to extend the voltage range of regulators, including the simple 3-terminal type, is to raise the common terminal off ground with a zener (Fig. 6.49). In this circuit D_1 adds

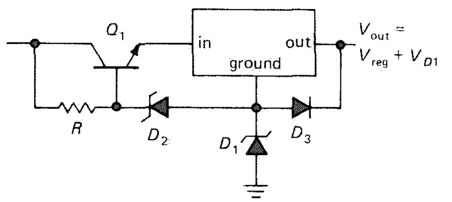


Figure 6.49

its voltage to the normal output of the regulator. D_2 sets the drop across the regulator via follower Q_1 and provides protection during short circuit because of D_3.

□ *Optically coupled transistor*

There is another way to handle the problem of transistor breakdown ratings in high-voltage supplies, especially if the pass transistor can be a relatively low voltage unit because of fixed (known) output voltage. In such cases only the driver transistor has to withstand high voltage, and even that can be avoided by using optically coupled transistors. These devices, which we will talk about further in connection with digital interfacing in Chapter 9, actually consist of two units electrically isolated from each other: a light-emitting diode (LED), which lights up when current flows through it in the forward direction, and a phototransistor (or photo-Darlington) mounted in close proximity in an opaque package. Running current through the diode causes the transistor to conduct, just as if there were base current. As with an ordinary transistor, you apply collector voltage to put the phototransistor in the active region. In many cases no separate base lead is actually brought out. Optically coupled devices are typically insulated to withstand several thousand volts between input and output.

Figure 6.50 shows a couple of ways to use an optically coupled transistor in a high-voltage supply. In the first circuit, phototransistor Q_2 shuts off pass transistor Q_3 when the output rises too high. In the second version, for which only the pass-transistor circuitry is shown, phototransistor Q_2 increases the output voltage when driven, so the error-amplifier inputs should be reversed. Both circuits generate some output current through the pass-transistor biasing circuit, so some load from output to ground is needed to keep the output voltage from rising under no-load conditions. The output-sensing voltage divider can do the job, or a separate "bleeder" resistor can be connected across the output, which is always a good idea anyway in a high-voltage supply.

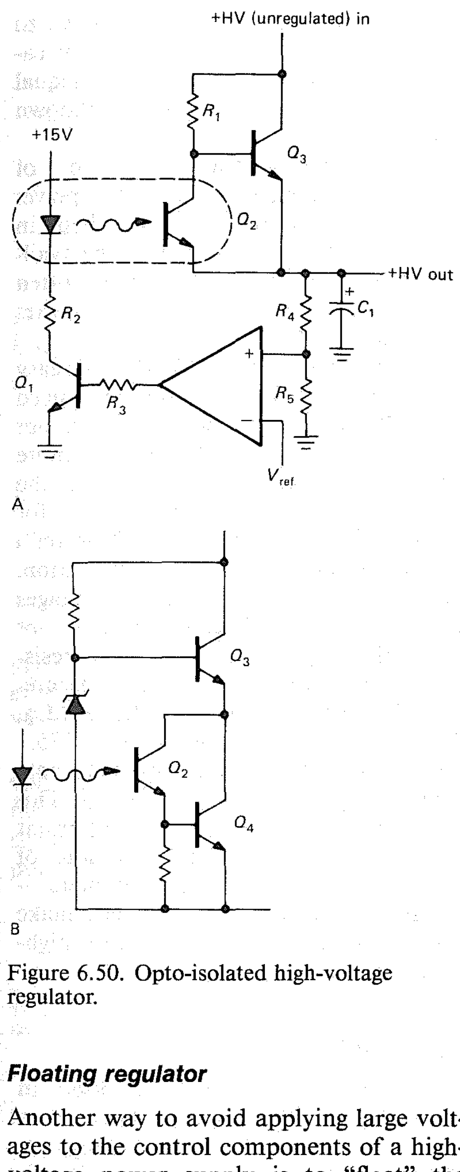


Figure 6.50. Opto-isolated high-voltage regulator.

□ *Floating regulator*

Another way to avoid applying large voltages to the control components of a high-voltage power supply is to "float" the control circuitry at the pass-transistor potential, comparing the drop across its own voltage reference with the drop down to ground. The excellent MC1466 regulator chip is intended for this kind of application, which normally requires an auxiliary low-current floating power supply to provide dc (20–30V) for the chip itself. The output voltage is limited only by the pass transistors and the isolation (transformer insulation breakdown voltage) of the auxiliary supply. The MC1466 features very good regulation and precise current-limiting circuitry and is well suited for accurate "laboratory" power supplies. A warning, however: The MC1466, unlike more recent regulator designs, does not include on-chip thermal protection.

An elegant way to rig up a floating regulator is provided by the LM10 op-amp plus voltage-reference combination, a remarkable breakthrough in chip technology from the legendary Widlar (see Section 4.13) that will operate from a single 1.2 volt supply. Such a chip can be powered from the base-emitter drop of a Darlington pass transistor! Figure 6.51 shows an example. If you like analogies, think of a giraffe who measures his height by looking at the distance to the ground, then stabilizes it by craning his neck accordingly. The TL783 from Texas Instruments is a 125 volt IC regulator that works this way; for lower-current applications it replaces the discrete circuitry of Figure 6.51.

□ *Transistors in series*

Figure 6.52 shows a trick for connecting transistors in series to increase the breakdown voltage. Driver Q_1 drives series-connected transistors Q_2 – Q_4, which share the large voltage from Q_2's collector to the output. The equal base resistors are chosen small enough to drive the transistors to full output current. The same circuit works with MOSFETs as well, but be sure to provide reverse-gate-protection diodes, as shown (you don't have to worry about *forward* gate breakdown, because the MOSFETs should turn on vigorously long before gate-channel breakdown). Note that the bias resistors produce some output current even when the transistors are cut off,

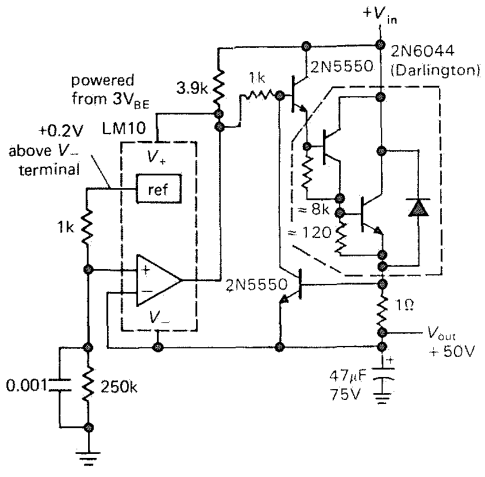


Figure 6.51. High-voltage floating regulator.

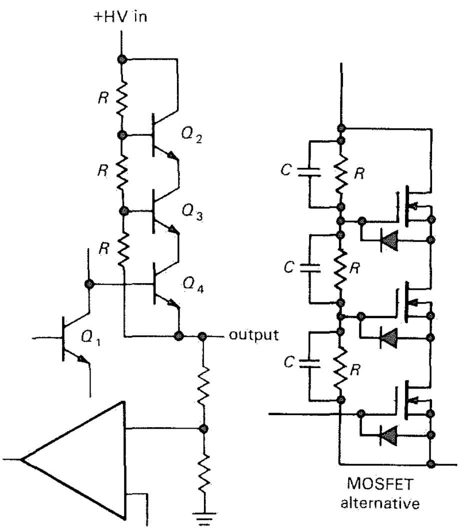


Figure 6.52. Connecting transistors in series to raise breakdown voltage.

so there must be a minimum load to ground to prevent the output from rising above its regulated voltage. It's often a good idea to parallel the divider resistors with small capacitors, as indicated, in order to maintain the divider action at high frequencies; choose a capacitor value large enough to swamp differences in transistor input capacitance, which otherwise cause unequal division, reducing overall breakdown voltage.

Series-connected transistors can, of course, be used in circuits other than power supplies. You'll sometimes see them in high-voltage amplifiers, although the availability of high-voltage MOSFETs often makes it unnecessary to resort to the series connection at all.

In high voltage circuits like this, it's easy to overlook the fact that you may need to use 1 watt (or larger) resistors, rather than the standard $\frac{1}{4}$ watt type. A more subtle trap awaits the unwary, namely the maximum *voltage* rating of 250 volts for standard $\frac{1}{4}$ watt composition ("carbon") resistors, regardless of power dissipation. Carbon resistors run at higher voltages show astounding voltage coefficients, not to mention permanent changes of resistance. For example, in an actual measurement (Fig. 6.53) a 1000:1 divider (10Meg, 10k) produced a division ratio of 775:1 (29% error!) when driven with 1kV; note that the *power* was well within ratings. This non-ohmic effect is particularly important in the output-voltage-sensing divider of high-voltage supplies and amplifiers – beware! Companies like Victoreen make resistors in many styles designed for high-voltage applications like this.

□ *Regulating the input*

Another technique sometimes used in high-voltage supplies, particularly those intended for low currents, is to regulate the input rather than the output. This is usually done with high-frequency dc-to-dc switching supplies, since attempting to regulate the 60Hz ac input will result in poor regulation and plenty of residual ripple. Figure 6.54 shows the general idea. T_1 and associated circuitry generate unregulated dc at some manageable voltage, say 24

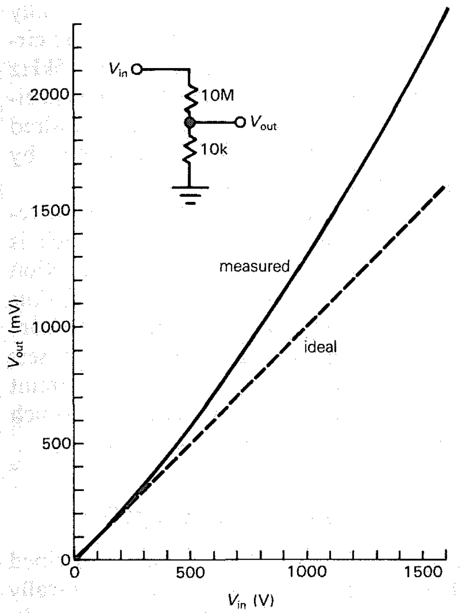


Figure 6.53. Carbon composition resistors exhibit a reduction in resistance above 250 volts.

volts; alternatively, batteries might provide the dc input. This powers a high-frequency square-wave power oscillator, with its output full-wave-rectified and filtered. This filtered dc is the output, a sample of which is fed back to control the oscillator's duty cycle or amplitude in response to the output voltage. Since the oscillator runs at high frequency, the response is rapid, and its rectified waveform is easy to filter, especially since it is a full-wave-rectified square wave. T_2 must be designed for high-frequency operation, since ordinary laminated-core power transformers will have excessive core losses. Suitable transformers are built with iron powder, ferrite, or "tape-wound" toroidal cores and are much smaller and lighter than conventional power transformers of the same power rating. No high-voltage components are used, except, of course, for the output bridge rectifier and capacitor.

The astute reader may experience a sense of *déjà vu* while reading the last paragraph. In fact, it describes switching regulators (Section 6.19) in nearly all respects. The one significant difference is that switching supplies usually use inductors as energy-storage devices, whereas the input-regulated high-voltage supply uses T_2 as a "normal" (albeit high-frequency) transformer. In common with switching supplies, these high-voltage supplies display high-frequency ripple and noise.

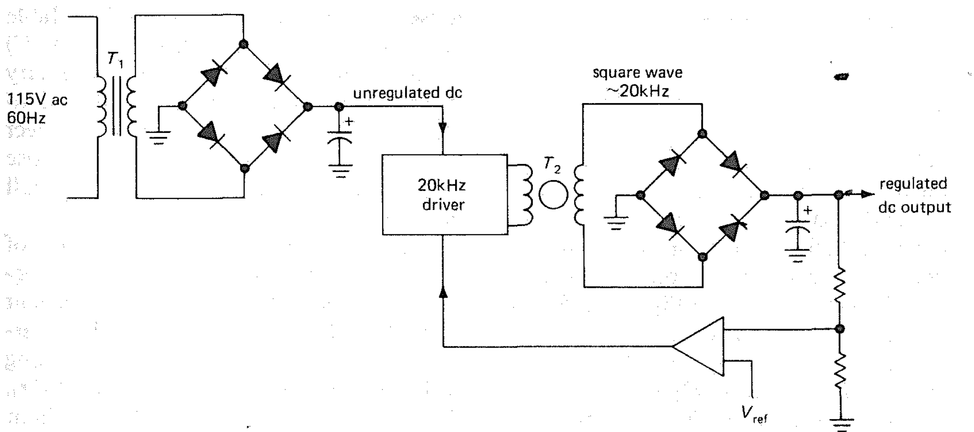


Figure 6.54. High-voltage switching supply.

Video flyback supplies

A variation on the conventional flyback switching regulator (Fig. 6.43A) is commonly used to generate the high dc voltages (10kV or more) needed in television and cathode-ray-tube (CRT) video displays. As we'll see, this circuit is especially clever, because it also generates the horizontal sweep signal used to drive the deflection coils.

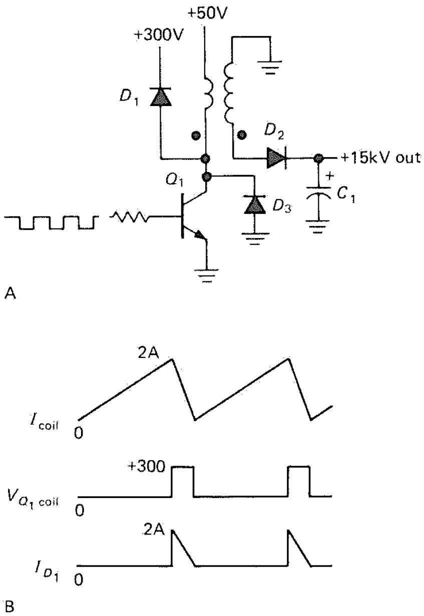


Figure 6.55. Video flyback high-voltage supply.

The basic idea is to use a transformer with a large turns ratio, driving the primary with a saturated transistor, just like a conventional flyback circuit. The output is taken from the secondary, rectified to generate high-voltage dc; see Figure 6.55. Q_1 is driven by wide pulses, pulling the primary to ground. It may be self-excited or driven by an oscillator. D_1 is a "damper" diode that prevents Q_1's collector from rising too high during the flyback. D_2, connected to the high-voltage secondary winding, rectifies the output, typically 10kV–20kV at a few microamps. The circuit is operated at frequencies of 15kHz or more, which means that filter capacitor C_1 can be as small as a few hundred picofarads (check this for yourself, by calculating the ripple).

Note that the collector-current waveform is a linearly rising ramp, which is often used to drive the magnetic deflection coils (called the "yoke") of the CRT, thus producing the linear horizontal raster scan. In such cases the oscillator frequency sets the horizontal scan rate. A related circuit is the so-called blocking oscillator, which generates its own excitation pulses.

□ 6.21 Low-noise, low-drift supplies

The regulated supplies we have described thus far are pretty good – they typically have ripple and noise below a millivolt, and drift with temperature of 100ppm/°C or so. This is more than adequate for just about everything you will ever need to power. However, there are times when you may need better performance, and you can't get it with any available regulator ICs. The solution is to design your own regulator circuit, using the best available IC references (in terms of stability and noise; see, e.g., the REF101KM in Table 6.7). This kind of stability (<1ppm/°C) is far better than the tempco of ordinary metal-film resistors (50ppm/°C), for example; so you must use great care to select op-amps and passive components whose errors and drifts do not degrade overall performance.

Figure 6.56 shows a complete design of an exceptional low-noise, low-drift dc regulated supply. It begins with the excellent REF10KM from Burr-Brown, which guarantees better than 1ppm/°C tempco, along with very low noise (6μV pp, 0.1–10Hz). Furthermore, it achieves this without thermostatic control, which helps keep the subsurface zener noise low. The reference

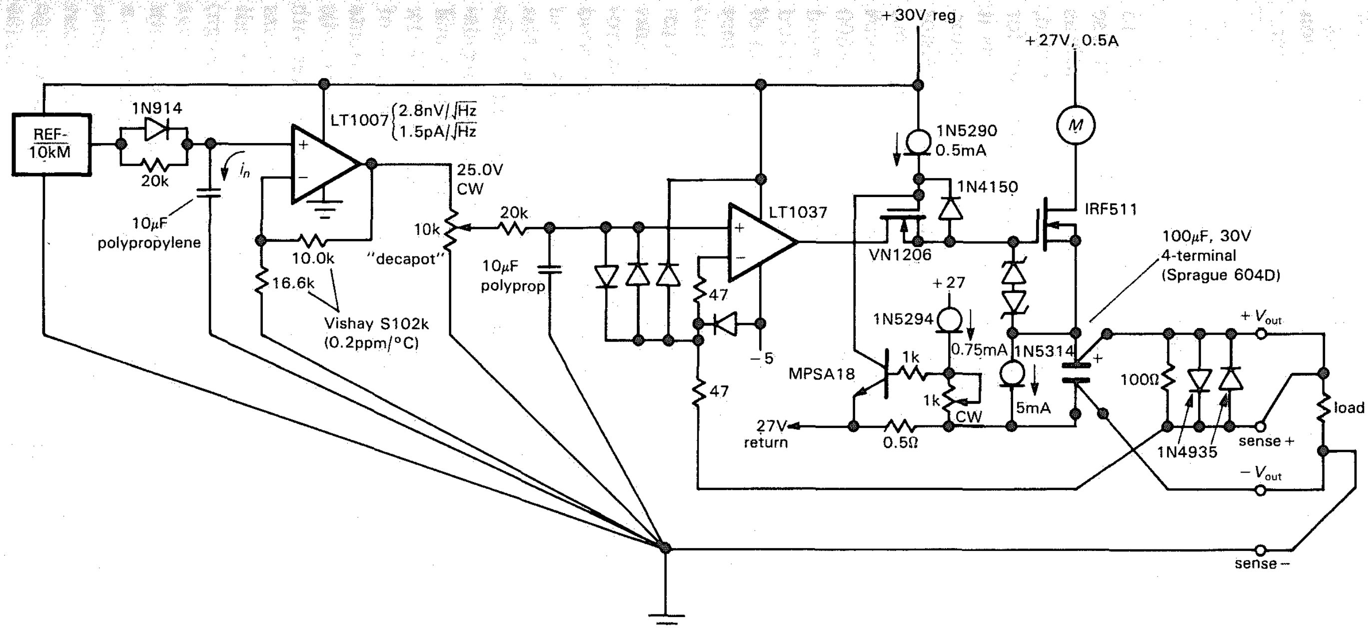


Figure 6.56. Ultrastable low-noise power supply.

is followed by a low-pass filter, to reduce the noise further. The large capacitor value is needed to suppress current noise from the op-amp: the value shown converts the current noise ($1.5\text{pA}/\sqrt{\text{Hz}}$ at 10Hz) to a voltage noise of $2.4\text{nV}\sqrt{\text{Hz}}$, comparable with the op-amp's e_n. A polypropylene capacitor is used because the capacitor leakage (more precisely, *changes* in leakage over time and temperature) must be less than 0.1nA in order to avoid microvolt drifts in output voltage. The reference is boosted to +25 volts by the op-amp, whose feedback resistors have ultra-low tempco (0.2ppm/°C, max); note the +30 volt supply voltage. The resultant +25.0 volt reference drives a voltage divider to produce the desired output voltage, which is then low-pass-filtered a second time, again using a low-leakage capacitor. Because a potentiometer is used to divide the reference voltage, resistor tempco isn't as critical here – it's a *ratiometric* measurement.

The rest of the circuit is simply a follower, using a precision low-noise error amplifier to compare the output voltage from a power MOSFET series pass transistor. A decompensated op-amp has been used, since the large output capacitor provides the dominant pole for compensation. Note the unusual current-limit circuit and the liberal use of constant-current "diodes" (really JFETs) to provide operating bias. Note also the use of "sense" wires to sample the voltage across the load. In a precision circuit like this it is important to pay careful attention to ground paths, since, for example, a 100mA load current flowing through 1 inch of #20 wire produces a voltage drop of $100\mu\text{V}$ – which is a 100ppm error for a 1 volt output! The circuit shown has excellent performance and surpasses the typical noise and drift figures given earlier by at least a factor of 100. According to EVI, Inc. (Columbia, MD), which kindly provided the circuit, it produces noise and hum below $1\mu\text{V}$, tempco below 1ppm/°C, output impedance below $1\mu\Omega$, and drift below 1ppm/working day.

We will talk more about such precision and low-noise design in the next chapter.

□ 6.22 Micropower regulators

As we've hinted earlier, it's possible to design battery operated circuits that use very low quiescent current, often as little as tens of microamps. That's what's needed, of course, to make the circuit run for months or years on a small battery, as it must if it is a wristwatch or calculator. For example, an alkaline 9 volt transistor battery is exhausted after supplying about 400mA-hours; thus you can run a $50\mu\text{A}$ circuit with it for about a year (8800 hours). If such *micropower* circuits need regulated voltages, you clearly can't afford to squander the 3mA quiescent current of a 78L05, since that would degrade battery life to less than a week!

The solution is either to design a micropower regulator from discrete components or use one of the ICs intended for micropower applications. Luckily, some good ICs have come along in recent years. One of the best is the LP2950 series from National, available as a TO-92 (small transistor package) 3-terminal fixed 5 volt regulator (LP2950ACZ-5.0) or as a multiterminal adjustable 1.2–30 volt regulator (LP2951). Both versions have a quiescent current of $75\mu\text{A}$. For even lower quiescent currents there are the ICL7663/4 (or MAX663/4), adjustable regulators of both polarities with $4\mu\text{A}$ quiescent current. We will discuss micropower regulators, along with all aspects of battery-powered circuit design, in Chapter 14.

As an example of what you can do with discrete design, we show in Figure 6.57 a micropower circuit, designed for possible use in a lithium-battery-powered heart pacemaker, that converts an input voltage in the range +5 volts down to +3 volts (as

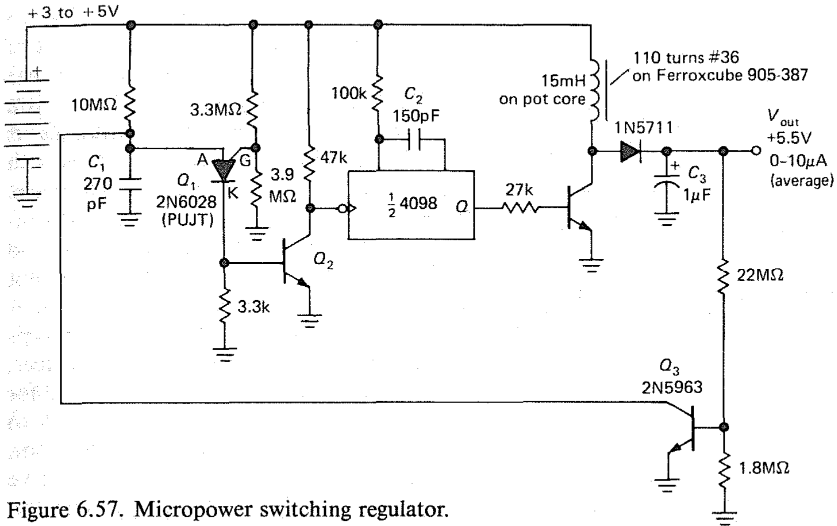


Figure 6.57. Micropower switching regulator.

the battery ages) to a regulated +5.5 volt supply. The power supply has a quiescent current of 1μA and provides line and load regulation of about 5%, with 85% conversion efficiency under full load for all battery voltages. As we remarked when discussing switching supplies, a conventional linear supply using an oscillator, doubler, and series pass regulator would be far less efficient because of regulator losses following the higher unregulated dc voltage. The flyback technique is effectively like a variable-ratio voltage multiplier, which yields extremely high efficiency, making it an attractive technique for micropower applications.

The 2N6028 programmable unijunction transistor (PUJT) is a versatile relaxation oscillator component. Its sense terminal (the anode) draws no current until its voltage exceeds the gate programming voltage by a diode drop, at which point it goes into heavy conduction from anode to cathode, discharging the capacitor. The resulting positive pulse at Q_2's base pulls Q_2's collector to ground, triggering the 4098, a device known as a "monostable multivibrator" (see Section 8.20), which generates positive pulses of constant width at its output terminal labeled Q.

In this circuit, Q_3 senses the output voltage and robs charging current from C_1, reducing the energy-transfer pulsing rate of the inductor as necessary to maintain the desired output voltage. Note the large resistor values throughout the circuit. Temperature compensation is not an issue here because the circuit operates in a stable 98.6°F mobile oven. (Warning: We remind the reader to look again at the "Legal notice" in the Preface.)

6.23 Flying-capacitor (charge pump) voltage converters

In Section 6.19 we discussed switching supplies, with their bizarre ability to produce a dc output voltage *larger* than the dc input, or even of opposite polarity. We mentioned there that flying-capacitor voltage converters let you do some of the same

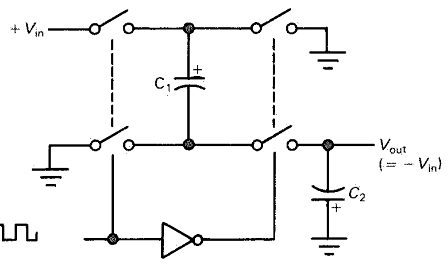


Figure 6.58. Flying-capacitor voltage inverter. C_1 and C_2 are external 10μF tantalum capacitors.

things. What is this strange "flying capacitor"?

Figure 6.58 shows a simplified circuit of the 7662 CMOS IC introduced by Intersil, and widely second-sourced. It has an internal oscillator and some CMOS switches, and it requires a pair of external capacitors to do its job. When the input pair of switches are closed (conducting), C_1 charges to V_{in}; then, during the second half cycle, C_1 is disconnected from the input and connected, upside-down, across the output. It thus transfers its charge to C_2 (and the load), producing an output of approximately $-V_{in}$. Alternatively, to use the 7662 to create an output of $2V_{in}$ you can arrange things so that C_1 charges as before, but then gets hooked in series with V_{in} during the second (transfer) half cycle.

This flying-capacitor technique is simple and efficient and requires few parts and no inductors. However, the output is not regulated, and it drops significantly under load currents greater than a few milliamps (Fig. 6.59). Also, like most CMOS devices, it has a limited supply voltage range; for the 7662, V_{in} can only range from 4.5 to 20 volts (1.5V to 10V for its predecessor, the 7660). Finally, unlike the inductive step-up or inverter circuits, which can generate any output voltage at all, the flying capacitor voltage converter can only generate discrete multiples of the input voltage. In spite of these drawbacks, flying-capacitor voltage converters can be very useful in some circumstances, for example to power a bipolarity op-amp or serial port (see Chapters 10 and 11) on a circuit board that has only +5 volts available.

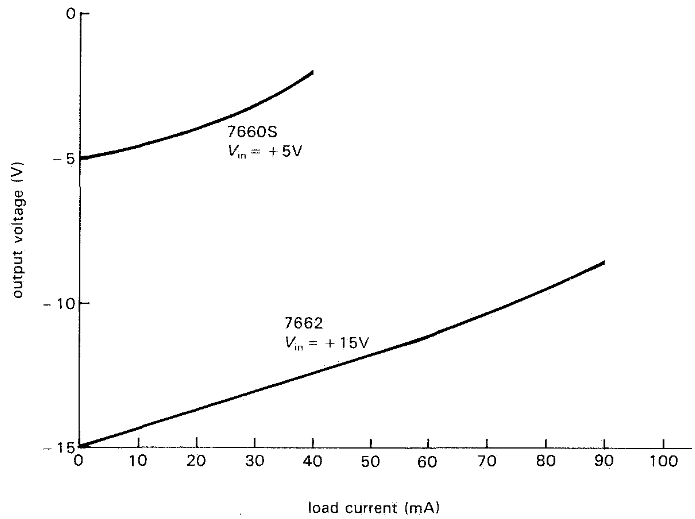


Figure 6.59. The output voltage of a flying-capacitor inverter drops significantly under load.

There are some other interesting flying-capacitor chips. The MAX680 from Maxim is a dual supply that generates ±10 volts (up to 10mA) from +5 volts (Fig. 6.60). The similar LT1026 from LTC operates to ±20 volts output (up to 20mA) and uses smaller capacitors (1μF instead of 20μF). The LT1054 from LTC combines a flying-capacitor converter with a linear regulator to provide a stiff regulated output up to 100mA (at lower efficiency, of course). The MAX232 series and the LT1080 combine a ±10 volt switched-capacitor supply with an RS-232C digital serial port (see Chapter 11), eliminating the need for bipolarity supplies in many computer boards; some chips in the MAX232 series even have built-in capacitors. And the LTC1043 is an uncommitted flying-capacitor building block, which you can use to do all kinds of magic. For example, you can use a flying capacitor to transfer a voltage drop measured at an inconvenient potential (e.g., a current-sensing resistor at the positive supply voltage) down to ground, where you can easily use it. The LTC1043 data sheet has 8 pages of similar clever applications.

6.24 Constant-current supplies

In Sections 2.06 and 2.14 we described some methods for generating constant currents within a circuit, including voltage-programmed currents with floating or grounded loads and various forms of current mirrors. In Section 3.06 we showed how to use FETs to construct some simple current-source circuits, including "current-regulator diodes" (a JFET with gate tied to source) such as the 1N5283 series. In Section 4.07 we showed how to get improved performance (at low frequencies, anyway) by using op-amps to construct current sources. And in Section 6.15 we mentioned the convenient LM334 3-terminal current source IC. There is often a need, however, for a flexible constant-current supply, which can supply substantial voltage and current, as a complete instrument. In this section we will look at some of the more successful circuit techniques.

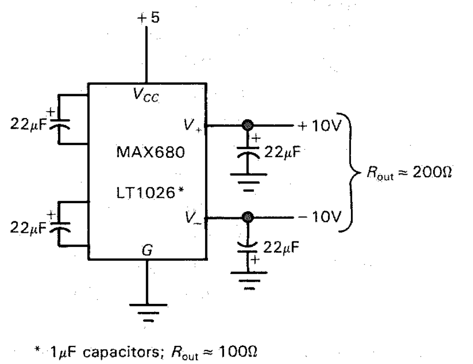


Figure 6.60. Flying-capacitor dual supply. The LT1026 is similar, but has $R_{\text{out}} \approx$ 100 ohms and requires only 1μF capacitors.

□ *Three-terminal regulator*

In Section 6.18 we showed how you can use a 3-terminal adjustable regulator to make a delightfully simple current source. The 317-type regulator, for example, maintains a constant 1.25 volts (bandgap) between its output and its "ADJ" pin; by putting a resistor across these pins, you form a 2-terminal constant-current device (Fig. 6.38), which can be used as a sink or source. Performance degrades with less than about 3 volts across the circuit, since the regulator itself has a dropout voltage near 2 volts.

This type of current source is suitable for moderate to high currents: The LM317 has a maximum current of 1.5 amps and can operate with up to 37 volts drop. Its high-voltage cousin, the LM317HVK, can withstand 57 volts drop. Higher-current versions are available, e.g., the LM338 (5A) and LM396 (10A), although these have lower voltage ratings. Three-terminal regulators won't work as current sources below about 10mA, the worst-case quiescent current. However, note that the latter is not a source of current error, since it flows from input pin to output pin; the much smaller current that flows out of the ADJ pin (50μA, nominal) varies about 20% over the operating temperature range and is negligible by comparison.

In ancient times, before 3-terminal adjustable regulators were available, people sometimes used 5 volt *fixed* regulators (e.g., the 7805) as current sources in a similar arrangement (substituting "GND" for "ADJ"). This is an inferior circuit, because at low output currents the regulator's quiescent current (8mA max) contributes a large error, and at high currents the 5 volt drop across the current-setting resistor results in unnecessary power dissipation.

□ *Supply-line sensing*

A simple technique that yields good performance involves constructing a conventional series pass regulator, with current sensing at the input to the pass transistor (Fig. 6.61). R_2 is the current-sensing

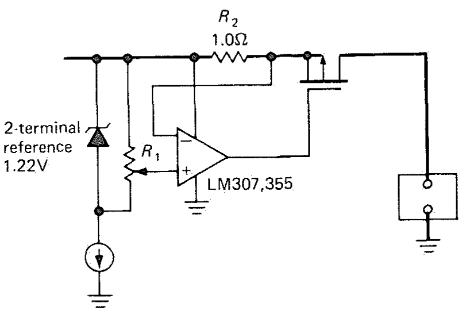


Figure 6.61. Input-rail current sensing.

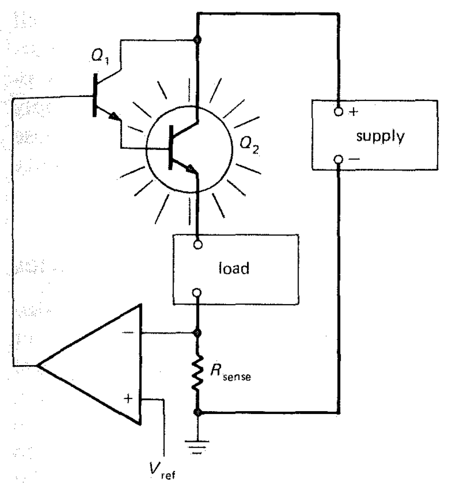
resistor, preferably a low-temperature-coefficient type. For very high current or high-precision applications, you should use a 4-wire resistor, intended for current-sensing applications, in which the sensing leads are connected internally. The sensing voltage does not then depend on the connection resistance of the joints to the current-carrying leads, which for clarity are drawn with heavy lines in this schematic.

For this circuit you must use an op-amp that has an input common-mode range all the way to the positive supply (the 307, 355, and 441 have this virtue), unless, of course, you power the op-amp with a more positive auxiliary supply. The MOSFET in this circuit could be replaced by a *pnp* pass transistor; however, since the output current would then include the base current, you should use a Darlington connection to minimize that error. Note that an n-channel output transistor (connected as a follower) can be used instead of the p-channel shown, if the input connections to the op-amp are reversed. However, the current source will then have an undesirably low output impedance at frequencies approaching f_T of the op-amp loop, since the output is actually a source follower. This is a common error in current-source design, since the dc analysis shows correct performance.

□ *Return-line current sensing*

A good way to make a precise current source is to sense the voltage across a precision resistor directly in series with the load, since this makes it easier to meet the simple criterion for eliminating current-source errors due to base drive currents; the base drive current must either pass through both the load and sense amplifier, or pass through neither. However, to meet this criterion it is necessary to "float" either the load or the power supply by at least the voltage drop across the current-sensing resistor. Figure 6.62 shows a couple of circuits that use floating loads.

The first circuit is a conventional series pass circuit, with the error signal derived from the drop across the small resistor in the load's return path to ground. The high-current path is again drawn with bold lines. The Darlington connection is used here



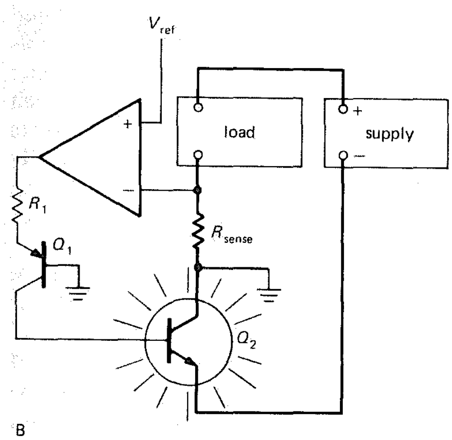


Figure 6.62. Return-line current sensing.

not to avoid base-current error, since the actual current through the load is sensed, but rather to keep the drive current down to a few milliamps so that ordinary op-amps can be used for the error amplifier. The sensing resistor should be a precision power resistor of low temperature coefficient, preferably a 4-wire resistor. In the second circuit the regulating transistor Q_2 is in the ground return of the high-current supply. The advantage here is that its collector is at ground, so you don't have to worry about insulating the transistor case from the heat sink.

In both circuits, R_{sense} will normally be chosen to drop a volt or so at typical operating currents; its value is a compromise between op-amp input offset errors, at one extreme, and a combination of reduced current-source compliance and increased dissipation at the other. If the circuit is meant to operate over large ranges of output current, R_{sense} should probably be a set of precision power resistors, with the appropriate resistor selected by a range switch.

□ *Grounded load*

If it is important for the load to be returned to circuit ground, a circuit with floating supply can be used. Figure 6.63 shows two examples. In the first circuit, the funny-looking op-amp represents an error amplifier with a high-current buffer output, run from a single split supply; it could be something as simple as a 723 (for currents up to 150mA) or one of the high-current op-amps listed in Table 4.4. The high-current supply has a common terminal that floats relative to circuit ground, and it is important that the error amplifier (or at least its buffer output) be powered from the floating supply so that base drive currents return through R_{sense}. An additional low-current supply with grounded common would be needed if other op-amps, etc., were in the same instrument. A negative reference (relative to circuit ground) programs the output current. Note the polarity at the error-amplifier inputs.

The second circuit illustrates the use of a second low-power supply when an ordinary low-current op-amp is used as error amplifier. Q_1 is the outboard pass transistor, which must be a Darlington (or MOSFET), since the base current returns through the load, but not through the sense resistor. The error amplifier is now

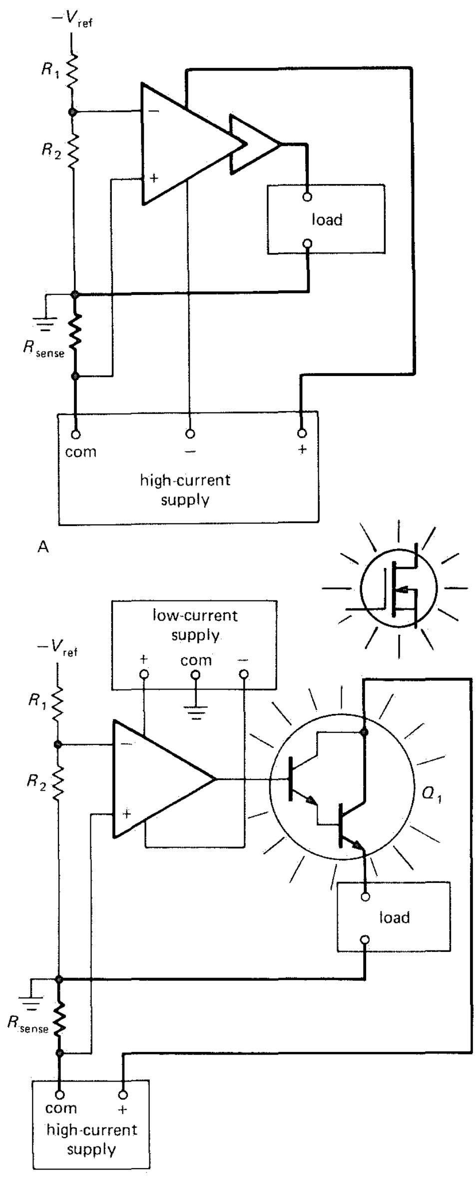


Figure 6.63. Current sources for grounded loads, employing floating high-current supplies.

powered from the same split supply with grounded common that powers the rest of the instrument. This circuit is well suited as a simple bench-instrument current source, with the low-current split supply built in and the high-current supply connected externally. You would choose the latter's voltage and current capability to fit each application.

6.25 Commercial power-supply modules

Throughout the chapter we have described how to design your own regulated power supply, implicitly assuming that is the best thing to do. Only in the discussion of line-operated switching power supplies did we suggest that the better part of valor is to swallow your pride and buy a commercial power supply.

As the economic realities of life would have it, however, the best approach is often to use one of the many commercial power supplies sold by companies such as ACDC, Acopian, Computer Products Inc., Lambda, Power-One, and literally hundreds more. They offer both switching and linear supplies, and they come in four basic packages (Figure 6.64):

1. Modular "potted" supplies: These are low-power supplies, often dual (±15) or triple (+5, ±15), packaged in "potted" modules that are usually 2.5"×3.5", and about 1" thick. The most common package has stiff wire leads on the bottom, so you can mount it directly on a circuit board; you can also screw it to a panel, or plug it into a socket. They are also available with terminal-strip screw connections along one side, for chassis mounting. A typical linear triple supply provides +5 volts at 0.5 amp and ±15 volts at 0.1 amp and costs about $100 in small quantity. Linear modular supplies fall in the 1–10 watt range, switchers in the 15–25 watt range.

2. "Open-frame" supplies: These consist of a sheet-metal chassis, with circuit board, transformer, and power transistors mounted in full view. They're meant to go

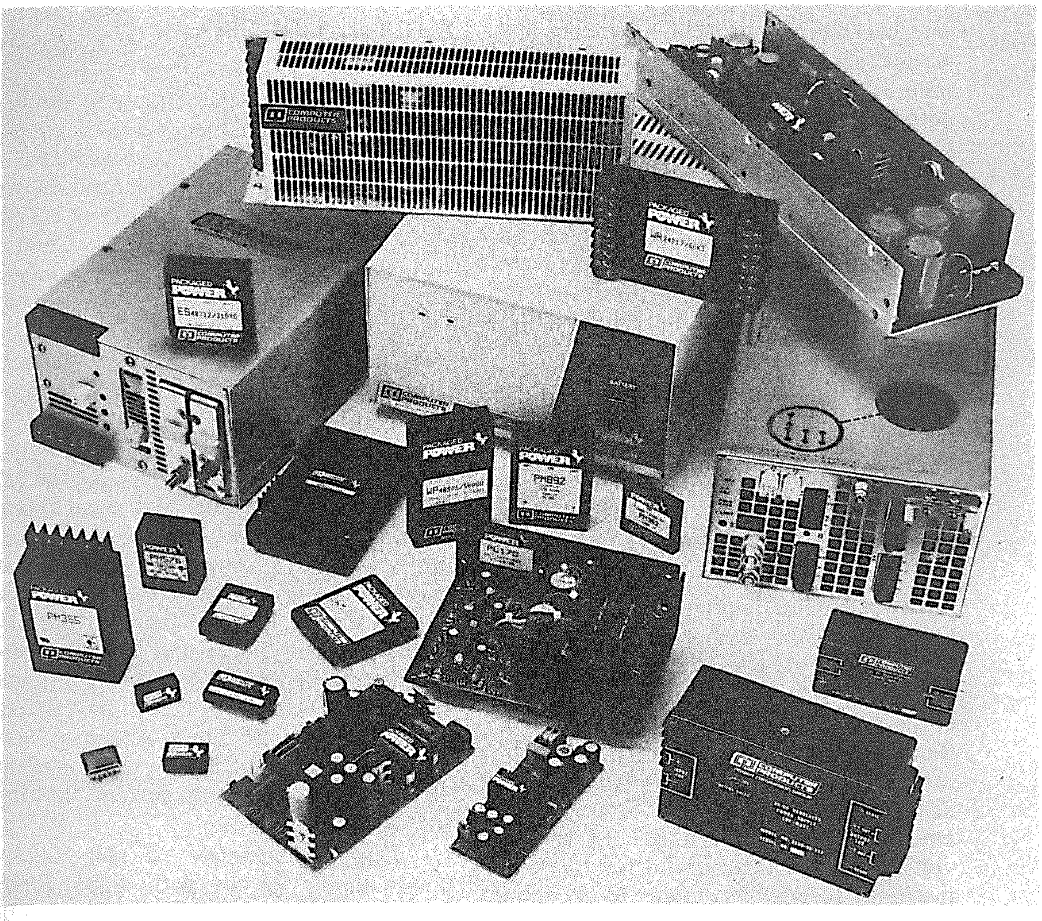


Figure 6.64. Commercial power supplies come in a variety of shapes and sizes, including potted modules, open-frame units, and fully enclosed boxes. (Courtesy of Computer Products, Inc.)

inside a larger instrument. They come in a wide range of voltages and currents and include dual and triple units as well as single-output supplies. For example, a popular triple open-frame linear provides +5 volts at 3 amps and ±15 volts at 0.8 amp and costs $75 in small quantity. Open-frame supplies are larger than potted modules, and you always screw them to the chassis. Open-frame linears fall in the 10–200 watt range, switchers in the 20–400 watt range. Open-frame supplies at the low end of the power range may have all components mounted directly on a circuit board, with no metal frame at all. As with the potted supplies, you are expected to provide switches, filters, and fuses for the ac line voltage circuits.

3. Fully enclosed supplies: These supplies have a full metal enclosure, usually perforated for cooling, and usually free of the protruding power transistors, etc., that you find on an open-frame supply. They can be mounted externally, because their full enclosure keeps fingers out; you can also mount them inside an instrument, if you want. They come with single and multiple outputs, in both linear and switchers. Fully enclosed linears fall in the 15–750 watt range, switchers in the 25–1500 watt range.

4. Wall plug-in power supplies: These are the familiar black plastic boxes that come with small consumer electronic gadgets and plug directly into the wall. They actually come in three varieties, namely (a) step-down ac transformer only, (b) unregulated dc supply, and (c) complete regulated supply; the latter can be either linear or switcher. For example, Ault has a nice series of dual (±12V or ±15V) and triple (+5V and ±12V or ±15V) linear regulated wall-plug-in supplies. These save you the trouble of bringing the ac line power into your instrument, and keep it light and small. Some of us think that these convenient supplies are getting a bit *too* popular, though, as measured by the cluster of wall plug-ins found feeding at the outlets in our house! Some "desktop" models have two cords, one each for the ac input and dc output. Some of the switching units allow a full 95 to 252 volts ac input range, useful for traveling instruments. We'll have more to say about wall plug-ins in Section 14.03, when we deal with low-power design.

SELF-EXPLANATORY CIRCUITS

6.26 Circuit ideas

Figure 6.65 presents a variety of current ideas, mostly taken from manufacturers' data sheets.

6.27 Bad circuits

Figure 6.66 presents some circuits that are guaranteed not to work. Figure them out, and you will avoid these pitfalls.

ADDITIONAL EXERCISES

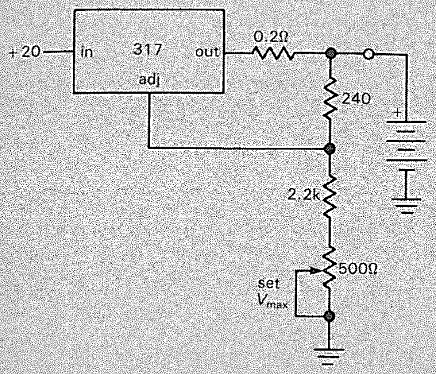
(1) Design a regulated supply to deliver exactly +10.0 volts at currents up to 10mA using a 723. You have available a 15 volt (rms) 100mA transformer, diodes by the bucketful, various capacitors, a 723, resistors, and a 1k trimmer pot. Choose your resistors so that they are standard (5%) values and so that the range of adjustment of the trimmer will be sufficient to accommodate the production spread of internal reference voltages (6.80V to 7.50V).

(2) Design +5 volt 50mA voltage regulators, assuming +10 volt unregulated input, using the following: **(a)** zener diode plus emitter follower; **(b)** 7805 3-terminal regulator; **(c)** 723 regulator; **(d)** 723 plus outboard *npn* pass transistor; use foldback current limiting with 100mA onset (full-voltage current limit) and 25mA short-circuit current limit; **(e)** a 317 3-terminal adjustable positive regulator; **(f)** discrete components, with zener reference and feedback. Be sure to show component values; provide 100mA current limiting for (a), (c), and (f).

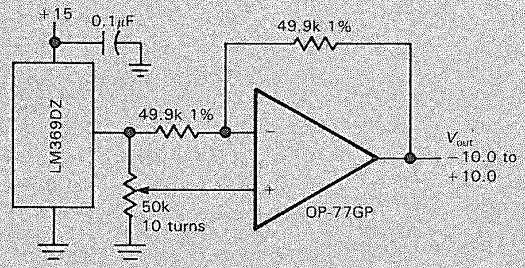
(3) Design a complete +5 volt 500mA power supply for use with digital logic. Begin at the beginning (the 115V ac wall socket), specifying such things as transformer voltage and current ratings, capacitor values, etc. To make your job easy, use a 7805 3-terminal regulator. Don't squander excess capacitance, but make your design conservative by allowing for ±10% variation in all parameters (powerline voltage, transformer and capacitor tolerances, etc.). When you're finished, calculate worst-case dissipation in the regulator.

Next, modify the circuit for 2 amp load capability by incorporating an outboard pass transistor. Include a 3 amp current limit.

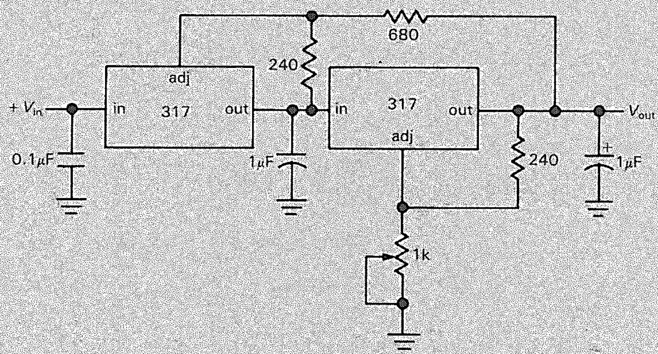
Circuit ideas



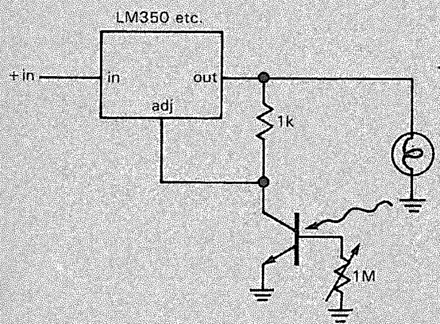
A. 12V battery charger



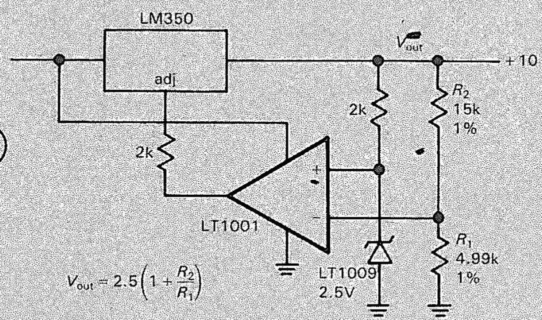
B. adjustable stable bipolar voltage reference



C. tracking preregulator

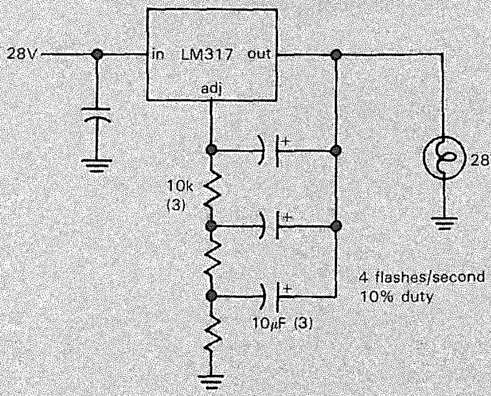


D. automatic incandescent bulb light regulator

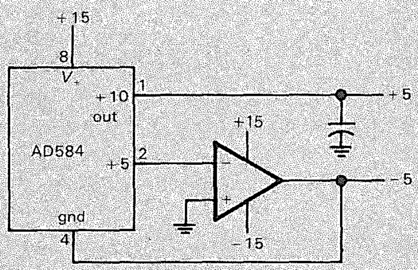


E. precision power voltage source

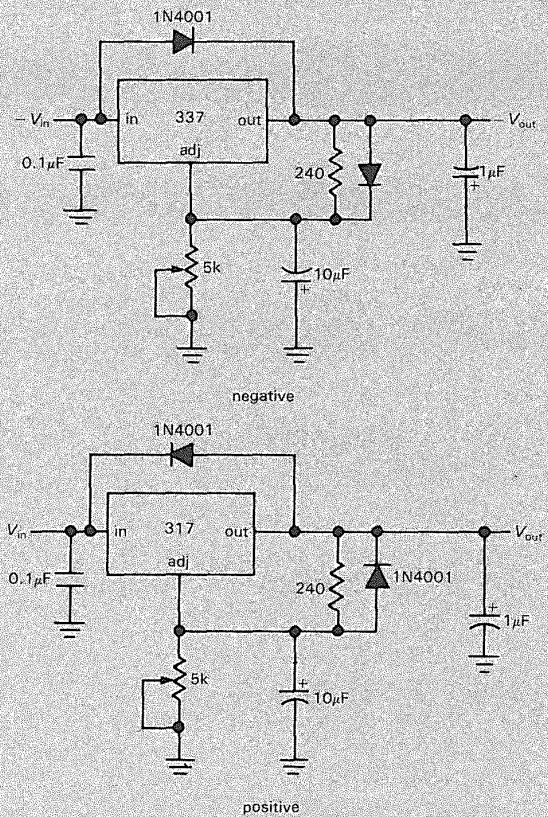
Figure 6.65



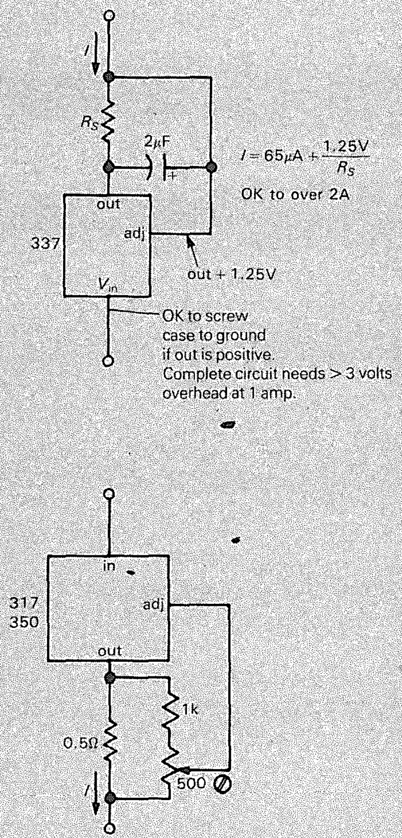
F. lamp flasher (from NSC317 data sheet)



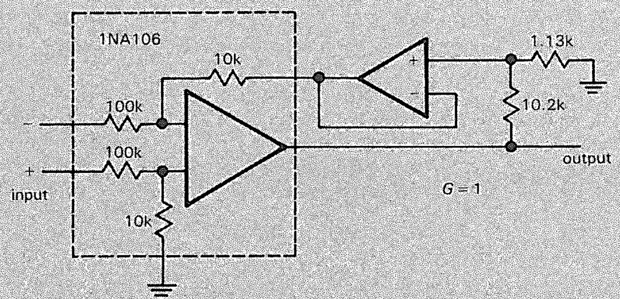
G. ±5 volt reference from one 2-output reference



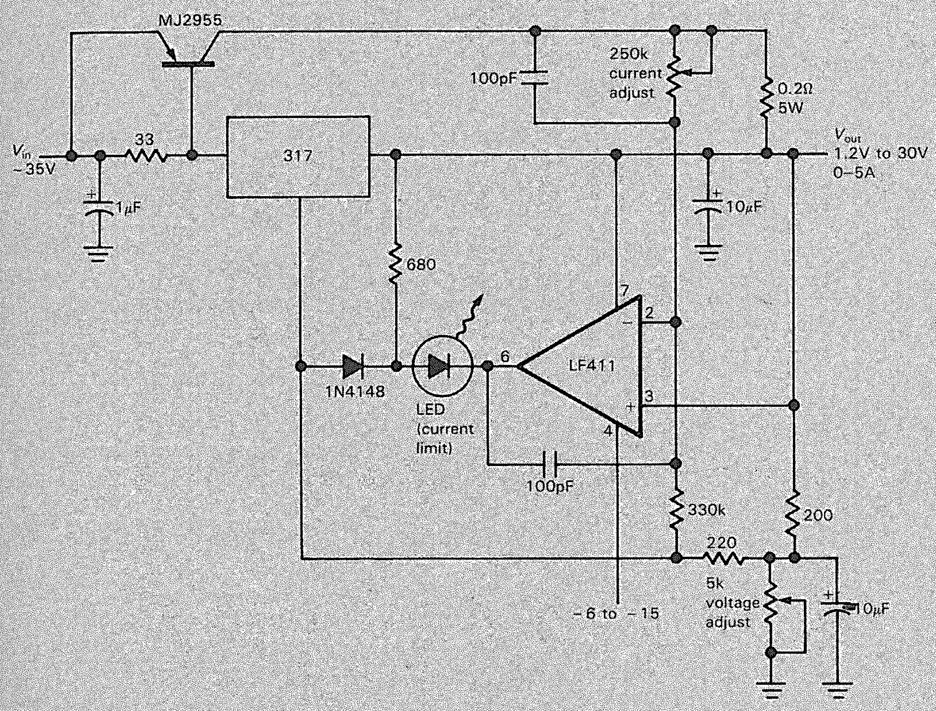
H. 3-terminal regulators with improved ripple rejection
(diodes protect against input/output shorts)



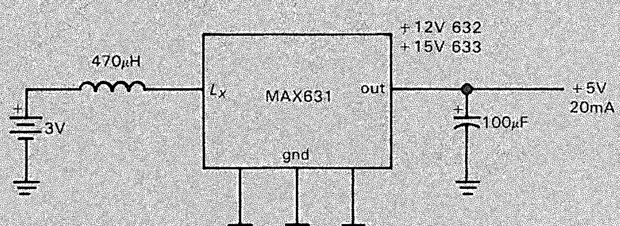
I. power current sources



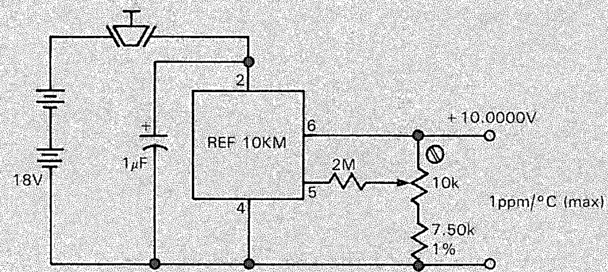
J. ±100V common-mode range differential follower



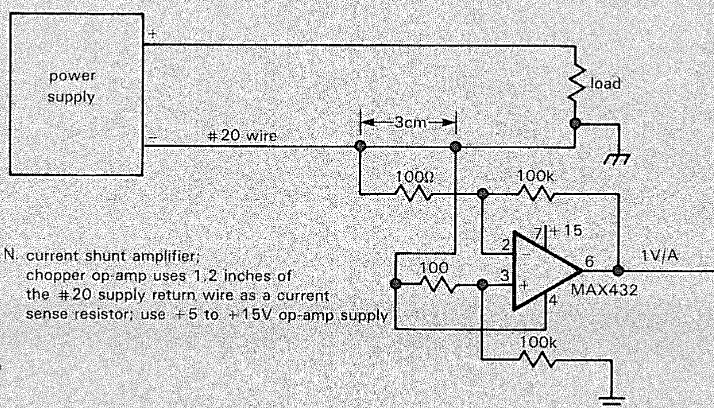
K. constant voltage/constant current supply



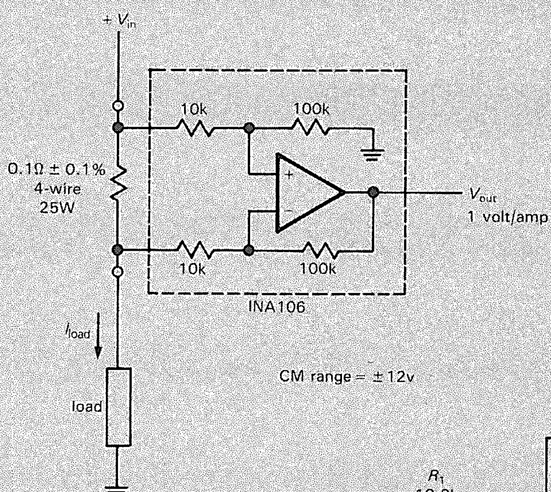
L. "world's simplest" dc–dc converter



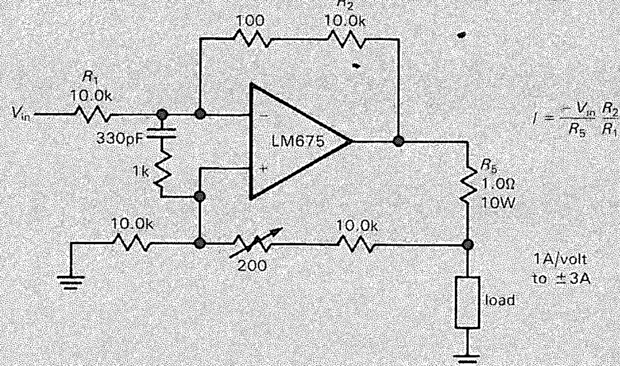
M. portable 10ppm voltage reference



N. current shunt amplifier;
chopper op-amp uses 1.2 inches of
the #20 supply return wire as a current
sense resistor; use +5 to +15V op-amp supply

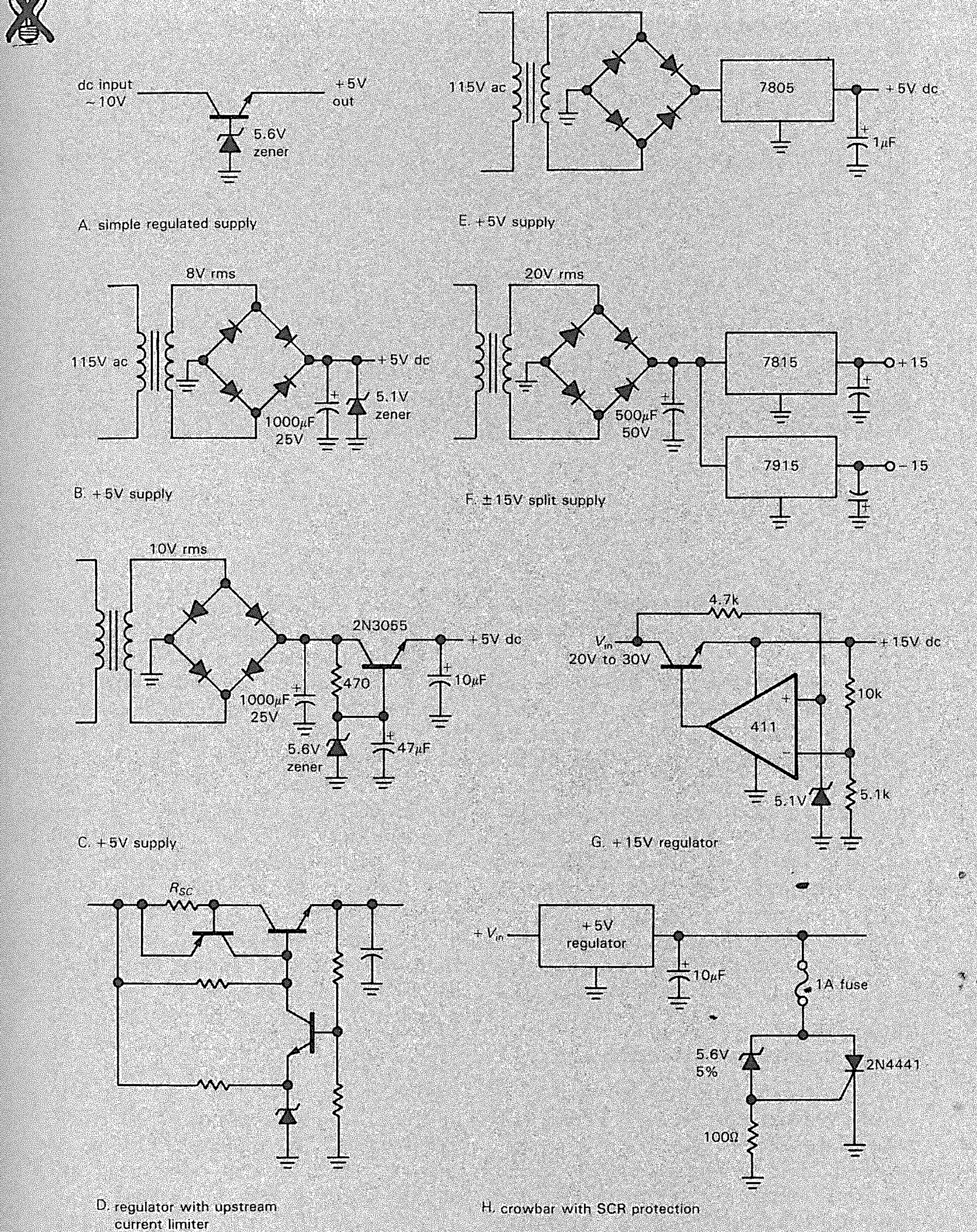


O. high-side current monitor



P. high-current bipolarity current source

Bad circuits



A. simple regulated supply

E. +5V supply

B. +5V supply

F. ±15V split supply

C. +5V supply

G. +15V regulator

D. regulator with upstream current limiter

H. crowbar with SCR protection

Figure 6.66

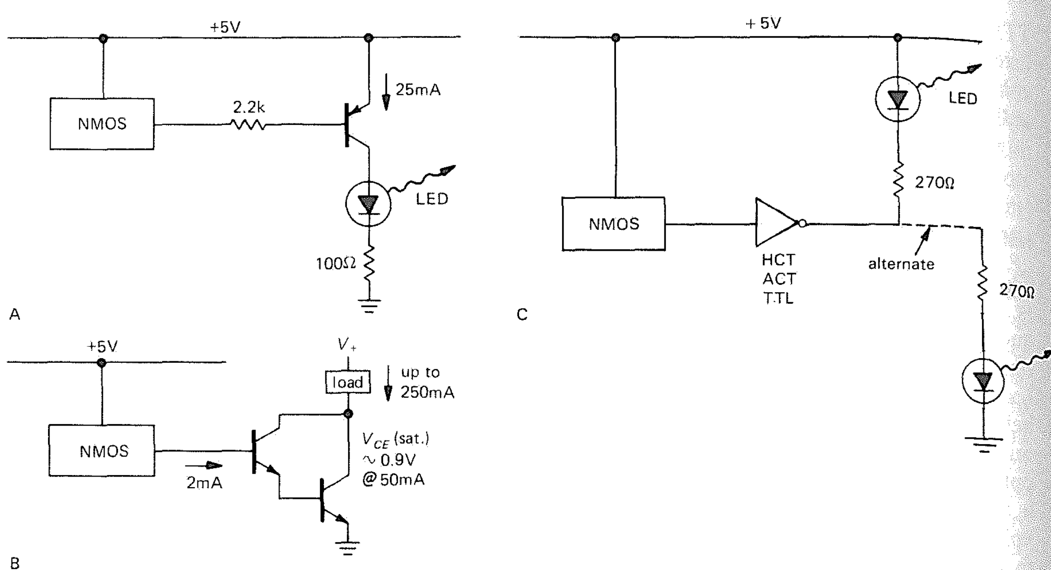


Figure 9.20. NMOS logic outputs driving loads.

the base of a grounded emitter Darlington without damage. A typical NMOS output would source 2mA into the Darlington's base at +1.5 volts, giving an output sink capability of 250mA at 1 volt with an IC like the 75492 hex Darlington array. The ULN series from Sprague has several hex and octal Darlington arrays in DIP packages.

Finally, you can always buffer NMOS outputs with an HCT or ACT (or even TTL) buffer, since NMOS outputs are fully compatible with those families' inputs. The buffered outputs can then sink current from a load; with HCT or ACT you can equally well *source* current, since the CMOS families have the same output drive capability whether sinking or sourcing.

9.10 Opto-electronics

In the last two chapters we have been using LED indicators and LED numeric display devices in various circuit contexts, as we needed them. LEDs belong to the general area of *opto-electronics*, which includes displays based on other technologies as well, notably liquid crystals, fluorescence, and gas discharge. It also includes optical electronics used for purposes other than indicators and displays: light-coupled isolators ("opto-isolators"), solid-state relays, position sensors ("interrupters" and "reflective sensors"), diode lasers, array detectors ("charge-coupled devices," CCDs), image intensifiers, and a variety of components used in fiber optics.

Although we will continue to conjure up miscellaneous magic devices as we need them, this seems like a good place to pull together the area of opto-electronics, since it is related to the logic interface problems we have just been discussing.

Indicators

Electronic instruments look nicer, and are more fun to use, if they have pretty little colored lights on them. LEDs have replaced all earlier technologies for this purpose. You can get red, yellow, and green indicators, and you can get them in many packages, the most useful of which are (a) panel-mounting lights, and (b) printed-circuit board (PCB) mounting types. The catalogs present you with a bewildering variety of them, differing mostly in size, color, efficiency, and illumination angle. The latter deserves some explanation: A "flooded" LED has some diffusing stuff mixed in, so the lamp looks uniformly bright over a range of viewing angles; that's usually best, but you pay a price in brightness.

An LED looks electrically like a diode, with a forward drop of about 2 volts (they're built with gallium arsenide phosphide, which has a larger band gap, and hence a larger forward drop, than silicon). Typical flooded panel-type LED indicators look good at 10mA forward current; on a board inside an instrument you can usually get away with 2–5mA, particularly if you use a narrow-angle LED.

Figure 9.21 shows how to drive LED indicators. Most of the circuits are obvious, but note that bipolar TTL is poor at *sourcing* current, so you always arrange the circuit so that logic LOW turns on the LED; by comparison, CMOS families have symmetrical drive capability. NMOS circuits not only share the feeble sourcing of bipolar TTL but also tend to have rather limited current-sinking capability; it's best to interpose a buffer (an HCT gate is good), or perhaps a discrete MOSFET. Note also that some LED indicators come with internal current-limiting resistors (or even internal constant-current circuits) – with these you omit the external resistor.

You can get little arrays of indicators – sticks of 2, 4, or 10 LEDs in a row – designed for PCB mounting. The latter are actually intended for linear "bar-graph" readouts. They come in upright or right-angle mounting. You can also get panel-mounting indicators that combine a red LED and green LED in one uncolored package. These make an impressive panel, with lights changing color to indicate good/bad conditions.

We've used LED indicators from manufacturers such as Dialight, General Instrument, HP, Panasonic, Siemens, and Stanley. The latter specializes in lamps of unusually high efficiency; you can usually locate their exhibit at electronics shows by the dazed look (and incipient sunburn) of recent visitors.

Displays

A *display* means an opto-electronic device that can show a number ("numeric" display); a hexadecimal digit, namely 0–9 and A–F ("hexadecimal display"), or any letter or number ("alphanumeric display"). The dominant display technologies today are LEDs and LCDs (liquid-crystal displays). LCDs are the newer technology, with significant advantages for (a) battery-operated equipment, owing to its very low power dissipation, (b) equipment for use outdoors or in high ambient light levels, (c) displays that require custom shapes and symbols, and (d) displays with many digits or characters. LEDs, by comparison, are somewhat simpler to use, particularly if you only need a few digits or characters. They also come in three colors, and they look good in subdued light, where their good contrast makes them easier to read than LCD displays.

For displays of many characters – say a line or two of text – gas discharge ("plasma") display panels compete with LCDs, particularly if you care about clarity and contrast. They do require significant power, however, so LCD displays are usually preferred for battery-powered applications.

LED displays. Figure 9.22 shows the choices you have in LED displays. The original 7-segment display is the simplest and can display the digits 0–9 and the hex extension (A–F), albeit somewhat crudely (the hex letters are displayed as

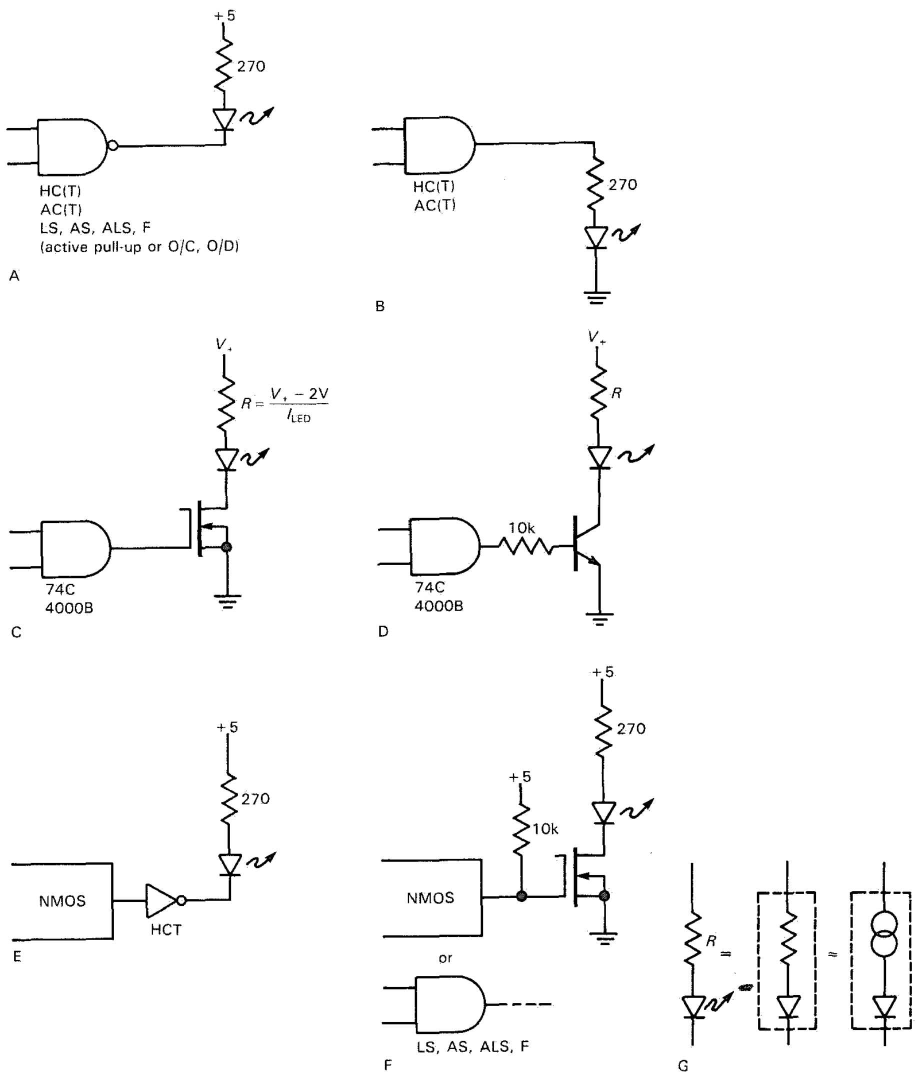


Figure 9.21. Driving LED indicators.

"AbcdEF"). You can get single-character 7-segment displays in many sizes, and in "sticks" containing 2, 3, 4, or 8 characters (generally intended to be "multiplexed" – the characters displayed one at a time in rapid sequence). Single-character displays bring out leads for the 7 segments and the common electrode; the two flavors are thus "common cathode" and "common anode." Multiple-character sticks bring each character's common electrode out, but tie the corresponding segments together, which is what you want for multiplexing.

Sixteen-segment and 5×7 dot-matrix displays are available in two varieties: "dumb" displays that bring out the

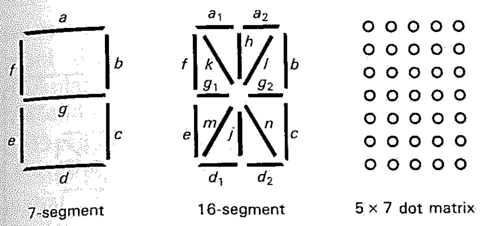


Figure 9.22

segments and common leads (the same as with 7-segment displays) and "smart" displays that do the hard work of decoding and driving for you.

Rather than speak further in generalities, let's look at some examples (Fig. 9.23). The first circuit shows how to drive a single-digit 7-segment common-cathode LED display. The 'HC4511 is a "BCD-to-7-segment latch-decoder-driver," able to source about 15mA while holding its active outputs at +4.5 volts. The series resistors make sure the segment current is limited to that value, with a forward diode drop of 2 volts. You can get arrays of equal-value resistors in convenient SIPs (single in-line package).

You only need a single decoder-driver chip, even if you are displaying multiple digits, as long as you multiplex the display, i.e., illuminate only one displayed digit at a time. Figure 9.23B shows the idea, in this case using an LSI 4-digit counter chip with built-in 7-segment multiplexed drivers. The 74C925 asserts its segment drivers (active HIGH, with plenty of drive capability) for each digit in turn, simultaneously asserting an active HIGH on the corresponding digit output A–D. The rest of the circuitry is self-evident, except perhaps for the unsavory manner in which the digit outputs are pinned a diode drop above ground; luckily, the 74C925 specifies proper operation with this circuit, since the digit outputs are buffered and current-limited.

Figure 9.24A shows how to drive a single hexadecimal display, implemented with a 5×7 dot matrix. The HP 5082-7340 is an example of a "smart" display, with built-in latch, decoder, and driver. All you have to do is assert the 4-bit data, wait at least 50ns, then bring the latch enable HIGH. If you don't want to use its latch, just keep the enable LOW. In Figure 9.24B we've shown one of Siemens' "intelligent" (smarter than "smart"?) displays, in this case a 4-character stick that uses 16-segment displays. This display is intended to look like memory to a microprocessor, something we'll learn about in the next two chapters. To make a long story short, you just assert any 7-bit character and its position ("address," 2 bits), then assert WR′ (write) while making sure the chip is enabled. The data then get stored

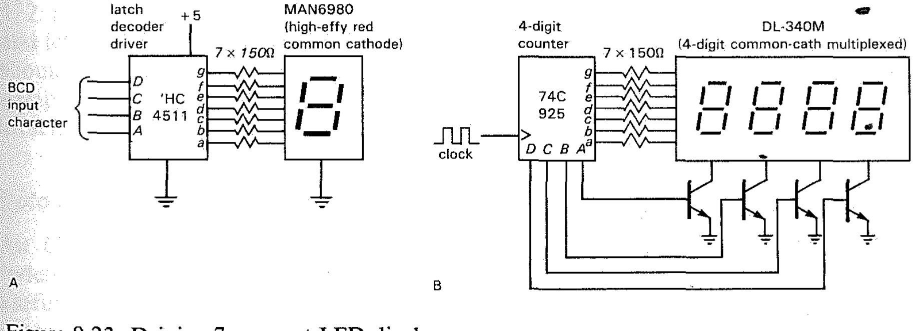


Figure 9.23. Driving 7-segment LED displays.
A. Single-digit.
B. Multiplexed.

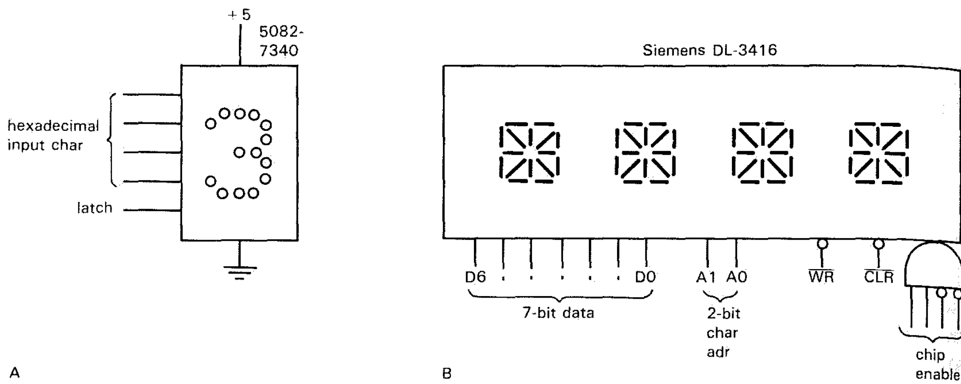


Figure 9.24. Integrated displays.
A. Single-character, dot matrix.
B. 4-character, 16-segment, addressable.

internally, and the corresponding display position changes to the new character. Figure 9.25 shows the character set that can be displayed.

If you want to use a dumb display (perhaps it isn't available with intelligence), but you're spoiled by the simplicity of these intelligent displays, you can simply interpose a chip like the 8-digit Intersil ICM7218/28, which looks like memory to the microprocessor, and which drives a dumb LED display stick with appropriate segment and digit drives. Another alternative is to let the microprocessor do all the work of figuring out which segments and digits to drive, using bits of its "parallel ports" to drive the appropriate lines. This will make more sense to you after you've digested the two microprocessor chapters (Chapters 10 and 11).

LCD and gas-discharge displays. Much of what we have said about LED displays applies also to LCD displays. However, there are some important differences. For one thing, LCDs must be driven by an *ac* waveform; otherwise their liquid guts are ruined. So LCD driver chips usually have

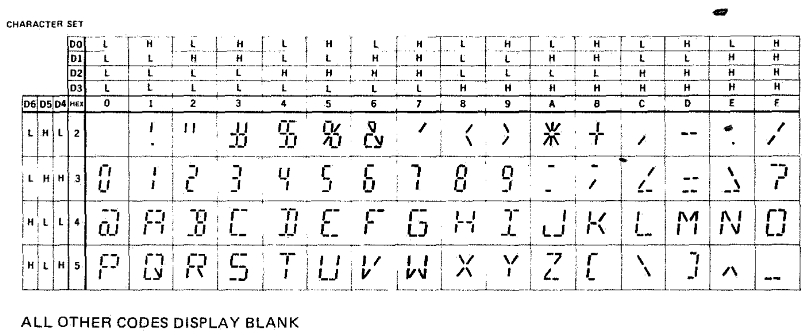


Figure 9.25. Display codes for Siemens DL-3416 16-segment display. (Courtesy of Siemens Components, Inc.)

some way to generate a square-wave segment drive, synchronized to the LCD backplane waveform. An example is the 'HC4543, the LCD cousin of the 'HC4511 LED latch/decoder/driver.

Another difference with LCD displays is that you don't often see single-character displays. Instead, they come in rather large panels that can display a line or two of text. Luckily, the manufacturers realize how complicated things can get, and therefore they provide complete displays that are more than intelligent, they're positively at the genius level. In general, you're talking to these things with a microprocessor, so (as with the display in Figure 9.24) the display is configured to look like a block of memory. Whatever you write gets displayed. Some of the fancier displays go even further, with the ability to store several messages and communicate via serial ports. Look in the EEM (see bibliography) for manufacturers.

Gas-discharge displays feature those handsome red-orange characters you see on some higher-priced portable computers. They require high-voltage drivers, and the manufacturers generally provide the driver circuitry. You can get single- and multiple-digit displays, and you can get large multi-character panels complete with memory and convenient interface. Examples of the latter are the multi-line displays by Cherry, with battery-backed memory that can hold 512 messages, interleave real-time data, and let you edit its memory. You probably shouldn't think of these as *displays*, but as computers that happen to have a display attached!

Opto-couplers and relays

An LED emitter, combined with a photodetector in close proximity, forms a very useful object known as an *opto-coupler* or *opto-isolator*. In a nutshell, opto-couplers let you send digital (and sometimes analog) signals between circuits with separate grounds. This "galvanic isolation" is a good way to prevent ground loops in equipment that drives a remote load. It is essential in circuits that interact with the ac power mains. For example, you might want to turn a heater on and off from a digital signal provided by a microprocessor; in this case you would probably use a "solid-state" relay, which consists of an LED coupled to a high-current triac. Some ac-operated switching supplies (e.g., the supply used in the IBM PC-AT) use opto-couplers for the isolated feedback path (see Section 6.19). Similarly, designers of high-voltage power supplies sometimes use opto-couplers to get a signal up to a circuit floating at high voltage.

Even in less exotic situations you can take advantage of opto-isolators. For example, an opto-FET lets you switch analog signals with no charge injection whatsoever; the same goes for sample-and-hold circuits and integrators. Opto-isolators can keep you out of trouble when driving industrial current loops, hammer drivers, etc. Finally, the galvanic isolation of opto-isolators comes in handy in high-precision or low-level circuits: For example, it is difficult to take full advantage of a 16-bit analog-to-digital converter, because the digital output signals (and noise on the digital ground to which you connect the converter's output) get back into the analog front end. You can extricate yourself from "noise city" with optical isolation of the digital half.

Opto-couplers typically provide 2500 volts (rms) isolation, 10^{12} ohms insulation resistance, and less than a picofarad coupling between input and output.

Before looking at actual opto-couplers, let's take a quick look at photodiodes and phototransistors. Visible light causes ionization in silicon, producing charge pairs in the exposed base region; this mimics exactly the effect of an externally applied base current. There are two ways to use a phototransistor: (a) You can use it as a *photodiode*, connecting only to the base and collector leads; in that case the detected photocurrent will typically be a few percent of the LED drive current. A photodiode generates its photocurrent whether or not you apply a bias voltage; thus you can hook it directly into an op-amp summing junction (a virtual short circuit), or you can back-bias it (Fig. 9.26A,B). (b) If you let the photodiode current act as a base current, you get normal transistor current multiplication, with a resulting I_{CE} that is typically 100 times larger; in this case you must bias the transistor, as in Figure 9.26C. You pay for the increased current with slower response, because of the open base circuit. You can add a resistor from base to emitter to improve the speed; however, this produces a threshold effect, since the phototransistor doesn't begin to conduct until the photodiode current is large enough to produce a V_{BE} across the external base resistor. In digital applications the threshold can be useful, but in analog applications it is an undesirable nonlinearity.

Figure 9.26D–S shows typical examples of nearly every kind of opto-coupler you are likely to encounter. The earliest (and simplest) is typified by the 4N35, an LED-phototransistor pair with 40% (min) current transfer ratio (CTR) as a phototransistor, and sluggish $5\mu s$ turn-off time (t_{OFF}) into a 100 ohm load. The figure shows how to use it: A gate and resistor generate current-limited 8mA drive, and a relatively large collector resistor guarantees saturated switching of the output between logic levels. Note the use of a Schmitt-trigger inverter, a good idea here because of the long switching times. You can get LED-phototransistor pairs with CTRs of 100% or more (e.g., the MCT2201, with CTR=100% min), and you can get LED-photo-Darlingtons, as shown; they're even slower than phototransistors! To get improved speed, the manufacturers sometimes use separate photodiode and transistor, as shown in the 6N136 and 6N139 opto-transistor and opto-Darlingtons.

These opto-couplers are nice, but somewhat annoying to use because you have to supply discrete components at both input and output. Furthermore, the input loads ordinary logic gates to their maximum capability, and the passive pullup output suffers from slow switching and mediocre noise immunity. To remedy these deficiencies the silicon wizards bring us "logic" opto-couplers. The 6N137 in Figure 9.26I goes halfway, with diode input and logic output; you still need plenty of input current (specified as 6.3mA, min, to guarantee output switching), but you get clean logic swings at the output (albeit open-collector), and speeds to 10Mbit/sec. Note that you must supply +5 volts to the internal output circuitry. The newer 74OL6000 series from General Instrument (Fig. 9.26J) does what you really want: It accepts logic-level inputs, and produces logic-level outputs, with both totem-pole and open-collector types available; these opto-couplers operate to 15Mbit/sec. Because of the internal logic circuitry at both input and output, both sides of the chip require logic supply voltages.

Figure 9.26 continues with some variations on the LED-phototransistor theme. The IL252 hooks a pair of LEDs back-to-back, so you can drive it with ac. The IL11 uses a long isolation gap (and package) to obtain 10kV rms voltage standoff, compared with the usual 2.5kV rms value for all the other couplers shown. The H11C4 is an opto-*SCR*, useful for switching high voltages and currents. The MCP3023 replaces the unidirectional SCR with a *triac*, which is a bidirectional SCR; it can drive an ac load directly, as in Figure 9.15N. When driving ac loads, it's best to switch on the load during a zero crossing of the ac waveform, in order to avoid putting spikes onto the power line. This is easily done, with an opto-triac containing

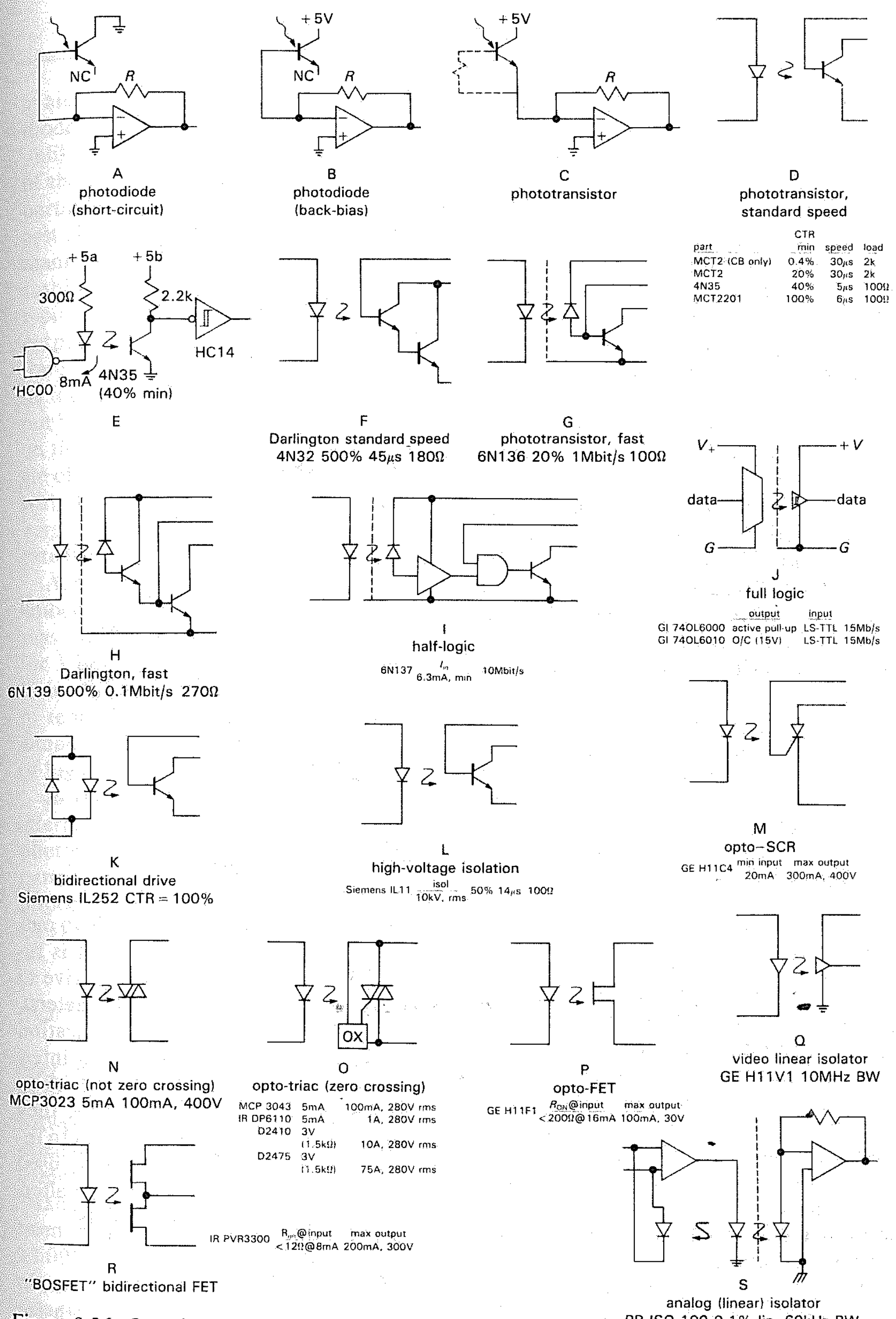


Figure 9.26. Opto-isolators.

internal "zero-voltage-switching" circuitry (which blocks triac drive until the next zero crossing); the small MCP3043 shown uses such circuitry, as do the higher-current "solid-state relays" shown. The DP6110 "ChipSwitch" from IR comes in a 16-pin DIP (with all but 4 pins missing), while the mighty D2410 and D2475 are packaged in 1.75″×2.25″×1" power modules intended for mounting to a heat sink.

The remaining opto-couplers in Figure 9.26 can be used for linear signals. The H11F series of opto-FETs can be used as an isolated variable resistor or isolated analog switch. There are no problems of compatible voltage levels, SCR latchup, or charge injection. You might use one of these in a sample/hold or integrator. The PVR series of "BOSFETs" are similar, but with a pair of power MOSFETs hooked in series as the output element. They are intended primarily to switch ac loads directly, in the manner of an opto-triac. The H11V1 is a linear video isolator, with 10MHz bandwidth. And the ISO-100 from Burr-Brown is a clever analog isolation element, in which the LED couples to *two* matched photodiodes; one is used in a feedback loop to linearize the coupled response to the second photodiode.

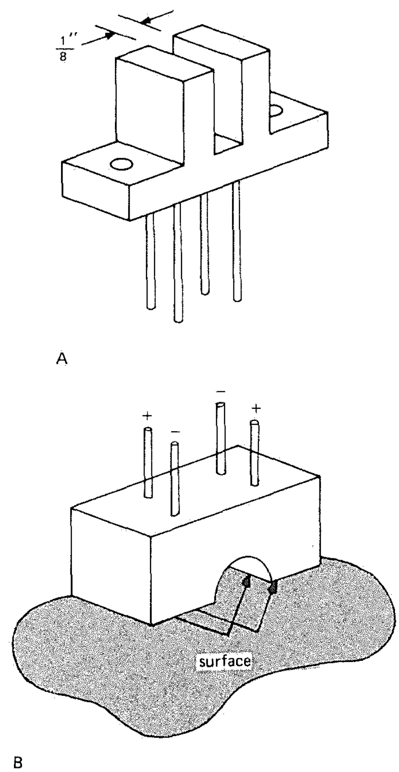


Figure 9.27. A. Optical interrupter.
B. Reflective object sensor.

□ *Interrupters*

You can use LED-phototransistor technology to sense proximity or motion. An "optical interrupter" consists of an LED coupled to a phototransistor across a 1/8 inch slot. It can sense the presence of an opaque strip, for example, or the rotation of a slotted disk. An alternative form has the LED and photodetector looking in the same direction, and it senses the presence of a reflective object nearby (most of the time, anyway!). Take a look at Figure 9.27. Optical interrupters are used in disk drives and printers, to sense the end of travel of the moving assembly. You can get optical "rotary encoders" that generate a quadrature pulse train (two outputs, 90° out of phase) as the shaft is rotated. These provide a nice alternative to resistive panel controls (potentiometers). See Section 11.09. In any application where you're considering an optical interrupter or reflective sensor, take a look at Hall-effect sensors as an alternative; they use solid-state magnetic-field sensors to indicate proximity. They're commonly used in automobile ignition systems as an alternative to mechanical breaker points.

□ *Emitters and detectors*

We've already mentioned LEDs, both for displays and opto-couplers. A recent development in opto-electronics is the availability of the inexpensive solid-state diode *laser*, a true coherent source of light, unlike the diffuse LED. You can see one of these when you open the top door of a portable compact-disk player. Diode lasers cost about $20, sold by the companies that manufacture consumer electronics (Matsushita, Mitsubishi, Sharp, and Sony). A typical diode laser generates 10mW of light output at 800nm (invisible, in the "near infrared") when powered by 80mA through its forward diode drop of 2 volts. The output beam emerges directly from a tiny spot on the GaAlAs chip, diverging with an angle of about 10°–20°; it can be collimated with a lens to form a parallel beam or a very small focal spot. LED lasers are used extensively in optical fiber transmission.

Another recent emitter technology is embodied in high-density linear LED arrays, 300 emitters/inch or more, intended for LED printers. If the semiconductor technology is successful, these are likely to replace laser printers, because they are simpler and more reliable and ultimately of higher resolution.

In the detector arena there are several alternatives to the simple photodiode and phototransistor that we discussed earlier, particularly when speed or sensitivity is important. We will discuss PIN diodes, CCDs, and intensifiers in Section 15.02.

DIGITAL SIGNALS AND LONG WIRES

Special problems arise when you try to send digital signals through cables or between instruments. Effects such as capacitive loading of the fast signals, common-mode interference pickup, and "transmission-line" effects (reflections from impedance mismatching, see Section 13.09) become important, and special techniques and interface ICs are often necessary to ensure reliable transmission of digital signals. Some of these problems arise even on a single circuit board, so a knowledge of digital transmission techniques is generally handy. We will begin by considering on-card problems. Then we will go on to consider the problems that arise when signals are sent between cards, on data buses, and finally between instruments via twisted-pair or coaxial cables.

9.11 On-board interconnections

Output-stage current transient

The push-pull output circuit for TTL and CMOS ICs consists of a pair of transistors going from V_+ to ground. When the output changes state, there is a brief interval during which both transistors are ON; during that time, a pulse of current flows from V_+ to ground, putting a short negative-going spike on the V_+ line and a short positive spike on the ground line. The situation is shown in Figure 9.28. Suppose that IC_1 makes a transition, with a momentary large current from +5 to ground along the paths as indicated [with 74Fxx or 74AC(T)xx circuits the current might reach 100mA]. This current, in combination with the inductance of the ground and V_+ leads, causes short voltage spikes relative to the reference point, as shown. These spikes may be only 5ns to 20ns long, but they can cause plenty of trouble: Suppose that IC_2, an innocent bystander located near the offending chip, has a steady LOW output that drives IC_3 located some distance away. The positive spike at IC_2's ground line appears also at its output, and if it is large enough, it gets interpreted by IC_3 as a short HIGH spike. Thus, at IC_3, some distance away from the troublemaker IC_1, a full-size bona fide logic output pulse appears, ready to mess up an otherwise well-behaved circuit. It doesn't take very much to toggle or reset a flip-flop, and this sort of ground-current spike can do the job nicely.

The best therapy for this situation consists of (a) using hefty ground lines throughout the circuit, even to the extent of

LOW-POWER DESIGN

CHAPTER 14

INTRODUCTION

Lightweight hand-held instruments, data loggers that make measurements at the ocean bottom, digital modems that power themselves from the "holding current" in the telephone line - these are just a few of the applications that invite (or demand) low-power electronic design techniques. Within such instruments you often find examples of all the areas of design already treated in this book - regulated supplies, linear circuits (both discrete and op-amp), digital circuits (almost invariably CMOS) and associated conversion techniques, and, increasingly, microprocessor circuits of considerable complexity. Although we have occasionally discussed power consumption and speed/power trade-offs (when comparing logic families, for example), the design of micropower electronic instruments involves special techniques and cautions throughout and requires a chapter all its own.

We will begin by considering a range of applications where low-power consumption is desirable or essential. They're not all of the exotic transmitter-attached-to-wild-moose variety, and in fact there are plenty of places where ac line power is plentiful, but where battery power is more convenient or performs better. Next we will review the power sources that make sense for low-power electronics, beginning with the ubiquitous "primary" (nonrechargeable) cells (alkaline, mercury, silver, lithium) and the closely related "secondary" (rechargeable) cells (nickel-cadmium, lead-acid, "gel" cells). We've hounded the manufacturers of batteries mercilessly for their latest technical literature, and so we've got really useful battery comparison data - energy content, discharge characteristics, effects of temperature, discharge rate, and storage conditions, etc. We'll help you pick the right battery for your application.

Batteries aren't the only power sources to consider, so we'll continue by talking about those little black plastic "wall-plug-in" modules you get with the consumer electronics gizmos you can buy. Wall plug-ins are remarkably inexpensive and are available as simple transformers, or unregulated dc supplies (transformer-rectifier-capacitor), or as complete regulated dc supplies; they also come in dual- or triple-supply varieties. Solar cells are useful in some unusual applications, so we'll discuss them too. Finally, we'll mention the use of signal currents, for example the dc current you find on the telephone line, or the ac relay energizing current in a thermostat or doorbell, to activate a micropower instrument.

We'll then talk about low-power design at the same level of detail that we've used in the rest of the book, mirroring a similar progression of subjects: regulators and references, then linear circuitry (discrete and op-amp), digital circuits and conversion techniques, and finally microprocessors and memories. There are, in addition, techniques that have no counterpart in ordinary design, for example "power switching," in which a normal circuit is rendered micropower by applying dc for extremely short intervals; for example, a microprocessor data logger might be powered for 20ms every 60 seconds. We'll talk about some of those tricks, and the curious pitfalls that await the uninitiated.

Finally, we'll talk about packaging – including the small plastic instrument cases that are widely available, complete with back door for access to the batteries. Micropower instruments are generally much easier to package than conventional instruments are, since they're usually lightweight, they don't produce heat, and they don't need the usual paraphernalia of power cords, line filters, and fuses. Micropower design is different and fun, and it presents new challenges for the jaded electronic designer. Read on to learn more about this exciting (but often neglected) subject!

We hate to say it, but this chapter could be passed over in a first reading of the book.

14.01 Low-power applications

We've collected together most of the reasons that might motivate you to do low-power circuit design. They're listed here, in no particular order.

Portability

You can't carry it around with you if it has a power cord running off to a wall outlet. "It" could be a commercial product such as a calculator, wristwatch, hearing aid, "walkabout" tape recorder or receiver, paging radio, or digital multimeter. Or it might be a custom portable instrument, for example a small transmitter used to study herd migration and physiology. Because batteries have a finite energy content, you've got to keep power consumption low in order to have reasonable battery life with acceptable battery weight. A multimeter that runs 1000 hours on a single 9 volt battery will outsell a competing unit that requires four D cells and gives only 100 hours of service. A portable transmitter for animal migration studies is useless if it runs only two days on a fresh set of batteries. Thus, low-power circuitry is at a premium in instruments designed for portability. In the special case of extremely small instruments (e.g., a wristwatch), the tiny energy content of the self-contained batteries dictates micropower design, with total current drains of just a few microamps.

Isolation

Instruments that are powered from ac line current are not suitable for some kinds of "floating" measurements at high potential. For example, you might want to measure microamp charged-particle-beam currents at the +100kV terminal of a particle accelerator. It may well be that you can't make the measurement by lifting the low-voltage end off ground (as in Fig. 4.79), because the power transformer of the high-voltage supply causes significant 60Hz currents to be capacitively coupled to the high-voltage supply through its transformer (or perhaps because of corona discharge and other high-voltage leakage effects, which add spurious current as measured at the ground return). If you try to build an ac-powered instrument to measure the current, say by using a differential amplifier connected across a precision resistor in the high-voltage lead, the power supply of your instrument will have to use a special power transformer rated at 100kV insulation breakdown so that the measuring circuit (op-amps, readouts) can float at 100kV. Since such a transformer is almost impossible to find, this is a good place for a circuit powered by a battery (or possibly by a solar cell, illuminated by light beam across the potential gap), intrinsically isolated from the power line and from earth ground.

In the foregoing example, an ac-powered circuit would have another problem, namely that it would impress some 60Hz ripple onto the circuit it was attempting to measure, due to capacitive coupling and leakage of 60Hz currents through its power transformer. So the power transformer would have to be of special design to ensure low inter-winding capacitance and low leakage currents. This problem of coupled 60Hz ripple can also crop up in conventional circuits that deal with signals at very low levels, for example weak audio signals. Although such problems can usually be solved by careful design, the isolation provided by a battery-powered preamp can be a real advantage in these situations.

Little power available

Dialers, modems, remote data-acquisition systems that send through the telephone lines – these are examples of instrument designs that can be powered by the holding current of the telephone line itself (it's about 50V dc open-circuit, driven from an impedance of about 600Ω; you must load it so that the dc voltage is below 6V, in order for the telephone company to think you're "off hook," and therefore to maintain the connection). Likewise, "smart" thermostats for heating systems often use NiCd rechargeable cells for their dc power, charged during intervals when the relay is not activated by the low-current ac then available (usually a 24V ac transformer in series with a relay coil of a few hundred ohms resistance).

This same trick of powering your circuit with ac signaling currents could also be used with doorbell circuits, and any other application where low-voltage ac relays are used. Another example of extracting power from a signaling current is the use of "industrial sensor current loops," in which a dc current in the range 4mA to 20mA (or, sometimes, 10mA to 50mA) is used to send analog sensor measurements over a two-wire system. Modules using this standard typically permit a voltage drop of 5 to 10 volts; hence the opportunity to power remote instrumentation from the signal current itself.

For these applications you have available a power supply delivering currents of the order of a few milliamps across a few volts, which is enough to power relatively complex low-power circuits. It is certainly attractive to attempt signal-current-driven low-power design, given the more cumbersome alternative of separate ac power sources for the same instrument.

A final example of a power source that limits you to a few milliwatts is the use of solar cells to power instruments (and/or charge their batteries). There are wristwatches and inexpensive pocket calculators built this way, and they have the advantage of (a) staying sealed and (b) remaining inexpensive, respectively.

No power available

Battery operation really becomes essential when there's nothing else available. Examples include physical oceanography, where you may wish to deploy a set of sensors on the ocean bottom for six months, quietly logging ocean currents, sediments, salinity, temperature, and pressure, as well as environmental studies, where remote measurements of pollutants at inaccessible sites are required. In these applications you usually want extended operation on a set of batteries, sometimes up to a year or more; hence the need for careful micropower design.

There are other situations where ac power is available, but not convenient. Household examples include smoke detectors and wall clocks.

Minimizing heat management

Digital circuitry constructed with ECL or Schottky technologies can easily consume 10 watts or more per board, and a system of several such boards requires forced air cooling. On the other hand, the newer high-speed CMOS logic families (with names such as 74AC*xx* and 74ACT*xx*) deliver performance comparable to that of their advanced Schottky cousins, with negligible static power consumption and greatly reduced dynamic power consumption (Figs. 8.18 and 9.2). That means smaller power supplies, closed dirt-free enclosures (no fan), and better long-term reliability.

The same considerations apply to linear design, making low power consumption a desirable objective to keep in mind in almost any application, even when plenty of power is available.

Uninterruptibility

Momentary power interruptions often cause microprocessor-based instruments to re-initialize themselves, computers to crash, etc. A nice solution is the use of an uninterruptible power supply (UPS), usually in the form of a battery-powered dc-to-ac inverter with 115 volt 60Hz output, able to switch on automatically within a few milliseconds of a power interruption. Uninterruptible power supplies are available with power ratings of many kilowatts. The big ones are expensive and bulky; however, there are compact units, powered by a small bank of lead-acid gel cells (see Section 14.02), for systems that use less than a kilowatt of ac power. For truly low-power systems, a small UPS inverter or direct dc battery backup (as in Fig. 1.83) is convenient and is a good reason to practice low-power design.

POWER SOURCES

14.02 Battery types

The Duracell "Comprehensive Battery Guide" lists 133 off-the-shelf batteries, with descriptions like zinc-carbon, alkaline manganese, lithium, mercury, silver, zinc-air, and nickel-cadmium. There are even subclasses, for example Li/FeS_2, Li/MnO_2, Li/SO_2, $Li/SOCl_2$, and "lithium solid state." And from other manufacturers you can get sealed lead-acid and gel-type batteries. For the truly exotic application you might even want to consider fuel cells or radioactive thermal generators. What are all these batteries? How do you choose what's best for your portable widget?

The foregoing list divides into so-called *primary* and *secondary* batteries. Primary batteries are designed for a single discharge cycle only, i.e., they're nonrechargeable. Secondary cells (NiCd, lead-acid, and gel-type in the foregoing list), by comparison, are designed to be recharged, typically from 200 to 1000 times. Among primary batteries, you usually make your choice of chemistry based on trade-offs among price, energy density, shelf life, constancy of voltage during discharge, peak current capability, temperature range, and availability.

Once you've picked the right battery chemistry, you figure out which battery (or series combination of batteries) has enough energy content for the job.

Fortunately, it's pretty easy to eliminate most of the batteries in the catalogs, if you follow our first suggestion: *Avoid hard-to-get batteries.* Besides being hard to find, they're usually not fresh. So it's usually better to stick with the varieties available at the drugstore, or perhaps photography store, even if it results in somewhat less than optimum design. We particularly recommend the use of commonly available batteries in the design of any consumer electronic device; as consumers ourselves, we shun those inexpensive marvels that use exotic and expensive batteries. (Remember those early smoke detectors that used an 11.2V mercury battery?)

□ *Primary batteries*

Now for details. Table 14.1 compares the characteristics of the various primary cells, and Table 14.2 and Figure 14.1 give actual numbers for the most popular cells.

The old-fashioned "dry cell" with a cat on the outside is a LeClanche cell. Inside it's as primitive as you might guess, with a carbon rod stuck down into a cathode mixture of manganese dioxide, carbon, and ammonium and zinc chloride electrolytes. There's a cylindrical separator made of flour-and-starch paste, then a zinc anode outer can. The top is sealed on with wax and asphalt seals, designed to vent the innards if too much pressure builds up. These cells are the cheapest you can buy, but you don't get too much for your money. In particular, the voltage drops and the impedance rises steadily as the battery is used; furthermore, the battery's capacity drops drastically if used at high currents.

The "heavy-duty" dry cells are similar, but with a higher proportion of zinc chloride and correspondingly different mechanical design to accommodate greater gassing. Although their total energy content is only slightly greater than that of LeClanche cells, these cells are considerably better in delivering most of their rated capacity even when operated at high currents. For example, a LeClanche D cell delivers 4.2 amp-hours (Ah) into a 150 ohm load, 1.2Ah into 15 ohms, and 0.15Ah into 1.5 ohms; the equivalent zinc chloride cell delivers 5.6, 5.4, and 1.4 amp-hours, respectively. The zinc chloride cell also shows less dropoff of capacity at low temperatures.

The alkaline manganese cell, generally sold as simply "alkaline," is better still in high-current-discharge and low-temperature operation. It is inside-out, compared with zinc-carbon, having the powdered-zinc negative anode and potassium hydroxide electrolyte in the middle, surrounded by a manganese-dioxide-and-carbon outer positive cathode. For comparison with the numbers above, an alkaline D cell delivers 10 amp-hours into 150 ohms, 8Ah into 15 ohms, and 4Ah into 1.5 ohms. Because of its particular chemistry, an alkaline battery maintains a low and slowly increasing internal resistance as it discharges, compared with the rapidly rising internal resistance of both types of zinc-carbon cells. It also works better at low temperatures. Alkaline batteries have a longer shelf life than LeClanche or zinc chloride. As Figure 14.1 suggests, the cell voltage-versus-discharge curve for all three types of batteries lets you easily estimate the condition of the battery. Figure 14.2 shows comparative performance for the three kinds of "dry cells."

Mercury, silver oxide, lithium – these are the real premium cells, with greatly superior performance compared with alkaline and zinc-carbon. The mercury cell uses an amalgamated zinc anode, mercuric-oxide-plus-carbon cathode, and sodium or potassium hydroxide electrolyte. It excels in constancy of open-circuit

TABLE 14.1. PRIMARY BATTERIES

Type	Advantages	Disadvantages
Zinc-carbon (LeClanche) (standard "dry cell")	least expensive widely available	lowest energy density (1–2Wh/in^3) sloping discharge curve poor high-current performance impedance increases as discharged poor low-temperature performance
Zinc-carbon (zinc chloride) ("heavy duty" dry cell)	less expensive than alkaline better than LeClanche at high curr and low temp	low energy density (2–2.5Wh/in^3) sloping discharge curve
Alkaline manganese ("alkaline" dry cell)	moderate cost better than zinc chloride at high curr and low temp maintains low impedance as discharged moderate energy density (3.5Wh/in^3) widely available	sloping discharge curve
Mercury	high energy density (7Wh/in^3) flat discharge curve good at high temperatures good shelf life low and constant impedance open-circuit voltage 1.35V±1%	expensive poor at low temp (<0˚C)
Silver oxide	high energy density (6Wh/in^3) flat discharge curve good at high & low temp (to -20˚C) excellent shelf life	expensive
Lithium oxyhalide	high energy density (8Wh/in^3) highest energy density per unit weight flat discharge curve excellent at high & low temp (to -55˚C) extraordinary shelf life (5–10 yrs @70˚C) light weight high cell voltage (3.0V)	expensive
Lithium solid-state	high energy density (5–8Wh/in^3) excellent at high & low temp (-40˚C to 120˚C) unbelievable shelf life (>20y @ 70˚C) light weight	expensive low current drain only

TABLE 14.2. BATTERY CHARACTERISTICS

Type	R_{int} (Ω)	V_{oc} (V)	Capacity[a] continuous, to 1V/cell (mAh)	@ (mA)	(mAh)	@ (mA)	Size (in)	Weight (gm)	Connec[b]	Comments
9V "1604"										
Le Clanche	35	9	300	1	160	10	0.65x1x1.9	35	S	
Heavy Duty	35	9	400	1	180	10	"	40	S	
Alkaline	2	9	500	1	470	10	"	55	S	280mAh@100mA
Lithium	18	9	1000	25	950	80	"	38	S	Kodak Li-MnO_2
1.5V Alkaline										
D	0.1	1.5	10000	10	8000	100	1.3Dx2.2L	125	B	4000mAh @ 1A
C	0.2	1.5	4500	10	3200	100	1.0Dx1.8L	64	B	
AA	0.4	1.5	1400	10	1000	100	0.55Dx1.9L	22	B	
AAA	0.6	1.5	600	10	400	100	0.4Dx1.7L	12	B	
Mercury										
625	–	1.35	250[c]	1	250[c]	10	0.62Dx0.24L	4	B	
675	10	1.35	190[c]	0.2	–	–	0.64Dx0.21L	2.6	B	
431	–	11.2	1000[c]	25	–	–	1.0Dx2.9L	115	S	
Silver										
76	10	1.55	180	1	–	–	0.46Dx0.21L	2.2	B	
Li-Oxyhalide										
D	–	3.9	14000[d]	175	10500[d]	350	1.3Dx2.3L	113	B,T	$SOCl_2/BrCl$
D	–	3.95	14000[d]	175	12000[d]	1000	"	110	B,T	SO_2Cl_2/Cl_2
D	–	3.5	9500[d]	175	8500[d]	1000	"	120	B,T	$SOCl_2$
Li solid	–	4.0	350[d]	1μA	175[d]	0.1	1.2Dx0.23L	16	T	high impedance
Ni-Cd										
D	0.009	1.3	4000[c]	800	3500[c]	4000	1.3Dx2.3L	130	B	Saft/Powersonic
9V	0.84	8.1	100[c]	10			0.65x1x1.9	35	S	
Pb-acid										
D	0.006	2.0	2500[e]	25	2000[e]	1000	1.3Dx2.6L	180	T	

(a) see Fig. 14.1 for discharge curves. (b) B - button; S - snap; T - solder tabs. (c) to 0.9V/cell. (d) to 2.5V/cell. (e) to 1.75V/cell

voltage (1.35V, stable to 1%) as well as constancy of voltage during discharge (a "flat discharge curve"); see Figure 14.1. It performs well at temperatures up to 60°C, but performance is seriously degraded below −10°C.

The silver oxide cell is similar to the mercury cell, but with the mercuric oxide replaced by silver oxide. It, too, has a very flat discharge curve, but with higher open-circuit voltage (1.6V) and improved performance at low temperatures (to −20°C).

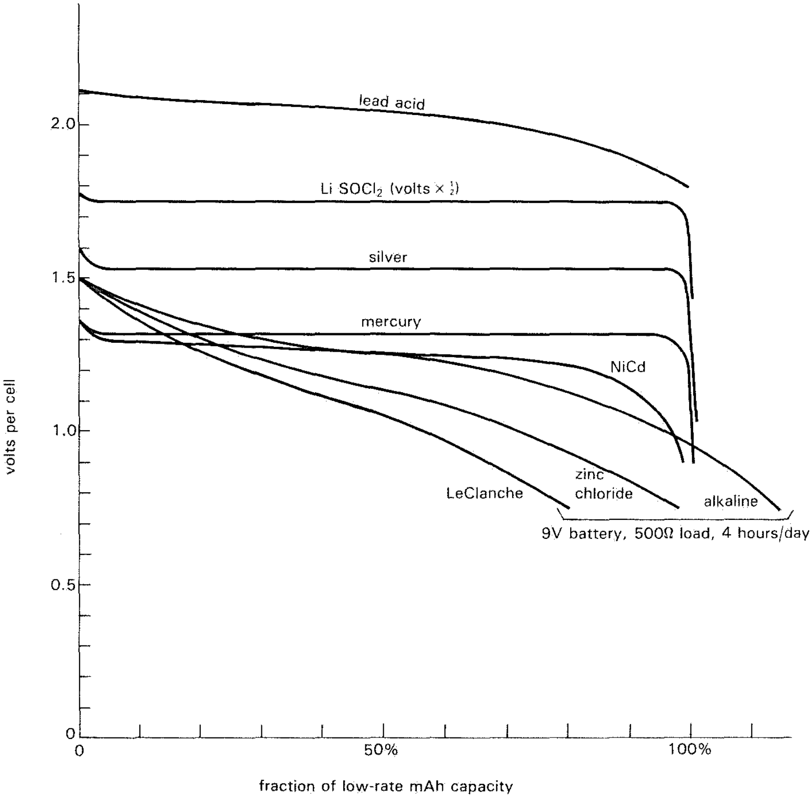
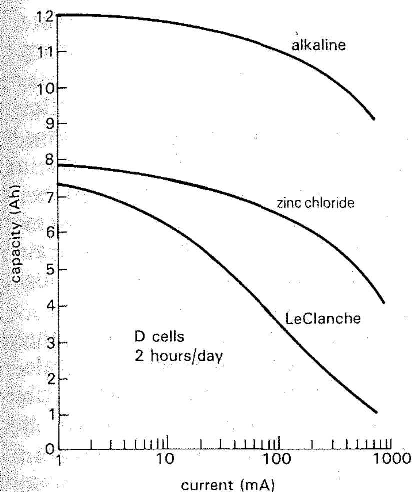


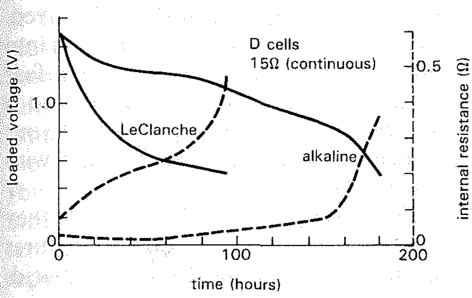
Figure 14.1. Discharge curves for primary batteries. (This and subsequent figures in this chapter are adapted from battery literature by Arco Solar, Duracell, Electrochem Industries, Eveready, Gates, Kodak, PowerSonic, Solavolt, and Yuasa.)

Lithium cells are the newest of the commercially available primary cells, with several different chemistries available. They have the highest energy density per unit weight. They are the best performers at very high and low temperatures, and they have extraordinary shelf life at all temperatures. For example, a D-size lithium thionyl chloride ($Li/SOCl_2$) cell delivers more than three times the energy (10Ah at a terminal voltage of 3.5V) of an alkaline D cell, with comparable size and weight. Lithium batteries will operate down to −50°C and up to 70°C (see Fig. 14.3), with 50% of their room-temperature service at −40°C, a temperature at which other primary batteries cease to function at all. Lithium batteries have shelf lives of 5 to 20 years at room-temperature and can be stored for 1–2 years at 70°C, a temperature that makes other batteries wither. They have a flat discharge curve. Their long shelf life and 3–3.5 volts make lithium batteries ideal for on-board CMOS memory backup.

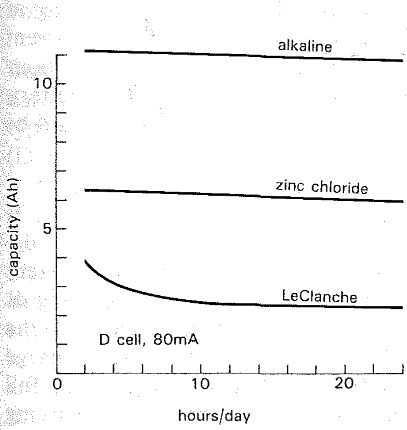
Each lithium chemistry system has its own peculiarities. For example, lithium thionyl chloride batteries have a tendency to develop an electrode passivation that raises their internal resistance enormously; it can be "burned off" by momentary operation at high current. Lithium sulfur dioxide has been implicated in some battery explosions:



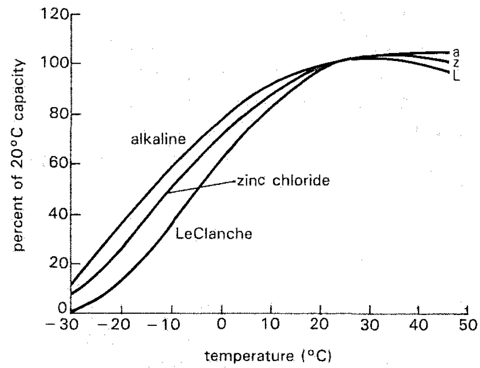
A. D-cell capacity versus load current



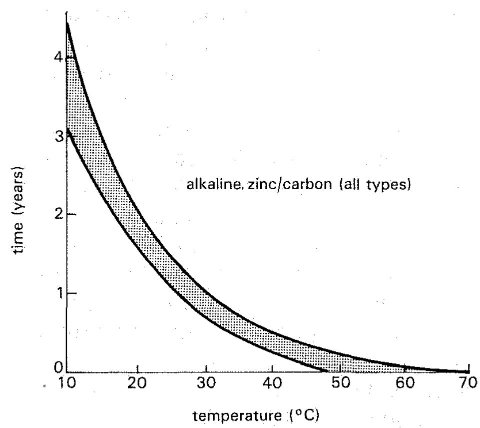
B. D-cell voltage and internal resistance during discharge



C. D-cell capacity versus duty cycle



D. capacity versus temperature



E. shelf life (10% loss of capacity) versus temperature

Figure 14.2. Zinc "dry-cell" performance comparison.

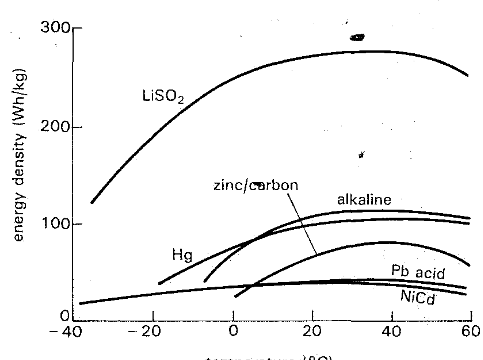


Figure 14.3. Energy density versus temperature for primary cells.

Warning: There have been incidents in which lithium batteries have exploded, in some cases causing severe personal injury. Having warned you about this, we will not be responsible for any calamities you might experience at the hands of a lithium battery.

□ *Secondary batteries*

For use in electronic equipment, your choices are (a) nickel-cadmium ("nicad") or (b) sealed lead-acid. Both have lower energy content than primary cells (Table 14.2), but they *are* rechargeable. Nicad cells provide 1.2 volts, are generally available in the 100mAh–5Ah range, and work down to −20°C (and up to +45°C); lead-acid batteries provide 2 volts per cell, are generally built to provide 1 to 20 amp-hours, and will work down to −65°C (and up to +65°C). Both types have relatively flat discharge curves. Lead-acid batteries have low self-discharge rates and are claimed to retain two-thirds of their charge after a year's storage at room temperature (though our experience leads us to be skeptical); nicad batteries have relatively poor charge retention, typically losing half their charge in 4 months (which we do believe!) (see Fig. 14.4). A nicad D cell provides 5Ah (at 1.2V), whereas a lead-acid D cell provides 2.5Ah (at 2V); the comparable alkaline cell provides 10Ah at 1.5 volts.

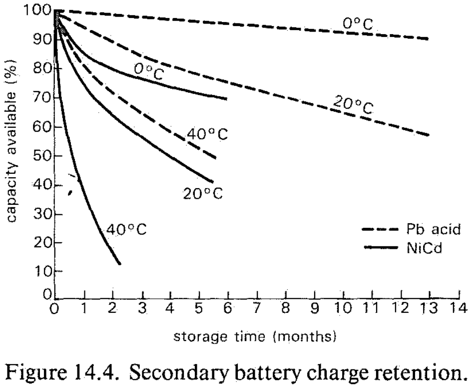


Figure 14.4. Secondary battery charge retention.

Both nicad and sealed lead-acid batteries claim to be good for 250–1000 charge/discharge cycles (more if they are only partially discharged each time; less if completely discharged, or charged/discharged rapidly). Nicads have an overall life expectancy of 2–4 years if held at a constant trickle charge current (see below); the comparable life for sealed lead-acid batteries held at constant "float" voltage is claimed to be 5–10 years.

It's worth pointing out that these rechargeable batteries really are sealed; they won't drip mysterious and terrible chemicals. In particular, although the name "lead-acid" conjures up images of husky car batteries with corroded terminals and leaking acid, the sealed types really are clean batteries: You can run them in any position, they don't drip or ooze, and they're generally well-behaved. In our experience you can design them into real electronic instruments without fear of your circuit boards dissolving into a white crusty plague, or the bottom of your expensive enclosure becoming awash with foul-smelling bilge juices.

Secondary batteries die young if they aren't charged correctly. The procedures are different for nicad and lead-acid. It's conventional to designate charging rates in terms of the ampere-hour capacity of the battery; for example, charging at "C/10" means applying a charging current equal to one-tenth of the ampere-hour capacity of the fully charged battery. For the nicad D cell above, that would be 500mA.

□ *Nickel-cadmium.* Nicad cells are designed to be charged at constant current and to withstand continuous charging at C/10. Because of inefficiencies in the charge/discharge cycle, you have to charge at this rate for 14 hours to guarantee a full charge; you can think of this as charging the battery 140%.

Although it's OK to overcharge nicad batteries indefinitely at C/10, it's better to switch over to a "trickle" charge, typically at C/30 to C/50. However, nicads are funny, with a "memory" effect, so that a trickle rate may fail to revive a fully discharged battery; a minimum of C/20 is recommended.

There are applications where you can't wait a whole day for nicads to recharge. Nicad literature gives you permission to charge normal cells at a "high rate" of C/3 to C/10, if you don't do it too much. Up to three days at C/3 is about the limit. There may be some venting of gases under these conditions, in contrast to "normal" charging at C/10 in which internally evolved oxygen gets recombined within the cell. There are special "fast-charge" nicad cells designed to be charged at C/1 to C/3 in a special charger that senses the fully charged condition by monitoring cell temperature (they have internal chemistry that makes them heat rapidly once they are charged). Unlike the situation with lead-acid batteries, you can't reliably determine when a nicad is fully charged by monitoring terminal voltage, because it changes with repetitive cycling, temperature, and rate. Nicads should not be charged by a constant voltage, nor held "floated" at a fixed voltage.

You can buy handy little nicad chargers from several companies, including the battery manufacturers themselves. They typically let you charge all the popular sizes (D, C, AA, and 9V).

Nicads have a pathology all their own. If you're like us, you probably take it personally when your rechargeable calculator dies during a tax audit. As the graphs show (Fig. 14.5), nicads have "memory" effects, so that the first discharge after a long period of charging may be poor. They are intolerant of reverse polarity; thus, the first cell to discharge suffers horribly if a series string of them is fully discharged. Likewise, nicads shouldn't be connected in parallel. You'll find people promoting various snake remedies, such as periodic "deep discharge," or shock therapy in the form of a substantial electrolytic capacitor discharged across a moribund nicad. Although we're skeptical of the latter, periodic deep discharge is important for nicad health.

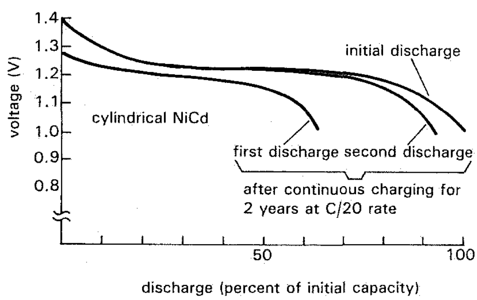


Figure 14.5. Nicad battery restored to good health by "deep discharge."

□ *Lead-acid.* These versatile batteries can be charged by applying a current-limited constant voltage, a constant current, or something in between. With current-limited constant-voltage charging, you apply a fixed voltage (typically between 2.3V and 2.6V per cell); the battery initially draws a high current (up to 2C), but tapers down as it charges, eventually leveling off to a trickle current that maintains the battery in a fully charged state. A higher applied voltage gives you a faster charge, at the expense of greater required charger current and reduced overall battery life. A simple implementation is to use a 3-terminal regulator like the 317 to supply a current-limited fixed voltage. The battery's charge can be held indefinitely by maintaining a fixed "float" voltage between 2.3 and 2.4 volts per cell (corresponding to a trickle current of C/1000 to C/500).

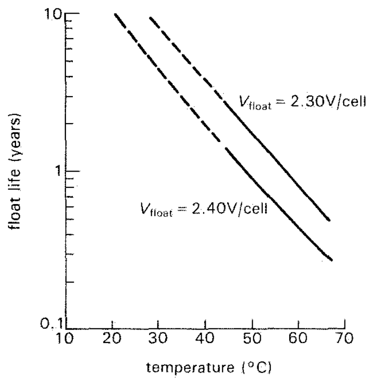


Figure 14.6. Excessive float voltage reduces life of lead-acid batteries.

Figure 14.6 shows the trade-offs. These charging and floating voltages are mildly temperature-dependent and should be adjusted by −4mV/°C for operation at extremes of temperature.

With constant-current charging (which is not often used) you apply a fixed current, typically C/5 to C/20; the battery voltage rises gradually as the battery charges, then increases dramatically as full charge is reached. At this point (indicated by a terminal voltage of 2.5V/cell) you must reduce the current, typically to a fixed C/500 rate, which will maintain full charge indefinitely. Sealed lead-acid batteries will give 8–10 years of service while being charged at a C/500 rate.

A nice lead-acid charging method is the so-called two-step technique (Fig. 14.7): After a preliminary "trickle" charge, you begin with a high-current "bulk-charge" phase, applying a fixed high current I_{max} until the battery reaches the "overcharge voltage," V_{OC}. You then hold the voltage constant at V_{OC}, monitoring the (dropping) current until it reaches the "overcharge transition current," I_{OCT}. You then hold a constant "float voltage," V_F, which is less than V_{OC}, across the battery. For a 12 volt 2.5Ah lead-acid battery, typical values are $I_{max} = 0.5$ amp, $V_{OC} = 14.8$ volts, $I_{OCT} = 0.05$ amp, and $V_F = 14.0$ volts.

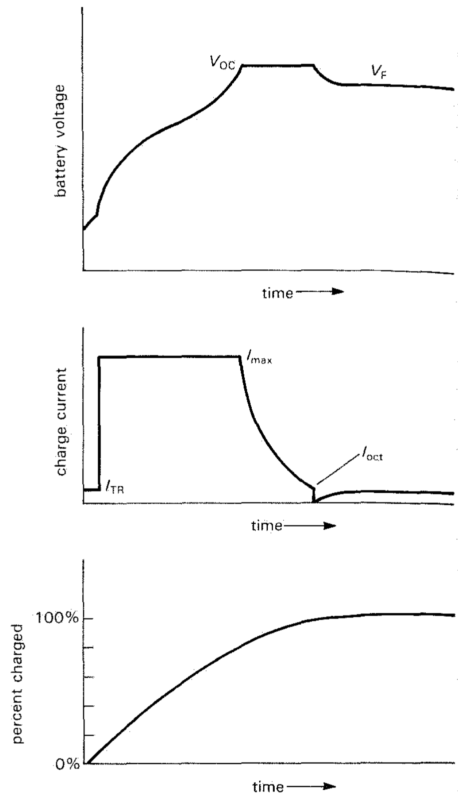


Figure 14.7. Recommended lead-acid battery charging cycle ("two step").

Although this all sounds rather complicated, it results in rapid recharge of the battery without damage. Unitrode makes a nice IC, the UC3906, that has just about everything you need to do the job. It even includes an internal voltage reference that tracks the temperature characteristics of lead-acid cells and requires only an external *pnp* pass transistor and four parameter-setting resistors.

Battery availability and recommendations

As we said at the outset, it's really a good idea to design your instrument to use a popular and readily available battery.

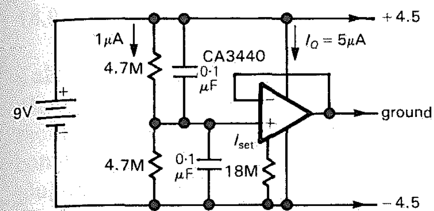


Figure 14.8. Creating a split supply from a single battery.

Tops on the list are the 9 volt "transistor" batteries, known generically as NEDA 1604 (1604, LeClanche; 1604D, heavy duty; 1604A, alkaline; 1604M, mercury; 1604LC, lithium; 1604NC, nicad). You can buy 9 volt alkaline batteries in every corner grocery store (or open-air market) in the world. Op-amps work well on ±9 volt supplies; you can even use ±4.5 volts if you use a resistive divider and a follower to generate a midpoint "ground" (Fig. 14.8; discussed further in Section 14.08). There are nice little plastic instrument cases, complete with 9 volt battery compartment underneath, available from many manufacturers at very attractive prices. We recommend using alkaline, rather than zinc-carbon, because of the improved electrical characteristics discussed earlier. Kodak's new "Ultralife" lithium 9 volt battery looks like a real winner, with 1000mAh capacity, long shelf life (80% retention after 10 years), and flat discharge curve (Fig.14.9); they wisely used 3 cells, not 2, so its terminal voltage

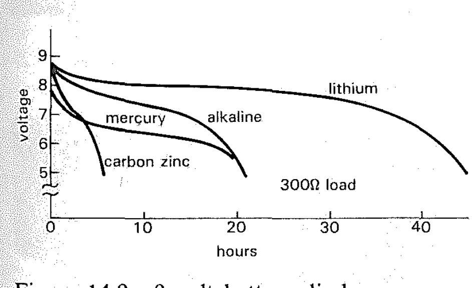


Figure 14.9. 9 volt battery discharge curves; "lithium" is the 3-cell Kodak "Ultralife."

is close to 9 volts, the same as alkaline. Our preliminary measurements on some early samples showed a rather high internal impedance, however.

The familiar alkaline AA, C, and D cells have more energy capacity (and correspondingly lower internal impedance) than the 500mAh of the 1604A (3, 9, and 20 times as much, respectively), and they're just as easy to get. But they are somewhat less convenient because of the problem of holding and connecting reliably to a group of series-connected cells. Everyone has noticed that if you shake a dim flashlight, it usually gets brighter. The problem is compounded by the tendency of some types (alkaline, mercury) to grow white deposits on the terminals (this is officially known as "salting").

Nicads are also available (though not in every drugstore) in the standard battery sizes (AA, C, D, and 9V), for applications where it makes sense to use secondary batteries. But you get only about 25%–50% the energy capacity, and reduced battery voltage (1.2V versus the alkaline's 1.5V per cell).

Lithium batteries are available in the same standard battery sizes, although they provide 3 volts or more per cell. Most manufacturers also provide them with solder tabs for more reliable connections; this makes good sense, considering their extended shelf life. Lithium cells are also available in flat "button" shapes with solder tabs, for use as CMOS memory backup, or to power calendar clock chips. Kodak's 9 volt lithium battery has a nice twist, namely gold-plated snap tabs, for more reliable connections. See our earlier warning on lithium-battery detonation.

Sticking to batteries you can get in any small town, in most photo stores you'll find a selection of mercury, silver, and lithium cells. They're meant to go into cameras (and calculators and watches) and are generally of the "button" variety. For

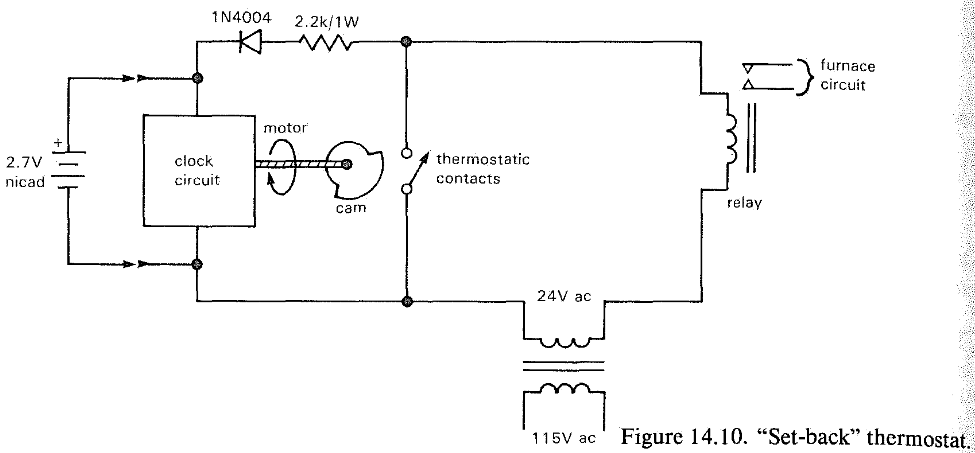


Figure 14.10. "Set-back" thermostat.

example, there's the popular 625 mercury button, hardly larger than a coat button and good for 250mAh. The smaller 76 silver oxide cell (and energetically equivalent type 675 mercury cell) has an interesting twist, namely an offspring lithium 3 volt cell (NEDA 5008L) of the same diameter and twice the height, intended to substitute for a pair of the 1.5 volt cells. At this voltage you can run CMOS logic directly, as well as low-voltage op-amps like the LM10 and the ICL7610 series and TI's versatile "LinCMOS"series of op-amps (the TLC251–254 series) and comparators (TLC372/4, TLC339/393, and TLC3702/4).

If your application requires the rechargeability and high peak currents of sealed lead-acid batteries, or some exotic form of primary battery, you'll generally have to deal with the battery manufacturers or their distributors. Look for names like Gates, Powersonic, and Yuasa for lead-acid. Duracell and Eveready dominate the primary-cell market. All of these companies have helpful and extensive data books on batteries and battery lore.

In the next few sections we will consider alternative power sources – wall-plug-in modules, solar cells, and signal currents – for low-power equipment. It's worth remembering that each of these power sources can be used to charge secondary batteries. For example, the popular "set-back" thermostats that turn down the heat at night use the high-impedance 24 volt ac relay signaling current to charge nicads and keep the clock running during periods of relay-ON (Fig. 14.10).

Table 14.3 summarizes our advice on the relative merits of various primary-cell battery types.

TABLE 14.3. PRIMARY-BATTERY ATTRIBUTES

	9V alkaline	1.5V alkaline	Mercury	Silver	Lithium
Properties					
Inexpensive	•	•	–	–	–
Available	•	•	–	–	–
Wide temp range	–	–	–	–	•
Stable voltage	–	–	•	•	•
Reliable contacts	•	–	–	–	•
Good at high current	•	•	–	–	•
Long shelf life	–	–	–	–	•
Miniature	–	–	•	•	•
Applications					
Linear circuits	•	–	–	–	–
Low voltage CMOS	–	•	•	•	•
4000-series CMOS	•	–	–	–	•
CMOS backup	–	–	–	–	•

14.03 Wall-plug-in units

Calculators, modems, tape recorders, telephone dialers, small measurement instruments – more and more low-power devices come with those familiar square black wall-plug-in power units (Fig. 14.11). Although they're usually labeled to match the instrument they power, you can easily get them in a variety of ratings, both in large quantities and small (distributors like Radio Shack and Digi-Key each stock a few types). The best news is the incredibly low price: A 9 volt 500mA (unregulated dc) wall plug-in costs about $2.50 in quantity.

Wall plug-ins are a good way to power small instruments that need more power than you can get from batteries, or that need to keep their rechargeable batteries charged. They're cheaper than internally mounted discrete or modular power supplies, and by using them you save space and keep heat (and high voltage) outside the instrument. Furthermore, they usually satisfy UL and CSA safety requirements, which is important if you want to market an instrument without the lengthy UL approval process.

Wall plug-ins come in three flavors: plain step-down transformers, filtered but unregulated dc supplies, and complete regulated supplies (both linear and switchers). All come in a variety of voltages and currents, and the regulated supplies even come in useful combinations like +5 volts at 1 amp and ±15 volts at 250mA. They have the usual features of IC regulators, namely current limiting and thermal shutdown, as well as optional overvoltage crowbar. You can get them with three-prong (grounding) wall plugs and with various output connectors; many of the larger units are also available as free-standing desk units, with an ac power cord. One word of caution: There is no standardization of connector type and voltage ratings. In fact, there is not even standardization of polarity! So it is effortless to blow out an instrument by plugging the wrong wall unit into it. Beware!

An extensive line of high-quality wall power units is manufactured by Ault (Minneapolis, MN). For inexpensive imported units, look at the catalogs of Condor (Sunnyvale, CA) or Multi Products

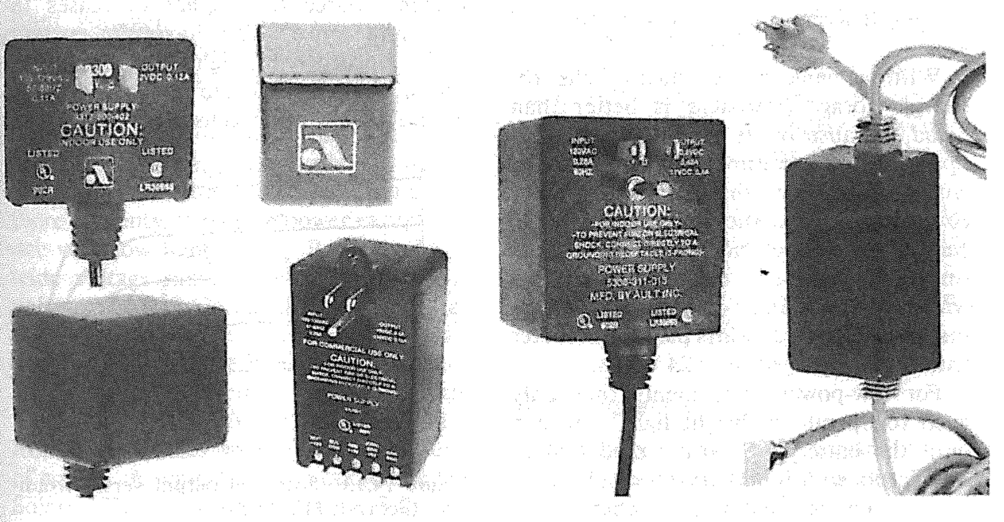


Figure 14.11. Wall-plug-in power supplies. (Courtesy of Ault, Inc.)

International (Cedar Grove, NJ). Check the EEM (see Bibliography) for addresses and additional manufacturers.

□ 14.04 Solar cells

A combination of lead-acid or nicad battery plus silicon solar cells forms a good power source for a moderate-power instrument that is to be deployed at a remote site for extended periods. For example, you might want to tether a buoy that makes ocean measurements and transmits them periodically. If the average power consumption is 1 watt, primary batteries become prohibitively bulky (you'd have to use 500 alkaline D cells to last 1 year). Full sunlight delivers about 1kW per square meter after traversing the atmosphere; after accounting for the inefficiencies of solar cells (they're about 10% efficient when operating into the correct load) and the daylight and weather cycle at middle northern latitudes (where you average 100W/m^2 in winter, 250W/m^2 in summer), you can average about 25 watts (in July) or 10 watts (in January) per square meter from good quality solar cells, which cost about $800 in 1986. In peak sunlight, such a solar module delivers 100 watts to a matched load.

With a bank of secondary cells for energy storage (lead-acid is better than nicad, because of its long life and wide operating temperature range), you can withdraw nearly this average power *continuously*; lead-acid cells are typically 70%–80% efficient, so, all factors (including weather) considered, you can withdraw something like 8 watts per square meter (winter) to 20 watts per square meter (summer), averaged over 24 hours.

For low-power instruments that only need to operate in bright light, you can omit the battery. Solar-powered CMOS calculators with liquid crystal displays are a boon to battery haters everywhere.

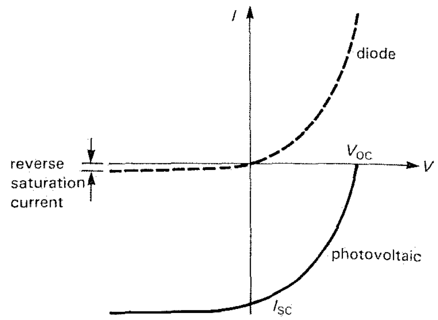


Figure 14.12. Solar-cell output voltage versus load current is simply a displaced diode V–I curve.

□ *VI characteristics*

Silicon solar cells have a simple and very useful volt–ampere characteristic. It turns out that the open-circuit voltage is almost independent of light level and averages about 0.5 volt per cell; the V–I curve is simply a displaced diode curve (Fig. 14.12). Typical solar panels consist of 36 series-connected cells, for an open-circuit voltage of about 18 volts. The terminal voltage stays nearly constant as load current is increased up to a maximum current, at which point the solar module becomes roughly a constant current source for further decreases in load impedance. The maximum current scales linearly with light level, giving a set of characteristic curves as shown in Figure 14.13. Solar cells work best when

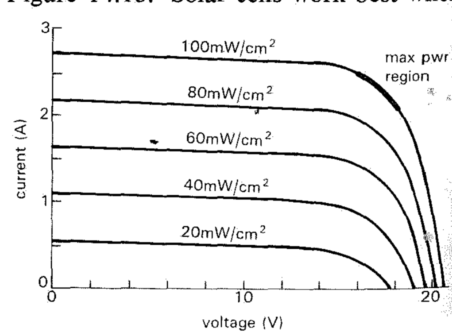


Figure 14.13. Solar-cell output versus irradiance (Solavolt MSVM4011).

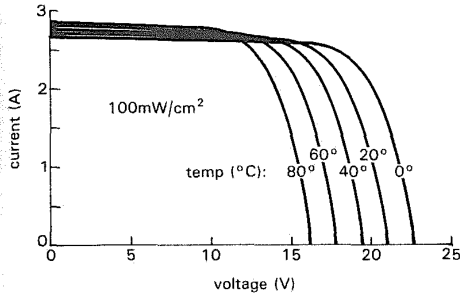


Figure 14.14. Solar-cell output versus temperature (Solavolt MSVM4011).

cold, since the open-circuit voltage drops with increasing temperature (Fig. 14.14).

For a given light level, maximum power is delivered when the operating point has maximum product *VI*; in other words, the point on the *V–I* curve that touches a family of hyperbolas (constant product *VI*) plotted on the same axes. Roughly speaking, that's at the knee of the *V–I* curve. Since the load impedance that corresponds to the knee changes rapidly with light level, you can't expect to maintain an optimum load (which would be a load impedance increasing inversely with light level, or, put another way, a load that draws a current proportional to light level, at roughly constant voltage). However, for low-power applications it isn't essential that the load extract maximum power – all that matters is that the load be powered under normal lighting conditions. That's the case for those solar-powered calculators, whose CMOS circuits draw so little current that there is plenty of reserve power except under very low light levels. Because of the wide voltage range of 74C/4000B "high-voltage" CMOS (3V to 18V), and the fact that solar cells have an open-circuit voltage that is relatively independent of light level, you don't need to use any voltage regulators; just power the CMOS directly from the module, with bypass capacitors, of course. A typical small module like the Solarex SX-2 provides 290mA at 8.5 volts in sunlight and has an open-circuit voltage of 11 volts; you could use it, unregulated, for high-voltage CMOS, or, with a regulator, for any +5 volt logic family.

For any application that uses secondary cells for energy storage, it's worth noting the rather good match of solar-cell *VI* characteristics to the charging requirements of lead-acid cells. A solar module provides roughly constant charging current into a discharged battery, changing over to a constant-voltage "float" as the battery voltage rises at the end of charging. The temperature coefficient of open-circuit voltage (−0.5%/°C) is a fair match to the recommended float-voltage tempco for lead-acid batteries (−0.18%/°C). So some suppliers make solar modules that are intended to charge lead-acid batteries directly, for example the Arco M65 (2.9A@14.5V). The more usual way to match solar modules to the charge/float characteristics of lead-acid batteries is with a series or shunt regulator circuit designed for the job. Many solar modules are designed to work this way, with 20 volts open-circuit voltage and matching regulator module for charging 12 volt batteries. The regulators switch over from charging to temperature-compensated floating, with automatic load disconnection if the battery voltage drops too low. These systems are available for multiples of 12 volt systems (24V, 36V, 48V, etc.), and you can get accessories such as 60Hz inverters (to make ac), or dc-operated refrigerators, attic ventilators, etc.

Some of the bigger names in solar modules and systems are Arco Solar (Chatsworth, CA), Mobil-Solar (Waltham, MA), Solarex (Rockville, MD), and Solavolt (Phoenix, AZ).

14.05 Signal currents

Don't forget about the possibility of using signaling currents to power a micropower instrument. Four of the more

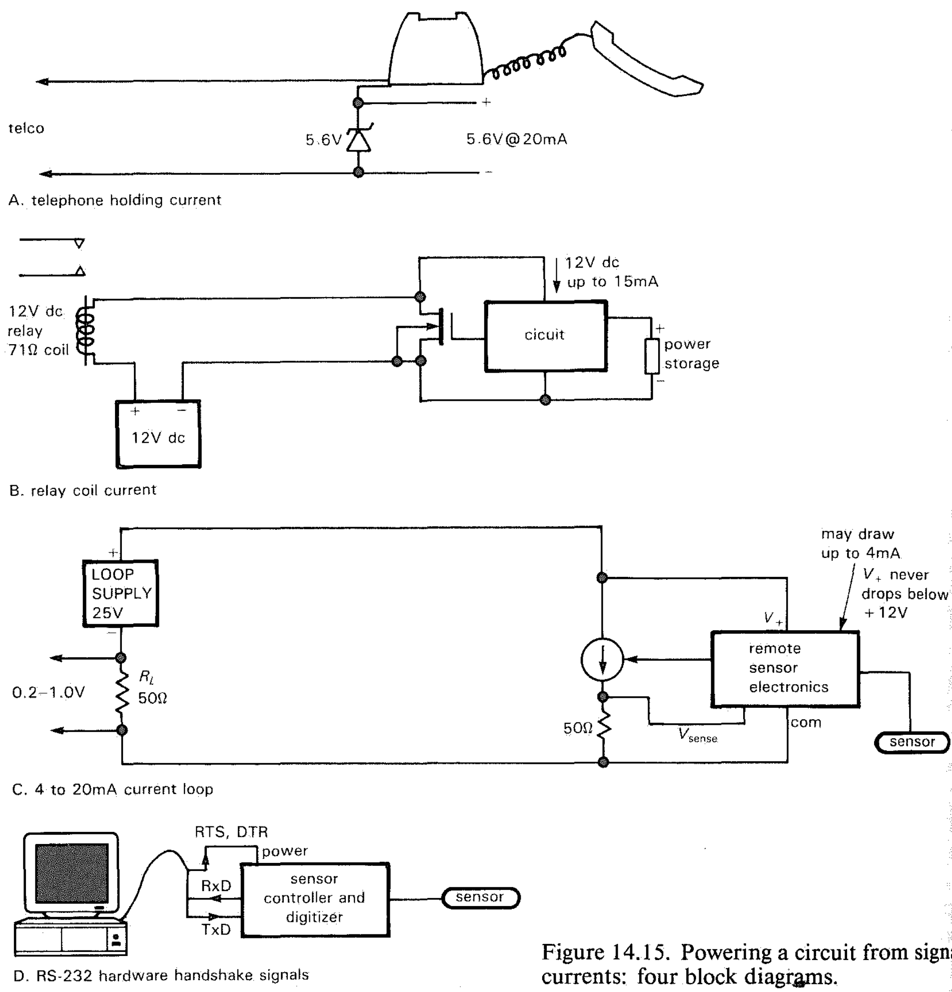


Figure 14.15. Powering a circuit from signal currents: four block diagrams.

common opportunities (Fig. 14.15) are (a) the dc holding current flowing through a telephone circuit that is "off hook," (b) the ac or dc voltage available from a relay circuit when it is not energized, (c) the 4–20mA dc current used for industrial-sensor current-loop signaling, and (d) the serial-port RS-232C bipolarity "handshake" signals (RTS, DSR, etc.). In the first two schemes, your source of power is available only part of the time – power disappears when the phone is hung up ("on hook") or when the relay circuit is energized by your instrument. If you need power continuously, you'll have to use rechargeable batteries, charged during periods of available power; for very low current loads, another possibility is a high-capacitance (up to 5 farads) "double-layer" capacitor, the same type used for CMOS memory retention.

Each of these power sources has rather strict limits on voltage compliance or maximum current. Here are their characteristics, and some hints on the parasitic use of these power sources.

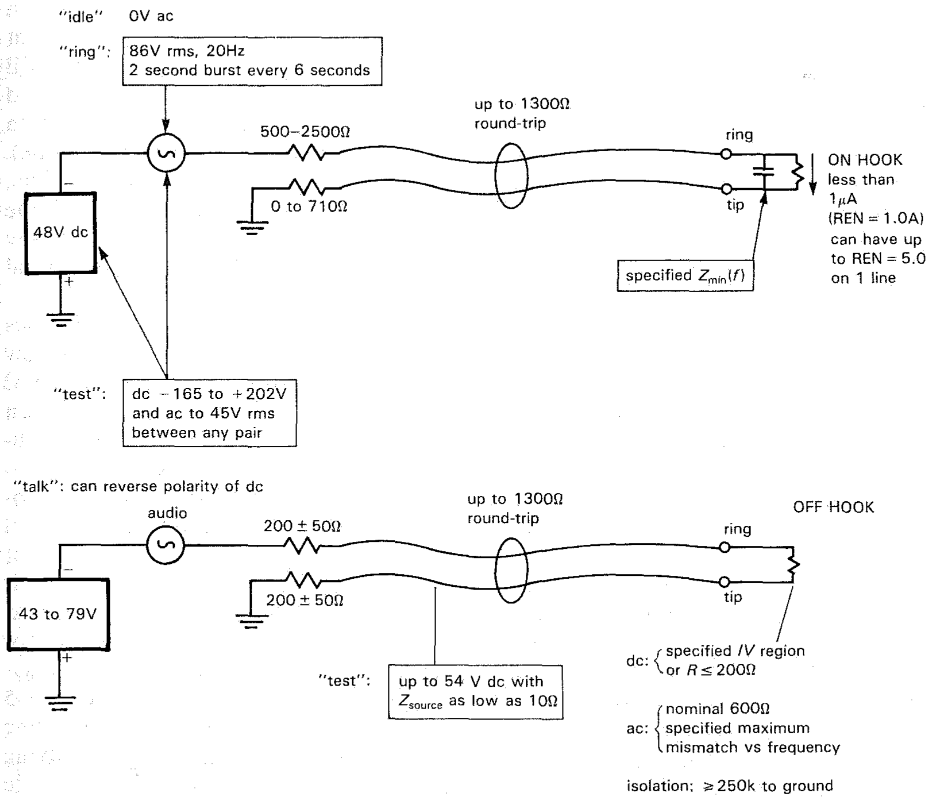


Figure 14.16. Telephone operating and test states.

☐ *Telephone-line power*

There are several different states that the phone line can be in, depending upon what your phone is doing, and what the phone company is doing to you. The central office (or nearby equivalent) applies various dc (and ac) voltages to the two-wire phone loop (labeled "tip" and "ring") during these various stages of call progress (Fig. 14.16). In the *idle* state, the telephone company central office applies −48 (±6) volts dc in series with 500 to 2500 ohms to the "ring" line, and terminates the "tip" line to ground with 0 to 710 ohms. In addition, there is typically up to 1300 ohms of external line resistance between the central office and you (the "subscriber"). When you go off-hook, the central office goes into *dialing* mode, applying a dial tone and a dc level of −43 to −79 volts in series with 200 ohms (±50Ω) on "ring," and terminating "tip" with the same impedance to ground. The same dc voltage and source impedances are present in the *talking* state (after the connection is made), although the telephone company may, at its discretion, reverse the polarity of the dc voltage applied to "ring." Of course, in the talking state you also have audio signals superposed on the dc, which is the whole purpose of the telephone!

There are two other states. During *ring*, the phone company applies 86 (±2) volts rms, at 20Hz, on top of the usual −48 volt dc bias. As with the dc, the ac ringing signal is applied to the "ring" lead. The official ringing specification is 2 second bursts at 6 second intervals. During *test* mode, the phone company applies various ac and dc test signals to make sure that the network is working properly. They can apply dc voltages in the range −165 to +202 volts, and ac voltages up to 45 volts rms, between any pair of conductors (ring, tip, ground) in the on-hook state, and up to 54 volts dc with source impedance as low as 10 ohms in the off-hook state. The phone company also specifies a range of lightning-induced "high-voltage surges" that you may find on your phone lines. They're typically a few thousand volts, capable of sourcing a few hundred amps; equipment connected to the phone line is supposed to have transient suppressors so that it will survive such pulses. In addition, the phone company specifies "very high voltage surges" that may occur from a nearby lightning strike. These may reach 10kV and 1000 amps, and the idea is that even if your equipment gets fried, nobody should get hurt. So the specification says that the equipment shouldn't shoot out pieces of stuff, catch fire, or electrocute anyone.

Permissible loads are specified in terms of "ringer equivalence number" (REN). Typical phones have an REN of 1.0A, which corresponds to (a) an on-hook dc resistance of 50MΩ, and on-hook ac impedance that stays above a specified curve of impedance versus frequency (satisfied by keeping |Z| > 125k from 4Hz to 3.2kHz, though it can be much lower over certain frequencies and voltages), (b) an off-hook dc characteristic that stays within the acceptable region of Figure 14.17 (or that measures 200Ω or less), and (c) an off-hook impedance that approximates 600 ohms from 200Hz to 3.2kHz (this is actually specified in terms of reflection when driven by a 600Ω signal source: at least 3.5dB from 200Hz to 3.2kHz and 7dB from 500Hz to 2.5kHz). Loads connected to the phone lines must be dc-isolated from ground (50MΩ on-hook, 250k off-hook). A total REN up to 5.0A is permissible, i.e., a load impedance as low as 1/5 the above values. The telephone company requires you to notify them of your total REN loading.

From the foregoing data it is obvious that the subscriber is not supposed to draw current in the on-hook state, and the 50MΩ minimum dc bridging resistance (for an REN of 1.0A) is really a leakage specification – 50MΩ corresponds to 1μA. Nevertheless, with careful design and component selection you can maintain CMOS circuitry (digital or analog) in a quiescent state with a few microamps, and have it "wake up" when the line goes off-hook. Use a small tantalum electrolytic (or "double-layer" memory-retention capacitor, available in tiny packages up to 5 farads!) to keep things going during dialing or other transients. In the off-hook talking state, you are guaranteed 6 volts dc (7.8V after a few seconds) at a minimum of 26mA (see Fig. 14.17), which is enough

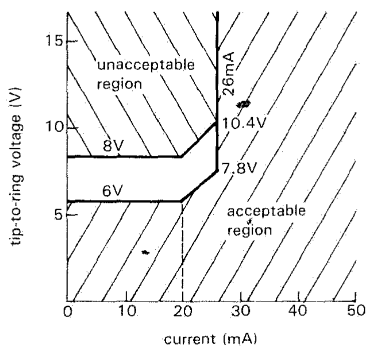


Figure 14.17. Allowable load conditions for "off-hook" telephone. The unshaded region is allowable beginning 1 second after going off-hook. (Adapted from Bell System Tech. Ref., Pub. 47001.)

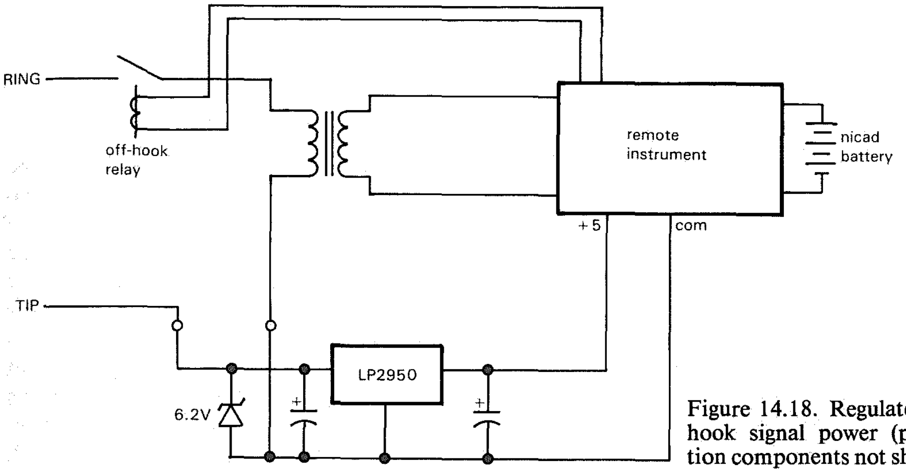


Figure 14.18. Regulated off-hook signal power (protection components not shown).

to operate lots of micropower circuitry; it's really comparable to the kind of power you get from a 9 volt battery. With a micropower low-dropout regulator like the LP2950 (80μA quiescent, 0.4V dropout at 100mA load) you can supply regulated 5 volts to digital circuitry, as in Figure 14.18. If you can be sure of having regular periods of off-hook operation, you can use the extra current then available to keep rechargeable batteries charged. For example, if you have an hour of off-hook operation per day, you can draw nearly a milliamp continuously.

Warning: Before designing any device for direct connection to the telephone system, be sure to get the relevant specifications. You must be in compliance with FCC regulations, which include test and approval procedures. Do not assume that the specifications quoted in this chapter are correct.

□ *Relay circuits*

Those popular "set-back" thermostats that turn down the heat at night, and turn it back up a half hour before you wake up, use the small ac power that can be extracted from the relay circuit without making the relay close. Typical mechanical control relays draw 100mA or more at the coil rating of 24 volts ac and can be relied on to stay open at 10% of the normal coil current. So you can have 10mA or so, at nearly 24 volts ac, to power your instrument. Be sure to include rechargeable cells (or perhaps a memory-retention capacitor, if that's all that's needed), because the power source goes away when you close the contacts to energize the relay. Figure 14.10 shows the idea.

Industrial current loops

In industrial environments there is a standard for current-loop signaling, in which a remote sensor (a thermocouple, say; see Section 15.01) sends back its measurements by converting them to an analog current, which then flows through a loop. The dc bias for the loop is usually provided at the receiving end (Fig. 14.15C). There are two standards, namely full-scale ranges of 4 to 20mA and 10 to 50mA. The 4–20mA standard is more popular and usually uses a dc bias of 24 volts dc (though sometimes higher). For simplicity it is often desirable to use the signaling current to power the electronics at the remote end.

For this purpose you can use part of the loop bias for power. Commercially available current-loop modules generally specify that the data recipient has to provide a maximum load resistance R and minimum dc bias V_s such that $(V_s - 12\text{V})/R_s$ is equal to the full-scale current. In other words, the remote module can drop up to 12 volts while still applying full-scale loop current. Of course, the module has to keep running when sending a loop current corresponding to minimum output. So the bottom line is that you always have available at least 12 volts at 4mA to power your equipment; you may have more, but don't count on it. That's plenty for even rather complex circuits, if you practice careful micropower design.

RS-232 serial-port signals

The RS-232C/D standard specifies bipolarity data and control signals of substantial drive capability (see Section 10.19); you can use one of the control signals (or even a data signal!) to run a low-power circuit. Officially an output must be able to assert ±5 volt to ±15 volt levels into a 3k to 7k load resistance. The RS-232 drivers in common use typically have an output impedance of a few hundred ohms, and current limit at 5 to 15mA. To become a parasite on this power source, you have to arrange your software to keep a known control line in a known (and stable) state. You can even use a *pair* of control lines, if available, to get split supply voltages (±5V, min). Remember that control signals (RTS, DTR, etc.) are asserted HIGH, which is the opposite of the data signals.

Since there's usually plenty of commercial ac power available around a computer, you aren't really doing anything miraculous by sucking the life forces from the 25-pin *D* connector. However, for a simple serial-port hang-on circuit it is an elegant source of power. You can get commercial network interfaces and modems that work this way.

POWER SWITCHING AND MICROPOWER REGULATORS

14.06 Power switching

You can tame your usual microprocessors, regulators, and other power-hungry components into a micropower application, if the design permits the circuit to be turned off (or put into a low-current standby state) most of the time, and only occasionally run at full current. For example, an oceanographic data logger might make a 10 second salvo of observations (temperature, pressure, salinity, ocean currents) once each hour for a 6 month period. Only the real-time clock need run continuously, with the analog signal conditioning circuitry, microprocessors, and data-recording media shut off except during actual data logging.

Even if you take pains to use micropower design techniques, you may still be forced to use some high-current devices, for example, if you need to use high-speed transducers or high-current actuators. You may need to use some specialized LSI digital circuits, op-amps, filters, or other circuits that are simply not available in low-power versions. In all these cases it is necessary to switch off power to the high-current portions of the circuit except when they must operate.

Such "power switching" can be the simplest form of micropower design, since ordinary design techniques with ordinary components can be used throughout. You've got to make sure the circuit "wakes up" gracefully (a linear circuit should be designed to avoid embarrassing momentary states, for example driving its outputs into saturation; a fully shut down microprocessor circuit would usually do a complete "cold boot"). Likewise, the circuit should be designed to shut down in an orderly manner.

There are several ways to do the power switching (Fig. 14.19):

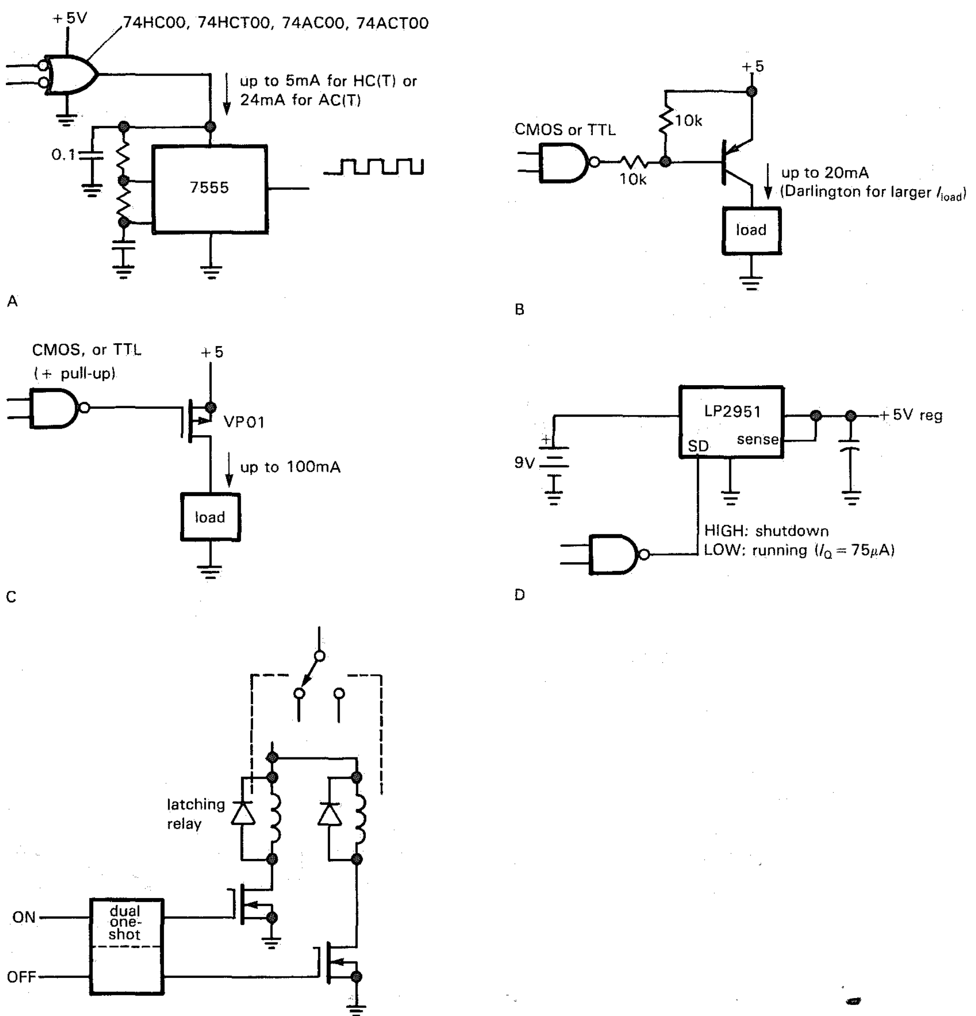


Figure 14.19. Power-switching methods.

1. If the switched components run at less than 5mA or so, you can power them directly from a CMOS logic output. The HC/HCT families can supply 5mA with only 0.5 volt drop below the positive rail; for higher current, several outputs can be used in parallel. The AC/ACT CMOS families are good for 24mA.

2. Use a power transistor, operating as a saturated switch (not a follower) to minimize forward drop (thus *pnp* for a positive supply). The necessary base drive, chosen conservatively large to guarantee saturation, is a disadvantage, though it will probably be smaller than the current used by the switched circuit.

3. Use a power MOSFET. As with bipolar transistors, use as a switch, not a follower (thus *p*-channel for a positive supply). MOSFETs are easy to drive and have no gate current in either state.

4. Many of the low-power regulators include a "shutdown" input, with very low quiescent current in the standby mode (see Section 14.07). You can do power switching by commanding such a regulator into the active state.

5. Use a mechanical relay, perhaps a *latching* relay. There's a good variety now available in DIPs and tiny metal cans, and they offer zero voltage drop, high overload capability, and the ability to switch bipolarity (or even ac) voltages. In addition, latching relays require no holding current. Be sure to use a diode to protect the relay driver from inductive spikes (Fig. 1.95).

Current limiting

It is essential to limit the inrush current in a power-switched circuit, for two reasons: The high peak currents that would result from switching a battery (bypassed with a capacitor) into a load (similarly bypassed) could destroy the switch; this is true even for a small mechanical relay, which is most likely to fail by having its contacts fuse shut. Furthermore, the momentary collapse of the battery voltage during a high current-switching transient can cause volatile memory and other circuitry being held in a standby state to lose information (Fig. 14.20).

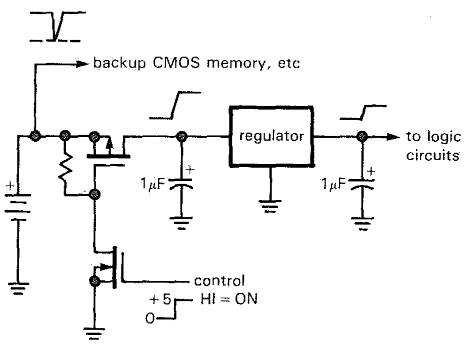


Figure 14.20. Inrush current can cause transient loss of battery voltage.

Several approaches are shown in Figure 14.21. As long as the switch can handle the transient, you can decouple the negative-going dip from the maintaining

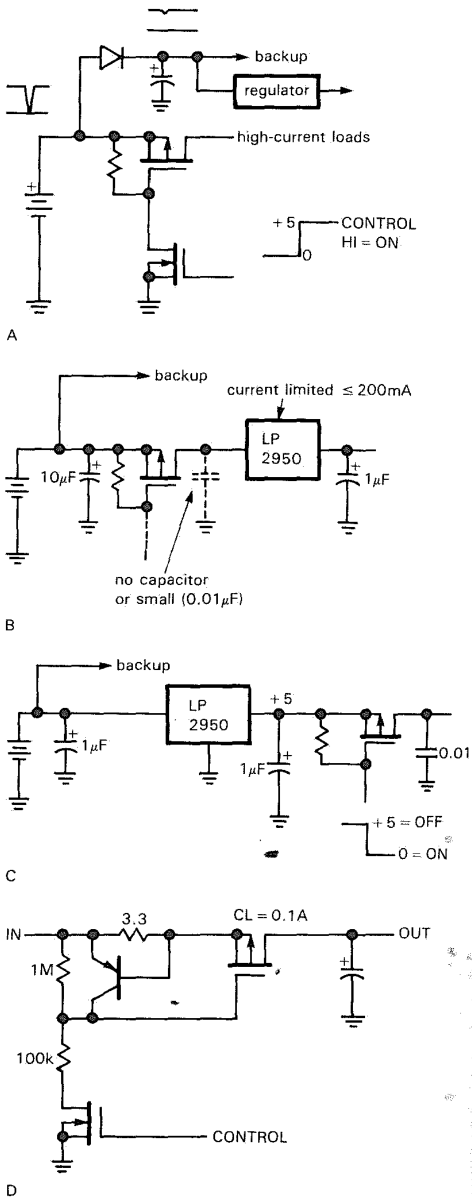


Figure 14.21. Four cures for inrush-current transients.

regulators with a diode, as in option 1. Alternatively, do the switching upstream of a current-limited regulator (keep its input bypass capacitor small), as in option 2, or put the switch after the regulator (option 3). The latter method isn't as good, because of degraded supply stiffness due to the switch's R_{ON}. Another method is to use upstream current-limited switching (option 4), in this case with 150mA current limit, to prevent collapse of V_{batt}.

14.07 Micropower regulators

Until recently it had been difficult to find voltage regulator ICs with microamp quiescent currents capable of substantial output currents. The choice was (a) the Intersil 7663/4 or (b) build your own! Fortunately the situation is improving. Here is the current selection:

ICL7663/4; MAX663/4/6 (Intersil; second-sourced by Maxim and others). These are multiterminal positive and negative regulators, with 1.5–16 volts operating range and maximum quiescent current of 10μA. The bad news is that they are slow (due to a "starved" servo amplifier; use lots of bypass capacitance) and only good for a few milliamps of load current (they're stiffer at higher input voltage, being CMOS devices); for example, with +9 volts input the output impedance is typically 70 ohms.

LP2950/1 (National). These are positive regulators, available as a 3-terminal +5 volt regulator (2950) and an 8-terminal adjustable regulator (2951). Quiescent current is 80μA at zero load current, rising to 8mA at 100mA load current. These regulators use *pnp* pass transistors for low dropout voltage (80mV max at 100μA, 450mV max at 100mA) and are designed so that the quiescent current does not soar when the input voltage dips below dropout (a common disease of bipolar transconductance regulators). This last feature is particularly useful for battery-powered instruments that can continue to function with a low battery. The 2951 includes a shutdown input and dropout-detector output.

LT1020 (Linear Technology). This is a multiterminal positive regulator with 40μA quiescent current, 2.5–35 volt output range, and 125mA maximum current. The *pnp* pass transistor gives low dropout voltage (20mV typ at 100μA, 500mV typ at 125mA). There is a shutdown input and dropout-detector output.

TL580C (Texas Instruments). This is a dual positive micropower *switching* regulator with 2.5–24 volt output range and 140μA quiescent current. Like all switching regulators, you get high efficiency (up to 80%) over a range of battery voltages and the flexibility to have output voltages greater than the unregulated input voltage.

MAX630 series (Maxim). These are micropower switching regulators, in a nice variety of options. The MAX630 is an adjustable (2V to 18V) positive step-up regulator (i.e., $V_{out} > V_{in}$), while the MAX634 is an inverting switcher (i.e., positive input, negative output). The MAX631-3 are fixed-voltage (5, 12, 15V) positive step-up switchers, with MAX635-7 being the inverting equivalents. The MAX638 is an adjustable positive step-down ($V_{out} < V_{in}$) switcher. All are capable of output currents of a few hundred milliamps, with quiescent currents around 100μA and efficiencies around 80%.

MAX644 series (Maxim). These micropower switching regulators are designed to generate a +5 volt output when powered from a one- or two-cell battery. This clever design uses a two-part switching upconverter: One section runs continually, providing a low-current (<0.5mA) +5 volt output; it also supplies +12 volts dc needed to switch the MOSFET used for the high-current (up to 50mA) +5 volt output. In standby mode (low-current output only) the quiescent current is 80μA. The MAX644 is designed for 1.5 volt nominal input and operates down to input voltages of 0.9 volt.

TABLE 14.4. LOW-POWER REGULATORS

Type	Mfg[a]	Pins	I_Q typ (μA)	Pol	V_{out} (V)	V_{in} (V)	Dropout V (V)	@ I (mA)	Shutdown	Tempco typ (ppm/°C)	Comments
ICL7663	IL+	8	4	+	1.5–16	1.5–16	0.8	20[b]	•	200	MAX663, 7663S improved
MAX664	MA+	8	6	–	1.3–16	2–16.5	0.2	20[b]	•	100	also ICL7664
MAX666	MA	8	6	+	1.3–16	2–16	0.9	40	•	100	MAX663 + dropout detec
LT1020	LT	14	40[c]	+	0–35	5–36	0.5	125	–	1%	dropout detector
LP2950	NS	3	75[c]	+	5	5–30	0.45	100	–	20	dropout I_Q = 110μA
LP2951	NS	8	75[c]	+	1.2–29	2–30	0.45	100	•	20	
MAX630	MA	8	70	+	V_{in} to 18	2–16.5	-	375	•	-	switching steput
MAX635-7	MA	8	80	±	–5,–12,–15	+2 to 16.5	-	375	–	-	switching inverter
MAX634	MA	8	100	±	to -20	2–16.5	-	375	•	-	switching inverter
MAX631-3	MA	8	135	+	•5,12,15	1.5–V_{out}	-	325	–	-	switching stepup
MAX638	MA	8	135	+	< V_{in}	2–16.5	-	375	•	-	switching stepdown
TL580C	TI	8	140	+	2.5–24	2.4–30	-	100	–	-	
LM10	NS+	8	300	+	1–40	1.1–40	0.4	20	–	30	
LM2931	NS+	5	400	+	1.2–25	to 26V	0.2	150	•	-	TO-220
LM2931-5	NS+	3	400	+	5	5.2–26	0.2	150	–	-	TO-92; 2931CT is adjustable
TL750L05	TI	3	1000	+	5	5.2–26	0.6[d]	150	–	-	TO-92; TL751 has shutdown
LM317L	NS+	3	2500[e]	+	1.2–37	to 40V	2	100	–	0.7%	TO-92
LM337L	NS+	3	2200[e]	–	1.2–37	to –40V	2	100	–	0.7%	TO-92
78Lxx	FS+	3	3000	+	5,12,15	to 30V	2	100	–	-	TO-92
79Lxx	FS+	3	2000	–	–5,–12,–15	to –35V	2	100	–	-	TO-92; LM320L also
LM330	NS+	3	3000[c]	+	5	5.3–26	0.6	150	–	-	TO-220
LM2930	NS+	3	4000	+	5	5.3–26	0.6	150	–	-	TO-220; LM2935 also

(a) see footnote to Table 4.1. (b) for V_{in} = 9V. (c) no load. (d) over full temp range. (e) I_L (min).

In addition, there are several "low-power" regulators (78L05, LM330, LM317L, LM2930/1), characterized by quiescent currents of a few milliamps. These are useful for instruments with some external power source, for example solar cells or telephone holding current. Also, don't overlook the possibility of using a micropower voltage *reference*, rather than a regulator, if its voltage happens to be what you want. For example, the REF-43 from PMI is a 3-terminal 2.5 volt reference with 250μA maximum quiescent and excellent characteristics.

Look at Table 14.4 (which also includes the regulators above) for characteristics of most available micropower regulators.

Negative supplies

With the exception of the ICL7664/MAX664, all of the linear micropower regulators are positive polarity only (though the LT1020 can be used to make a dual supply). If you need negative supply voltages, there are (besides the feeble 7664) several possibilities, namely (a) a "flying-capacitor" voltage converter chip like the 7662, (b) a discrete realization of a flying-capacitor voltage converter, using complementary power MOS transistors, (c) a voltage converter using a CMOS oscillator chip like the 7555 (that's a CMOS 555) or the output of any CMOS logic gate that is driven by a square wave, (d) a switching supply, with inductive energy storage, or (e) the use of a single positive supply, with an op-amp-generated ground reference part way between ground and the positive rail. Let's take them in turn:

1. The 7662 (and its predecessor 7660) is a CMOS IC introduced by Intersil and widely second-sourced (see Section 6.22). It has an oscillator and CMOS switches (Fig. 6.58), and with a few external capacitors you can use it to generate either $-V_{\text{supply}}$ or $+2V_{\text{supply}}$, when powered by a positive voltage V_{supply}. Like most CMOS devices, it has a limited supply voltage range; for the 7662, V_{supply} can range only from 4.5 to 20 volts (1.5V to 10V for the 7660). The output is not regulated, and it drops significantly for load currents greater than a few milliamps. In spite of these drawbacks, it can be very useful in special circumstances, for example to power an RS-232C driver chip on a board that otherwise runs on a single +5 volt supply. The MAX680 and LT1026 are flying-capacitor dual supplies that generate ± 10 volts (up to 10mA) from +5 volts (Fig. 6.60). There are also combination voltage converter and RS-232 driver/receivers available as single ICs, the LT1080 and MAX230–239 series. If your application requires RS-232 ports, you may be able to use the dual supply voltages generated by one of these RS-232 driver ICs to power your analog electronics.

2. To generate a larger negative voltage, you can use discrete MOS transistors in a flying-capacitor circuit (Fig. 14.22). The particular example shown idles at a few microamps and generates up to 30mA.

3. Figure 14.23 shows a simpler method, again somewhat limited in voltage range, using the CMOS 7555 timer chip. You can power the 7555 from a positive supply in the range 2 to 18 volts, thus generating up to -15 volts or so. With a voltage multiplier (see Section 1.28) you can, of course, generate higher voltages, with correspondingly poorer regulation. If you have some CMOS logic in your circuit, you can use the output of a CMOS gate instead of the 7555. However, if you're using a high-performance CMOS family such as HC/HCT or AC/ACT, then you are limited to 5 volt logic swings, whereas the older 4000 or 74C series permit 15 volt swings, albeit at lower current.

4. As we explained in Chapter 5, with inductive energy storage you can make switching supplies for which the output voltage is higher than the input, or much lower, or even negative, all with efficiencies of 75% or so, independent of input voltage. This is obviously useful in

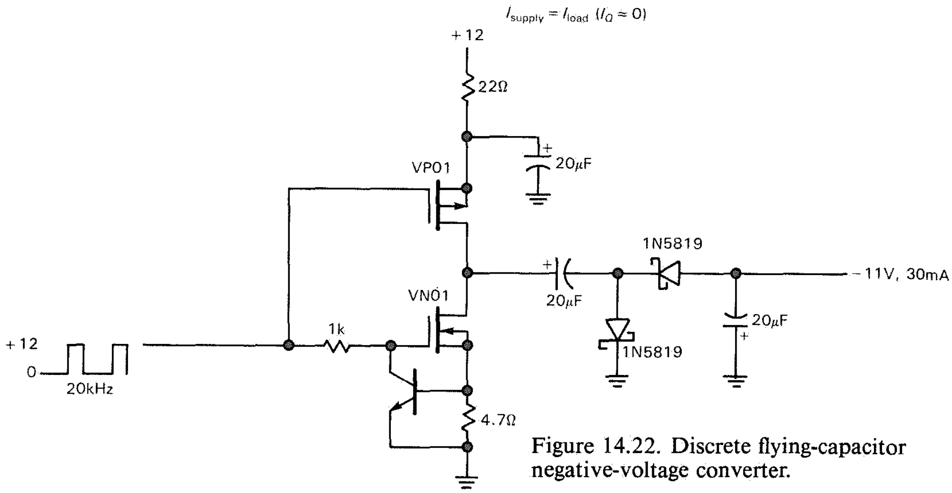


Figure 14.22. Discrete flying-capacitor negative-voltage converter.

micropower design, where the unregulated dc may be supplied by batteries whose voltage drops off with use. Switching supplies for micropower applications can be designed to maintain high efficiency even when unloaded (unlike ordinary high-current switchers), by using circuitry that shuts the oscillator off until the output drops, at which point it supplies a single charging pulse, then goes to sleep again. Figure 14.24 shows a +5 volt supply constructed with the low-power MAX631.

5. You may not need a separate negative supply, even if you are using op-amps with bipolarity output swings, etc. For example, you might generate a +4.5 volt ground reference (using a resistive divider and micropower op-amp follower) for an op-amp circuit running from a single 9 volt battery. Let's look at this method in some more detail.

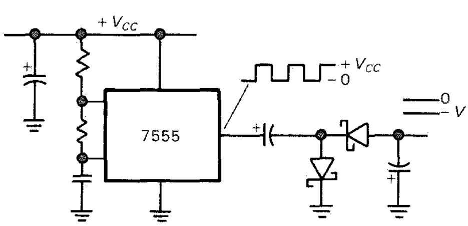


Figure 14.23. Negative-voltage generator from a positive square wave.

14.08 Ground reference

Most of our op-amp circuits in Chapter 3 used symmetrical power supplies, usually ±15 volts, because of the flexibility of dealing with signals near ground. As we mentioned in Section 4.22, however, it is possible to use only a single supply, by generating a reference voltage that substitutes for the ground potential of the usual bipolarity op-amp power supplies. When your power supply is a battery, there's an added incentive to keep things simple, preferably by using a single 9 volt battery.

The easiest way to generate an analog "common" is to split the battery voltage with a resistive voltage divider, then use a micropower op-amp follower to generate the low-impedance common. To the outside world that common voltage is "ground," with both ends of the battery floating; see Figure 14.8.

In the example circuit, we've chosen a 3440 CMOS programmable op-amp, biased to run at 5μA quiescent current. The divider's unusually large resistors keep its contribution to the current drain small, with capacitive bypassing to keep the impedance low at the midpoint, which otherwise would be susceptible to hum and pickup of other signal

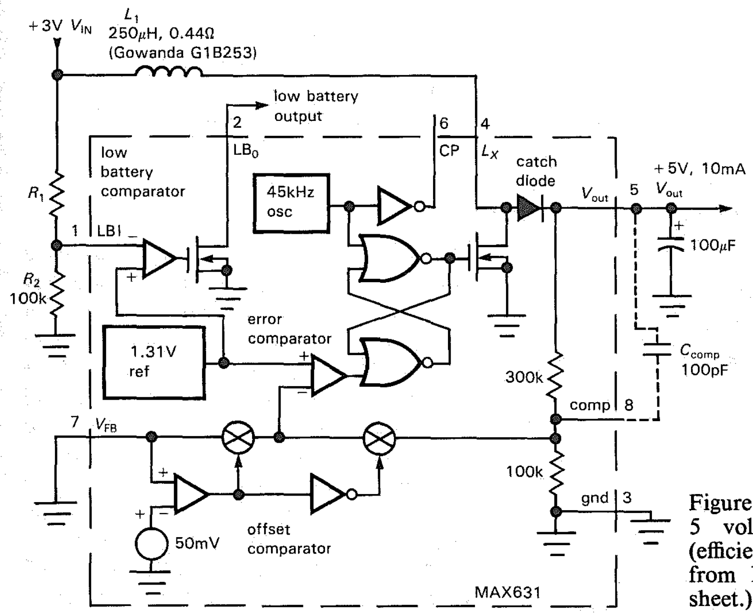


Figure 14.24. Low-power 5 volt switching regulator (efficiency = 74%). (Adapted from Maxim MAX631 data sheet.)

frequency interference. The 3440 is a good choice in this application, because it can sink or source substantial currents (up to a few milliamps) even when biased at 1μA; this property is not shared by all programmable op-amps, many of which have poor sourcing capability when operated at micropower levels. For example, the LM346 operating at 5μA can source only about 0.1mA, although it can sink 20mA (look ahead to Fig. 14.32).

Note that the reference voltage doesn't have to be half of the battery voltage; it may be best to split the supply unsymmetrically, to allow maximum signal swing. (We'll have an example in Section 14.12.) In some instances it may be preferable to put it at a fixed voltage from one rail, perhaps determined by a precision micropower voltage reference. That rail is then a regulated supply with respect to the common reference.

Output impedance

There are some situations in which you don't even need to use an op-amp to generate the ground reference. For example, if the reference voltage goes only to op-amp inputs (which would have been connected to ground in the usual split-supply configuration), then a high-impedance resistive divider, bypassed to maintain a low impedance at signal frequencies, will usually suffice.

In the more usual case, however, the ground reference must present a low impedance, both at dc and at signal frequencies. For example, some ICs may use it as their negative rail; it might be the common point for low-pass filters, biasing networks, loads, etc. Look at almost any normal split-supply circuit and you'll find dc and signal currents flowing into and out of ground. As in the example above, be sure the op-amp you choose to generate the ground reference has the source and sink capability the circuit needs. Micropower op-amps tend to have rather high open-loop output impedances (Fig. 7.16), so at high frequencies (where there isn't much loop gain) the impedance of the ground reference may rise to several thousand ohms.

The obvious cure is to bypass the ground reference (Fig. 14.25A), but this is likely to cause ringing or even oscillation because of the lagging phase shift of the bypass capacitor in combination with the op-amp's relatively large output impedance, all of which is inside the feedback loop. Figure 14.25B shows one cure – a decoupling resistor of a few hundred ohms, which, however, raises the impedance at dc since it is outside the feedback loop. With two more parts, Figure 14.25C does the trick, maintaining dc feedback (via R_2) and stability at the same time.

Whatever method you choose, make

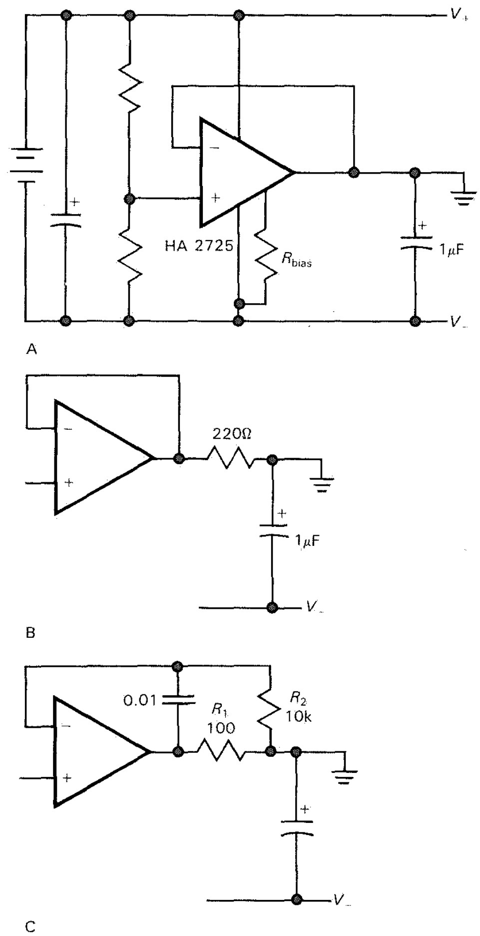


Figure 14.25. Bypassed split-supply generators.

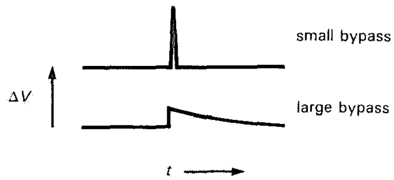


Figure 14.26

sure you test it under various steady-state and transient load conditions. A good way to test for transient behavior is to watch the voltage waveforms while applying a low-frequency square-wave load. There are some op-amps (e.g., the HA2725 and MC3476) that can drive directly into a large capacitive load without stability problems; apparently in these cases the external capacitor reflects back onto the internal compensation capacitor, pushing down the dominant pole in a "brute-force" compensation. In many cases, however, you're more likely to wind up with a pair of nearby lagging phase shifts, which spells trouble.

Note that the choice of bypass capacitor value may involve some subtlety: For a load-induced spike of fixed charge injection into the ground reference node (i.e., a fixed amp-second product), a larger bypass capacitor will keep transient ground noise smaller, but will have a longer recovery time than a small capacitor (Fig. 14.26). For a high-gain low-speed circuit that may be worse, perhaps producing slow exponential recovery instead of harmless little spikes at the output.

When designing ground reference circuits, don't overlook the reference voltage outputs that are sometimes provided on other ICs. For example, the LM322 timer provides a stable 3.15 volt output. Other chips that have external access to internal voltage references are A/D converters, V/F converters (e.g., the 331, with its 1.89V reference), and chips like the LM10, which has a 200mV reference and amplifier, in addition to an uncommitted op-amp. Figure 14.27 shows several buffered reference schemes.

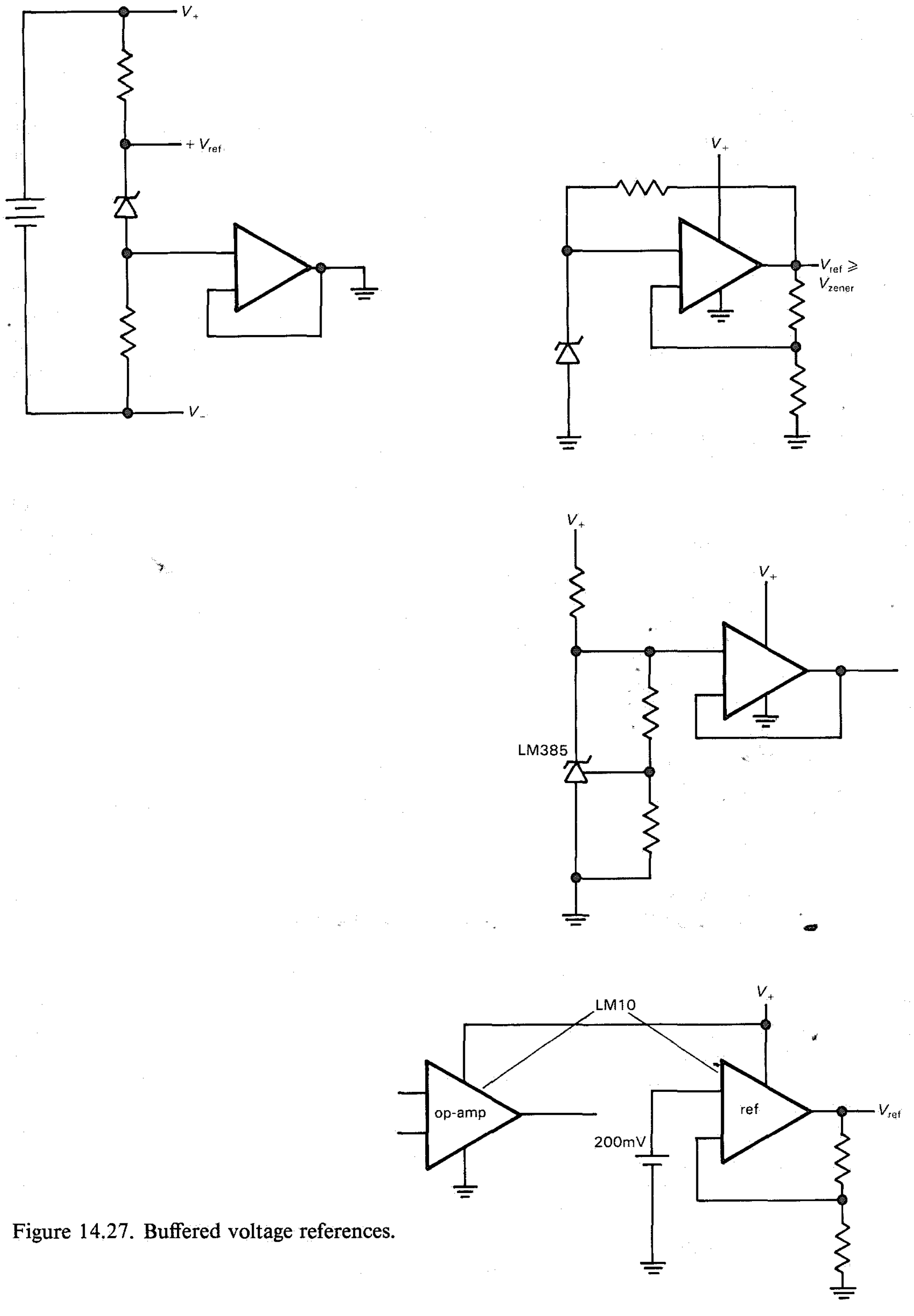


Figure 14.27. Buffered voltage references.

14.09 Micropower voltage references and temperature sensors

Most zener and bandgap references are relatively power-hungry and are not suitable for micropower circuits. As Table 6.7 demonstrates, most 3-terminal references run at about a milliamp, and most 2-terminal zener-like references are specified at similar operating currents.

Fortunately, there are several voltage references intended for micropower applications. The LM385 series includes a programmable 2-terminal bandgap reference (LM385, 1.24V–5.30V) and two fixed voltage references (LM385-1.2, 1.235V and LM385-2.5, 2.50V). The fixed-voltage models are specified to operate at currents down to 10μA, with dynamic impedances of 1 ohm at 40μA and 100μA, respectively. The minimum current of the programmable version goes from 10μA to 50μA, depending on voltage. All versions are available with tempcos down to 30ppm/°C. The ICL7663/4 regulators (Section 14.07) can be used as 3-terminal references, with typical quiescent current of 4μA and dynamic output impedance of 2 ohms. The ICL8069 is a 2-terminal bandgap reference that operates down to 50μA (where the dynamic impedance is 1Ω), with tempco down to 50ppm/°C. The AD589 has the same characteristics, with improved tempco (down to 10ppm/°C). The LT1004 from Linear Technology is like the LM385-1.2, while their LT1034 is a dual 2-terminal reference (1.2V and 7.0V) with minimum operating currents of 20μA and tempco of 20ppm/°C for its 1.2 volt reference; the 7 volt reference should be operated at 100μA (min), but is quieter than the bandgap references.

For better tempco at not-quite-micropower currents, there's the LM368 3-terminal reference, available in 5, 6.2, and 10 volt versions (0.05% accuracy). It draws 300μA, has low output impedance over frequency, and is available with tempcos down to 10ppm/°C. Even better is the REF-43, a 3-terminal 2.5 volt positive reference with 0.05% accuracy and 3ppm/°C tempco (max). It has low Z_{out} (0.1Ω), excellent regulation (2ppm/V_{in}, max), output current to 10mA, and quiescent current of 250μA, max.

Table 14.5 lists currently available micropower references.

Finally, there are micropower ICs that convert temperature to current or to voltage. The AD590 and AD592 are 2-terminal current sources that run on 4 volts to 30 volts, and give a current of 1μA/°K (e.g., 298.2μA at 0°C). The LM334 is similar, but with a programming pin to set the conversion factor; the operating range is 1μA to 10mA. The LM34 (Fahrenheit) and LM35 (Centigrade) are 3-terminal temperature sensors with voltage output (thus 0V at 0°F or 0°C, and 10mV/°F or °C, respectively) and quiescent current of 100μA. The LM335 is a 2-terminal IC zener with a breakdown voltage of 10mV/K (e.g., 2.982V at 0°C), operable down to 400μA. See Section 15.1 for additional information.

LINEAR MICROPOWER DESIGN TECHNIQUES

Thus far we have treated power sources, power-switching techniques, regulators, and references for the design of micropower instruments. Now, following the progression of topics in the rest of the book, we turn to the design of linear and digital circuits themselves. We will begin with a discrete linear circuit example (a high-gain micropower audio amplifier), then proceed to micropower op-amp design techniques. That will be followed by sections on digital and microprocessor design, and finally some comments on packaging techniques for low-power instruments.