The Art of Electronics

Third Edition

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THE ART OF ELECTRONICS

Third Edition

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To Vida and Ava

In Memoriam: Jim Williams, 1948–2011

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VOLTAGE REGULATION AND POWER CONVERSION

CHAPTER 9

The control and conversion of power – power engineering – is a rich and exciting subfield of electrical engineering and electronic design. It encompasses applications ranging from high-voltage (kilovolts and upward) and high-current (kiloamperes and upward) dc transmission, transportation, and pulsing, all the way down to low-power fixed and portable (battery-operated) and micropower (energy-harvesting) applications. Perhaps of most interest to us in the context of circuit design; it includes the production of the voltages and currents needed in electronic circuit design.

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 powerline ripple (120 Hz or 100 Hz). Fortunately, it is easy to construct highly stable power supplies, by 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,[1] a battery, or some other source of dc input) and a few other components.

In this chapter we will see how to construct voltage regulators by 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 usually 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 heatsinking 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 we 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 plenty of uses outside of power-supply design (for example in analog–digital conversion).

We begin with the *linear* regulator, in which feedback controls conduction in a series voltage-dropping "pass transistor" to hold constant the output voltage. Later we treat the important topic of *switching* regulators, in which one or more transistors are switched rapidly to transfer energy, via an inductor (or capacitor) to the load, again with voltage-regulating feedback. In a nutshell, linear regulators are simpler and generate "cleaner" (i.e., noise-free) dc output; switchers (the nickname for switching regulators and converters) are more compact and efficient (Figure 9.1), but noisier and usually more complex.

It would be wrong to leave the impression that voltage regulators are used exclusively in ac-powered dc supplies. In addition to their use in creating stable dc voltages from the ac powerline, voltage regulators are used widely also to produce additional dc voltages from an existing *regulated dc* voltage within a circuit: it's common to see, for example, a regulator that accepts an existing +5 V input and generates a +2.5 V or +3.3 V output; this is easily done with a linear regulator, in which feedback controls the voltage drop to maintain constant (and reduced) output voltage. Perhaps more surprising, you can use a switching regulator to convert a given dc input to a *larger* output voltage, to an output voltage of opposite polarity, or to a constant current (for example, to drive a string of LEDs). These applications are particularly relevant to battery-powered devices. The more general term *power converter* is often used in such applications, which include also creating an ac output from a dc input.

[1] Sometimes the transformer can be omitted; this is most commonly done in *switchmode* power supplies (SMPSs), see §9.6.

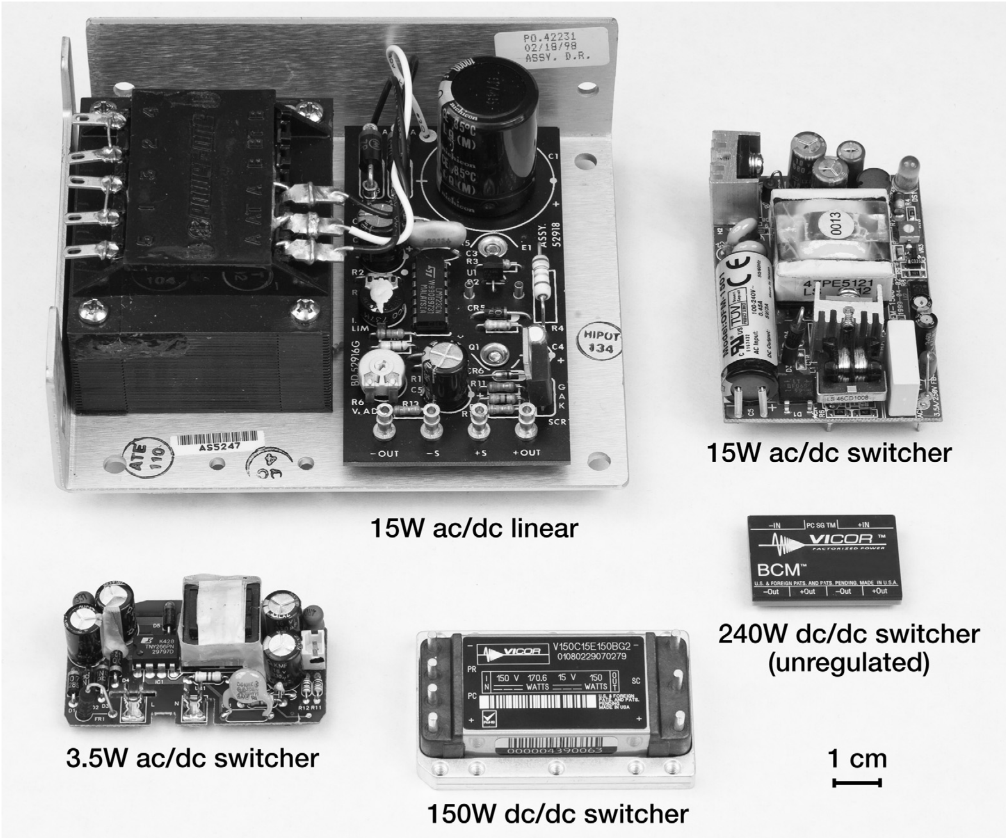

Figure 9.1. Switching power supplies ("switchers") are smaller and more efficient than traditional linear regulated power supplies, but the switching operation generates some unavoidable electrical noise.

9.1 Tutorial: from zener to series-pass linear regulator

To get started, let's look at the circuits in Figure 9.2. Recall that a zener diode is a voltage regulator, of sorts: it draws negligible current until the voltage across it is brought close to its zener voltage V_Z, at which point the current rises abruptly (look back at Figure 1.15 for a reminder). So a zener (or 2-terminal zener-like *reference* IC, see §9.10.2) biased through a resistor from a dc voltage greater than V_Z, as in Figure 9.2A, will have approximately V_Z across it, with the current set by the resistor:[2] $I_{\text{zener}} = (V_+ - V_Z)/R$. You can connect a load to this relatively stable output voltage; then, as long as the load draws less than I_{zener} (as just calculated), there will be some remaining zener current, and the output voltage will change little.

The simple resistor-plus-zener is occasionally useful as is, but it has numerous drawbacks: (a) you cannot easily change (or even choose precisely) the output voltage; (b) the zener voltage (which is also the output voltage) changes somewhat with zener current; so it will change with variations in V_+ and with variations in load current;[3] (c) you've got to set the zener current (by choice of R) large enough so that there's still some zener current at maximum load; this means that the V_+ dc supply is running at full current all the time, generating as much heat as the maximum anticipated load; (d) to accommodate large load currents[4] you would need a high-power zener; these are hard to find, and rarely used, precisely because there are much better ways to make a regulator, as we'll see.

Exercise 9.1. Try this out, to get a sense of the problems with

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[2] With the exact I versus V curve of the zener in hand, you could determine the voltage and current exactly, using the method of *load lines*; see Appendix F, and §3.2.6B.

[3] These are called *line* and *load* variations, respectively.

[4] Or, more precisely, large *variations* in load current, and/or in V_+ dc input voltage.

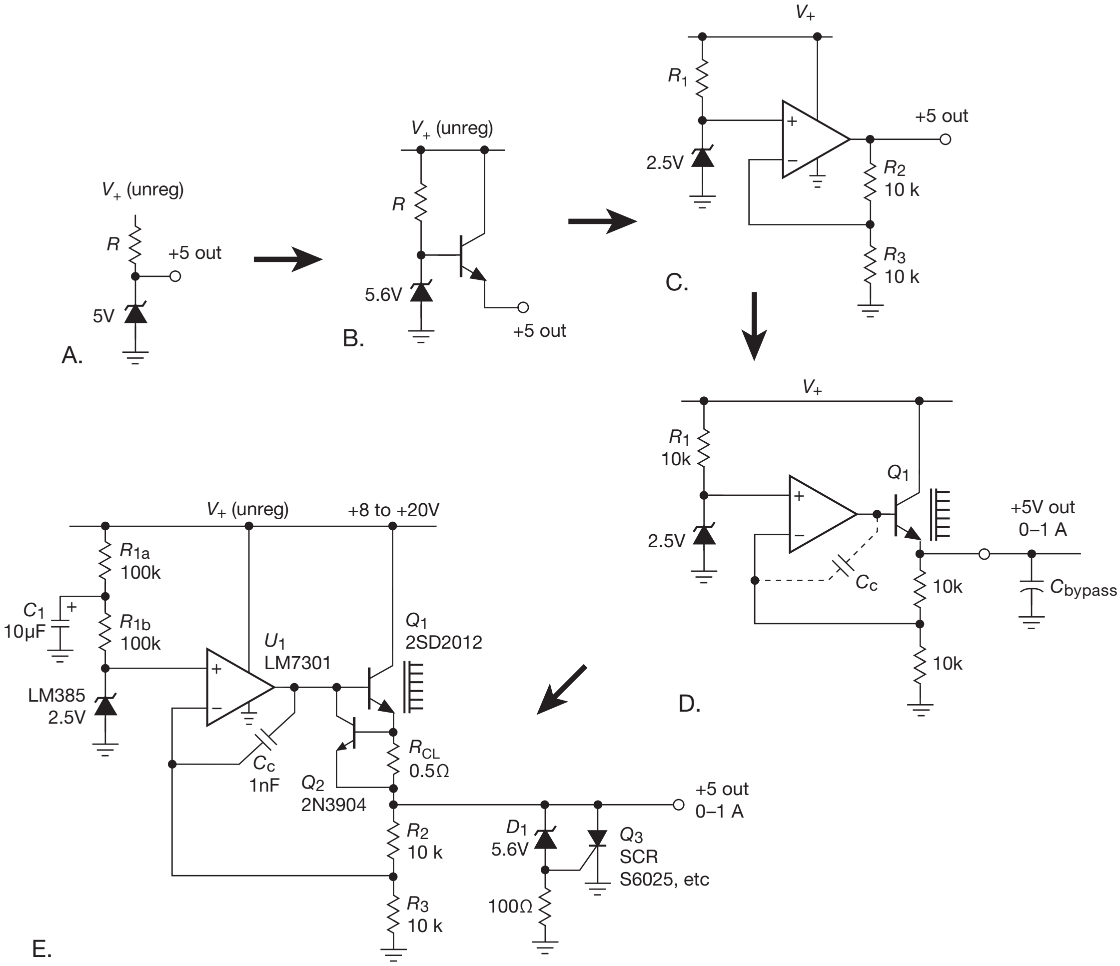


Figure 9.2. Evolving the (discrete-component) series-pass linear voltage regulator.

this simple regulator circuit: imagine we want a stable +5 V dc output, to power a load that can draw from zero to 1 A. We've built an unregulated dc supply (using a transformer, diode bridge, and capacitor) that puts out approximately +12 V when unloaded, dropping to +9 V at 1 A load. Those voltages are "nominal" and can vary ±10%.

(a) What is the correct resistor value, *R*, for the circuit of Figure 9.2A, such that the minimum zener current, under "worst-case" conditions, is 50 mA?

(b) What is worst-case (maximum) power dissipation in *R* and in the zener?

Contrasted with this approach – with its requirement for a 10 W zener at the desired output voltage, and nearly 10 W of power dissipation in each component, even at zero load – we'll see that it is a routine task to make a regulated power supply, with adjustable output voltage, without the need for a power zener and with 75% or better efficiency over most of the load-current range.

9.1.1 Adding feedback

We could improve the situation somewhat by tacking an emitter follower onto a zener (Figure 9.2B); that lets you run at lower zener current, and low quiescent dissipation when unloaded. But the output regulation is still poor (because V_{BE} varies with output current), and the circuit still does not allow adjustment of output voltage.

The solution is to use a zener (or other voltage reference device; see §9.10.2) as a low-current voltage reference, against which we compare the output. Let's take it in a few easy steps.

A. Zener plus "amplifier"

First we solve the problem of *adjustability* by following the zener reference with a simple dc amplifier (Figure 9.2C). Now the zener current can be small, just enough to ensure a stable reference. For typical zeners this might be a few milliamps, whereas for an IC voltage reference, 0.1–1 mA will usually suffice. This circuit lets you adjust the output voltage: $V_{\text{out}} = V_Z(1+R_2/R_3)$. But note that you are limited to having $V_{\text{out}} \geq V_Z$; note also that the output voltage comes from an op-amp, so it can at most reach V_+, with an output current limited by the op-amp's I_{out}(max), typically 20 mA. We will overcome both these limits.

B. Adding outboard pass transistor

More output current is easy – just add an *npn* follower, to boost the output current by a factor of β. You might be tempted to just hang the follower on the op-amp's output, but that would be a mistake: the output voltage would be down by a V_{BE} drop, roughly 0.6 V. You could, of course, adjust the ratio R_2/R_3 to compensate. But the V_{BE} drop is imprecise, varying both with temperature and with load current, and so the output voltage would vary accordingly. The better way is to close the feedback loop around the pass transistor, as in Figure 9.2D; that way the error amplifier sees the actual output voltage, holding it stable via the circuit's loop gain. The inclusion of the output emitter follower boosts the op-amp's I_{out}(max) by the beta of Q_1, giving us an available output current of an ampere or so. (We could use a Darlington, instead, for more current; another possibility is an *n*-channel MOSFET.) Q_1 will be dissipating 5–10 W at maximum output current, so you'll need a heatsink (more on this in §9.4.1). And, as we'll see next, you'll also need to add a compensation capacitor C_C to ensure stability.

C. Some important additions

Our voltage regulator circuit is nearly complete, but lacks a couple of essential features, related to loop stability and overcurrent protection.

Feedback loop stability

Regulated power supplies are used to power electronic circuits, typically festooned with many bypass capacitors between the dc rails and ground. (Those bypass capacitors, of course, are needed to maintain a pleasantly low impedance at all signal frequencies.) Thus the dc supply sees a large capacitive load, which, when combined with the finite output resistance of the pass transistor (and overcurrent sense resistor, if present), causes a lagging phase shift and possible oscillation. We've shown the load capacitance in Figure 9.2D as C_{bypass}, a portion of which might be included explicitly (as a real capacitor) in the power supply itself.

The solution here, as with the op-amp circuits we worried about earlier (§4.9), is to include some form of *frequency compensation*. That is most simply done (as it is within op-amps) with a Miller feedback capacitor C_C around the inverting gain stage, as shown. Typical values are 100–1000 pF, usually found experimentally ("cut-and-try") by increasing C_C until the output shows a well-damped response to a step change in load (and then doubling that, to provide a good margin of stability). The IC regulators we'll see later will either include internal compensation, or they'll give you suggested values for compensation components.

Overcurrent protection

The circuit as drawn in Figure 9.2D does not deal well with a short-circuit load condition.[5] With the output shorted to ground, feedback will act to force the op-amp's maximum output current into the pass transistor's base; so that I_B of 20–40 mA will be multiplied by Q_1's beta (which might range from 50 to 250, say), to produce an output current of 1 A to 10 A. Assuming the unregulated V_+ input can supply it, such a current will cause excessive heating in the pass transistor, as well as interesting forms of damage to the misbehaving load.

The solution is to include some form of overcurrent protection, most simply the classic current-limiting circuit consisting of Q_2 and R_{CL} in Figure 9.2E. Here R_{CL} is a low-value *sense resistor*, chosen to drop approximately 0.6 V (a V_{BE} diode drop) at a current somewhat larger than the maximum rated current; for example, we might choose $R_{\text{CL}} = 5\,\Omega$ in a 100 mA supply. The drop across R_{CL} is applied across Q_2's base–emitter, turning it on at the desired maximum output current; Q_2's conduction robs base current from Q_1, preventing further increase of output current. Note that the current-limit sense transistor Q_2 does not handle high voltage, high current, or high power; it sees at most two diode drops from collector to emitter, the op-amp's maximum output current, and the product of those two, respectively. During an overcurrent load condition, then, it typically would have to handle $V_{\text{CE}} \leq 1.5$ V at $I_C \leq 40$ mA, or 60 mW; that's peanuts for any general-purpose small-signal transistor.

Later we'll see variations on this simple overcurrent protection theme, including methods that limit to an

[5] Engineers like to refer to various bad situations such as this under the general rubric of *fault conditions*.

adjustable and stable current limit, and the technique known as *foldback current limiting* (§9.13.3).

Zener bias; overvoltage crowbar

We've shown two additional wrinkles in Figure 9.2E. First, we split the zener biasing resistor R_1 and bypassed the midpoint, to filter out ripple current. By choosing the time constant ($\tau = (R_{1a}\|R_{1b})C_1$) to be long compared with the ripple period of 8.3 ms, the zener sees ripple-free bias current. (You wouldn't bother with this if the dc supply V_+ were already free of ripple, for example a regulated dc supply of higher voltage.) Alternatively, you could use a current source to bias the zener.

Second, we've shown an "overvoltage crowbar" protection circuit consisting of D_1, Q_3, and the 100 Ω resistor. Its function is to short the output if some circuit fault causes the output voltage to exceed about 6.2 V (this can happen easily enough, for example if the pass transistor Q_1 fails by having a collector-to-emitter short, or if a humble component like resistor R_2 becomes open-circuited.). Q_3 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 for example in Figure 9.48. We will treat overvoltage crowbar circuits in more detail in §§9.13.1 and *9x.7*.

Exercise 9.2. Explain how an open circuit at R_2 causes the output to soar. What voltage, approximately, would then appear at the output?

9.2 Basic linear regulator circuits with the classic 723

In the preceding tutorial we evolved the basic form of the linear *series pass regulator*: voltage reference, pass transistor, error amplifier, and provisions for loop stability and overvoltage–overcurrent protection. In practice you seldom need to assemble these components from scratch – they are available as complete integrated circuits. One broad class of IC linear regulators might be thought of as flexible *kits* – they contain all the pieces, but you've got to hook up a few external components (including the pass transistor) to make them work; an example is the classic 723 regulator. The other class of regulator ICs are complete, with built-in pass transistor and overload protection, and requiring at most one or two external parts; an example is the classic 78L05 "3-terminal" regulator – its three terminals are labeled *input*, *output*, and *ground* (and that's how easy it is to use!).

9.2.1 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.[6] Although you might not choose it for a new design nowadays, it is worth looking at in some detail, because more recent regulators work on the same principles. Its block diagram is shown in Figure 9.3. As you can see, it is really a power-supply *kit*, containing a temperature-compensated voltage reference (7.15 V±5%), 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.

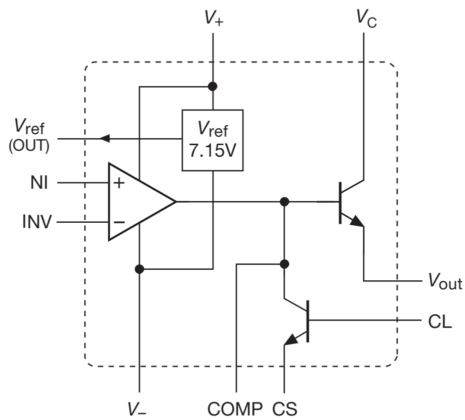


Figure 9.3. The classic μA723 voltage regulator.

The 723's internal *npn* pass transistor is limited to

[6] Building on the 723's success, other manufacturers introduced "improved" versions, such as the LAS1000, LAS1100, SG3532, and MC1469. However, while the 723 lives on, the improved versions are all gone! The 723 is "good enough," *very* inexpensive (about $0.15 in quantity), and is popular in many commercial linear power supplies, where the easily adjusted current limit is especially useful. It also has less noise than most modern replacements. And we like it for its pedagogic value.

150 mA, and it can dissipate about 0.5 W maximum. Unlike newer regulators, the 723 does not incorporate internal shutdown circuitry to protect against excessive load current or chip dissipation.

A. 723 regulator example: $V_{out} > V_{ref}$

Figure 9.4 shows how to make a positive voltage regulator with the 723 for output voltages greater than the reference voltage; it is the same circuit topology as the tutorial's Figure 9.2E. All the components except the three resistors and the two capacitors are contained in the 723. With this circuit, a regulated supply with output voltage ranging from V_{ref} up to the maximum allowable output voltage (37 V) 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. This is a bit large by contemporary standards, where the dropout voltage is typically 2 V, and much less for *low dropout* (LDO) regulators, as we'll see in §9.3.6. Note also that the 723's relatively high reference voltage means that you cannot use it in a power supply whose unregulated dc input is less than +9.5 V, its specified minimum V_+; this shortcoming is remedied in a large selection of regulators that use a lower-voltage *bandgap reference* (1.25 V or 2.5 V). And while we're complaining, we note that the reference is not exactly sterling in its initial accuracy – the production spread in V_{ref} is 6.8 to 7.5 volts – which means that you must provide for output-voltage trimming, by making R_1 or R_2 adjustable; we'll soon see regulators with excellent initial accuracy, for which no trim is needed.

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, to ensure stability against oscillations. In general, it is a good idea to bypass power-supply leads to ground liberally throughout a circuit, using a combination of ceramic types (0.01–0.1 μF) and electrolytic or tantalum types (1–10 μF).[7]

[7] The ceramic capacitors provide low impedance at high frequencies, whereas the larger electrolytics provide energy storage, and also damping of oscillations (via their internal equivalent series resistance, or ESR).

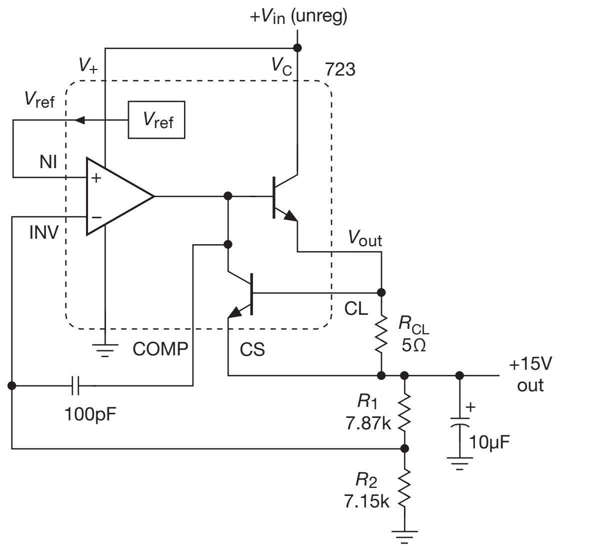


Figure 9.4. 723 regulator: configuration for $V_{out} > V_{ref}$, with 100 mA current limit.

B. 723 regulator example: $V_{out} < V_{ref}$

For output voltages less than V_{ref}, you just put the voltage divider on the reference (Figure 9.5). Now the full output voltage is compared with a fraction of the reference. The values shown are for a +5 V output. With this circuit configuration, output voltages from +2 V 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, as specified on the datasheet. Note again that the unregulated input voltage must never drop below +9.5 V, the voltage necessary to power the reference.

For this example we've added an external pass transistor, in a Darlington configuration with the 723's small internal pass transistor, to get beyond the latter's 150 mA current limit. An external transistor is needed, also, because of power dissipation: the 723 is rated at 1 watt at 25°C (less at higher ambient temperatures; the 723 must be "derated" at 8.3 mW/°C above 25°C in order to keep the junction temperature within safe limits). Thus, for instance, a 5 volt regulator with +15 V input cannot deliver more than about 80 mA to the load. Here the external power transistor Q_1 will dissipate 14 W for V_{in}=12 V and maximum load current (2 A); that requires a *heatsink*, 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 later in the chapter.[8] A trimmer

[8] And for a table of bipolar power transistors see Table 2.2 on page 106.

potentiometer has been used so that the output can be set accurately to $+5\,\mathrm{V}$; 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.

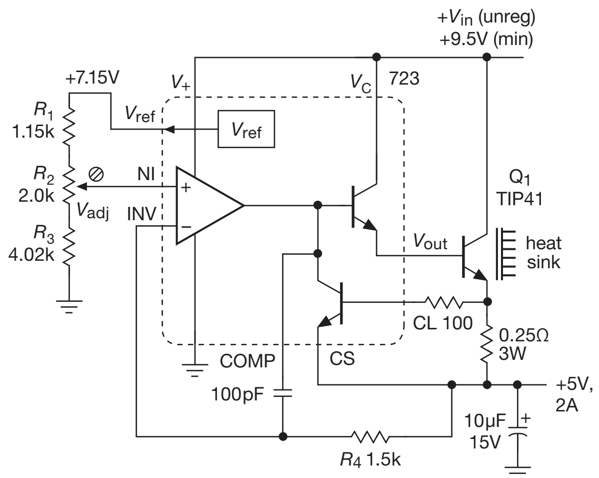


Figure 9.5. 723 regulator: configuration for $V_{\mathrm{out}} < V_{\mathrm{ref}}$, with 2 A current limit.

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

C. Pass-transistor dropout voltage

One problem with this circuit is the high power dissipation in the pass transistor (at least 10 W 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., $+12\,\mathrm{V}$), the minimum unregulated input to the external pass transistor can be as little as 1.5 V or so above the regulated output voltage (i.e., two V_{BE}'s).[9]

9.2.2 In defense of the beleaguered 723

Lest we leave the wrong impression, we hasten to remark that rumors of the death of the vintage 723 regulator are greatly exaggerated.[10] We have been using dozens of linear regulated power supplies manufactured by Power One for more than three decades without a single failure. All of them use the humble 723 regulator chip, as do other OEMs ("original equipment manufacturers"). Here are some reasons not to overlook this remarkable design of the legendary Bob Widlar:

- very low cost, \$0.17 (in qty 1000)
- many, many manufacturers
- fully settable current limit, including foldback
- good for pedagogy (that's why it's here!)
- the power dissipation is not in the control IC
- quiet voltage reference, plus can add filter
- works with *npn* or *pnp* pass transistors
- easily configured for negative outputs

9.3 Fully integrated linear regulators

The overall regulator circuit of Figure 9.5 has ten components, but only three terminals (IN, OUT, and GROUND), thus suggesting the possibility of an integrated solution, with on-chip voltage-setting resistors and with integrated components for current-limiting and loop compensation – a *3-terminal* regulator. The 723 is approaching half-century vintage (though still going strong!), during which the semiconductor industry has not been sleeping: contemporary linear regulator ICs generally integrate all regulator functions on-chip, including overcurrent and thermal protection, loop compensation, high-current pass transistor, and preset voltage divider for commonly used output voltages. Most of these regulators come also in adjustable versions, for which you provide only the voltage-setting resistor pair. And, with an additional terminal or two, you can get a "shutdown" control input and a "power-good" status output. Finally, a large and growing population of low-dropout regulators addresses low voltage applications, of increasing importance in low-power and portable electronic devices. Let's look at the choices favored for contemporary design.

[9] A trick you can use to reduce the minimum headroom to a single V_{BE} is to replace Q_1 with a *pnp* pass transistor (tie its emitter to V_{in} and drive its base from the V_{c} pin of the 723), forming a Sziklai pair rather than a Darlington (see §2.4.2A and Figure 2.77). If the input comes from an unregulated dc supply, however, you will always have to allow at least a few volts of headroom, because worst-case design dictates proper operation even at 105 Vac line input.

[10] Paraphrasing Mark Twain's famous remark, upon opening the newspaper and reading his obituary.

9.3.1 Taxonomy of linear regulator ICs

As a guide to the following sections, we've organized the universe of integrated linear voltage regulators into a few distinct categories, here simply listed in outline form. For each category we've listed typical example part numbers of devices that we are fond of, and use often. Read on for explanations of when and how to use them, a description of their distinguishing features, and some important cautions.

3-terminal fixed

pos: 78xx
neg: 79xx

3-terminal adjustable

pos: LM317
neg: LM337

3-term "lower dropout" (adj & fixed)

pos: LM1117, LT1083-85

3-term fixed & 4-term adj "true LDO"

pos: LT1764A/LT1963 (BJT); TPS744xx (CMOS)
neg: LT1175, LM2991 (BJT); TPS7A3xxx (CMOS)

3-term current reference

pos: LT3080

9.3.2 Three-terminal fixed regulators

The original (and often good enough) 3-terminal regulator is the 78xx series (Figure 9.6), originated by Fairchild in the early 1970s.[11] It is factory trimmed to provide a fixed output, in which the voltage is specified by the last two digits of the part number, and can be any of the following: 05, 06, 08, 09, 10, 12, 15, 18, or 24. These regulators can supply up to 1 A of output current, and come in power packages (TO-220, DPAK, D^{2}PAK) that you attach to a heatsink or to an area of circuit-board copper. If you don't need much current, use the 78Lxx/LM340Lxx series, which come in small transistor packages, either surface-mount or TO-92 (through-hole). For negative output voltage use the 79xx/79Lxx (or LM320/320L) series. Figures 9.6 and 9.7 show, in simplified form, what's inside these inexpensive ($0.30) regulators.

Figure 9.8 shows how easy it is to make a +5 V regulator, for example, with one of these ICs. Here we've added also a 7905 negative regulator to create a −5 V regulated output from a more negative unregulated dc input. The bypass capacitors at the outputs ensure stability; they also improve transient response, and maintain a low output impedance at high frequencies (where the regulator's loop gain is low).[12] The input bypass capacitors are also needed

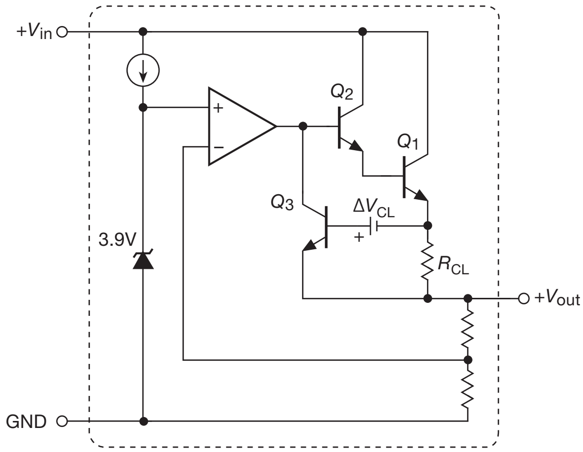


Figure 9.6. Simplified 78xx fixed 3-terminal positive voltage regulator. All components are internal, so only a pair of bypass capacitors is required (as in Figure 9.8). R_{CL}, the current-sensing resistor, is 0.2 Ω, and develops somewhat less than a diode drop at full current; its drop is supplemented by an internal bias ΔV_{CL}, to turn on current-limiting transistor Q_3.

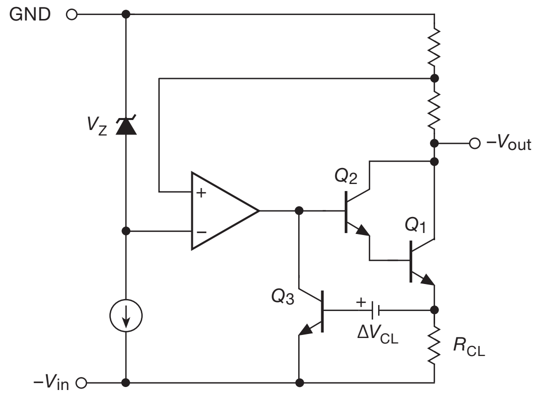


Figure 9.7. 79xx fixed 3-terminal negative voltage regulator.

[11] The LM340 series from National is essentially the same.

[12] A regulator's datasheet will always specify minimum required capacitance. It may go into considerably more detail, in cases where stability is an important issue, for example with low-dropout regulators (see later discussion). Note the larger capacitance values in the negative regulator circuit: they are needed to ensure stability, because the 7905 regulator's output comes from the collector of a common-emitter amplifier output stage (whose gain depends on load impedance), rather than from the emitter follower output stage of the 7805 positive regulator (whose gain is near unity); the larger bypass capacitor kills the loop gain at high frequency, preventing oscillation.

for stability; the values shown are the minimum suggested in the datasheets. However, if the input supply or output load is bypassed close to the regulator, the corresponding capacitors can be omitted.

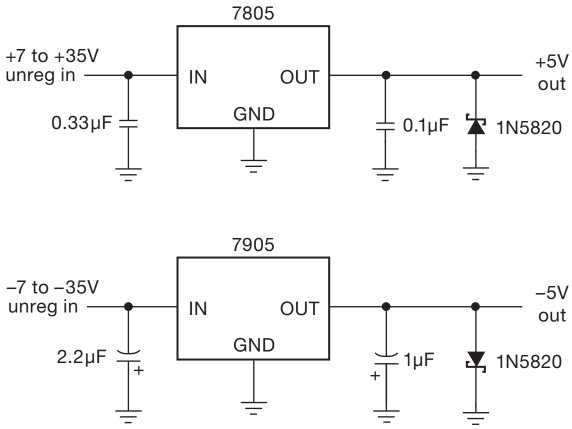


Figure 9.8. ±5 V regulated dc from a 7805/7905 regulator pair.

This regulator example includes a pair of reverse-protection Schottky (low-forward-drop) diodes, always a good idea when you have supplies of both polarities powering a circuit. Without the diodes, one of the supplies can bring the other into reverse output voltage, via the load; this reversed supply polarity can cause failure in the load (from transistors or ICs that are subjected to reverse supply voltage), or in the regulator (which may even go into a latchup condition). You often see the diodes omitted; don't get into this lazy habit!

These regulators have 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 §9.4.2) by reducing available output current for large input–output voltage differentials. These regulators are inexpensive and easy to use, and they make it practical to design a system with many printed circuit boards (PCBs) in which the unregulated dc is brought to each board and regulation is done locally on each circuit card. Table 9.1 lists the characteristics of a representative selection of 3-terminal fixed regulators.

Three-terminal fixed regulators come in some highly useful variants. There are low-power and micropower versions (e.g., the LM2936 and LM2950, with quiescent current in the microampere range), and there are the very popular LDO regulators, which maintain regulation with only a few tenths of a volt input–output differential (e.g., the LT1764A, TPS755xx, and micropower LM2936, with typical dropout voltages ≈0.25 V). We'll discuss LDOs after taking a look at the very useful 3-terminal *adjustable* regulator.

Table 9.1 7800-Style Fixed Regulators[a]

Part #[c]	V_{in} max (V)	V_{out}^{d} nom (V)	Tol (±%)	I_Q typ (mA)	I_{out} max (A)	Cost qty 25 ($US)
78L05	35	5	5	3	0.1	0.29
78L15	35	15	4	3	0.1	0.31
7805	35	5	4[b]	5[e]	1.0	0.47
7824	40	24	4[b]	5	1.0	0.49
79L05	−35	−5	5	2	0.1	0.30
79L15	−35	−15	4	2	0.1	0.30
7905	−35	−5	4[b]	3	1.0	0.47
7924	−40	−24	4[b]	4	1.0	0.56

Notes: (a) often called '7800 and '7900 series, e.g., "LM7800-series." L series available in TO-92, SO-8 and SOT-89 packages; regular series available in TO-220, DPAK, D2PAK, and TO-3. Some use buried-zener ref, some use bandgap. (b) A-suffix types are ±2% tol. (c) prefixes: uA, LM, MC, KA, NCP, L, NJM, etc. (d) L-series: 2.6 to 24V, regular: 5 to 24V. (e) some lower, 3.3mA to 4mA

9.3.3 Three-terminal adjustable regulators

Sometimes you want a nonstandard regulated voltage (say +9 V, to emulate a battery) and can't use a 78xx-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"!

These convenient ICs are typified by the classic LM317 originally from National (Figure 9.9). This regulator has no ground terminal; instead, it adjusts V_{out} to maintain a constant 1.25 V (internal "bandgap" reference, §9.10.2) from the output terminal to the "adjustment" terminal. Figure 9.10 shows the easiest way to use it. The regulator puts 1.25 V across R_1, so 10 mA flows through it. The adjustment terminal draws very little current (50–100 μA), so the output voltage is just

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

In this case the output voltage is +3.3 V, with an untrimmed accuracy of ≈3% (from the ±2% internal 1.25 V reference and the 1% resistors). If you want accurate settability,

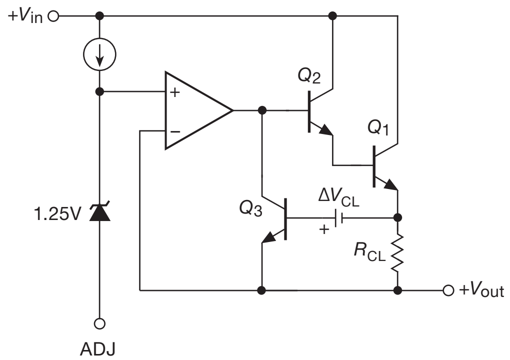


Figure 9.9. LM317 three-terminal adjustable positive voltage regulator.

replace the lower resistor with a 25 Ω trimmer in series with a 191 Ω fixed resistor, to narrow the trimmer's adjustment range to ±6%. If you want instead a wide adjustment range, you could replace the lower resistor with a 2.5k trimmer, for an output range of +1.25 V to +20 V. Whatever the output voltage, the input must be at least 2 V higher (the dropout voltage).

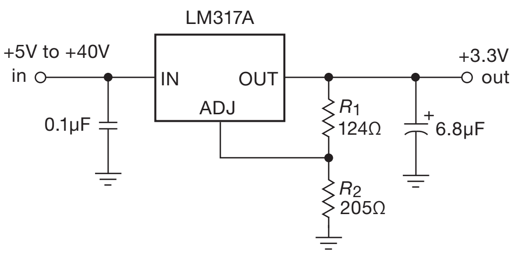


Figure 9.10. +3.3 V positive regulator circuit.

When using this type of regulator, choose your resistive divider values small enough to allow for a 5 μA change in adjustment pin current with temperature: many designers use 124 Ω for the upper resistor, as we've done, so that the divider alone sinks the chip's specified minimum load current of 10 mA. Note also that the current sourced out of the adjustment pin may be as large as 100 μA (the worst-case spec). The output capacitor, though not necessary for stability, greatly improves transient response. It's a good idea to use at least 1 μF, and ideally something more like 6.8 μF.

The LM317 is available in many package styles, including the plastic power package (TO-220), the surface-mount power package (DPAK and D^{2}PAK), and many small transistor packages (both through-hole TO-92, and a half dozen tiny surface-mount styles). In the power packages it can deliver up to 1.5 amps, with proper heatsinking; the low-power variant (317L) is rated to 100 mA, again limited by power dissipation. The popular LM1117 variant, also available from multiple manufacturers, improves on the dropout voltage of the classic 317 (1.2 V versus 2.5 V), but you pay a price (literally): in the TO-220 package it costs about $0.75 versus the 317's $0.20; it also has a more limited voltage range (see Table 9.2), and, in common with many low-dropout regulators, it requires a larger output capacitor (10 μF minimum).

Exercise 9.4. Design a +5 V 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). We've listed their properties in Table 9.2 on page 605. Read the datasheets carefully before using these parts, noting bypass capacitor requirements and safety diode suggestions. Note also that the rated maximum output currents generally apply at lower values of V_{in}–V_{out}, and can drop to as little as 20% of their maximum values as V_{in}–V_{out} approaches $V_{in(max)}$; the maximum output current drops also with increasing temperature.[13]

An alternative for high load currents is to add an outboard transistor (§9.13.4), though a high-current switching regulator (§9.6) is often a better choice. The LM317 family regulators are "conventional" (as opposed to low-dropout) linear regulators; typical dropout voltages are ≈2 V.

As with the fixed 3-terminal regulators, you can get lower-dropout versions (e.g., the popular LM1117, with 1.3 V maximum dropout at 0.8 A, or the heftier LT1083-85 series, with comparable dropout at currents to 7.5 A), and you can get micropower versions (e.g., the LP2951, the adjustable variant of the fixed 5V LP2950; both have I_Q=75 μA); see Figure 9.11. You can also get *negative* versions, though there's less variety: the LM337 (Figure 9.12) is the negative cousin of the LM317 (1.5 A), and the LM333 is a negative LM350 (3 A). There's more discussion ahead in §§9.3.6 and 9.3.9; see particularly Figure 9.24.

[13] As discussed later in §9.4.1, the junction temperature $T_J = P_{diss}(R_{\Theta JC}+R_{\Theta CS}+R_{\Theta SA})+T_A$, where the R_Θ are the thermal resistances from junction to case, case to heatsink, and heatsink to ambient. In situations with good heatsinking, you may choose to use a regulator of higher current rating and larger package style (e.g., the LM338K in its TO-3 metal can package) in order to take advantage of the much lower thermal resistance $R_{\Theta JC}$ (1°C/W versus 4°C/W for the LM317T in its TO-220 package). The larger parts also offer more relaxed safe-operating-area (SOA) constraints, e.g., at $V_{in}-V_{out}$=20 V the LM338 allows 3.5 A of output current, versus 1.4 A for the LM317.

Anatomy of a 317

The classic LM317, designed around 1970 by the legendary team of Widlar and Dobkin[14] has endured for more than four decades. Indeed, the generic 317 (along with the complementary LM337) has become the go-to part for linear regulators of modest current capability (to $\sim$1 A) in situations where you've got a few volts of headroom. And it has spawned a host of imitators and look-alikes, spanning a range of voltages, currents, and package styles, with some variants of lower dropout-voltage; see Table 9.2.

Its design exhibits a nice elegance, for example by combining the functions of error amplifier and zero-tempco bandgap reference. It was also one of the first regulators to include thermal overload and safe-area protection. Figure 9.13 is a simplified circuit of its essential innards, with part designations following the schematic diagram in the National Semiconductor (TI) datasheet.

The transistor pair Q_{17} and Q_{19} forms the bandgap voltage reference, operating at equal currents from the $Q_{16}Q_{18}$ mirror. Because Q_{19} has $10\times$ larger emitter area (or 10 emitters), it operates at 1/10 the current density of Q_{17}, thus a $V_{\rm BE}$ that is smaller by $(kT/q)\log_e 10$, about 60 mV (§*2x.3.2*). That sets its current (via R_{15}) to be $I_{\rm Q19}=\Delta V_{\rm BE}/R_{15}=25\,\mu$A, and thus the pair's total current is 50 μA.[15] Note that the current has a linear dependence on absolute temperature (because the drop across R_{15} is $\propto T_{\rm abs}$) – it is "PTAT" (proportional to absolute temperature).

Now for the classic "bandgap reference" temperature compensation: the positive tempco of current is exploited to cancel Q_{17}'s negative tempco of $V_{\rm BE}$, which is nominally about 600 mV and goes as $1/T_{\rm abs}$, or -2.1 mV/°C (§2.3.2). Cancellation occurs when R_{14} is chosen to drop a comparable 600 mV at the nominal 50 μA, thus a +2.1 mV/°C tempco – voilà: zero tempco at a reference voltage of $\sim$1.2 V (the extrapolated bandgap energy of silicon).

The bandgap reference is also the error amplifier: Q_{17}'s collector sees a high-impedance (current-source) load, buffered by three stages of emitter follower (in the full schematic there are five) to the output pin; so even with its relatively low transconductance ($g_m \sim 1/R_{14}$) there's plenty of loop gain in the error amplifier (whose input is the ADJ pin, offset by $V_{\rm ref}$, relative to $V_{\rm out}$).

Resistor R_{26} senses the output current, for current limiting via Q_{21}. A bias that depends on $V_{\rm in}$–$V_{\rm out}$ is added (the battery symbol), for safe-operating-area protection. Additional components add hysteretic overtemperature shutdown (Q_{21} is paired with a *pnp* to make a latch). One final note: the Widlar–Dobkin duo also created the LM395 and LP395 protected-transistor ICs; these include the current and thermal limiting from the 317, but without the bandgap reference. They call it, modestly, the "Ultra-Reliable Power Transistor." The '395 transistor's base is the base of the *pnp* transistor Q_{15} in Figure 9.13. This yields a roughly 800 mV base-to-emitter voltage, with a 3 μA pullup base current. It's a great idea, but an LM395T costs about \$2.50, whereas an LM317T costs about \$0.50. So we use the '317 as our "pretty reliable power transistor," with its -1.2 V base-to-emitter voltage and 50 μA pullup base current, as for example in Figures 9.16 and 9.18.

9.3.4 317-style regulator: application hints

The LM317-style adjustable 3-terminal regulators are delightfully easy to use, and there are some nice tricks you can use to make them do more than simply create a fixed dc output voltage. There are also some basic cautions to keep in mind. In Figure 9.14 we've sketched some helpful circuit ideas.

Herewith a quick tour (keyed to the figure parts), taking them in order.

A: The regulator requires some minimum load current, because the operating current for the internal circuitry returns through the load. So if you want it to work clear down to zero external load, you should choose the upper feedback resistor R_1 small enough, i.e., so that $V_{\rm ref}/R_1 \geq I_{\rm out(min)}$ for the worst-case (maximum) value of $I_{\rm out(min)}$. For $V_{\rm ref}=1.25$ V and the classic LM317's $I_{\rm out(min)}=10$ mA, R_1 should be no larger than 125 Ω.[16] Of course, you could instead use a larger value of R_1 and add a load resistor to

[14] See Robert Widlar, "New developments in voltage regulators," JSSC, SC-6, pp 2–9, 1971, and US patent 3,617,859: "Electrical regulator apparatus including a zero temperature coefficient voltage reference circuit," filed 23 March 1970, issued 2 November 1971.

[15] The typical tolerance on resistor values in the planar silicon process (which is good for resistor *ratios*, but not absolute values) is ×0.5 to ×2, so the nominal 50 μA current sourced out of the ADJ pin can actually range from 25 μA to 100 μA.

[16] In contradiction to the LM117/317 datasheet's many circuit examples,

Table 9.2. 3-Terminal Adjustable Voltage Regulators ("LM317-style")[a]

Part #	Packages[z] TO-92	SOIC	TO-220	TO-3, TO-3P	D-PAK	SOT-223	V_{in} max (V)	I_{out}^{v} max (A)	V_{DO}^{h} max (V)	I_{out} min[b] (mA)	C_{out} min (μF)	V_{ref} (V)	± (%)	I_{adj} typ (μA)	Temp stab[c] typ (%)	Ripple reject 120Hz typ (dB)	Regulation Line typ (%)	Load[e] typ (%)	Cost qty 25 ($US)	Comments
Positive																				
LM317L	•	•	-	-	-	-	40	0.1	2.5^t	5	0.1	1.25	4	50	0.5	80^g	0.15	0.1	0.34	TO-92 lo-power '317
LM1117[n]	-		•	-	•	•	20	0.8	1.2	5	10	1.25	1	52	0.5	73	0.035	0.2	0.88	low V_{DO} '317, popular
NCP1117	-	-	-	-	•	•	20	1.0	1.2	5	10	1.25	1	52	0.5	73	0.04	0.2	0.40	higher current '1117
LMS8117A	-	-	-	-	d	•	20	1.0	1.2	5	10	1.25	1	60	0.5	75	0.035	0.2	0.92	higher current '1117
LM317[k]	-	-	•	•	•	•	40	1.5^p	2.5^t	10	0.1	1.25	4	50	0.6	80^g	0.01	0.1	0.15	orig, cheap, popular
LT1086CP	-	-	•	•	-	-	30	1.5	1.5	10	22^u	1.25	1	55	0.5	75	0.02	0.1	2.67	low dropout
LM350T	-	-	•	•	-	-	35	3	2.5^t	10	1	1.25	4	50	0.6	65	0.1	0.1	0.49	3A monolithic
LT1085CT	-	-	•	-	•	-	30	3	1.5	10	22^u	1.25	1	55	0.5	75	0.02	0.1	4.50	3A low dropout
LT1084CP	-	-	•	•	-	-	30	5	1.5	10	22^u	1.25	1	55	0.5	75	0.02	0.1	5.34	5A low dropout
LM338T	-	-	•	•	-	-	40	5	2.5^t	5	1	1.24	4	45	0.6	80	0.1	0.1	1.62	5A monolithic
LT1083CP	-	-	-	•	-	-	30	7.5	1.5	10	22^u	1.25	1	55	0.5	75	0.02	0.1	9.80	7.5A low dropout
Positive, high-voltage																				
LM317HV	-	-	•	•	-	-	60	1.5	2.0^t	12	0.1	1.25	4	50	0.6	80^g	0.01	0.1	2.17	high-voltage '317
LR12	•	•		-	•	-	100	0.05	12	0.5	0.1	1.20	5	10	1	60	0.003	1.4	1.39	Supertex
TL783C	-	-	•	-	•	-	125	0.7	10	15	1	1.27	5	83	0.3	76	0.02	0.15	1.62	TI, MOSFET
LR8	•	-	-	-	•	•	450	0.01	12	0.5	1	1.20	5	10	1	60	0.003	1.4	0.72	Supertex
Negative																				
LM337L	•	•	-	-	-	-	40	0.1	-	5	1	1.25	4	50	0.65	80^g	0.02	0.3	0.65	low-power (neg 317L)
LM337	-	-	•	•	•	•	40	1.5^p	2.0^t	10	1	1.25	3	65	0.6	77^g	0.02	0.3	0.28	negative 317

Notes: (a) all have V_{out} range from V_{ref} to $V_{in(max)}-V_{ref}$. (b) minimum current to operate the IC. (c) ΔV_{out} (%) for $\Delta T_J = 100°C$. (d) D^2PAK. (e) for 10% to 50% I_{max}. (f) at 5V. (g) with V_{adj} bypass cap. (h) maximum dropout voltage at I_{max}. (k) JRC's NJM317F has isolated tab. (n) also with prefixes like TLV, LD, and REF. (p) for TO-220 and D-PAK packages. (u) 10μF min if low ESR tantalum; also requires 10μF input bypass. (v) maximum I_{out} at low $V_{in}-V_{out}$, e.g., ΔV<10V; see text. (z) the metal case or tab (for TO-220, TO-3, D-PAK) is connected to V_{out} for positive regulators, and to V_{in} for negative regulators.
Beware differing pinouts: positive versus negative, and variants like LR8 and LR12.

make up the difference; but then you incur some additional uncertainty of output voltage, owing to the adjustment pin current of ∼50 μA; see **E** below.

B: The standard 317-style regulator circuit (as in Figure 9.10) can adjust only down to V_{ref}. But you can trick a 317 into going down to zero by returning the lower leg of the output divider (R_2) to a negative reference. Be sure to sink enough current to bias that reference into conduction, as shown.

C: You can use a MOSFET switch (or a low R_{ON} analog switch) to shunt additional fixed resistors across the voltage-setting lower resistor, allowing selection of output voltage under logic-level control.

where the value of R_1 is 240 Ω. This design error most likely originated with illustrative circuit examples for the more tightly spec'd LM117 on the same datasheet, whose worst-case $I_{out(min)}$ is 5 mA (half that of the LM317). It's been 40 years, and no one at the factory seems to have noticed!

D: Alternatively, you can program the output voltage by applying a dc voltage via the ADJ pin; the output voltage will be V_{ref} greater. The programming voltage could be generated by a pot, as shown, or by a DAC. If programmed as in the circuit fragment shown, you would need to ensure that the external load satisfies the minimum load current specification (5 or 10 mA for most devices; see Table 9.2). You also need to take into consideration the effect of ADJ pin bias current through the larger than usual impedance, in this example rising to more than 1 kΩ at the pot's mid-position; see **E**, next.

E: The ADJ pin sources ∼50 μA (see the box titled "Anatomy of a 317"), which causes the output voltage to become

$$V_{OUT} = V_{ADJ}\left(1+\frac{R_2}{R_1}\right)+I_{ADJ}R_2, \qquad (9.1)$$

where the last "error" term is caused by the ADJ pin current. For the worst-case I_{ADJ}=100 μA and nominal R_1=125 Ω

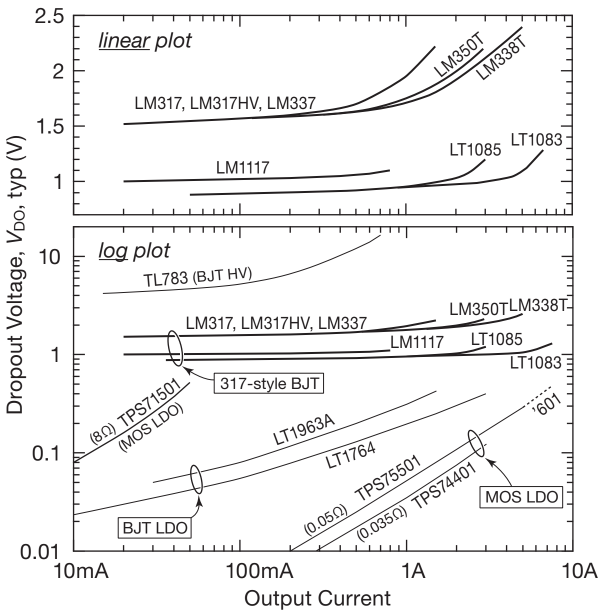


Figure 9.11. Typical dropout voltage (V_{IN}–V_{OUT}) versus load current for 317-style three-terminal regulators (bold curves). Representative low-dropout and high-voltage regulators are included for comparison. See also Figure 9.24.

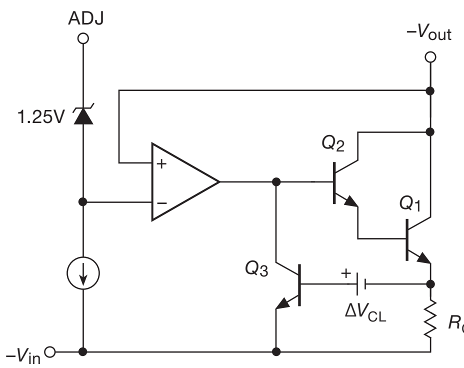


Figure 9.12. LM337 three-terminal adjustable negative voltage regulator. The common-emitter output stage requires at least 1 μF bypassing at input and output to ensure stability.

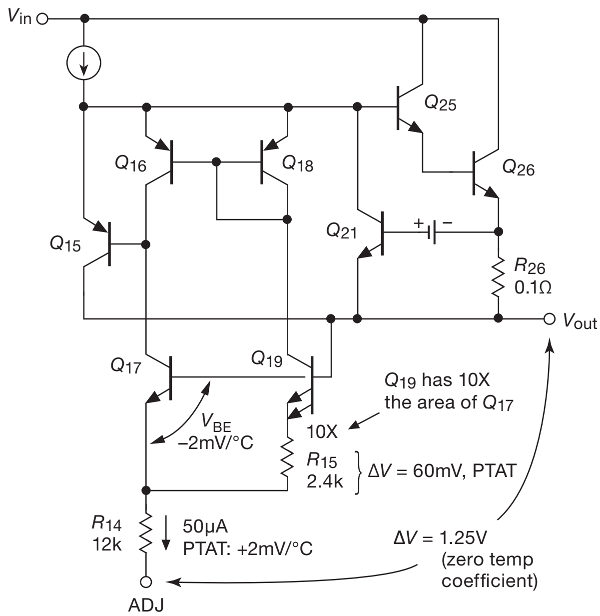


Figure 9.13. Simplified circuit of the 317-style linear regulator, illustrating its internal temperature-compensated bandgap reference; see the box "Anatomy of a 317."

this amounts to a 1% increase in output voltage, above and beyond the initial V_{ref} uncertainty (usually 1% or 4%; see Table 9.2).[17] The current-induced error increases linearly with divider impedance, as indicated in the plot (which assumes a V_{ref} tolerance of 4%, a worst-case ADJ pin current of 100 μA, and no correction for the adjustment current, i.e., ignoring the last term of eq'n 9.1).

F: A linear regulator can be damaged by fault conditions in which bypass capacitors discharge suddenly through the regulator circuit, causing destructive peak currents. Diode D_2 prevents the output bypass capacitor from discharging through the regulator if the input is shorted; it never hurts to include such a diode, and it is definitely indicated for higher output voltages. Likewise, add diode D_1 if an optional noise reduction capacitor C_1 is used, to protect against input or output shorts.

G and **H:** You can extend the output-voltage ramp-up time[18] by bypassing the ADJ terminal with a large value

[17] If you care, you can calculate your resistor values from eq'n 9.1, using the datasheet's *typical* value for I_{ADJ}; that reduces the worst-case error by typically a factor of two (the ratio of maximum to typical I_{ADJ}).

[18] Why would you do that? Perhaps this short story, from our research lab, will provide some motivation: we built a ±15 kV supply, using a pair of Spellman MP15 dc–dc HV converters (+24 V input, +15 kV and −15 kV maximum outputs, 10 W), powered from a commercial ac-powered +24 Vdc switching supply. We mounted the HV output connectors (type SHV; see Figure 1.125) a safe 2″ apart. Imagine our surprise, then, when we powered it up, and a huge spark jumped between the connectors; must have been at least 50 kV! Scary. And *worrisome* – can this thing (and its load) survive repeated startups? Our first attempted cure was to ensure that the MP15's 0–10 V control voltage was set to zero volts at startup. No joy. Finally we added an LT1085 three-terminal regulator with controlled ramp to the +24 V input, and, voilà, no more lightning.

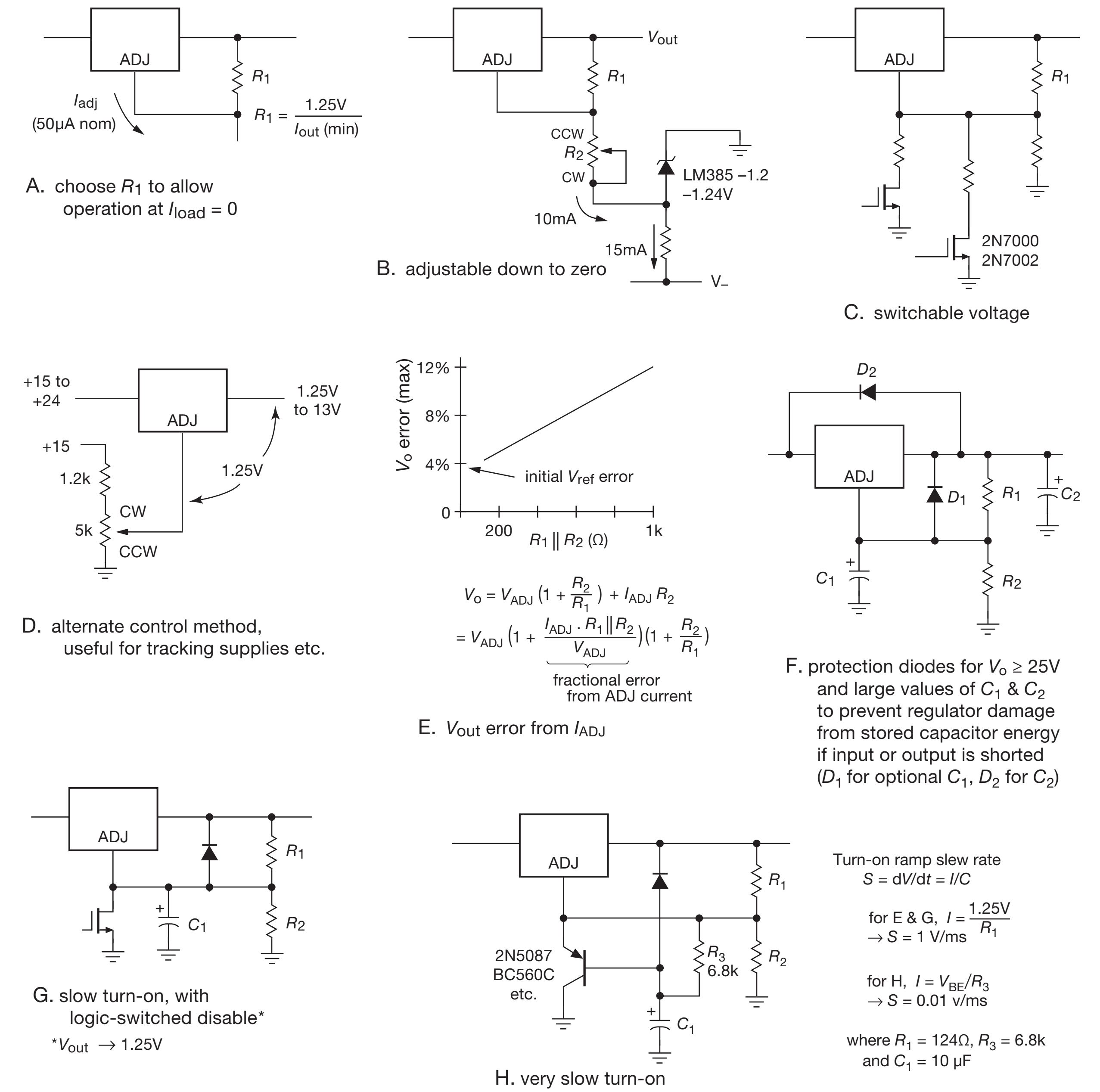


Figure 9.14. Application hints for the LM317-style three-terminal adjustable regulator, described in §9.3.4.

capacitor (be sure to add the protection diode; see **F** above). In both circuits the capacitor ramps up with a constant current, as indicated. Because R_1 is small, the capacitor value can become uncomfortably large (e.g., 100 μF for a 10 ms/V ramp with R_1=125 Ω), so you may want to add a follower, as in **H**. Note that these circuits do not ramp from zero output voltage – in **G** it jumps to V_{ref} (1.25 V) before ramping, and in **H** it jumps initially to $V_{ref}+V_{BE}$ (about 1.8 V). For the same reason the "disable" switch in **G** brings the output down only to V_{ref}.

Exercise 9.5. Draw a circuit (with component values), following the scheme of Figure 9.14C, to power a 12 V (nominal) dc cooling fan within an instrument: when little cooling is needed the circuit should provide +6 V (at which the fan runs, but quietly), but when more cooling is needed a logic-level signal (call it HOT) that is provided to your circuit will go HIGH (i.e., to +5 V) to switch on a MOSFET (as in the figure), at which point your circuit should increase the fan voltage to +12 V.

9.3.5 317-style regulator: circuit examples

Before moving on to the subject of low-dropout regulators, let's take a look at a few useful real-world examples that are easily handled with 317-style 3-terminal adjustable regulators: a 0 to ±25 V dual-tracking adjustable bench supply, a proportional fan speed control, and two ways to create an adjustable high-voltage dc supply.

A. Laboratory dual-tracking bench supply

It's nice to have an adjustable dual supply on the bench, for example with matched ("dual tracking") outputs that go from 0 to ±25 V at currents to 0.5 A. You can buy them for a few hundred dollars or so; but you can make one pretty easily with a pair of 3-terminal adjustable regulators. Figure 9.15 shows how, beginning with the unregulated dc inputs. The positive regulator is an LM317 in a TO-220 package (suffix T), with an appropriately sized heatsink ($R_{\Theta JC}$~2°C/W; see §9.4). To get adjustability down to zero volts we use the trick of Figure 9.14B (output sense divider return to −1.25 V). For the tracking negative output we negate accurately the voltage at U_1's ADJ pin to program the LM337 negative regulator.

A few details: we added a noise-suppression capacitor C_1 to U_1 (along with a protection diode), and we used a precision low-noise op-amp to generate the inverted control voltage (thus no capacitor is needed at U_2's ADJ pin). Resistors R_1 and R_6 provide the regulators' 10 mA minimum load current, but note that op-amp A_2 must be able to source 10 mA, and, likewise, R_5 must sink enough current to power the op-amp, sink the 10 mA through R_2, and bias the zener Z_1. Schottky diodes D_1 and D_2 protect against reverse polarity, for example from a load bridging both rails. Finally, if there is some way in which the dc inputs could be shorted abruptly to ground, diodes should be connected between input and output terminals of each regulator (as in Figure 9.14F), to protect them from the fault current flowing back from the output bypass capacitors (including whatever you've got in the powered external circuit); a couple of 1N4004 rectifiers would be fine here.

Exercise 9.6. Zener reference Z_1 (actually it's a low-current shunt regulator) has a specified current range of 50 μA to 20 mA. Show that this circuit respects those limits, by calculating the zener current at both maximum and minimum negative input voltage (i.e., at −38 V and at −28 V). Assume the dual op-amp's supply current is in the range of 3 mA to 5.7 mA.

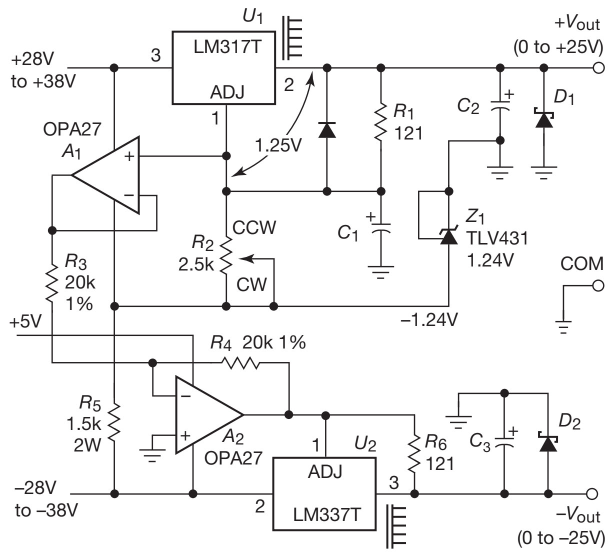


Figure 9.15. Laboratory dual-tracking bench supply, 0 to ±25 V, implemented with 317-style three-terminal linear regulators.

B. Proportional fan control

On–off (or high–low) fan control, as in Exercise 9.5, is simple; but you can do better than "bang–bang" control by tailoring the fan's drive voltage (thus fan speed) to hold the heatsink at a given elevated temperature setpoint. Figure 9.16 shows how to use an LM317T as a power driver, exploiting its internal protection (overtemperature, overcurrent) and its simple ADJ-pin control scheme. Here we've used the forever-popular LM358 op-amp as an integrator of the error signal from a bridge, one leg of which is a negative

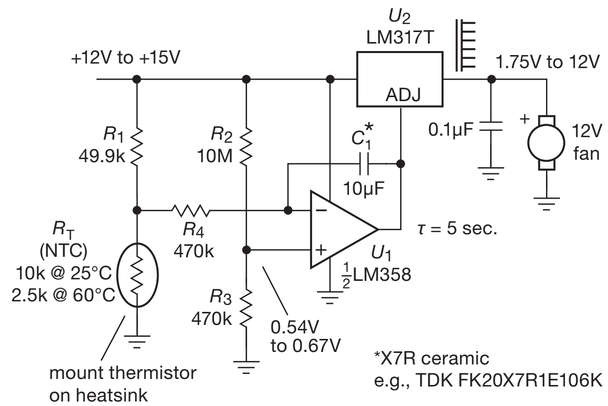


Figure 9.16. Controlling fan speed with analog feedback from a thermistor sensor, with 60°C setpoint. Fully analog control eliminates switching noise, and variable-speed operation minimizes acoustic noise.

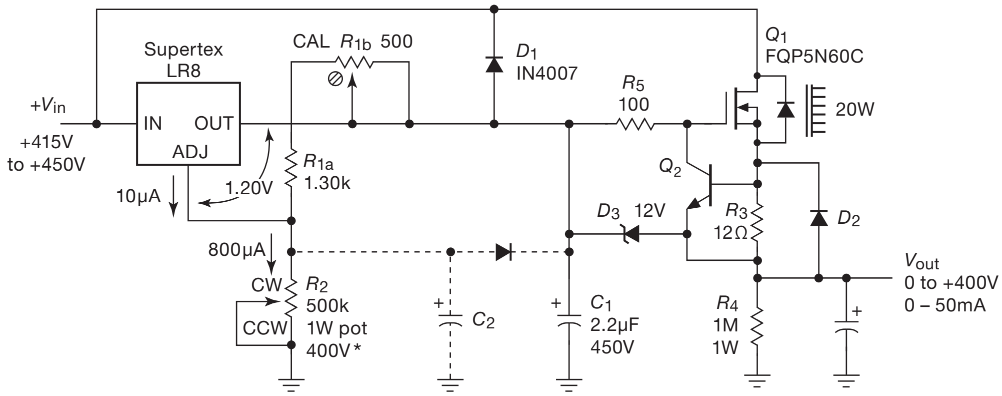


Figure 9.17. High-voltage adjustable supply I, 0 to 400 V, implemented with a low-current LR8 plus outboard MOSFET follower. * See text for R_2's voltage rating.

temperature coefficient (NTC) thermistor. The inputs are balanced at the 60°C setpoint, above which the integrator's output, and therefore the fan voltage, is driven positive. The integrator time constant R_4C_1 should be chosen somewhat longer than the thermal time constant from the heat source(s) to the thermistor sensor, to minimize feedback-loop "hunting."

We chose the LM358 for its low cost (as little as \$0.10 in unit quantity), single-supply operation (input and output to the negative rail), and robust tolerance to supply voltages well beyond +15 V. But its 50 nA worst-case input offset current forced us to use a rather large integrating capacitor. Ideally one would like an inexpensive single-supply op-amp with bias-compensated or FET inputs. Happily, there's the unusual single-supply OPA171, whose offset current is down in the tens of picoamps even at elevated temperature, and which operates over the full supply voltage range of 3 V to 36 V; it costs about a dollar (but you come out ahead, with the less expensive 0.47 μF capacitor that goes with a 10 MΩ resistor R_4).

C. High-voltage supply I: linear regulator

Figure 9.17 shows a simple circuit that extends the output current of the LR8 high-voltage 3-terminal regulator from Supertex. This part operates to 450 V, but it's limited to 10 mA of output current, and further limited by power dissipation (~2 W in the D-PAK power package, depending on circuit-board foil pattern; see Figure 9.45) when operating at nearly its full rated voltage differential.

The regulator's minimum load current is specified as 0.5 mA, so we operate it at slightly less than 1 mA, with its output driving a high voltage power MOSFET follower. Feedback is local to the LR8, so the follower introduces an offset (slightly load dependent) that can be approximately trimmed to zero by R_{1b}. Q_2 provides current limiting, and diodes D_1–D_3 protect against the many injuries that are possible in a high voltage supply. Add capacitor C_2 (along with a protection diode) for noise reduction. The output-setting pot R_2 dissipates 320 mW at maximum output voltage, so it's wise to use a part rated at 1 W or more. Be sure to check the pot's specifications for a *voltage* rating, as well; for this application you could use a Bourns 95C1C-D24-A23 or a Honeywell 53C3500K.

D. High-voltage supply II: dc–dc switching converter

Another approach to generating regulated high-voltage dc is to use a switching dc–dc converter module. These come with a huge range of output voltages (to tens of kilovolts) and with selectable output polarity. You can get a wide variety of these with built-in regulator circuitry, intended to be powered by a low-voltage dc input (+5 V, or +15 V, etc.), and with the output voltage programmed via a variable resistor or by a low-voltage dc programming input. These are handy for generating the bias for a photomultiplier tube, avalanche photodiode detector, channel plate multiplier, or other devices that need a stable high-voltage low-current bias.

A less-expensive approach is to use a bare-bones dc–dc converter module that lacks internal regulation; for these the output voltage scales with input voltage – they are sometimes called "proportional" dc–dc converters. Typical units take a 0–12 V dc input, with output ranges going from as little as 100 V to as much as 25 kV, at power ratings from a fraction of a watt to 10 watts or so. Figure 9.18 shows how to do this, using one of EMCO's 3 W proportional converter modules. The output voltage is controlled by R_3, with the low end of the voltage range set by the LM317's 1.25 V minimum output. The optional current limit modification protects the converter when the output is overloaded.[19] The total cost of the external circuit components here adds up to less than $0.75 in unit quantity – you can't beat that, for cheap!

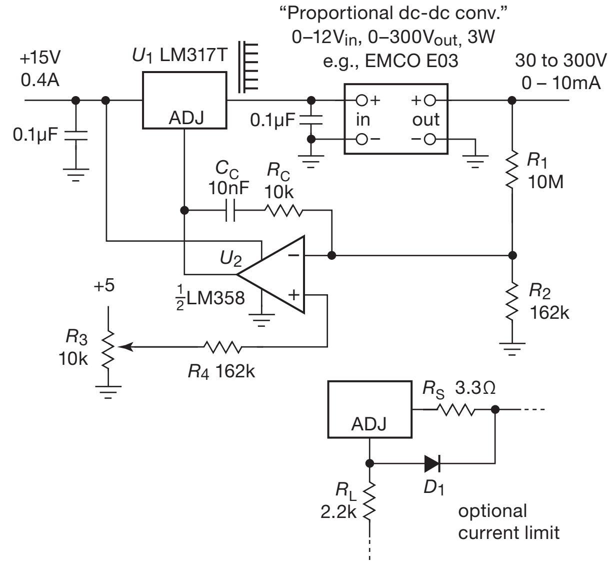


Figure 9.18. High-voltage adjustable supply II, implemented with a proportional dc–dc converter module powered by an LM317 with feedback control to the ADJ pin.

9.3.6 Lower-dropout regulators

There are applications where the ∼2 V dropout voltage (i.e., minimum input–output voltage differential) of these regulators is a serious limitation. For example, in a digital logic circuit you might need to create a +3.3 V supply from an existing +5 V; or (worse), a +2.5 V supply from an existing +3.3 V supply rail. Another application might be a portable device needing +5 V, and operating from a 9 V alkaline battery; the latter begins life at about 9.4 V, declining at end of life to 6 V or 5.4 V (depending on whether you subscribe to 1.0 V/cell or 0.9 V/cell as the definition of a fully exhausted battery). For these situations you need a regulator that can operate with a small input–output differential, ideally down to a few tenths of a volt.

One solution is to abandon linear regulators altogether and use instead a switching regulator (§9.6), which deals differently with voltage differential. Switchers (the nickname for switching, or switchmode, regulators) are popular in such applications; but they have problems of their own (particularly in terms of switching noise and transients), and you may prefer the placid calm, and simplicity, of a linear regulator.

Look again at the conventional linear regulators in Figures 9.6 and 9.9; the ∼2 V dropout is caused by the two cascaded V_{BE} drops of the Darlington output follower, plus another V_{BE} across the current-limiting sense resistor. The solution (inspired by Sziklai's complementary replacement for a conventional Darlington; see §2.4.2A) is to use a different output-stage topology, and a different current-limit scheme.

Figure 9.19 shows a partial solution. This design retains an *npn* output follower, but substitutes a *pnp* driver transistor; the latter can run close to saturation, eliminating one of the V_{BE} drops. Further, the current-limit sense resistor has been relocated in the collector ("high-side current sensing"), where it does not contribute to the dropout voltage (as long as its drop is less than a V_{BE} at current limit, which is easy to manage if a comparator is used to sense maximum current, as shown). With this circuit topology the LT1083-85 regulators (rated at 7.5 A, 5 A, and 3 A, respectively) achieve a typical dropout of 1 V at their maximum current.

We've used this series of regulators in many designs, with good success. Electrically they mimic the classic LM317 three-terminal adjustable regulator, with an internal 1.25 V referenced to the output pin. However, as with most low-dropout designs, these regulators are fussy about bypassing: the datasheet recommends 10 μF at the input, and at least 10 μF (tantalum) or 50 μF (aluminum electrolytic) at the output. If the ADJ pin is bypassed for noise reduction (see §9.3.13), the datasheet recommends tripling the output bypass capacitor value.

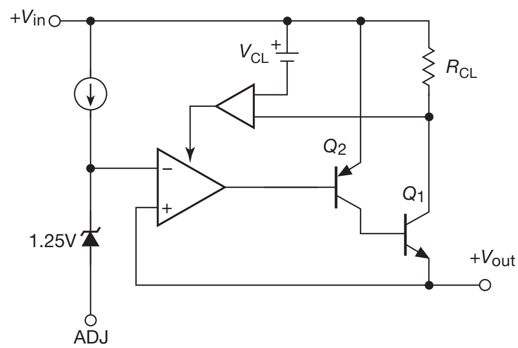


Figure 9.19. LT1083-85 series three-terminal positive regulators with reduced dropout voltage.

[19] We administered a torture test to an unprotected converter and measured an input current of 1.2 A.

9.3.7 True low-dropout regulators

The dropout voltage can be reduced further, by replacing the *npn* (follower) output stage with a *pnp* (common emitter) stage (Figure 9.20A). That eliminates the V_{BE} drop, the dropout voltage now being set by transistor saturation. To keep the dropout voltage as low as possible, the current-limit circuit eliminates the series sense resistor, using instead a fractional output-current sample, derived from a second collector on Q_1. This is less accurate, but "good enough," given that its job is merely to limit destructive currents: the datasheet for the 3 A LT1764A, for example, specifies a current limit of 3.1 A (minimum) and 4 A (typical).[20]

Many contemporary low-voltage regulators use MOSFETs, rather than bipolar transistors. The analogous LDO circuit is shown in Figure 9.20B. Like the bipolar LDO, these parts tend to be rather fussy about bypassing. For example, the TPS775xx regulators set requirements on both the capacitance and ESR of the output bypass: the capacitance must be at least 10 μF (with an ESR no less than 50 mΩ *and* no greater than 1.5 Ω), with the datasheet additionally showing regions of stability and instability in graphs plotting combinations of C_{bypass}, ESR, and I_{out}.

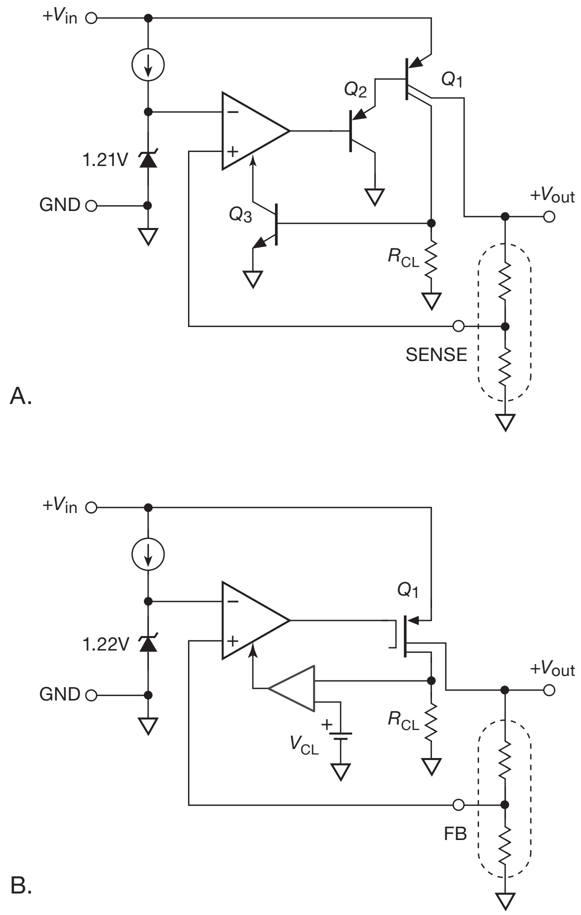


Figure 9.20. A. LT1764 (bipolar) and B. TPS75xxx (CMOS) positive LDO regulators.

9.3.8 Current-reference 3-terminal regulator

All the regulators we've seen so far use an internal voltage reference (usually a 1.25 V "bandgap" reference), against which a voltage-divider fraction of the output is compared. The result is that you cannot have an output voltage less than that reference. In most cases this sets a lower bound of V_{out}=1.25 V (though a few can go down to 0.8 V, or even 0.6 V; see Table 9.3 on page 614).

Sometimes you want a lower voltage! Or you may want to have an adjustment range that goes clear down to zero volts. This has traditionally required an auxiliary negative supply, as for example in the book's previous edition's "laboratory bench supply" (its Figure 6.16).

A nice solution is the LT3080-style regulator, originated by Linear Technology (Figure 9.21). It is a 3-terminal adjustable positive regulator (with a fourth pin, in some package styles) in which the ADJ pin (called SET) sources an accurate *current* (I_{SET}=10 μA, ±2%); the error amplifier then forces the output to follow the SET pin. So, if you connect a resistor R from SET to ground, the output voltage will be simply $V_{\text{out}} = I_{\text{SET}}R$. The output-voltage range goes all the way to zero: when R=0, V_{out}=0.[21]

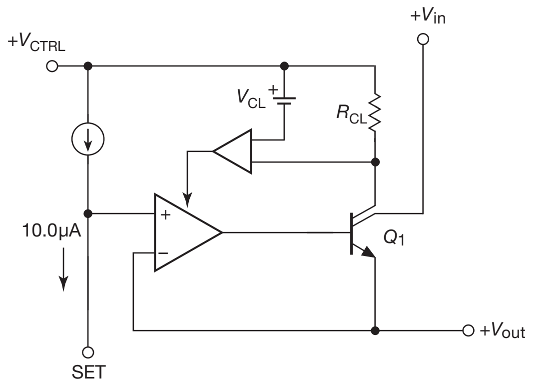


Figure 9.21. LT3080 "3-terminal" adjustable positive voltage regulator with precision current reference.

[20] The CMOS TPS755xx series of 5 A LDO regulators has a revealing current limit specification of 5.5 A (minimum), 10 A (typical), and 14 A (maximum).

[21] With one slight *gotcha*: the minimum load current is ~1 mA. So, for example, the output voltage into a 100 Ω load will not go below 0.1 V. To reach 0 V you need to sink a small current toward a negative supply.

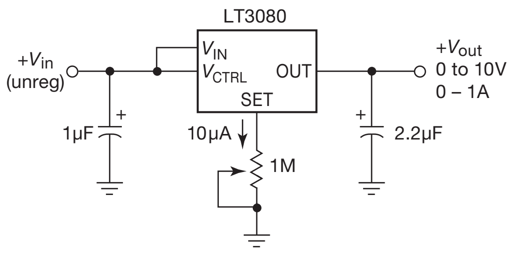


Figure 9.22. Positive regulator, adjustable down to 0 V.

Figure 9.22 shows the basic connection, here used to make a 0–10 V adjustable supply. The 3080-series architecture makes it easy to add an adjustable current limit, itself adjustable down to zero, as shown in Figure 9.23. The upstream regulator U_1 is by itself a 0–1 A current source; the cascaded regulators together act as a current-limited voltage source (or a voltage-limited current source, depending on the load).

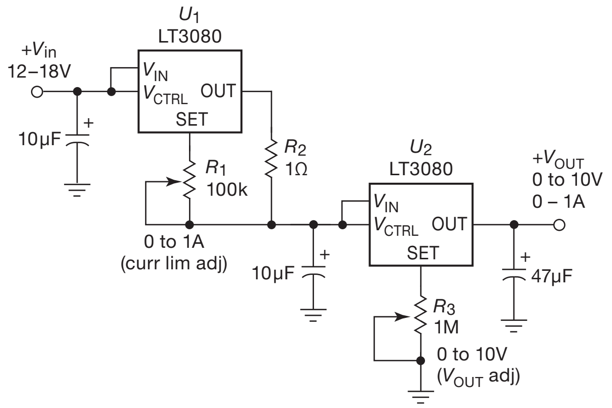


Figure 9.23. Adjustable "bench supply," with independent voltage and current limit controls.

The 3080-style regulator includes a V_{CTRL} pin in the packages with more than three pins (e.g., TO-220-5), which lets you run the internal control circuitry from a higher input voltage. When operated that way, the LT3080 is a true low-dropout voltage regulator, with a typical dropout voltage of 0.1 V at 250 mA load current. Its low impedance (emitter follower) output requires only 2.2 μF output bypassing, with no minimum ESR requirement.

9.3.9 Dropout voltages compared

To summarize the business of dropout voltage in these various regulator designs, we plotted in Figure 9.24 the dropout voltages of a representative regulator of each type. The curves are taken from "typical" dropout specifications in the datasheets, all at 40°C, and are scaled to the rated maximum current of each device. Three categories are clearly seen: conventional regulators with Darlington *npn* pass transistor (top three curves); lower-dropout regulators with *pnp* driver and *npn* output follower (middle curve); and true low-dropout regulators with *pnp* or *p*MOS output stage (bottom four curves). Note particularly the resistive behavior of the CMOS regulators (bottom two curves), where the dropout voltage is linear in output current, and goes to zero at low current.

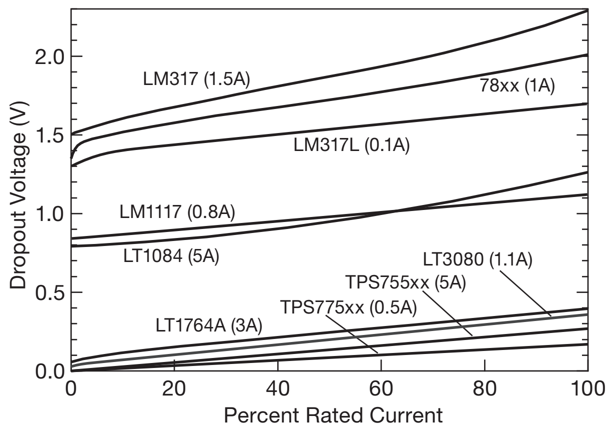


Figure 9.24. Linear regulator dropout voltage versus output current. The bottom pair of curves (TPS prefix) are CMOS; all others are bipolar. See also Figure 9.11.

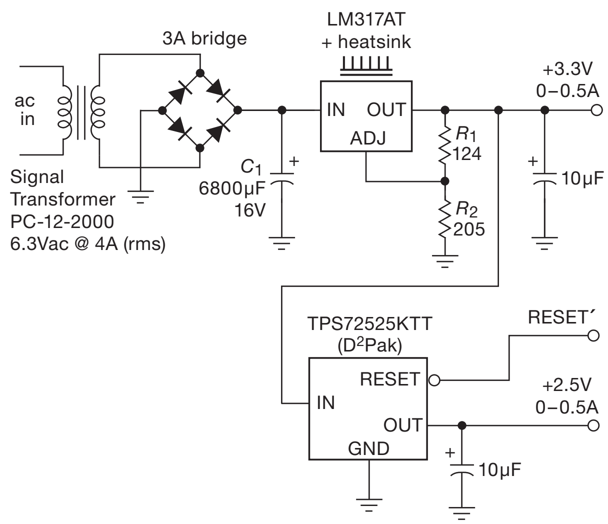


Figure 9.25. Dual low-voltage regulated supply.

9.3.10 Dual-voltage regulator circuit example

As an illustration, imagine we have a small digital circuit requiring +3.3 V and +2.5 V regulated supplies, each capable of supplying up to 500 mA. Figure 9.25 shows how to do this with a small PCB-mounted transformer driving an unregulated bridge rectifier, followed by a pair of linear regulators. The design is straightforward: (a) we began by choosing a transformer to deliver about +8 Vdc (unregulated) from Signal Transformer's nice selection; a unit with 6.3 Vrms is about right (its ac peak amplitude of $6.3\sqrt{2} \approx 9$ V is reduced by two diode drops); (b) we chose a conservative transformer current rating of 4 Arms, to allow for the extra heating caused by the relatively short current pulses in a bridge rectifier circuit ("small conduction angle," see §1.6.5); (c) the storage capacitor C_1 was then chosen (using $I = C dV/dt$) to allow ∼1 Vpp ripple at maximum load current, with a voltage rating adequate to allow for the worst-case combination of high line voltage and zero output load; (d) for the +3.3 V output we used a 3-terminal adjustable regulator (LM317A, in a TO-220 power package), mounted on a small heatsink (10°C/W, adequate for the ∼5 W maximum power dissipation; see §9.4; (e) finally, for the +2.5 V output we used a low-dropout CMOS fixed voltage regulator (the last two digits of the part number designate +2.5 V), taking its input from the regulated +3.3 V.

Several comments. (a) We have not shown details of the ac line input, including fuse, switch, and noise filter; (b) the bypass capacitor values shown are conservative (larger than the specified minimum), in order to improve transient response and provide robust stability; (c) the TPS72525 regulator includes an internal "supervisor" circuit providing a RESET′ output that goes LOW when the regulator falls out of regulation, often used to alert a microprocessor to save its state and shut down.

9.3.11 Linear regulator choices

Fixed or adjustable? 3-terminal or 4-terminal? Low dropout or conventional? How do you decide which type of integrated linear regulator to use? Here is some guidance.

- If you don't need low dropout, stick with a conventional regulator, either 3-terminal fixed (78xx/79xx style, Table 9.1) or 3-terminal adjustable (317/337 style, Table 9.2): they are less expensive, and they're stable with small-value bypass capacitors.
  - fixed: needs no external resistors; but limited voltage choices, and no trimmability.
  - adjustable: settable and trimmable, and fewer types to stock; but require a pair of external resistors.
- If you need adjustability down to zero volts, use a current-reference regulator (LT3080 style).
- If you need low dropout ($V_{\text{DO}} \leq 1$ V) you have many LDO choices (Table 9.3):
  - For input voltages ≥ 10 V, use bipolar types:
    - LT1083-85, LM1117, LM350, LM338 style (fixed or adj) for ∼ 1 V dropout;
    - LT1764A style (fixed or adj) for ∼ 0.3 V (but see §9.3.12);
  - For input voltages ≤ 10 V there are many MOSFET LDOs (fixed or adj).
- If you need high efficiency, high power density, voltage stepup, or voltage inversion, use a switching regulator/converter (§9.6).

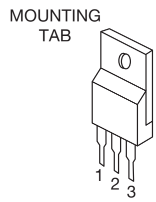
For high-current low-voltage applications, consider the use of a regulator that has separate control and pass-element input pins, like that shown in Figure 9.21. These are indicated in the "boost, bias pin" column of Table 9.3.

9.3.12 Linear regulator idiosyncrasies

These integrated regulators are genuinely easy to use, and, with their built-in overcurrent and thermal protection circuits, there's not much to worry about. Circuit designers should, however, be aware of the following idiosyncrasies.

A. Pinout variations

Our students succumb, with distressing frequency, to this gotcha: complementary polarity regulators, such as our favorite LM317 (positive) and LM337 (negative) 3-terminal adjustables, often have *different pinouts* (Figure 9.26). In the case of the 78xx/79xx fixed regulators, for example, this creates real mischief: the mounting tab is ground for the 78xx positive regulator (so you can screw it to the chassis, or solder it to the circuit board's ground plane), but for the 79xx negative regulator the tab is connected electrically to the input voltage – ground that and you're in real trouble![22]



"TO-220" package

	1	2	3	TAB
LM317	ADJ	OUT	IN	OUT
LM337	ADJ	IN	OUT	IN
78xx	IN	GND	OUT	GND
79xx	GND	IN	OUT	IN

Figure 9.26. Never assume that negative regulators have the same pinouts as their positive twins. In fact, don't assume *anything* without consulting the datasheet.

[22] This comes about because the IC's substrate (normally at the most negative voltage) is soldered to a metal mounting frame, which is the best pathway for heat removal.

Table 9.3 Low-dropout Linear Regulators[a]

Part #[n]	TO-220	DDPAK, DPAK	DIP	SOIC, MSOP, TSOP	SOT-223	SOT-23	TO-92	smaller	adj version avail	# fixed[ti]	Enable	Filter pin	Soft start	UVLO	Reverse block	Sense pin (or func'n)	boost, bias pin	PG or /reset pin	V_{in} min-max (V)	V_{out} (V)	I_{out} (A)	Acc'y 25C (%)	I_Q typ I_{out}=0 (mA)	V_n[b] typ (μVrms)	@V_{out} (V)	Regulation line max (ppm/V)	C_{out} min (μF)	ESR min,max (Ω)	Price[n] qty 25 ($US)
	Packages																												
LM317-style																													
LT3080[A]	5	5	-	8	3	-	-	8	•	-	-	-	•	-	x	-	•	-	1.2-36	0-35	1.1	1	1[h]	40	all	50	2.2	0,0.5	2.71
LT1086	3	3	-	-	-	-	-	-	•	5	-	-	•	-	Δ	-	-	-	3.5-25	1.25-25	1.5	1	10[h]	-	-	1000	22[f2]	?	2.67
LT3083A[A]	5	5	-	16	-	-	-	12	•	-	-	-	•	-	x	-	•	-	1.2-23	0-22	3	1	1[h]	40	all	10	10[c]	0,0.5	5.90
LT1083, 84, 85	3	3	-	-	-	-	-	-	•	-	-	-	•	-	-	-	-	-	2.7-25[g]	1.25-25	3,5,7.5	1	10[h]	150[e2]	5	300	22[f2]	0,-	5.20
LT1580	5,7	-	-	-	-	-	-	-	•	1	-	-	•	-	-	•	•	-	3-5.5	1.25-6	7	0.6	6	-	-	1000	22	0.1	3.74
LT1581	7	-	-	-	-	-	-	-	•	1	-	-	•	-	-	•	•	-	3.3-6	1.25-5	10	0.7	10[h]	-	-	300	22	0,1	8.93
High Voltage (≥10V)																													
TPS709xx	-	-	-	-	-	5	-	-	-	11	•	-	-	•	•	-	-	-	2.5-30	1.5-6.5	0.15	2[tt]	0.001	300	1.8	650	2.2	0,-	1.16
LT3008	-	-	-	-	-	8	-	6	•	6	•	-	-	-	□	•	-	-	2-45[o]	0.6-45	0.02	1	0.003	92	0.6	250	2.2	0,3	2.66
TLV704xx	-	-	-	-	-	5	-	-	-	3	-	-	-	-	Δ	-	-	-	2.5-24	3-5	0.15	2	0.003	-	-	500	1	0,-	0.81
LT3014B[B]	-	-	-	-	-	5	-	8	•	-	•	-	-	-	□	•	-	-	3-80[o]	1.22-60	0.02	1.6	0.007	115	1.22	125	0.47	0,3	2.14
LT1521	-	-	-	8	3	-	-	-	•	3	•	-	-	-	□	•	-	-	3.5-20	3-20	0.3	1.5	0.012	-	-	125	1.5	0,?	2.08
LM2936	-	3	-	8	3	-	3	-	-	3	q	-	-	-	□	-	-	-	4-26[o]	3.0-5	0.05	2	0.015	500	all	1000	22	0.3,6	1.03
TPS769, 70xx	-	-	-	-	-	5	-	-	•	9	•	-	-	-	Δ	v	-	-	2.7-10	1.2-5.5	0.05, 0.1	3[tt]	0.017	190[e3]	all	400	4.7	0.2,5	0.77
TPS789, 90xx	-	-	-	-	-	5	-	-	-	5	•	•	-	-	Δ	-	-	-	2.7-10	1.5-3.0	0.1, 0.05	3[tt]	0.017	56[f,e3]	3	1000	4.7	0.2,5	0.56
TPS788xx	-	-	-	-	-	5	-	-	-	2	•	•	s	-	Δ	-	-	-	2.7-13.5	2.5,3.3	0.15	3[tt]	0.017	56[f,e1]	all	1000	4.7	0.2,5	1.09
ADP667	-	-	8	8	-	-	-	-	•	1	•	-	-	-	x	v	-	•	3.5-16.5	1.3-16	0.25	4[tt]	0.02	-	-	500	10	0,-	2.98
LT1761	-	-	-	-	-	5	-	-	•	9	•	v2	-	-	□	-	-	-	1.8-20	1.22-20	0.1	1.3	0.02	20[f]	all	600	1	0.3	2.13
LM9076	-	5	-	8	-	-	-	-	-	2	q	-	-	-	□[y]	-	-	r	3.5-50[o]	3.3-5	0.15	1.5	0.025	-	-	1500	22[c]	0.1,3	1.31
LT1763	-	-	-	8	-	-	-	12	•	6	•	•	-	-	□	•	-	-	1.8-20	1.22-20	0.5	1	0.03	20[f]	all	50	3.3	0,3	2.30
ADP3331	-	-	-	-	-	6	-	-	•	-	•	-	-	-	x	•	-	•	2.6-12	1.5-11.75	0.2	0.7	0.034	95	3	240	0.47	0,-	1.25
TPS765, 66xx	-	-	-	8	-	-	-	-	•	8	•	-	-	-	Δ	v	-	p	2.7-10	1.2-5.5	0.15, 0.25	3[tt]	0.035	200[e3]	all	100[t]	4.7	0.3,20	1.50
TPS760, 61xx	-	-	-	-	-	5	-	-	-	5	•	-	-	-	Δ	-	-	-	3.5-16	3-5	0.05, 0.1	1	0.09	190[e3]	all	300	2.2	0.1,20	0.90
TPS763, 64xx	-	-	-	-	-	5	-	-	z	9	•	z	-	-	Δ	z	-	-	3.3-10	1.5-6.5	0.15	2	0.09	140[e3]	all	700	4.7	0.3,10	0.38
LP2950A[C]	-	3	-	-	-	-	3	-	-	3	-	-	-	-	x	-	-	-	3.1-30	3.0-5.0	0.1	0.5	0.075	430	all	35	2.2	0,5	***0.37***
LP2951A[C]	-	-	8	8	-	-	-	8	•	2	•	v2	-	-	x	•	-	•	1.5-30	1.4-29	0.1	0.5	0.2 m	100[f]	all	40	2.2	0,5	***0.29***
LP2981A[C]	-	-	-	-	-	5	-	-	-	9	•	-	-	-	Δ	-	-	-	2.1-16	2.5-5	0.1	0.75	0.095	160[e3]	all	140	3.3[c]	0.01,8	0.84
LP2985A	-	-	-	-	-	5	-	-	-	9	•	•	-	-	Δ	-	-	-	2.2-16	1.8-10	0.15	1	0.065	30[f]	all	140	2.2	0.005-0.5	0.82
LM2931	3	3	-	8	-	3	3	6	•	2	-	-	-	-	□	v	-	-	3.3-26[o]	3.0-24	0.1	5	0.4	500	5	1500	47	0.03,0.4	***0.65***
LM2930	3	3	-	-	-	-	-	-	-	2	-	-	-	-	□[y]	-	-	-	5.3-26[o]	5.0-8.0	0.15	6	4	140	5	800	10	0,1	0.65
LP2986A	-	-	-	8	-	-	-	-	•	3	•	v2	-	-	Δ	•	-	•	2.5-16	1.23-15	0.2	0.5[d]	0.1	160[e3]	all	140	4.7[c]	0.1,10	2.32
TPS7A4901	-	-	-	8	-	-	-	-	•	-	•	•	•	-	x	v	-	-	3-36	1.19-33	0.15	1.5	0.061	21[f]	5	32[t]	2.2[c]	0,0.2	2.14
TL750Lxx	3	3	8[y]	8	-	-	3	-	-	4	-	-	-	-	□[y]	-	-	-	6-26[o]	5,8,10,12	0.15	4	1.0	700	10	300	10	0,0.4	0.62
TPS72xx	-	-	-	8	-	-	-	-	•	5	•	-	-	-	Δ	•	-	p	3-10	1.2-10	0.25	2[tt]	0.18	300	5	1400	10	0.1,1.3	2.00
REG101, 102	-	-	-	8	5[y]	5	-	-	•	6	•	d	-	-	x	-	-		1.8-10	2.5-5.5	0.1, 0.25	1.5	0.4	18[f]	2.5	1000	0	-	2.22
REG103, 04	-	5	-	8[y]	5	-	-	-	•	5	•	d	-	-	x	-	-	•	2.1-15	1.3-5	0.5, 1	2	0.5	25[f]	2.5	1000	0	-	3.78
REG113	-	-	-	8	-	5	-	-	-	5	-	•	-	-	x	-	-		1.8-10	2.5-5	0.4	1.5	0.4	18[f]	2.5	1000	0	-	1.41
ADP3303	-	-	-	8	-	-	-	-	-	5	•	•	-	-	x	-	-	•	3.2-12	2.7-5.0	0.2	0.8	0.25	100	5	100[t]	0.47	0,-	2.50
MIC5205, 07	-	-	-	-	-	5	-	-	•	13[y]	•	d	-	-	□	-	-	-	2.5-16	1.24-15	0.15	1	0.08	260[nV]	all	120	2.2	0,5	0.41
MIC5209	-	5	-	8	3	-	-	8	•	7	q	d	-	-	□	-	-	-	2.5-16	1.24-15	0.5	1	0.08	300[nV]	all	500	2.2	1 rec	0.70
MIC5219	-	-	-	8	-	5	-	6	•	11	•	d	-	-	□	-	-	-	2.5-12	1.24-12	0.5	1	0.08	300[nV]	all	500	2.2	0.1-1	1.50
LP38691, 93	-	3[y]	-	-	4[y]	-	-	6	•	4	z	-	-	-	Δ	q	-	-	2.7-10	1.8-5	0.5	2	0.055	70	3.3	1000	1[c]	0,100	0.90
LT3021	-	-	-	8	-	-	-	16	•	3	•	-	-	-	□	•	-	-	0.9-10	0.2-9.5	0.5	2	0.11	300	1.2	650	3.3[c]	0,0.2	3.13
ADP3334	-	-	-	8	-	-	-	8	•	-	•	-	-	-	x	•	-	-	2.6-11	1.5-10	0.5	0.9	0.09	45	all	200	1	0,-	1.34
ADP3335, 6[D]	-	-	-	8	-	-	-	8	•	5	•	v2	-	-	x	-	-	-	2.6-12	1.5-10	0.5	0.9	0.08	47[f]	all	400[t]	1	0,-	2.13
ADP7102, 04	-	-	-	8	-	-	-	8	•	7	•	-	•	•	•	•	-	•	3.3-20	1.22-20	0.3, 0.5	0.8	0.4	15	3.3	150	1	0,0.2	3.22
TPS73xx	-	-	8	8	-	-	-	-	•	5	•	-	-	-	x	•	-	r	2.5-10	1.2-9.8	0.5	3	0.34	89	1.2	2000	10	0.1,4	2.25
LM2937	3	3	-	-	3	-	-	-	-	5	-	-	-	-	□	-	-	-	7-26[o]	5-15	0.5	3	2.0	150	5	2500	10	0,3	0.76
LP3878	-	-	-	8P	-	-	-	8	•	-	•	•	-	-	Δ	•	-	-	2.5-16	1.0-5.5	0.8	1	0.18	18[f]	all	140	10[c]	0,0.4	1.44
TPS775xx-78xx	-	-	-	8	-	-	-	-	•	6	•	-	-	-	Δ	v	-	r,p	2.7-10	1.2-5.5	0.5, 0.75	2[tt]	0.085	53[e1]	all	100[t]	10	0.05,1.5	2.44
TPS767, 68xx	-	-	-	8	-	-	-	-	•	8	•	-	-	-	Δ	v	-	r,p	2.7-10	1.2-5.5	1.0	2[tt]	0.085	55[e1]	all	100[t]	10	0.05,1.5	2.77
LM2940	3	3	16	-	-	-	-	8	-	6	-	-	-	-	□	-	-	-	7-26[o]	5-15	1.0	3	10	150	5	2000	22	0.1,1	1.71
TPS73801	-	-	-	-	5	-	-	-	•	-	•	-	-	-	•	v	-	-	2.2-20	1.2-20	1.0	1.5	1.0	45	1.2	230	10	0,0.3	2.34
LT1965	5	5	-	8	-	-	-	8	•	4	•	-	-	-	□	•	-	-	1.8-20	1.2-19	1.1	1.5	0.5	40		600	10	0,3	2.77
UCC381	-	-	-	8	-	-	-	-	•	2	•	-	v	•	•	•	-	-	3-9	1.2-8	1.0	1.5	0.4	-	-	150	0	-	4.61
TPS7A45xx	-	5	-	-	5	-	-	-	•	4	•	-	-	-	•	•	-	-	2.2-20	1.2-20	1.5	1	1.0	35	all	140	10	0,3	3.12
LT1963A[E]	5	5	-	8	3	-	-	-	•	4	•	-	-	-	□	•	-	-	1.9-20	1.21-20	1.5	1.5	1.0	40	all	240	10	0.02,3	3.57
LT1764A	3	3	-	16P	-	-	-	-	•	4	•	-	-	-	□	•	-	-	2.7-20	1.2-20	3	1.5	1.0	40	all	600	10	0.005,3	4.97
UCC383	3,5	5	-	-	-	-	-	-	•	2	•	-	v	•	•	v	-	-	3-9	1.2-8	3	1.5	0.4	-	-	400	0	-	4.32

	Packages																												
Part #[n]	TO-220	DDPAK, DPAK	DIP	SOIC, MSOP, TSOP	SOT-223	SOT-23	TO-92	smaller	adj version avail	# fixed[ti]	Enable	Filter pin	Soft start	UVLO	Reverse block	Sense pin (or func'n)	boost, bias pin	PG or /reset pin	V_{in} min-max (V)	V_{out} (V)	I_{out} (A)	Acc'y 25C (%)	I_Q typ I_{out}=0 (mA)	V_n[b] typ (μVrms)	@V_{out} (V)	Regulation line max (ppm/V)	C_{out} min (μF)	ESR min,max (Ω)	Price[n] qty 25 ($US)
<u>Low Voltage (≤8V)</u>																													
ADP121	-	-	-	-	-	5	-	4	•	8	•	-	-	•	x	v	-	-	2.3-5.5	1.2-3.3	0.15	1	0.01	52	2.5	300	0.7	0,1	0.87[u]
ADP150	-	-	-	-	-	5	-	4	-	8	•	-	-	•	x	-	-	-	2.2-5.5	1.8-3.3	0.15	1	0.01	9	2.5	500	1	0,1	0.80
ADP130[F]	-	-	-	-	-	5	-	-	-	5	•	-	-	•	x	-	•	-	1.2-3.6	0.8-2.5	0.35	1	0.025	61	2.5	1000	1	0,1	0.85
LTC1844	-	-	-	-	-	5	-	-	•	5	-	•	-	-	•	-	-	-	1.6-6.5	1.25-6	0.15	1.5	0.035	65[f]	all	2000	1	0,0.3	1.45[k]
MAX8510	-	-	-	-	-	-	-	5,8	•	10	•	d	-	-	x	-	-	-	2-6	1.45-6	0.12	1	0.04	11[f]	all	10[t]	1[c]	0,0.3	0.58[k]
ADP122,23	-	-	-	-	-	5	-	6	•	10	•	-	-	•	x	v	-	-	2.3-5.5	0.8-5	0.3	1	0.045	45	2.5	500	0.7	0,1	0.88
ADP124,25	-	-	-	8	-	-	-	8	•	8	•	-	-	•	x	z	-	-	2.3-5.5	0.8-5	0.5	1	0.045	45	2.5	500	0.7	0,1	1.14
MAX8867-68	-	-	-	-	-	5	-	-	-	8	•	•	-	-	□	-	-	-	2.5-6.5	2.5-5	0.15	1.4	0.085	30[f]	all	1500	1[c]	0,2	0.88[k]
MIC5255	-	-	-	-	-	5	-	-	-	12	•	•	-	-	Δ	-	-	-	2.7-6	2.5-3.5	0.15	1	0.09	30[f]	all	500	1[c]	0,0.3	0.39[k]
TPS731xx	-	-	-	-	-	5	-	-	•	9	•	d	-	-	•	v	-	-	1.7-5.5	1.2-5	0.15	0.5	0.4	21[f]	2.5	100[t]	0	-	1.08
TPS791,2,3xx	-	-	-	-	-	5,6	-	-	•	5[y]	•	•	-	•	Δ	v	-	-	2.7-5.5	1.2-5.2	0.1, 0.2	2[tt]	0.17	16[f]	all	1200	1[c]	0.01,1	1.15
NCP700B	-	-	-	-	-	5	-	6	-	4	•	•	-	-	x	-	-	-	2.6-5.5	1.8-3.3	0.2	2.5[tt]	0.07	10[f]	1.8	1000	1	0,1	0.33
TPS730xx	-	-	-	-	-	5,6	-	-	•	7	•	•	-	•	Δ	-	-	-	2.7-5.5	1.2-5.5	0.2	2[tt]	0.17	33[f,e1]	all	500	2.2[c]	0.01,1	0.65
MIC5249	-	-	-	8	-	-	-	-	-	6	•	•	-	-	x	-	-	•	2.7-6	1.8-3.3	0.3	1	0.085	-	-	3000	2.2[c]	0,0.3	0.98
TPS732, 36xx	-	-	-	-	5	5	-	8	•	8	•	d	-	-	•	v	-	-	2.2-5.5	1.2-5.5	0.25, 0.4	0.5	0.4	21[f]	2.5	100[t]	0	-	1.41
TPS794, 95xx	-	-	-	8	5	-	-	-	•	5	•	d,w	-	-	Δ	v	-	-	2.7-5.5	1.2-5.5	0.25, 0.5	3[tt]	0.17	33[f]	all	1200	1[c]	0.01,1	1.88
LP3879	-	-	-	8P	-	-	-	8	-	2	•	•	-	-	Δ	•	-	-	2.5-6	1.0-1.2	0.8	1	0.2	18[f,e1]	all	140	10[c]	0,0.2	1.26
LP3961, 64	5	5	-	-	3	-	-	-	•	4	•	-	-	-	Δ	z	-	z	2.5-7	1.2-5.5	0.8	1.5	3	150	all	30[t]	33[c]	0.2,5	1.83
TPS725xx	-	5	-	8	5	-	-	-	•	4	•	-	-	•	Δ	v	-	r	1.8-6	1.2-5.5	1.0	2[tt]	0.075	-	-	1500	0[c]	-	2.77
TPS726xx	-	5	-	-	5	-	-	8	-	5	•	-	-	•	Δ	-	-	p	1.8-6	1.2-2.5	1.0	2[tt]	0.075	-	-	1500	0[c]	-	3.37
TPS737xx[G]	-	-	-	-	5	-	-	8	•	3	•	d	-	-	•	v	-	-	2.2-5.5	1.2-5.5	1.0	1	0.4	68	2.5	100	1	0,-	1.64
TPS796xx	-	5	-	-	5	-	-	8	•	7	•	d	-	•	Δ	v	-	-	2.7-5.5	1.2-5.5	1.0	2[tt]	0.27	41[f,e1]	all	1200	1	0.01,1	2.44
LP3891, 92	5	5	-	8P	-	-	-	-	-	3	•	-	-	-	x	-	•	-	1.3-5.5	1.2-1.8	0.8, 1.5	1.5	4	150	1.8	100[t]	10	0.01,4	2.25
ADP3338, 39	-	-	-	-	3	-	-	-	-	7	-	-	-	-	x	-	-	-	2.7-8	1.5-5	1, 1.5	0.8	0.11	95	all	400[t]	1	0,-	2.30
LP3871 - 76	5	5	-	-	5	-	-	-	-	4	•	-	-	-	Δ	z	-	z	2.5-7	1.22-5.5	0.8,1.5,3	1.5	5	150	2.5	40[t]	10[c]	0.1,4	2.23
LP3881,2,3	5	5	-	8P	-	-	-	-	-	3	•	•	-	•	x	-	•	-	1.3-5.5	1.2-1.8	0.8,1.5,3	1.5	4	150	all	100[t]	4.7[c]	0.01,4	4.23
LP38851,2,3	7	7	-	8P	-	-	-	-	•	-	•	-	•	•	•	•	•	-	3.0-5.5	0.8-1.8	0.8,1.5,3	1.5	10	150	0.7	400	10	0,-	1.77
LP3855	5	5	-	-	5	-	-	-	•	-	•	-	-	-	Δ	•	-	-	2.5-7	1.22-5.5	1.5	1.5	3	150	2.5	35[t]	10[c]	0.1,5	3.86
TPS786xx	-	5	-	-	5	-	-	-	•	5	•	d	-	•	Δ	v	-	-	2.7-6	1.2-5.5	1.5	2[tt]	0.26	49[f,e1]	all	1000[t]	1[c]	0.01,0.5	4.69
XRP6272	-	5	-	8P	-	-	-	-	•	1	•	•	•	-	-	-	-	•	0.7-5	1.8-6	2	2	0.03	24[f]	all	3000	10[c]	0,-	1.19
LP38855 - 59	5	5	-	-	-	-	-	-	•	-	z	-	z	•	•	-	•	-	3.0-5.5	0.8-1.2	1.5, 3	1	10	150	all	1000[t]	10[c]	0,-	1.53
LP3853, 56	5	5	-	-	-	-	-	-	-	4	•	-	-	-	Δ	z	-	z	2.5-7	1.8-5	3	1.5	4	150	2.5	40[t]	10[c]	0.2,5	4.40
TPS744xx	-	7	-	-	-	-	-	25	•	-	•	-	•	•	Δ	v	•	-	0.9-5.5	0.8-3.6	3	1[tt]	2	40[e1]	2.5	500	0	0,-	4.58
TPS755 -59xx	5	5	-	-	-	-	-	-	•	4	•	-	-	•	Δ	v	-	d,z	2.8-6	1.2-5.5	3,5,7.5	2	0.125	35[e3]	1.5	1000	47[c]	0.2,10	6.27
TLE4275	5	5	-	14P	-	-	-	-	•	4	•	-	-	-	□	v	-	•	6-42	1.2-5	5	2[tt]	0.275	-	-	600	22[c]	0.3,5	***1.14***
<u>Negative LDOs</u>																													
MAX664	-	-	8	8	-	-	-	-	•	1	•	-	-	-	x	•	-	-	2-16	1.3-16	0.05	5[tt]	0.006	-	-	3500	0	-	2.78
LT1964	-	-	-	-	-	5	-	8	•	1	•	q	-	-	□	•	-	-	2.8-20	1.22-20	0.2	1.5	0.03	30[f]	all	3000	1	0,3	2.85
TPS7A3001	-	-	-	8P	-	-	-	-	•	-	•	•	-	-	x	•	-	-	3-35	1.18-33	0.2	1.5	0.055	18[f]	5	60	2.2[c]	0,-	3.89
MAX1735	-	-	-	-	-	5	-	-	•	3	•	-	-	-	x	•	-	-	2.5-6.5	1.25-5.5	0.2	1	0.085	160[e4]	all	1500	1	0,0.1	2.33
TPS723xx	-	-	-	-	-	5	-	-	•	1	•	d	-	-	x	d	-	-	2.7-10	1.2-10	0.2	1	0.13	60	all	400[t]	2.2	0,-	2.57
LT1175	5	5	8	8	3	-	-	-	•	-	•	-	-	-	x	•	-	-	4-20	3.8-18	0.5	1.5	0.045	-	-	150	0.1	0,10	3.78
UCC384	-	-	-	8	-	-	-	-	•	2	•	-	-	-	x	d	-	-	3.2-15	1.25-15	0.5	1.5	0.2	200[e2]	5	200	4.7	0,-	4.82
TPS7A3301	7	-	-	-	-	-	-	20	•	-	•	•	•	-	x	-	-	-	3-35	1.2-33	1	1.5	0.21	16[f]	1.2	60	10[c]	0,-	4.95
LM2991	5	5	-	-	-	-	-	-	•	-	•	-	-	-	x	•	-	-	3-25	3-24	1	2	0.7	200	3	400	1	0.025,10	1.88
LM2990	3	3	16	-	-	-	-	-	-	4	-	-	-	-	x	-	-	-	6-25	5-15	1	2	1	250	5	2000	10	0.025,10	1.88
LT3015	5	5	-	12	-	-	-	8	•	6	•	-	-	-	□	•	-	-	1.8-30	1.22-29	1.5	1	1.1	60	all	800	10	0,0.5	4.81
LT1185	5	5	-	-	-	-	-	-	•	-	-	-	-	-	x	-	-	-	4-30	2.5-25	3	1	2.5	-	-	100	2	0.1,2	4.40

Notes: (a) "Low Dropout" (LDO) regulators typically drop <100mV, except at the highest currents; sorted approximately by ascending I_Q and I_{out} capability, with some family grouping; the ADJ versions of some parts have less accuracy than the fixed versions (which can be laser trimmed). (b) 10Hz-100kHz, with no noise reduction cap, unless indicated otherwise. (c) input and other capacitors also required. (d) fixed voltage versions only. (e1) 100Hz-100kHz or 200Hz-100kHz. (e2) 10Hz-10kHz. (e3) 300Hz-50kHz. (e4) 10Hz-1MHz. (f) with NR cap. (f2) if the ADJ pin is bypassed, larger output capacitors are required, see datasheet. (g) output-input differential. (h) minimum load current. (k) qty 1k or more. (m) min or max. (n) ***bold italic*** indicates an inexpensive "jellybean." (nV) nV/√Hz, with 10nF filter cap. (o) "load-dump": withstands input voltage transients (or continuous inputs) to 60V (or 40V for some parts), and does not pass the spike through to the output; often associated with "automotive" parts. (p) power good. (q) some pkgs only. (r) delayed reset output for μC. (rec) recommended. (s) slew control. (t) typical. (ti) some of TI's parts have an internal EEPROM, allowing quick custom factory voltage programming. (tt) over operating temperature range. (u) unit qty. (v) adjustable version only. (v2) filter cap across feedback resistor. (w) ADJ and filter avail in 8-pin pkg. (x) reversed input voltage with excess current flow may damage part, a Schottky diode is recommended; although not marked "**Δ**," these parts may have a reverse-diode function. (y) check datasheet. (z) choose between two versions.
Δ = input-output back diode conducts, limit current; also, reverse supply not allowed, see note x. □ = protected against reverse polarity input.

Comments: **A:** current-programmed; $V_{control}$ dropout 1.6V max; 1mA min load. **B:** 100V for -HV version. **C:** -C suffix for looser tolerance and tempco, etc. **D:** ADP7104 suggested alternative. **E:** see also TI's TL1963A, $2.65. **F:** control and pass-transistor inputs. **G:** internal charge pump.

B. Polarity and bypassing

As we mentioned earlier, the negative versions of common positive regulators have a different output topology (an *npn* common emitter stage), and require larger bypass capacitors to prevent oscillation. Always "go by the book" (the datasheet, that is) – don't assume you know better. Also, be careful to connect the bypass capacitors with correct polarity (and see next).

C. Reverse polarity protection

An additional caution with dual supplies (whether regulated or not): 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 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. Even in the absence of a fault condition like this, asymmetrical loads can cause a polarity reversal when power is turned off. For these reasons it is wise to connect a power rectifier (preferably Schottky) in the reverse direction from each regulated output to ground, as we drew in Figure 9.8.

Some regulator ICs are designed to block any current flow if the input voltage is lower than the output; these are marked with a bullet dot symbol (•) in the "reverse block" column in Table 9.3. Other regulator ICs go further and also block current flow for reversed *input* polarity; these are marked with a square symbol (□).

D. Ground-pin current

A particular idiosyncrasy of bipolar low-dropout regulators with *pnp* output stages (Figure 9.20) is the sharp rise in ground-pin current when the regulator is close to dropout. At that point the output stage is near saturation, with greatly reduced beta, and therefore requires substantial sinking base current. This is particularly noticeable when the regulator is lightly loaded, or unloaded, when it would otherwise have only a small ground-pin or quiescent current. As an example, the bipolar LT1764A-3.3 (fixed 3.3 V LDO), driving a 100 mA load, has a normal ground-pin current of about 5 mA, rising to ∼50 mA at dropout. The no-load quiescent current shows similar behavior, spiking to ∼30 mA from its normal ∼1 mA.[23] Manufacturers rarely advertise this "feature" on the front page of their datasheets, but you can find it inside, if you look. It is of particular importance in battery-operated devices. *A caution*: the quiescent ground-pin currents listed in the I_Q column of Table 9.3 are for a light load and with the input voltage above dropout.

E. Maximum input voltage

Table 9.3 lists the maximum specified input voltage for more than a hundred LDO linear regulators. CMOS regulators are good choices for low-voltage designs, and they come in a dazzling variety of fixed voltages (and, of course, adjustable versions): for example, the TPS7xxxx series from Texas Instruments includes a dozen types, each of which comes in a choice of 1.2, 1.5, 1.8, 2.5, 3.0, 3.3, or 5.0 volt outputs. But be careful, because many of these CMOS regulators have a specified maximum input voltage of just +5.5 V.[24] Some CMOS regulators, however, accept up to +10 V input; for higher input voltage you've got to use bipolar regulators, for example the LT1764A or LT3012, with input voltage ranges of +2.7–20 V and

[23] The load-induced additional ground current would normally be I_{load}/β driving the base of the *pnp* pass transistor, but during dropout the ever-zealous feedback loop provides a base drive appropriate for the LDO's maximum load-current rating. Some designs carefully limit this drive current, whereas others detect the saturation condition and limit the current accordingly. You've got to pay close attention to this behavior of candidate LDOs when designing battery operated devices, if you want to maximize remaining operating time after the battery voltage has dropped below the LDO's input criteria. Alternatively you may want to select an LDO that uses a *p*-channel MOSFET pass transistor and does not exhibit increased ground current at high loads or during dropout. For example, a 5 V regulator like the LT3008-5 runs at 3 μA, but this soars to 30 μA if the battery drops below 5 V. However, a TLV70450 (with internal *p*MOS pass transistor) suffers no increase at all, continuing to draw 3 μA under the same conditions.

[24] Contemporary regulator ICs, with their nanometer-scale features, are more susceptible to overvoltage transients and the like, as compared with robust legacy parts with their relatively large bipolar transistors. We've seen painful experiences, for example when a carefully designed and tested small PCB, filled to the brim with tiny parts, suffers unexplained failures in the field or at the customer's test sites. Sometimes this is due to uncontrolled (and arguably improper) user-supplied input transients. Adding a transient voltage suppressor (see Chapter *9x*) at the input is a wise precaution for regulator ICs powered from an off-card dc source. Low-voltage ICs (such as those with 6 V or 7 V absolute-maximum ratings) are best powered from an on-card regulated 5 V, etc., rather than from external sources of power. (An exception might be made for a 3.7 V Li-ion cell.) Be very careful!

4–80 V, respectively. See §9.13.2 for an interesting way to extend the input-voltage range, to as much as +1 kV!

F. LDO Stability

It's worth repeating that low-dropout regulators can be quite fussy about bypassing (see the comments in §9.3.7) and that there are large differences among different types. For example, the LDO Selection Guide from Texas Instruments includes a C_{out} column, whose entries range from "No Cap" to "100 μF tantalum." The symptoms of instability may manifest themselves as incorrect, or even zero, output voltage. The latter symptom flummoxed one of our students, who replaced an LP2950 (fixed 5 V LDO) several times before the real culprit was identified: he used a 0.1 μF ceramic bypass capacitor, which is less than the specified 1 μF minimum, and also whose equivalent series resistance (ESR; see §*1x.3*) is too low, a hazard discussed in the Application Hints section of the regulator's datasheet.[25] A more serious symptom of oscillation is output *overvoltage*: we had a circuit with an LM2940 LDO (+5 V, 1 A) that was bypassed in error with 0.22 μF (instead of 22 μF); its internal oscillation caused the measured dc output to go to 7.5 V!

Table 9.3 has two columns to assist in selecting a part, C_{out} (min) and ESR (min,max). But these numbers should be considered a rough guide, and they do not capture everything you need to know to ensure proper operation. You'll find more guidance (e.g., contours of stable operation versus capacitance, ESR, and load current; see for example Figure 9.27) in the plots and applications section of the datasheet – study it carefully!

G. Transient response

Because voltage regulators must be stable into any capacitive load (the sum of all downstream bypass capacitance, often many μF), their feedback bandwidth is limited (analogous to op-amp "compensation"), with typical loop bandwidths in the range of tens to hundreds of kilohertz. So you rely on the output capacitor(s) to maintain low impedance

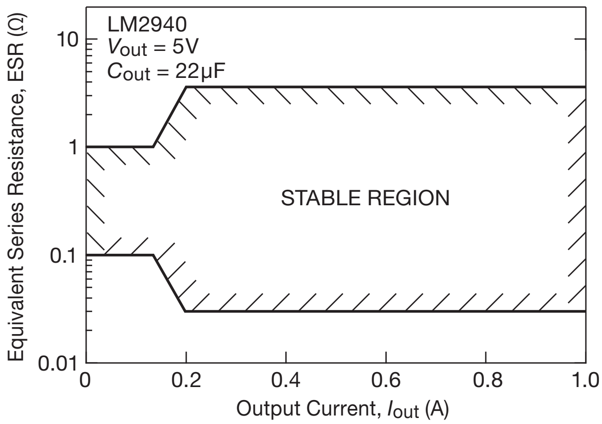


Figure 9.27. Low-dropout linear regulators can set rather fussy requirements for the ESR of the output capacitor, as seen for the LM2940; often you have to obey both minimum and maximum bounds – beware!

at higher frequencies. Or, to say it another way, the output capacitor(s) are responsible for holding the output voltage constant in the short term, in response to a step change in load current, until the regulator responds in the longer term. It's particularly important to include capacitors of low ESR (and equivalent series inductance, ESL) in the mix when you have low-voltage loads with abrupt current changes, as, for example, with microprocessors (which may generate steps of many amps).

We rigged up a 1 V 6 A LDO regulator, using the Micrel MIC5191 control chip, and measured the output response when we made abrupt load steps between 2 A and 4 A, and between 1 A and 5 A. We compared the transient response with two prototype configurations: (a) on a solderable protoboard, using mostly through-hole components; and (b) on a carefully laid-out[26] printed circuit board, using mostly surface-mount components, and with plenty of additional capacitance at both input and output.[27] Figures 9.28–9.31 show the measured step responses. The use of plentiful low-inductance surface-mount technology (SMT) capacitors and low resistance (and low inductance)

[25] In these words: "Ceramic capacitors whose value is greater than 1000 pF should not be connected directly from the LP2951 output to ground. Ceramic capacitors typically have ESR values in the range of 5 to 10 mΩ, a value below the lower limit for stable operation (see curve Output Capacitor ESR Range). The reason for the lower ESR limit is that the loop compensation of the part relies on the ESR of the output capacitor to provide the zero that gives the added phase lead. The ESR of ceramic capacitors is so low that this phase lead does not occur, significantly reducing phase margin. A ceramic output capacitor can be used if a series resistance is added (recommended value of resistance about 0.1 Ω to 2 Ω)."

[26] Expertly done by our student Curtis Mead.

[27] Specifically, for the through-hole configuration we used a 10 μF radial tantalum and two 0.1 μF ceramic bypass capacitors at the input, and a 47 μF ceramic (X5R) SMT plus another 10 μF radial tantalum at the output. For the surface-mount configuration we used a 560 μF radial aluminum polymer capacitor, a 100 μF SMT tantalum polymer capacitor, and two 22 μF ceramic (X5R, 0805) SMT capacitors at both input and output, plus two more 10 μF ceramic (X5R, 0805) SMT capacitors at the output.

power and ground foils produce a stunning improvement: the peak output transient dip drops by a factor of 10 (from ∼40 mV to ∼4 mV for the larger step amplitude), and the output recovers to within a fraction of a millivolt (compared with a ∼6 mV drop for the larger step amplitude).

A different kind of transient-response issue concerns input-voltage transients, and the amount of spike feedthrough to the regulated output. This is different from the "Regulation, line" column in Table 9.3, which lists dc (or low-frequency) rejection of input variations. Input capacitors are somewhat helpful in reducing input-transient effects, but larger output capacitors, especially with low ESR, are a better defense. A special case is a so-called automotive "load dump," a rapid input spike caused, for example, by accidental disconnection of the car battery (from a loose connection, or corrosion, or human error) while being charged by the alternator. This can cause the normal 13.8 V power rail to spike to amplitudes of 50 V or more, causing a spike at a regulator's output. Worse, it can destroy the IC by exceeding its maximum specified input voltage (the "V_{in} max" column of Table 9.3). Parts specifically designed to deal with load dumps are marked with an "o" note in the corresponding V_{in} entry.

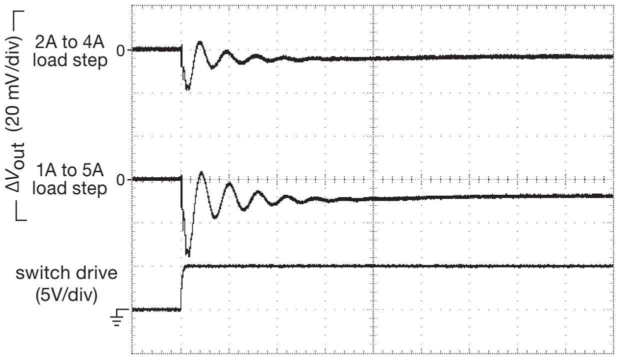


Figure 9.28. Output voltage response to a step increase in load current: 1 V 6 A LDO regulator, breadboarded primarily with through-hole components. Horizontal: 4 μs/div.

H. Noise

Linear regulators vary considerably in the level of output noise (that is, spectrum of output-voltage fluctuations). In many situations this may be unimportant, for example in a digital system, where the circuit itself is inherently noisy.[28] But for low-level or precision analog electronics where

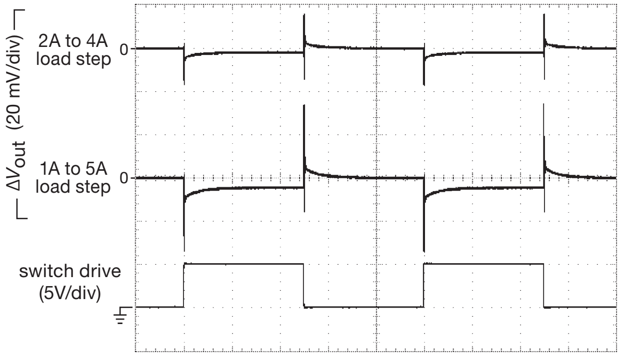


Figure 9.29. Same as Figure 9.28, with 'scope slowed to 400 μs/div to show full load cycle.

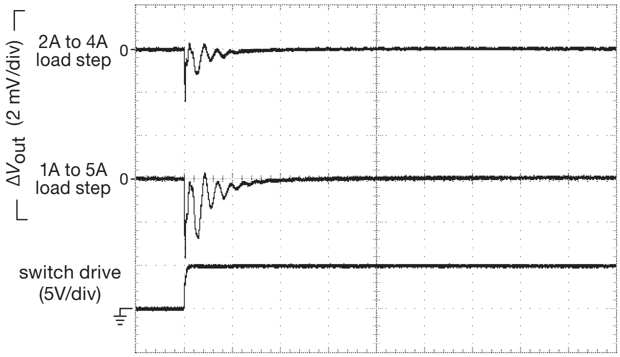


Figure 9.30. Same as Figure 9.28, but built on a printed circuit board using surface-mount capacitors. Note expanded vertical scale.

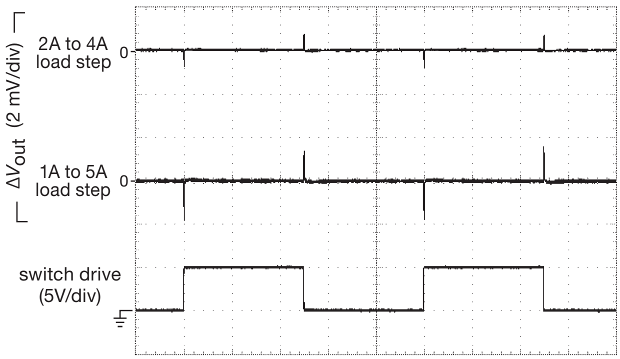


Figure 9.31. Same as Figure 9.30, with 'scope slowed to 400 μs/div to show full load cycle.

noise is important, there are regulators with superior noise specifications, for example the LT1764/1963 (40 μVrms,

[28] In such systems the additional noise of a switching regulator (§9.6) is usually irrelevant, and so switching converters are almost universally used to power digital circuits. They are particularly well suited, owing to their small size and high efficiency, and especially so for the low dc supply voltages (∼1.0–3.3 V) used in digital logic.

10 Hz–100 kHz) or the ADP7102/04 (15 μVrms). Additionally, some regulators provide access to the internal voltage reference, so that an external filter capacitor can be added to suppress all but the low-frequency end of the noise spectrum, for example the LT1964 negative regulator (30 μVrms with 10 nF capacitor); see §9.3.13.

Because manufacturers specify noise characteristics differently (bandwidth, rms versus peak-to-peak, etc.), it can be difficult to compare candidate parts. We've made an attempt in the $V_n@V_{out}$ columns of Table 9.3, but be sure to consult the footnotes (and then the datasheets).

I. Shutdown protection

Some regulator types can be damaged if they see a large capacitance at their output, and the input voltage is brought abruptly to zero (e.g., by a crowbar, or an accidental short-circuit). In that situation the charged output capacitance can source a destructive current back into the regulator's output terminal. Figure 9.32 shows how to prevent such damage, in this case with the popular LM317. Although many engineers don't bother with this nicety, it is the mark of a careful circuit designer. A similar hazard exists when an external bypass capacitor is used to filter the voltage noise of the regulator's reference; see §9.3.13.[29]

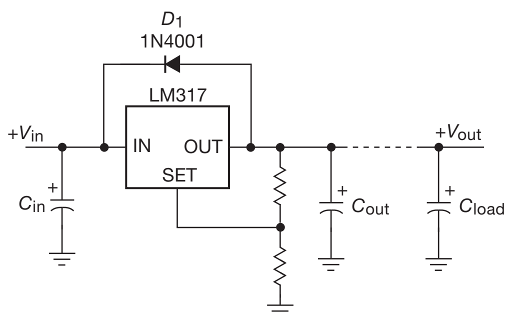


Figure 9.32. Diode D_1 protects the regulator if the input is suddenly grounded.

9.3.13 Noise and ripple filtering

The output noise from a linear regulator is caused by noise in the reference, multiplied by the ratio of V_{out}/V_{ref}, combined with noise in the error amplifier, and with noise and ripple at the input terminal that is not completely suppressed by feedback.[30] Some regulators let you add an external capacitor for lowpass filtering of the internal voltage reference and thus the dc output. Figure 9.33 shows several examples. In Figure 9.33A the ADJ pin of the LM317-style 3-terminal adjustable regulator is bypassed to ground; this provides significant noise improvement by preventing multiplication of the reference noise voltage by the factor $1 + R_2/R_1$ (the ratio of output voltage to 1.25 V reference voltage). It also improves the input ripple rejection ratio, from 65 dB to 80 dB (typical), according to the datasheet. Note the additional protection diode D_2, needed if the noise bypass capacitor C_1 is greater than 10 μF.

This scheme does not eliminate the reference noise, it just prevents it (an ac signal) from being "gained up" by the dc gain factor V_{out}/V_{ref}. The noise filtering in Figures 9.33B and C is more effective, because it filters the reference voltage directly. In Figure 9.33B the LT3080's SET pin, sourcing a stable 10 μA current, is converted to the output voltage by R_{SET}, filtered by C_1; the regulator's output is a unity-gain replica of this filtered voltage. With $C_1 = 0.1\,\mu$F, the reference noise is less than that of the error amplifier, producing an output noise of ~40 μVrms (10 Hz–100 kHz). Note that the noise-filtering capacitor slows the regulator's startup: a 0.1 μF capacitor in a 10 V regulator circuit (R_{SET}=1 MΩ) has a startup time constant $R_{SET}C_1$ of 100 ms.

Finally, Figure 9.33C shows a CMOS low-dropout regulator with a dedicated noise-reduction (NR) pin, for filtering directly the reference voltage presented to the error amplifier. With the recommended 0.1 μF capacitor the output noise voltage is ~40 μVrms (100 Hz–100 kHz). Regulators with this feature are marked in the "Filter pin" column of Table 9.3.

Prefiltering An effective way of reducing dramatically the output ripple at the powerline frequency (and its harmonics) is to prefilter the dc *input* to the regulator. This is also highly effective in attenuating broadband noise that may be present at the dc input; and it's easier than the alternative of increasing the regulator loop's gain and bandwidth. We discuss this in some detail in §8.15.1 ("Capacitance multiplier"), where we show the measured effects of prefiltering versus the brute-force approach of piling on lots of output capacitance (Figure 8.122).

See further discussion of noise in §9.10, in connection with voltage references.

[29] Many LDO regulator ICs include an internal diode sufficiently hardy to handle the reverse-discharge energy in modest (e.g., ≤10 μF) load capacitors. These are marked with a triangle symbol (△) in the "reverse block" column in Table 9.3. Other parts do not discharge the output capacitor if the input voltage is taken below the output; these ICs are marked with either a bullet dot (•) or a square (□).

[30] Table 9.3 includes a "Line regulation" column, but note that this is at dc and low frequencies where the loop gain is high; it is not necessarily indicative of high-frequency supply-noise rejection.

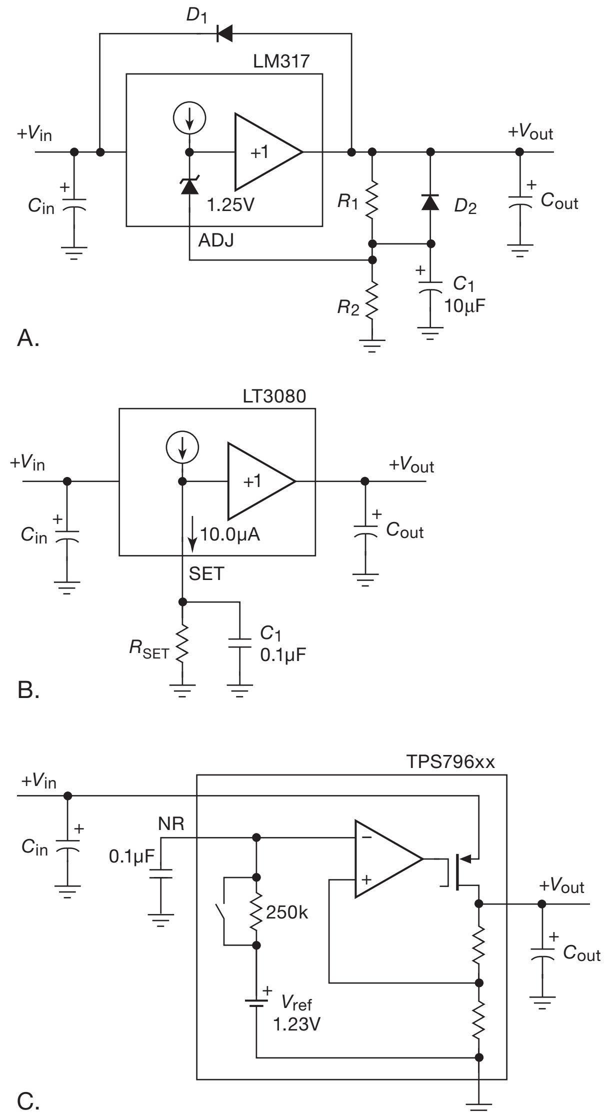


Figure 9.33. Reducing output-voltage noise (and improving transient line regulation) in linear regulators.

9.3.14 Current sources

A. Three-terminal regulators as current sources

A 3-terminal linear regulator can be used to make a simple current source by putting a resistor across the regulated output voltage (hence constant current $I_R = V_{reg}/R$), and floating the whole thing on top of a load returned to ground (Figure 9.34A). The current source is imperfect, however, because the regulator's operating current I_{reg} (which comes out the ground pin) is combined with the well-controlled resistor current to produce a total output current $I_{out} = V_{reg}/R + I_{reg}$. It's a reasonable current source, though, for output currents much larger than the regulator's operating current.

Originally this circuit had been implemented with a 7805, which has an operating current of ∼3 mA and additionally has the disadvantage of squandering a rather large 5 V (the lowest-voltage part in the 78xx series) to define the

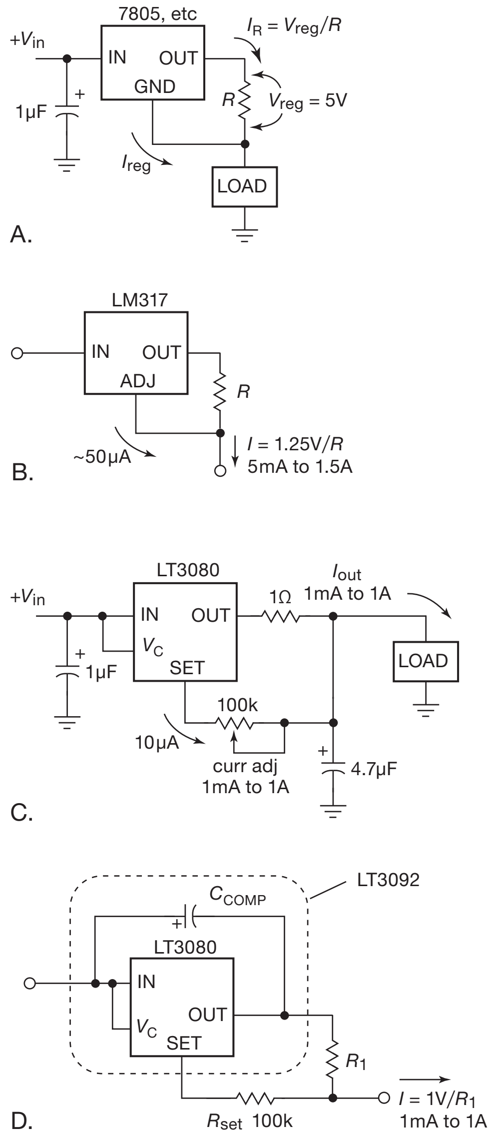


Figure 9.34. Three-terminal regulators used as current sources. The bypass and compensation capacitors can be eliminated in circuits C and D if the LT3080 is replaced by its internally compensated LT3092 variant (whose output current is limited to a maximum of 200 mA).

output current. Happily, with regulators like the LM317, this circuit (Figure 9.34B) becomes more attractive: only 1.25 V is used to set the current; and the regulator's operating current (∼5 mA) emerges from the *output* pin, and thus is accurately accounted for in setting $I_{out} = V_{ref}/R$. The only error term is the ADJ pin current of ∼50 μA, which gets added to the current through R: $I_{out} = V_{ref}/R + I_{ADJ}$. Because the minimum output current of 5 mA is 100 times larger, that is a small error even at the minimum output current, and smaller still at currents up to the regulator's maximum of 1.5 A. For this circuit, then, the output-current range is 5 mA–1.5 A. It requires a minimum voltage drop of 1.25 V plus the regulator's dropout voltage, or about 3 V; the maximum voltage *across the two terminals* is limited either to 40 V or (at higher currents) to the maximum junction temperature of 125°C (as determined by power dissipation and heatsinking), whichever is less.[31]

With the admirable LT3080-style regulator you can do better still, because its 10 μA SET-pin current reference lets you set the voltage across the current-setting resistor to be much less than the 1.25 V of a 317-style voltage-reference regulator. Its operating current is smaller as well (<1 mA), and the SET pin current (which gets added to the output current) is a stable and accurate 10.0 μA. Figure 9.34C shows how to make a (1-terminal) current source to ground with an LT3080, and Figure 9.34D shows how to make a 2-terminal "floating" current source, analogous to the LM317 current-source circuit. As with the latter, the voltage drop is limited to a maximum of 40 V (less at higher currents) at the high end; its lower dropout and low SET-derived reference voltage allows operation down to ∼1.5 V drop. The LT3092-series is a nice variant of the LT3080, designed specifically for use as a 2-terminal current source. It uses the same 10 μA reference current, and operates from 1.2 V to 40 V drop; its internal compensation is configured to require *no* external bypass or compensation capacitors. Based on the LT3092's datasheet plot of output impedance, the device's effective parallel capacitance is approximately 100 pF at 1 mA, 800 pF at 10 mA, and 6 nF at 100 mA.

Figure 9.35 shows, on a greatly expanded scale, the measured output currents of an LT3092 and an LM317, configured as 10 mA current sources. In our measurements the latter does a better job of maintaining constant current (versus voltage across it), but the LT3092 kicks in at a lower voltage.

Note that the current sources in Figures 9.34B and D are 2-terminal devices. Thus the load can be connected on either side. For example, you could use such a circuit to *sink* current from a load returned to ground, by connecting the load between ground and the input, and connecting the "output" to a negative voltage (of course, you could always use the negative-polarity 337, in a configuration analogous to Figure 9.34A.).

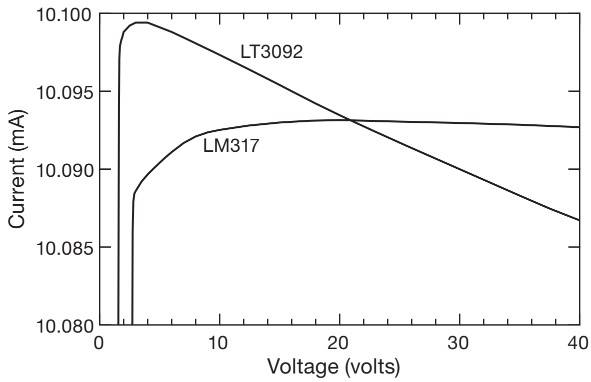


Figure 9.35. Measured current versus voltage drop for the current sources of Figures 9.34B and D, configured as 10 mA 2-terminal current sources. For the LM317, R_1=124 Ω; for the LT3092, R_1=20 Ω and R_{SET}=20k.

B. Lower currents

The above regulator-derived current sources are best suited for substantial output currents. For lower currents, or for higher voltages, there are some good alternatives.

LM334

It's worth knowing about the LM334 (originated by National Semiconductor), optimized for use as a low-power 2-terminal current source (Figure 9.36A). It comes in small-outline IC (SOIC) and TO-92 (transistor) packages and costs about $1 in small quantities. You can use it all the way down to 1 μA because the ADJ current is a small fraction of the total current; and it operates over a voltage range of 1–40 V. It has one peculiarity, however: the output current is temperature dependent – in fact, precisely proportional to absolute temperature (PTAT). So although it is not the world's most stable current source, you can use it as a temperature sensor! At room temperature (20°C, ∼293K) its tempco is about +0.34%/°C.[32]

[31] While we haven't experienced this ourselves, we've been told of possible issues with LM317-based current sources, such as long turn-on times, voltage retention, and poor voltage-compliance above a few kilohertz. It's always wise to test circuit performance fully (especially, uh, *creative* circuits).

[32] See also the discussion in §*2x.3*.

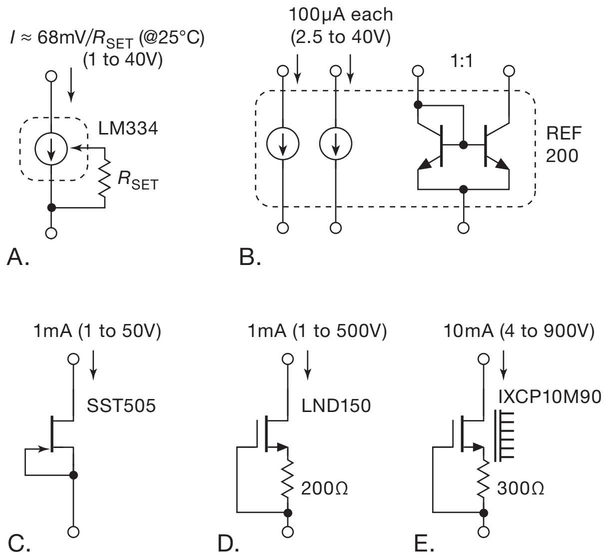


Figure 9.36. 2-terminal current source alternatives.

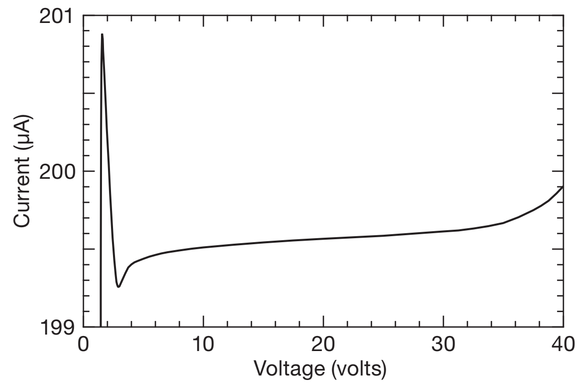


Figure 9.37. Measured current versus voltage for the REF200 two-terminal current source (parallel connection of the 100 μA pair).

REF200

The REF200 is another current-source IC worth knowing about (Figure 9.36B). It has a pair of floating high-quality 100 μA ($\pm$0.5%) 2-terminal current sources (output impedance >200 MΩ over a voltage range of 3.5 V–30 V). It comes in dual in-line packages (DIP) and SOIC packages and costs about \$4 in small quantities. Unlike the LM334, the REF200's current sources are temperature stable ($\pm$25ppm/°C, typ). It also has an on-chip unit-ratio current mirror, so you can make a 2-terminal current source with fixed currents of 50 μA, 100 μA, 200 μA, 300 μA, or 400 μA. The datasheet asserts that "applications for the REF200 are limitless," though we are skeptical. Figure 9.37 shows measured current versus voltage for the parallel connection of the 100 μA pair.

Discrete-component current sources

When thinking about current sources, don't forget about 2-terminal devices like

(a) the humble JFET "current-regulator diode" (§3.2.2), which makes a simple 2-terminal current source (Figure 9.36C) that operates nicely up to 100 V (we plotted measured current versus voltage in Figure 9.38);
(b) a discrete JFET (see Tables 3.1, 3.7, and 8.2), configured similarly as a 2-terminal current source;
(c) the analogous use of a depletion-mode MOSFET (see Table 3.6 on page 210) like the Supertex LND150 (Figure 9.36D), discussed on this page (§9.3.14C);
(d) the series of 2-terminal "Constant Current Regulator and LED Driver" devices from ON Semiconductor. These are inexpensive (\$0.10–\$0.20 in qty 100), and are offered with selected currents (e.g., NSI50010YT1G: 10 mA, 50 V; NSIC2020BT3G: 20 mA, 120 V) and in adjustable versions (e.g., NSI45020JZ: 20–40 mA, 45 V). The datasheets don't say much about what's inside these things, but depletion-mode FETs are the likely culprits.

Op-amp current-source configurations

If the application does not require a floating current source, then consider also

(d) the simple BJT current source (§2.2.6), drawn schematically in Figure 9.39A;
(e) the op-amp assisted BJT current source (§4.2.5), Figure 9.39B;
(f) the Howland current source (§4.2.5B), Figure 9.39C.

In these last three figures the bias voltage that programs the current is drawn as a floating battery; in a circuit implementation it would be a voltage relative to ground or to a supply rail, derived from a voltage reference.

C. High-voltage discrete current source

As mentioned above in the paragraph on discrete-component current sources, a simple source-biased depletion-mode MOSFET (Figures 9.36D and E) forms a pretty good 2-terminal current source. These parts come in convenient packages (TO-92, SMT, TO-220, D^2PAK), with voltage ratings to 1.7 kV; familiar examples are the LND150 and DN3545 from Supertex and the IXCP10M45S and IXCP10M90S from IXYS – see Table 3.6 on page 210. Because of the uncertainty in I_D versus V_{GS}, this sort of current source is not particularly precise or

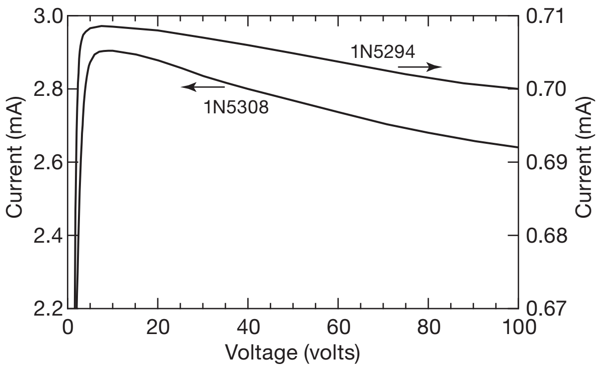


Figure 9.38. Measured current versus voltage for two members of the 1N5283 "current-regulator diode" (a JFET, actually) series.

predictable. But it's fine for noncritical applications, such as replacing a pullup resistor, and it has the advantage of operating to rather high voltages (500 V and 450 V for the Supertex parts; 450 V and 900 V for the IXYS). The IXYS datasheets call their product a "switchable current regulator." Figures 9.40 and 9.41 show some measured data for this simple circuit. The IXYS depletion-mode MOSFET line currently tops out at 1700 V (IXTH2N170). See §*3x.6* for a discussion of yet-higher-voltage versions (to 3 kV or more).

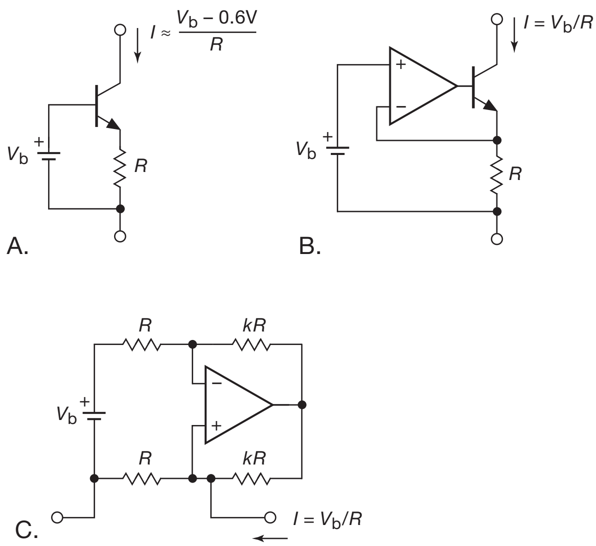


Figure 9.39. BJT and op-amp current sources, drawn in abbreviated form with a floating bias battery. For details see the relevant discussions in Chapters 2 (BJTs) and 4 (op-amps).

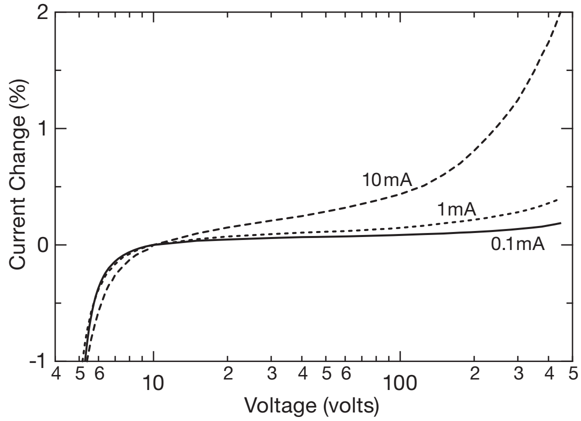


Figure 9.40. Measured current versus voltage for an IXCP10M45S depletion-mode power MOSFET, wired as a self-biased 2-terminal current source (as in Figure 9.36E).

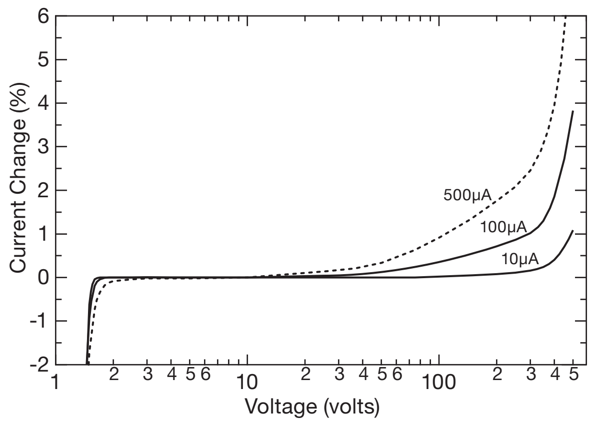


Figure 9.41. Measured current for the smaller LND150 depletion-mode MOSFET, a handy part for low-current applications (compare with Figure 9.40).

9.4 Heat and power design

Up to now we've been skirting the issue of *thermal management* – the business of dealing with the heat generated by transistors (and other power semiconductors) in which the power dissipation (the voltage drop times the current) is greater than a few tenths of a watt. The solution consists of some combination of passive cooling (conducting the heat to a heatsink or to the metal case of an instrument) and active (forced air or pumped liquid) cooling.

This problem is not unique to voltage regulators, of course – it affects linear power amplifiers, power-switching circuits, and other heat-generating components such as power resistors, rectifiers, and high-speed digital ICs. Contemporary computer processors, for example, dissipate many tens of watts and can be recognized by their attached finned heatsinks and blowers.

Linear voltage regulators get us into the topic "power electronics," because they are intrinsically inefficient: the full load current flows through the pass transistor, with a voltage drop at least adequate to prevent dropout. In the case of an unregulated dc input, as in Figure 9.25, that means a drop of at least a few volts; so with an amp of output current you've got at least a few watts... and you've got a problem. In the following sections we'll see how to solve it.

9.4.1 Power transistors and heatsinking

All power devices are packaged in cases that permit contact between a metal surface and an external heatsink. At the low-power end of the spectrum (up to a watt) the device may be cooled via conduction through its leads, soldered to a circuit board; the next step up is the surface-mount power packages with a larger tab (and with names like SOT-223, TO-252, TO-263, DPAK, and $\mathrm{D^2PAK}$), or more advanced packages like the "DirectFET" (see Figure 9.46). For power dissipation greater than about 5 watts the packages (with names like TO-3, TO-220, and TO-247) will have mounting holes for attachment to a substantial heatsink; and really high power semiconductors come in modules (like the "miniBLOC" or "Powertap" – see Figure 9.47) meant for mounting off-PCB. With adequate heatsinking the latter types can dissipate up to 100 watts or more. With the exception of "isolated" power packages, the metal surface of the device is electrically connected to one terminal (e.g., for bipolar power transistors the case is connected to the collector, and for power MOSFETs to the drain).

The whole point of heatsinking 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.[33] Heatsink 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, heatsink, etc., and the maximum ambient temperature in which the circuit is expected to operate. You then choose a heatsink large enough to keep the junction temperature well below the maximum specified by the manufacturer. It is wise to be conservative in heatsink design, because transistor life decreases rapidly at operating temperatures near or above maximum. Figure 9.42 shows a representative sample of heatsinks that we gathered from our lab's supply drawers.

Some people are cavalier about thermal design, and start worrying only if the component sizzles when they touch it with a wet finger. But it's far better to do it right initially! Read on...

A. Thermal resistance

To carry out heatsink calculations, you use *thermal resistance*, R_θ, defined as heat rise (in °C) divided by power transferred. For power transferred entirely by heat 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 heatsink, the total thermal resistance from transistor junction to the outside (ambient) world is the sum of the thermal resistance from junction to case $R_{\theta\mathrm{JC}}$, the thermal resistance from case to heatsink $R_{\theta\mathrm{CS}}$, and the thermal resistance from heatsink to ambient $R_{\theta\mathrm{SA}}$. The temperature of the junction is therefore

$$T_\mathrm{J} = T_\mathrm{A} + (R_{\theta\mathrm{JC}} + R_{\theta\mathrm{CS}} + R_{\theta\mathrm{SA}})P, \tag{9.2}$$

where P is the power being dissipated.

Let's take an example. The power-supply circuit of Figure 9.25, with 8 V unregulated dc input and full load (1 A) has a maximum LM317AT regulator dissipation of 4.7 W (4.7 V drop, 1 A). 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 100°C, well below its specified maximum of 125°C.

The allowable temperature difference from junction to ambient is thus 50°C, so the total thermal resistance from junction to ambient must be no more than $R_{\theta\mathrm{JA}} = (T_\mathrm{J} - T_\mathrm{A})/P = 10.6$°C/W. The specified thermal resistance from junction to case, $R_{\theta\mathrm{JC}}$, is 4°C/W, and the TO-220 power transistor package mounted with a heat-conducting pad has a thermal resistance from case to heatsink of about 0.5°C/W. So we've used up $R_{\theta\mathrm{JC}} + R_{\theta\mathrm{CS}} =$ 4.5°C/W of thermal resistance, leaving $R_{\theta\mathrm{SA}} = 6.1$°C/W for the heatsink. A quick scan of the ever-helpful DigiKey

[33] See Tables 2.2 and 3.4 for a selection of power transistors, including their maximum power dissipation assuming an (unrealistic) case temperature of 25°C. As we'll see, that's enough information to allow you to back out the thermal resistance $R_{\Theta\mathrm{JC}}$, from which you can figure out realistic values of maximum power dissipation, and thus appropriate heatsinking.

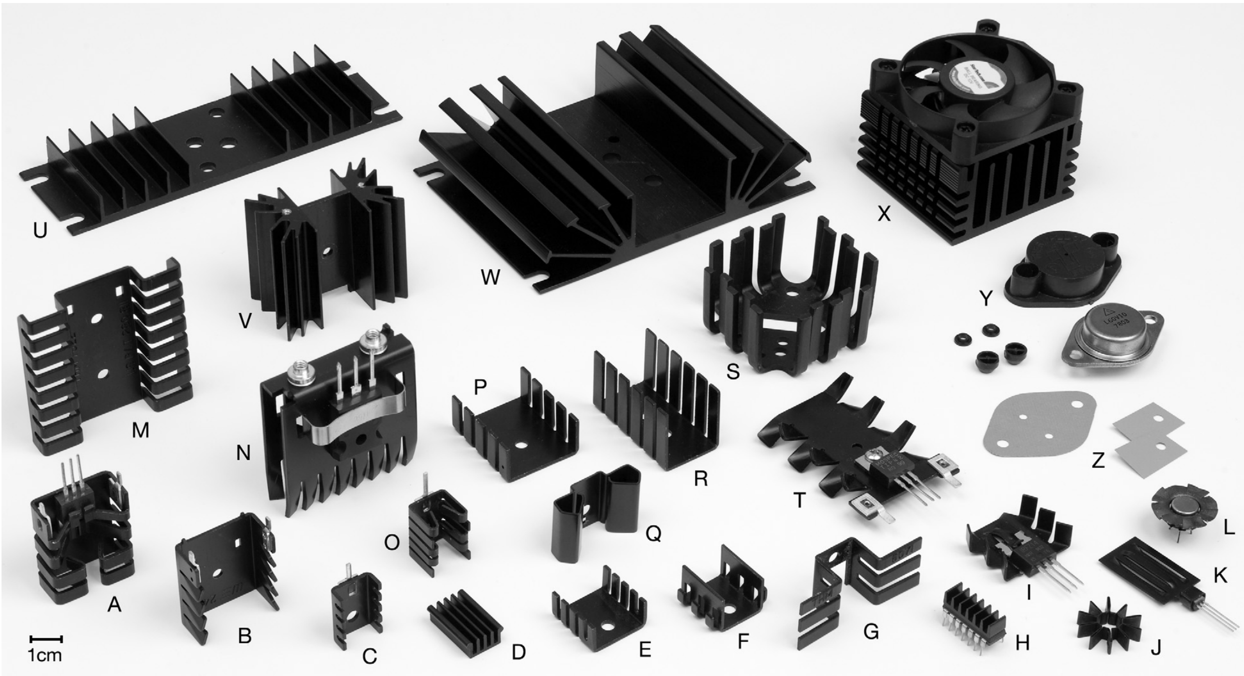


Figure 9.42. Heatsinks come in an impressive diversity, from little clip-on fins (I–L), to mid-sized PCB-mounting types (A–C, N, O, T), to large bolt-down units (U, W), to forced-air type used with microprocessors (X). The corresponding thermal resistance from sink to ambient, $R_{\Theta SA}$, ranges from about 50°C/W down to about 1.5°C/W. A TO-3 insulating cover is shown in (Y), along with shoulder washers and hole plugs; greaseless thermal insulating pads are shown in (Z). We've added alphabetic labels so readers can identify objects of interest when chatting on social media.

catalog finds many candidates, for example the Wakefield 647-15ABP "vertical board mounting" finned heatsink, with the requisite $R_{\theta SA} = 6.1$°C/W in still air ("natural convection"). They're priced at about \$2, with 2000 pieces in stock. With a "forced convection" of 400 LFM (linear feet per minute) we could use instead the smaller (and cheaper, about \$0.35) model 270-AB; DigiKey's got 4000 in stock today.

Here's a "sizzle-test" for checking for adequate heatsinking: touch the power transistor with a dampened finger – if it sizzles, it's too hot! (Be careful when using this "rule-of-finger" test to explore around high voltages.) More generally approved methods for checking component temperatures are (a) a contacting thermocouple or thermistor probe (these often come as standard equipment with handheld or benchtop digital multimeters); (b) special calibrated waxes that melt at designated temperatures (e.g., the Tempilstik® wax pencil kits from Tempil, Inc.); and (c) infrared non-contacting temperature probes,[34] for example the Fluke 80T-IR, which generates 1 mV/°C or 1 mV/°F (switchable), operates from −18°C to +260°C, is accurate to 3% of reading (or ±3°C, if greater), and plugs into any handheld or benchtop DMM.

B. Comments on heatsinks

1. Where very high power dissipation (several hundred watts, say) is involved, forced air cooling is usually necessary. Large heatsinks 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. In cases of such high thermal conductivity (low thermal resistance, $R_{\theta SA}$), you may find that the ultimate limit to power dissipation is in fact the transistor's own internal thermal resistance, combined with its attachment to the heatsink (i.e., $R_{\theta JC} + R_{\theta CS}$). This problem has been exacerbated in recent years by the evolution of smaller ("shrink") semiconductor chip sizes. The only solution

[34] Infrared ear thermometers use this method to measure core body temperature via infrared emission from the eardrum, evidently accurately enough for clinical purposes; e.g., the Braun ThermoScan takes a measurement in one second, with an accuracy claimed to be significantly better than 1°C.

here is to spread the heat among several power transistors (in parallel or in series). When paralleling power transistors you have to be careful to make sure they share the current equally – see §2.4.4 and Figure 3.117.

Similarly, when connecting transistors in series, make sure that their off-state voltage drops are evenly distributed.

3. Figure 9.43, adapted from the Wakefield Engineering heatsink literature, gives a rough estimate of the physical volume of heatsink required to achieve a given thermal resistance. Note that curves are given for still air (natural convection) and for two values of forced air flow. Don't take these curves too literally – we just gathered data from a half dozen representative heatsinks, and then drew trend lines through them; they're probably good to a factor of two, but don't rely on it (or, at least, don't complain to us later!).

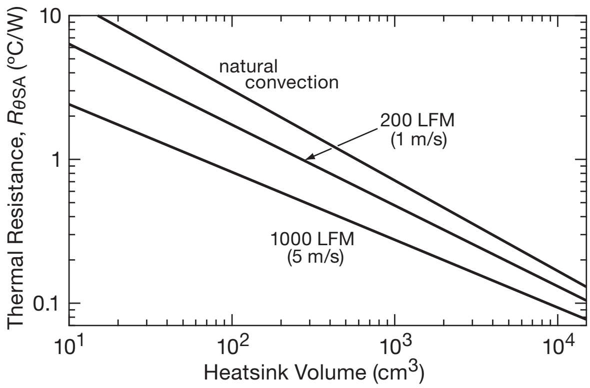


Figure 9.43. Rough guide to heatsink size needed for a given thermal resistance from sink to ambient ($R_{\theta SA}$).

4. When the transistor must be insulated from the heatsink, 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 transistor-shape cutouts made from mica, anodized (insulated) aluminum, beryllia (BeO), or polymer films such as Kapton®. Used with heat-conducting grease, these add from 0.14°C/W (beryllia) to about 0.5°C/W.

An attractive alternative to the classic mica-washer-plus-grease is provided by greaseless silicone-based insulators that are loaded with a dispersion of a thermally conductive compound, usually boron nitride or aluminum oxide ("Z" in Figure 9.42). 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. The electrically insulating types have thermal resistances of about 1–4°C/W for a TO-220 package footprint, comparable to values with the messy method; the non-insulating ("grease replacement") varieties do better – down in the 0.1–0.5°C/W for a TO-220 package. Bergquist calls its product line "Sil-Pad," Chomerics calls its "Cho-Therm," and Thermalloy calls its "Thermasil." We've been using these insulators, and we like them.

5. Small heatsinks are available that simply clip over the small transistor packages (like the standard TO-92 and TO-220, "I–L" in Figure 9.42). In situations of relatively low power dissipation (a watt or two) this often suffices, avoiding the nuisance of mounting the transistor remotely on a heatsink with its leads brought back to the circuit. In addition, there are various small heatsinks intended for use with the plastic power packages (many regulators, as well as power transistors, come in this package) that mount right on a PCB underneath the package. These are very handy in situations of a few watts dissipation; a typical unit is illustrated in Figure 2.3. If you've got vertical space over the PCB, it's often preferable to use a PCB-mounting stand-up heatsink (like A–C, N, O, or T in Figure 9.42), because these types take up less area on the PCB.

6. Surface-mount power transistors (such as the SOT-223, DPAK, and D^{2}PAK) carry their heat to the foil layer of a PCB via the soldered tab; here we're talking a few watts, not a hundred. You can see these packages in Figures 2.3 and 9.44. Figure 9.45 plots approximate values of thermal resistance versus foil area; these should be considered only a rough guide, because the actual heatsinking effectiveness depends on other factors such as the proximity of other heat-producing components, board stacking, and (for natural convection) board orientation.

7. 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 because a hot case will subject the other circuit components to high temperatures and thus shorten their lives.

8. If a transistor is mounted to a heatsink without insulating hardware, the heatsink must be insulated from the chassis. 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 heatsink may be attached directly to the chassis. But if the transistor is accessible from outside the instrument (e.g., if the heatsink is

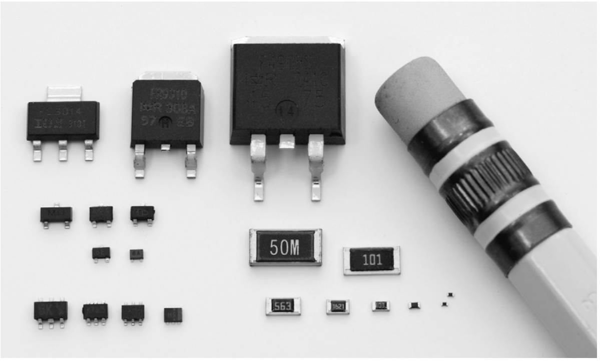


Figure 9.44. Power transistors come in convenient hand-solderable surface-mount packages that can dissipate up to several watts, through their mounting tab and leads, when soldered to a few square centimeters of foil area on a PCB (see Figure 9.45). The top three packages (SOT-223, DPAK, D^2PAK) are good for $\sim$3 W when mounted on 6 cm^2 of PCB foil area; the smaller packages in the row below can dissipate $\sim$0.5 W when similarly mounted. By way of comparison, the rectangular parts at lower right are chip resistors, tapering down from 2512 size to 0201 size (0603 metric).

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, "Y" in Figure 9.42) to prevent someone from accidentally coming in contact with it or shorting it to ground.

9. The thermal resistance from heatsink 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[35]); usually it is best to mount it on the rear of the instrument with fins vertical.
10. In the second edition of this book there is additional information: see Chapter 6 (Figure 6.6, page 315) on heatsinks, and Chapter 12 (Table 12.2 and Figure 12.17, page 858) on cooling fans.

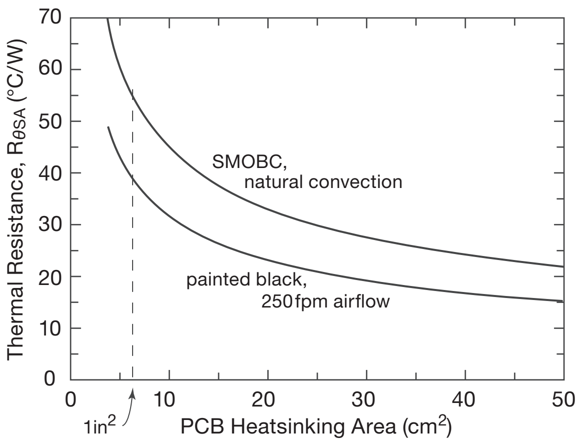


Figure 9.45. Approximate thermal resistance of isolated printed-circuit foil patterns. A soldermask layer (SMOBC means soldermask over bare copper) reduces the effectiveness, particularly when compared with forced air over exposed copper.

Exercise 9.7. An LM317T (TO-220 case), with a thermal resistance from junction to case of $R_{\Theta JC}$=4°C/W, is fitted with an Aavid Thermalloy 507222 bolt-on heatsink, whose thermal resistance is specified as $R_{\Theta SA}\approx$18°C/W in still air. The thermal pad (Bergquist SP400-0.007) specifies a thermal resistance of $R_{\Theta CS}\approx$5°C/W. The maximum permissible junction temperature is 125°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?

9.4.2 Safe operating area

The point of heatsinking is to keep junction temperatures within specified limits, given some ambient temperature and some maximum power dissipation, as just described. Of course, you must stay also within the specified voltage and current ratings of the power transistor. This is displayed graphically as a dc safe-operating-area (SOA) plot, on axes of transistor voltage and current, at some specified case temperature (usually an unrealistic T_C=25°C). For MOSFETs this plot (on logarithmic voltage and current axes) is bounded simply by straight lines representing maximum voltage, maximum current, and maximum power dissipation (at the specified T_C, as set by $R_{\theta JC}$ and $T_{J(max)}$) – see, for example, Figure 3.95.

There are two amendments to this basic picture.

A. Second breakdown

The bad news: in the case of *bipolar transistors* the SOA is further constrained by a phenomenon known as *second breakdown*, an important failure mechanism that you must keep in mind when designing power electronics with bipolar transistors. This is discussed in §3.6.4C, where the effect can be seen in the SOA plots in Figure 3.95 as a further reduction of allowable collector current at high voltages. Because MOSFET power transistors are largely immune to second breakdown, they are often favored over BJTs for power-regulator pass transistors (the exceptions are some newer small-geometry types, see IR App Note IN-1155).

[35] As a rough guide, you can expect roughly 20% increase in $R_{\theta SA}$ for "fins horizontal," 45% for "fins up," and 70% for "fins down."

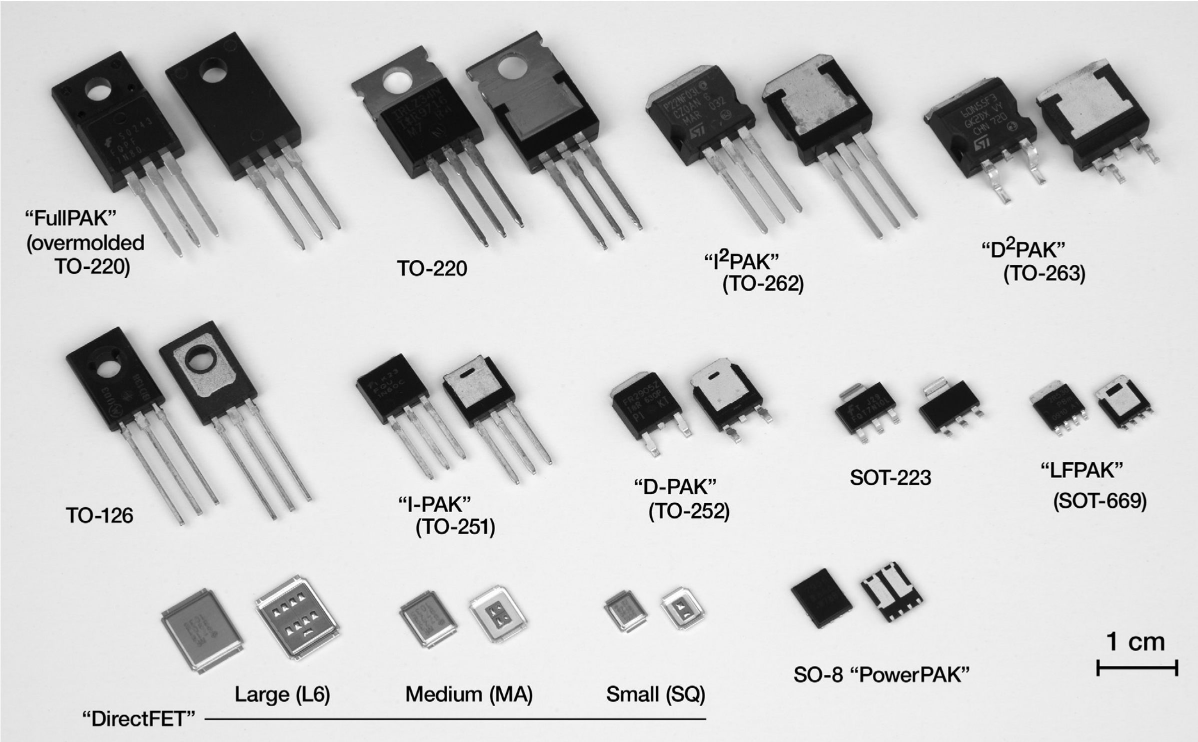


Figure 9.46. An extended selection of power packages, shown here and in Figure 9.47. The leadless packages in the bottom row require "reflow" solder techniques (trade in your soldering iron and wire solder for an oven and solder-paste dispenser!).

B. Transient thermal resistance

The good news: for pulses of short duration you can exceed the dc power-dissipation limit, sometimes by a large factor. That is because the mass of the semiconductor itself can absorb a short pulse of energy by heating locally ("heat capacity" or "specific heat"), limiting the temperature rise even if the instantaneous power dissipation is more than could be sustained continuously. This can be seen in the SOA plot (Figure 3.95), where the allowable power dissipation, for 100 μs pulses, is some 20 times greater than the dc value: an astonishing 3000 W versus 150 W. This is sometimes characterized on datasheets as a *transient thermal resistance* – a plot of R_θ versus pulse duration. The ability to dissipate very high peak power during short pulses extends to other electronic devices, for example, diodes, SCRs, and transient voltage suppressors. See §3.6.4C and the discussion in Chapter *9x*.

9.5 From ac line to unregulated supply

A regulated power supply that runs from ac line power begins[36] by generating "unregulated" dc, a subject we introduced in §1.6.2 in connection with rectifiers and ripple calculations. For the linear voltage regulators we've seen so far, the unregulated dc supply uses a transformer, both

[36] Well, it really begins back at the power plant! Perhaps worth knowing, though, is the situation at the wall plug: in the US the standard 3-prong outlet delivers its 120 Vrms ac across the "line" and "neutral" blades (the neutral is the slightly wider slot; it's on the upper left, if the outlet is oriented to look like a face), with the round safety ground returned to a good earth connection at the service entry. The power comes into the house as three wires from a center-tapped 240 V pole transformer, with the center tap (neutral) bonded to earth ground at the service entry. Any given 120 V outlet provides neutral (white wire) and one "live" phase (black); the outlets in a given room may be powered by one or the other phases. A 240 V appliance outlet brings both live phases, along with safety ground, to a different style socket (this is in contrast to European 220 or 240 V outlets, which provide line, neutral, and safety ground). The in-wall wiring consists of "Romex"-style plastic-insulated oval cable with solid copper conductors: AWG14 for a 15 A residential circuit, and AWG12 for a 20 A circuit.

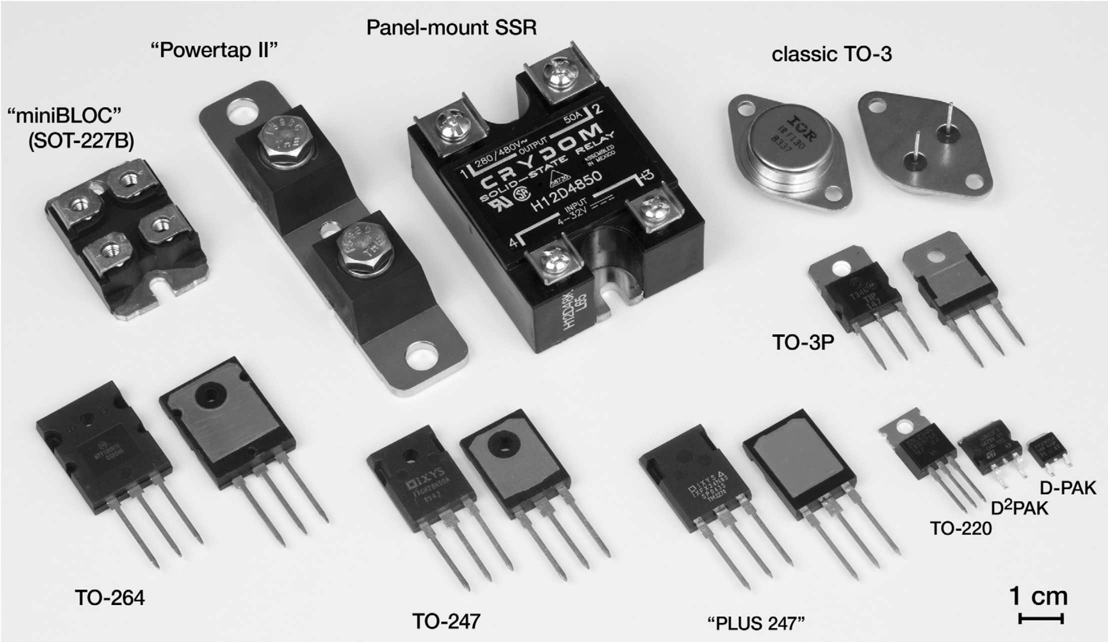


Figure 9.47. The larger cousins of the power packages in Figure 9.46, with three specimens from the latter shown for comparison. We're up in the tens to hundreds of watts with these when they are mounted on an appropriate heatsink.

to convert the incoming line voltage (120 Vrms in North America and a few other countries, 220 or 240 Vrms most everywhere else) to a (usually) lower voltage closer to the regulated output, and also to isolate the output from any direct connection to the hazardous line potentials ("galvanic isolation"); see Figure 9.48. Perhaps surprisingly, the switching power supplies we'll see shortly omit the transformer, generating line-derived dc at the powerline potential (∼160 Vdc or ∼320 Vdc), which powers the switching circuit directly. The essential galvanic isolation[37] is achieved instead with a transformer driven by the high-frequency switching signal.[38]

Transformer-isolated unregulated dc supplies are useful, also, for applications in which the stability and purity of regulated dc is unnecessary, for example high-power audio amplifiers. Let's look at this subject in more detail, beginning with the circuit shown in Figure 9.49. This is an unregulated ±50 volt (nominal) split supply, capable of 2 A output current, for a 100 watt linear audio amplifier. 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.

9.5.1 ac-line components

A. Three-wire connection

Always use a 3-wire line cord with ground (green or green/yellow) 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 powerline 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 (International Electrotechnical Commission) 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 we've used here. Note that ac wiring uses a nonintuitive color convention: black = "hot" (or "line"), white = neutral, and green = ground (or "protective earth").[39]

[37] There are occasions where isolation is not needed; see Chapter *9x* for a discussion of some low-power off-line power supplies (including a step-down technique that uses a capacitor instead of a transformer or inductor.)

[38] The advantage of this peculiar arrangement is that the transformer, running at a high frequency (20 kHz–1 MHz) is much smaller and lighter.

[39] IEC cords use brown = line, blue = neutral, and green/yellow = ground.

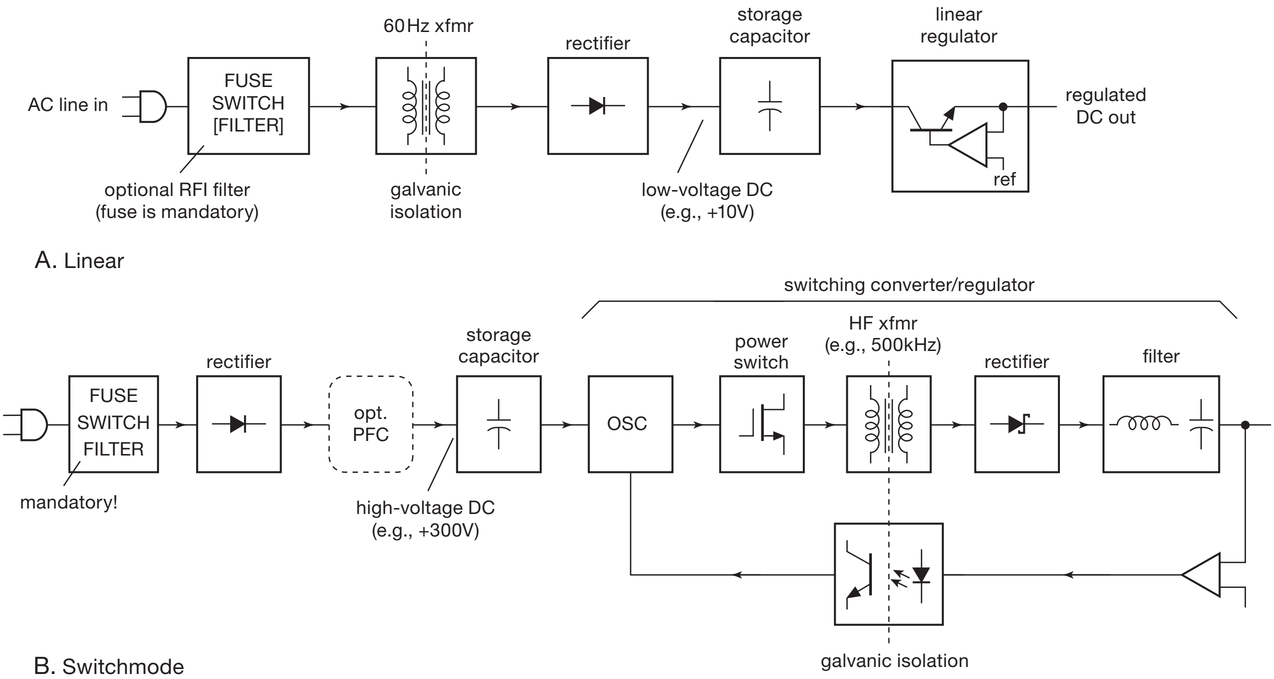


Figure 9.48. ac-line-powered ("offline") dc-output regulated supplies. A. In the linear supply the powerline transformer both isolates and transforms the input voltage. B. In the switching ("switchmode") converter the ac input is directly rectified to high-voltage dc, which powers the isolating switching converter. The PFC block performs power-factor correction, discussed later in §9.7.1C.

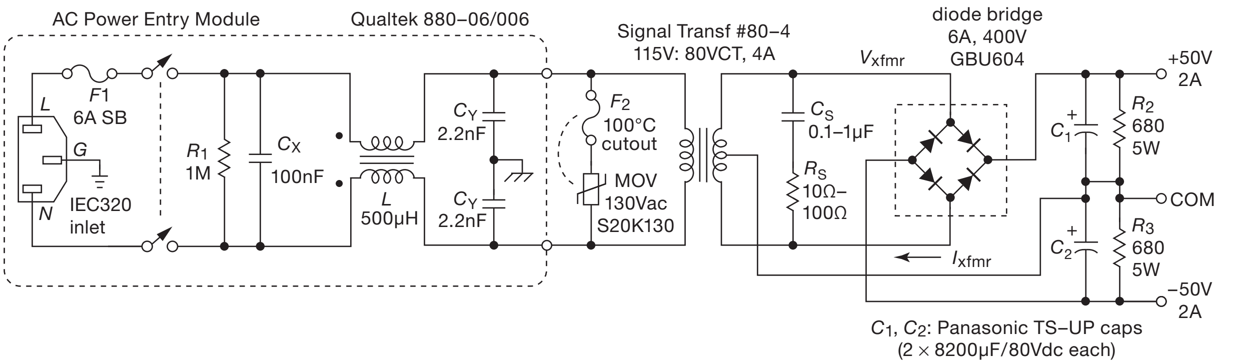


Figure 9.49. Unregulated ±50 V, 2 A power supply.

B. Fuse

A fuse, breaker, or equivalent function should be included in every piece of electronic equipment. A fuse holder, switch, and lowpass filter are often combined in the power entry module, but you can also wire them *à la carte*. The large wall fuses or circuit breakers (typically 15–20 A) in house or lab won't protect electronic equipment, because they are chosen to blow only when the current rating of the wiring in the wall is exceeded. For instance, a house circuit wired with 14-gauge wire will have a 15 A circuit breaker. Now, if a storage capacitor in our unregulated supply becomes short-circuited someday (a possible failure mode), the transformer might then draw 10 A primary current (instead of its usual 2–3 A). The house breaker won't open, but your instrument becomes an incendiary device, with its transformer dissipating over a kilowatt.

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 ("inrush current") 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 of this design[40] 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/AGC/MDL type 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 powerline. This means connecting the "hot" lead to the rear terminal of the fuse holder (the authors have learned this the hard way!). Commercial power-entry modules with integral fuse holders are cleverly arranged so that the fuse cannot be reached without removing the line cord.

C. Switch

In Figure 9.49 the switch is integral with the power entry, which is fine, but it forces the user to reach around the back to turn the thing on. When using a front-panel power switch, it's a good idea to put a line-rated capacitor (called X1 or X2) across its terminals, to prevent arcing. For similar reasons the transformer primary should have some bridging capacitance, which in this case is taken care of by the lowpass filter in the entry module.

D. Lowpass filter

Although they are often omitted, such filters are a good idea, because they serve the purpose of preventing possible radiation of radiofrequency interference (RFI) from the instrument via the powerline, as well as filtering out incoming interference that may be present on the powerline. These filters typically use an *LC* "π-section" filter (as in the figure), with the coupled inductor pair acting as a common-mode impedance. Power-line filters with excellent performance characteristics are available from several manufacturers, e.g., Corcom, Cornell–Dubilier, Curtis, Delta, Qualtek, and Schurter.[41] Studies have shown that spikes as large as 1 kV to 5 kV are occasionally present on the powerlines at most locations, and smaller spikes occur quite frequently. Line filters (in combination with transient suppressors, see below) are reasonably effective in reducing such interference (and thereby extending the life of a power supply and the equipment it powers).

E. Line-voltage capacitors

For reasons of fire and shock hazard, capacitors intended for line filtering and bypassing are given special ratings. Among other attributes, these capacitors are designed to be self-healing, i.e., to recover from internal breakdown.[42] There are two classes of line-rated capacitors: "X" capacitors (X1, X2, X3) are rated for use where failure would not create a shock hazard. They are used across the line (C_X in Figure 9.49; the common X2 type is rated for 250 Vac, with peak voltage of 1.2 kV). "Y" capacitors (Y1, Y2, Y3, Y4) are rated for use where failure would present a shock hazard. They are used for bypassing between the ac lines and ground (C_Y in Figure 9.49; the common Y2 type is rated for 250 Vac, with a peak voltage of 5 kV). Line-rated capacitors come in disc ceramic and in plastic-film flavors; the latter are usually a box geometry, with a flame retardant case. It's hard to miss these capacitors – they are usually festooned with markings proclaiming the various national certifications whose standards they meet[43] (Figure 9.50).

A *further* word about line-voltage capacitors: when the

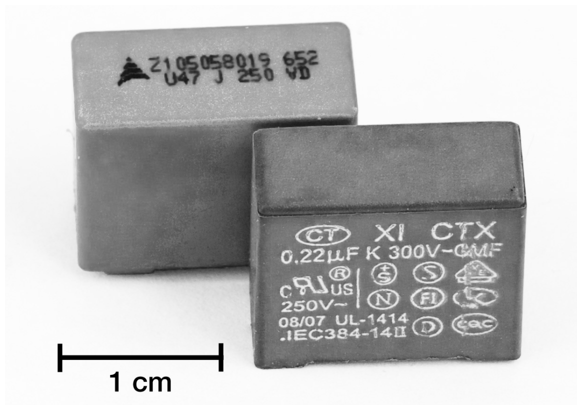


Figure 9.50. ac-line-rated capacitors flamboyantly display their safety ratings (right), compared with the minimalist decoration of a plain ol' film capacitor (left).

[40] Where the rectified input charges large storage capacitors at each voltage peak of the ac waveform. By contrast, switching power supplies with *power-factor correction* (PFC) nicely sidestep this problem; see §9.7.1C.

[41] Watch out, however, for misleading attenuation specifications: they are universally specified with 50 Ω source and load, because that's easy to measure with standard RF instrumentation, and not because it has any resemblance to the real world.

[42] For example, line-rated plastic film capacitors are constructed so that a perforating breakdown causes the metal plating near the hole to burn away, clearing the short-circuit.

[43] Here are some of them: UL, CSA, SEV, VDE, ENEC, DEMKO, FIMKO, NEMKO, SEMKO, CCEE, CB, EI, and CQC.

instrument is unplugged, the X capacitor may be left holding the peak ac line voltage, up to 325 V, which appears across the exposed power plugs! This can cause electrical shocks, and spark discharges. That's why there's a parallel discharge resistor, sized for a safe time constant of less than a second.[44] Here the Qualtek RFI filter module uses 1 MΩ, and the Astrodyne switching supply (§9.8) uses 540 kΩ. The latter continuously dissipates 100 mW when powered with 220 Vac line input, which could constitute one of the largest standby power losses in an "Energy Star" design. Power Integrations offers their CAPZero™ IC to solve this problem. This clever part works by looking for an ac-line-voltage reversal every 20 ms or less, and if it fails to see one it turns on, connecting two discharge resistors across the X capacitor.

Some designs have substantial high-voltage dc storage capacitors that need to be discharged when the power is off. Because of their large capacitance, the ac-sensing CAPZero would not work. Here you could use normally-on relay contacts, and energize the relay when external ac power is present. Or, if you don't like moving parts, a high-voltage depletion-mode MOSFET (see Table 3.6 on page 210) and a photovoltaic stack (see §12.7.5, Figure 12.91A) can do the job.

F. Transient suppressor

In many situations it is desirable to use a "transient suppressor" (or "metal-oxide varistor," MOV) as shown in Figure 9.49. The transient suppressor is 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 shunt hundreds of amperes of potentially harmful current in the form of spikes. Note the thermal cutout fuse: that protects in a situation in which the MOV starts to conduct partially (for example if the line voltage becomes highly elevated, or if an aged MOV exhibits lowered breakdown voltage from having absorbed large transients). Transient suppressors are made by a number of companies, e.g., Epcos, Littelfuse, and Panasonic. Effective transient suppression is an interesting challenge, and we discuss it further in Chapter *9x*.

G. Shock hazard

It is a good idea to insulate all exposed line-voltage connections inside any instrument, for example by using polymer heat-shrink tubing (the use of "friction tape" or electrical tape inside electronic instruments is strictly bush-league). Because most transistorized circuits operate on relatively low dc voltages (±15 V or less), from which it is not possible to receive a shock, the powerline 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, being 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.

9.5.2 Transformer

Now for the transformer. Never build an instrument to run off the powerline without an isolating transformer! To do so is to flirt with disaster. Transformerless power supplies, which have been popular in some consumer electronics (radios and televisions, particularly) because they're inexpensive, put the circuit at high voltage with respect to external ground (water pipes, etc.).[45] 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. It may be hard to find a transformer with the voltage and current ratings you need. We have found the Signal Transformer Company unusual, with their nice selection of transformers and quick delivery. And 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 on the voltage and current rating. If the unregulated supply is powering a linear regulator, then you want to keep the unregulated dc voltage low, in order to minimize the power 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 V above the regulated output voltage, for conventional regulators like the LM317; or 0.5–1 V for low-dropout types) or you may encounter 120 Hz dips in the regulated output; in the design you need to allow for low line voltage (10% below nominal, say – 105 Vac in the US) or even brownout conditions (20% below nominal). The amount of ripple in the unregulated output is involved

[44] We have seen entirely too many designs that omit this discharge resistor; not good!

[45] Non-isolated off-line supplies are commonly found in some types of self-contained electronics, such as a screw-in LED light bulb, a wall clock, a smoke alarm, a Wi-Fi surveillance camera, a toaster or coffeemaker, and so on. We discuss some of these in Chapter *9x*.

here, because it is the *minimum* input to the regulator that must stay above some critical voltage (see Figure 1.61), but it is the *average* input to the regulator that determines the transistor dissipation.

As an example, for a +5 V regulator you might use an unregulated input of +10 V at the minimum of the ripple, which itself might be 1 to 2 volts peak-to-peak. From the secondary voltage rating you can make a pretty good guess of the dc output from the bridge, because 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 (leakage inductance), 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 (105 V). 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 (the Triad F-90X series, for example), when available, for final adjustment of output voltage.

For the circuit shown in Figure 9.49, we wanted ±50 V output under full load. Allowing for two diode drops (from the bridge rectifier) we need a transformer with ∼52 V peak amplitude, or about 37 Vrms. Among the available transformer choices, the closest was the 40 Vrms unit shown, probably a good choice because of the effects of winding resistance and leakage inductance, which reduce slightly the loaded dc output voltage.

An important note: transformer current ratings are usually given as *rms* secondary current. However, because a rectifier circuit draws current only over a small part of the cycle (during the time the capacitor is actually charging), the secondary's rms current, and therefore the I^2R heating, will be significantly larger than the average rectified dc output current. So you have to choose a transformer whose rms current rating is somewhat larger (typically ∼ 2×) than the dc load current. Ironically, the situation gets worse as you increase capacitor size to reduce output ripple voltage. Full-wave rectification is better in this respect, because a greater portion of the transformer waveform is used. For the unregulated dc supply of Figure 9.49 we measured an rms current of 3.95 A at the transformer secondary when powering a 2 A dc load. The measured waveforms in Figure 9.51 show the pulsating nature of the current, as the rectified transformer output recharges the storage capacitors each half-cycle.

Naïvely you might expect that the conduction angle (fraction of the cycle during which the current flows) could be estimated simply by (a) calculating the capacitor's discharge between half-cycles, according to $I = C\,dV/dt$, then (b) calculating the time in the next half-cycle at which the rectified output exceeds the capacitor's voltage. However, this pretty scheme is complicated by the important effects of the transformer's winding resistance and leakage inductance, and the storage capacitor's ESR, all of which extend the conduction angle.[46] The best approach is to make measurements on the bench, perhaps informed by SPICE simulations using known or measured values of these parameters. In §9.5.4 we show results of such simulations.

9.5.3 dc components

A. Storage capacitor

The storage capacitors (sometimes called *filter* capacitors) are chosen large enough to provide acceptably low ripple voltage, with a voltage rating sufficient to handle the worst-case combination of no load and high line voltage (125–130Vrms).

At this point it may be helpful to look back at §1.7.16B, where we first discussed the subject of ripple. In general, you can calculate ripple voltage with sufficient accuracy by assuming a constant-current load equal to the average load current. (In the particular case that the unregulated supply drives a linear regulator, the load in fact is accurately 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 (the measured waveforms in Figure 9.51, taken with a resistive load, illustrate the validity of this approximation).

For the circuit shown in Figure 9.49, we wanted approximately 1 Vpp output ripple at full 2 A load. From $I{=}C\,dV/dt$ we get (with $\Delta t{=}8.33\,\text{ms}$) $C{=}I\Delta t/\Delta V = 16,700\,\mu\text{F}$. The nearest capacitor voltage ratings are 63 V and 80 V; we chose the latter, in an abundance of caution. The available 16,000 μF/80 V capacitors are somewhat large physically (40 mm diameter × 80 mm long), so we decided to put a pair of 8200 μF capacitors (35 mm × 50 mm) in parallel (using smaller capacitors in parallel

[46] Although a transformer with a large leakage inductance might seem advantageous (because it increases the conduction angle losslessly), it has the undesirable effect of degrading voltage regulation under load; it also introduces a phase lag in the input current relative to the voltage, thus reducing the power factor. Furthermore, leakage inductance causes nasty voltage spikes to appear, owing to diode reverse recovery, as described in §*9x.6*.

also reduces the overall series inductance of the capacitor). Good design practice calls for the use of storage capacitors whose ripple current rating is conservatively larger than the value estimated from the dc output current and the conduction angle. In the above circuit, for example, we designed for a maximum dc load current of 2 A, from which we estimated an rms current of about 4 A in both the transformer secondary and the storage capacitor. The particular capacitors shown in the figure have a ripple current rating of 5.8 Arms at 85°C for each 8200 μF capacitor of the parallel pair, thus 11.6 Arms when combined to make either C_1 or C_2. This is definitely conservative! You can also calculate the heating, from the ESR specification of 0.038 Ω (maximum) per capacitor: each parallel pair has an ESR no greater than 19 mΩ, which produces a heating power of $P = I^2_{\mathrm{rms}} R_{\mathrm{ESR}} \approx 0.15$ W in each capacitor.

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 of rms current to average current). It also increases stress on the rectifiers. But watch out for loose capacitance tolerance: although the capacitors we used here have a rated tolerance of ±20%, electrolytic storage capacitors can be as loose as +100%/−30%.

The resistors R_2 and R_3 across the output in Figure 9.49 serve two purposes: they provide a minimum load (to keep the unloaded output from "soaring"); and they act as "bleeders" to discharge the capacitors when the unloaded supply is turned off. 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.

B. Rectifier

The first point to be made is that the diodes used in power supplies (usually referred to as "rectifiers") are quite different from the small 1N914- or 1N4148-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 100 mA, 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 A to 25 A or more, and breakdown voltage ratings going from 100 V to 1000 V or more. They have relatively high leakage currents (in the range of microamps to milliamps) and plenty of junction capacitance. General-purpose rectifiers of the sort used in Figure 9.49 are not intended for high speed, unnecessary for operation at the powerline frequency of 60 Hz. By contrast, in *switching* power supplies it's necessary to use high-speed rectifiers because of the characteristic switching frequencies of 20 kHz–1 MHz; there the use of "fast recovery" or Schottky-barrier rectifiers (or MOSFETs used as "synchronous rectifiers"[47]) is universal.

Typical of general-purpose rectifiers are the popular 1N4001–1N4007 series, rated at 1 A, and the 1N5400–1N5408 series, rated at 3 A, with reverse-breakdown voltages ranging from 50 to 1000 volts. The 1N5817–1N5822 series of Schottky rectifiers come in axial lead packages, with current ratings of 1–3 A, and voltage ratings of 20–40 V. Rectifiers with higher current ratings require heatsinking, and come in packages similar to power transistors (TO-220, D^{2}PAK, stud-mount, etc). Examples are the MBR1545 and 30CTQ045 dual Schottky rectifiers (available in TO-220 or D^{2}PAK power packages), rated at 15 A and 30 A, respectively, at 45 V, and the MUR805 to MUR1100 6 A rectifiers (in TO-220 packages), with voltage ratings to 1 kV. Plastic-encapsulated bridge rectifiers are quite popular also, with lead-mounted 1 A to 6 A types, and heatsink mountable packages in ratings up to 35 A or more.[48]

C. Damping network

The series *RC* across the transformer secondary in Figure 9.49 is often omitted, but it shouldn't be. This simple linear unregulated dc supply has the surprising ability to generate substantial microsecond-scale voltage spikes, which can create strong 120 Hz interference and other forms of mischief. It turns out that a pair of non-ideal characteristics (transformer leakage inductance, combined with rectifier reverse recovery time) work together to create a train of periodic sharp spikes, whose amplitude can be tens of volts. This nasty effect is easily tamed with a series *RC* "snubber" network, as shown. There's some interesting stuff going on here; you can read more about it (and see a dramatic example) in §*9x.6*.

9.5.4 Unregulated split supply – on the bench!

We built the power supply of Figure 9.49, mostly out of curiosity to see how closely the actual device compared

[47] Sometimes called *active rectifiers*.

[48] An interesting option for implementing an efficient bridge rectifier is the use of four MOSFETs as synchronous switches; their gate control signals can be generated conveniently with an elegant device like the LT4320 "Ideal Diode Bridge Controller," which senses zero crossings and does the right thing on its gate-control output pins. Check out its datasheet.

with our predictions. Figure 9.51 shows the ac voltage and current at one end of the transformer secondary, and the positive dc output voltage, with the power supply driving ±2 A resistive loads. The waveforms are pretty much as expected: (a) the ripple voltage is about 0.8 Vpp, somewhat less than our 1 Vpp estimate; our calculation was conservative, though, because we assumed the storage capacitors had to supply output current for a complete half-period ($1/2f_{ac} \approx 8$ ms), whereas in reality recharging begins after ∼6 ms; (b) the dc output voltage (54 V) is somewhat higher than expected, probably because the transformer voltage rating is for the full-rated load current of 4 A, and also because the powerline voltage in our lab was 3% above nominal; with no load the output rose to 60 V, typical of unregulated supplies; (c) the transformer current is confined to a fairly narrow conduction angle (about 60° of each 180° half-cycle), as expected; during conduction the ac waveform at the transformer secondary is flattened by the heavy load current because of the combined effects of leakage inductance and winding resistance.[49]

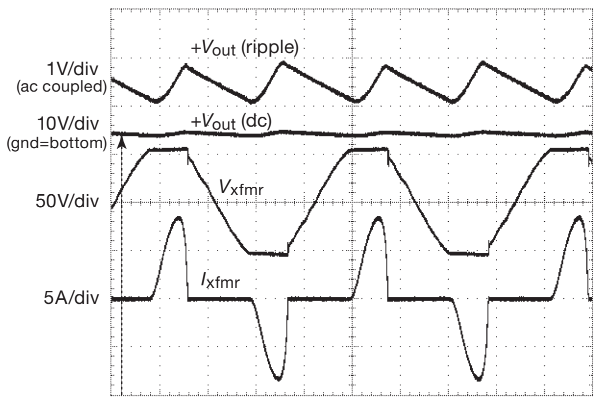


Figure 9.51. Measured waveforms for the unregulated dc power supply of Figure 9.49 driving ±2 A loads. Horizontal scale: 4 ms/div.

With a 2 A dc load on both outputs, the measured rms transformer current was 3.95 Arms. This doubling is caused by the shortened conduction angle: the *average* transformer current equals the dc output current, but the *rms* current is greater. This is sometimes described as a reduced *power factor* (the ratio of average input power to rms input power), an effect that is important in switching power supplies. With some cleverness it is possible to rectify incoming powerline ac to dc while maintaining nearly unity power factor, by means of a "power-factor correction" (PFC) input circuit; we'll explore this cleverness, briefly, in §9.7.1C.

And on the *computer*! (SPICE)

To explore the effects of component imperfections (winding resistance and leakage inductance in the transformer, series resistance in the capacitors) we ran a SPICE simulation (see Appendix J) of this circuit, beginning with measured parameters where possible (e.g., transformer resistance and inductances), values found in the SPICE libraries (e.g., rectifier forward voltage versus current) and plausible guesses for series resistance in the storage capacitors. With just a small amount of adjustment we obtained the simulation shown in Figure 9.52 (presented at the same scale factors as in Figure 9.51). The agreement is impressive (although the simulation somewhat underestimates the conduction angle, thus higher-than-measured transformer current).[50]

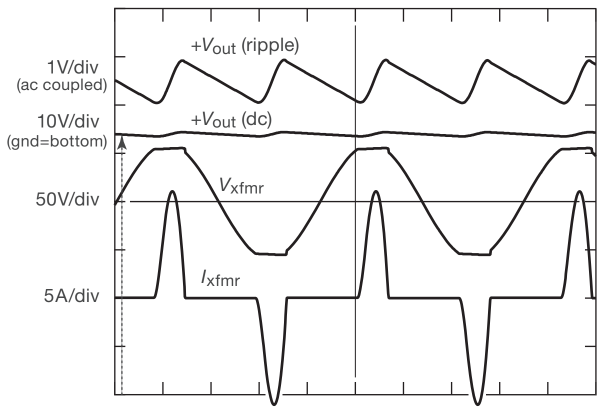


Figure 9.52. Waveforms from a SPICE simulation of the unregulated dc power supply of Figure 9.49, plotted on the same scales as those of Figure 9.51.

9.5.5 Linear versus switcher: ripple and noise

Coming next is the fascinating subject of *switching* regulators and power supplies. These have become dominant, owing to their combination of excellent efficiency, small

[49] For these waveform measurements we omitted Figure 9.49's $R_S C_S$ damping network to reveal the spike (and jump) that is visible on the transformer ac voltage waveform, caused by the combination of transformer leakage inductance and diode recovery time; see §9.5.3C and §*9x.6*.

[50] The dominant circuit parameters used are: transformer primary R=0.467 Ω, L_L=1.63 μH, L_M=80 mH, turns ratio of 0.365, transformer secondary R=0.217 Ω, $L_{L(sec)}$=20 μH, damping network C_s=0.5 μF, R_s=30 Ω, rectifier "KBPC806" (Vishay 8A, 600V bridge), storage capacitor C=14,000 μF, ESR=0.01 Ω, load resistors 27 Ω (each side).

size and weight, and low cost. However, all is not roses: the rapid switching process generates transients at the switching frequency and its harmonics, and these can be extremely difficult to filter effectively. We'll discuss this soon enough... but it's worth taking a look now at Figure 9.53, where the bad stuff on the outputs of two 5 V power supplies are compared.

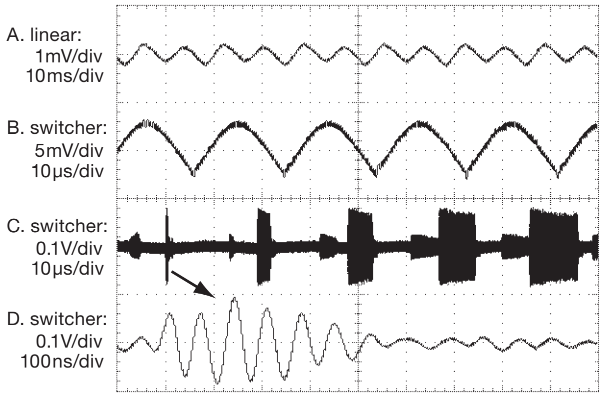


Figure 9.53. Comparing linear and switching power supply noise. All measurements are into a resistive load at 50% of rated current. A. Linear 5 V, 0.3 A supply, showing ~0.5 mVpp 120 Hz ripple. B. Switching 5 V, 2.5 A supply, measured directly across the output pins, showing ~6 mVpp ripple at the 50 kHz switching frequency (note scale change). C. Same switcher, but measured at a connected load 50 cm away (and with another factor of ×20 scale change), showing the large (~150 mV) switching spikes induced by high-frequency ground currents; note the frequency dithering seen in this persistent capture. D. Expanded trace of a single induced pulse, showing ringing at ~15 MHz.

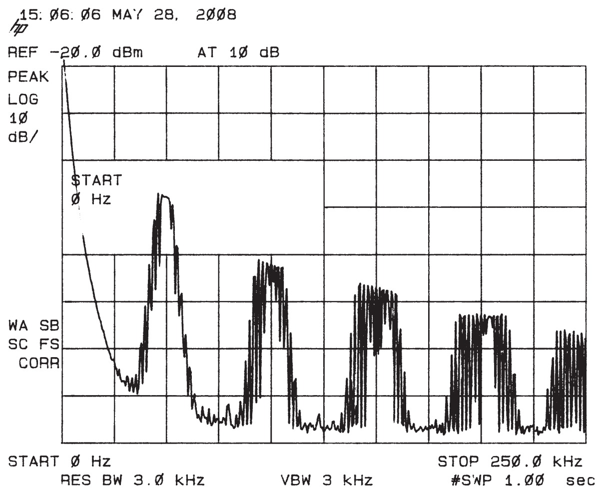


Figure 9.54. Averaged frequency spectrum of the switching power supply of Figure 9.53, showing the dithered ~50 kHz switching frequency and its harmonics.

9.6 Switching regulators and dc–dc converters

9.6.1 Linear versus switching

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 dc input is used, with feedback, to maintain constant output voltage (or perhaps constant current).[51] The output voltage is always lower in voltage than the input voltage, and significant power is dissipated in the control element, namely $P_{\text{diss}}=I_{\text{out}}(V_{\text{in}}-V_{\text{out}})$. As we've seen, the dc input to a linear regulator may be simply another (higher) regulated dc voltage within the system; or it may be unregulated dc that is derived from the powerline, via the by-now familiar transformer–rectifier–capacitor circuit.

Let's look a bit more at the issue of efficiency. Power supplies with *linear* regulators are necessarily inefficient, because the pass transistor carries the full load current, and it must have enough voltage drop to accommodate a worst-case combination of input ripple and low line voltage. The situation is exacerbated for low-output-voltage supplies: for example, a linear regulator to deliver +3.3 V at 10 A would use an unregulated dc voltage of perhaps +6 V to ensure adequate headroom; so you've got 27 W of pass-transistor dissipation while delivering 33 W to the load – that's 55% efficiency. You may not care that much about efficiency *per se*; but the wasted power has to be dissipated, which means a large heatsink area, blowers, etc. If you were to scale up this example to 100 A, say, you would have a serious problem removing the quarter of a kilowatt (!) of pass-transistor heat. You would have to use multiple pass transistors and forced-air cooling. The supply would be heavy, noisy, and hot.

There is another way to generate a regulated dc voltage (shown earlier in Figure 9.48B), which is fundamentally different from what we've seen so far – look at Figure 9.55. In this switching converter 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. When the switch is turned off, some

[51] 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.

or all[52] of this stored energy is transferred to a filter capacitor at the output, 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.[53]

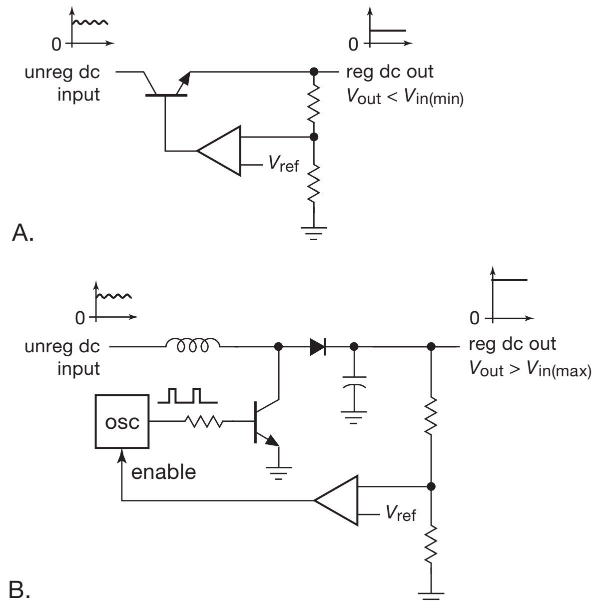


Figure 9.55. Two kinds of regulators: A. linear (series-pass); B. switcher (step-up, or "boost").

Advantages of switching converters

Switching regulators have unusual properties that have made them very popular:

(a) Because the control element is either off or saturated, there is very little power dissipation; switching supplies are thus very efficient, even when there is a large voltage difference between input and output. High efficiency translates to small size, because little heat needs to be dissipated.

(b) Switchers (slang for "switching power supplies") can generate output voltages *higher* than the unregulated input, as in Figure 9.55B; and they can just as easily generate outputs *opposite in polarity* to that of the input!

(c) The output storage capacitor can be small (in capacitance, and therefore in physical size), because the high operating frequency (typically 20 kHz–1 MHz) corresponds to a very short time interval (a few microseconds) between recharging.

(d) For a switching supply operated from the ac powerline input, the essential isolation is provided by a transformer operating at the switching frequency; it is *much* smaller than a low-frequency powerline transformer (see Figure 9.1).

The good news

The combination of small capacitor and transformer size, along with little power dissipation, permits compact, lightweight, and efficient ac-powered dc supplies, as well as dc-to-dc converters.[54] For these reasons, switching supplies (also known as *switchmode* power supplies, or SMPSs), are used almost universally in electronic devices such as computers, telecommunications, consumer electronics, battery-operated devices, and, well, just about everything electronic.

The bad news

Lest we leave too favorable an impression, we note that switching supplies do have their problems. The switching operation introduces "noise" into the dc output, and likewise onto the input powerline and as radiated electromagnetic interference (EMI); see Figures 9.53 and 9.54. Line-operated switchers (confusingly called "off-line") exhibit a rather large "inrush current" when initially powered on.[55]

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[52] All of the stored energy goes forward if the inductor current is allowed to go to zero ("discontinuous-conduction mode," DCM); you get only a portion of the stored energy in "continuous-conduction mode" (CCM), in which the inductor's current does not go to zero before the next conduction cycle.

[53] One could object that we're unfairly comparing a *step-up* switching converter circuit with an inherently "step-down" linear pass regulator. Indeed, the switching topology that is analogous in function to the linear regulator is the *buck* regulator (shown presently, in Figure 9.61A). But we like the shock value of the boost switching converter, because it's unexpected that you can even do that if you've lived exclusively in the linear world.

[54] Examples of the former include the little power "bricks" that are used for laptop computers, cellphones, and the like, as well as the more substantial power supplies built into desktop computers. Examples of the latter are the "point of load" dc–dc converters that you find clustered around the processor on a computer motherboard: the processor might require 1.0 Vdc at 60 A (!); to generate that enormous current you use a set of 12 V to 1.0 V step-down converters, right at the point of load, supplied by a lower current 12 V "bus."

[55] For an example we opened to a random page in the power-supply section of the DigiKey catalog, and found a little ac-input 5 W switcher (5 Vdc, 1 A) with a specified powerline inrush current of. . . (drumroll). . . 40 A – that's a peak power of 4*kilo*watts!

And switchers have suffered from a bad reputation for reliability, with occasional spectacular pyrotechnic displays during episodes of catastrophic failure.

The bottom line

Fortunately, switching supplies have largely overcome the drawbacks of their earlier brethren (unreliability, electrical and audible noise, inrush current and component stress). Because they are small, lightweight, efficient, and inexpensive, switchers have largely replaced linear regulators over the full range of load power (from watts to kilowatts) in contemporary electronics, and particularly in large commercial production. Linear supplies and regulators are still alive and well, however, particularly for simple low-power regulation and for applications requiring clean dc power; and this last feature – the absence of pervasive switching noise – can be of major importance in applications that deal with small signals.

9.6.2 Switching converter topologies

In the following sections we tell you all about switching regulators and power supplies (collectively called "switching converters"), in several steps.

- First (§9.6.3) we look briefly at *inductorless* converters, in which the energy is carried from input to output by capacitors, whose connections are switched with MOSFETs. These are sometimes called "charge-pump converters," or "flying-capacitor converters." These simple devices can double or invert a dc input voltage, and they're useful for relatively low current loads (up to $\sim$100 mA).
- Next (§9.6.4) we describe converter topologies that use inductors, beginning with the basic dc–dc non-isolated switching converter, of the sort you would use within a circuit, or with battery power. There are three basic circuit topologies, 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). All of these use an inductor for energy storage during the switching cycle.
- Next (§9.6.10) we look at dc–dc converters in which a transformer couples the input and output circuits. In addition to providing galvanic isolation (which may or may not be needed), the transformer is desirable when there's a large ratio between input and output voltages. That is because the transformer's turns ratio provides a helpful voltage conversion factor that is absent in the non-isolated (transformerless) designs. Transformer designs also let you produce multiple outputs, and of either polarity.
- Finally (§9.7) we describe how the isolated converter permits power-supply designs that run straight from the rectified ac powerline. These "offline" supplies are, of course, the bread and butter of most line-powered electronics. And they have their special problems, related to safety, interference, inrush current, power factor, and the like.

And, characteristically, we 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 humility, we won't leave you in any doubt!

9.6.3 Inductorless switching converters

The term "switching converter" usually means a power converter that uses inductors (and sometimes transformers), along with high-frequency transistor switches, to carry out voltage conversion. However, there is an interesting class of *inductorless* converters (also known as *charge-pump* converters, *switched capacitor* converters, or *flying-capacitor* converters) that can do some of the same tricks – generating an output voltage of opposite polarity, or an output voltage higher than the input. These converters are simpler and electrically quieter than converters with inductors, and they're handy when you need only a modest current (less than 100 mA or so). For example, you often have a source of +5 V (on a computer board, or a USB device), or perhaps +9 V from a battery, and you need a corresponding negative voltage because you want to run a dual-polarity op-amp. Just drop in a charge-pump inverter chip and two capacitors, and you're ready to go.[56]

Figure 9.56 shows how it goes: these devices have an internal oscillator and some CMOS switches, and they require a pair of external capacitors to do their job. When the input pair of switches is 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. If $C_2 \ll C_1$, then the output voltage goes nearly to $-V_{\text{IN}}$ in one cycle of operation. In the more typical case of $C_2 \geq C_1$ it takes a number of cycles, from cold start, for the output voltage to equilibrate to $-V_{\text{IN}}$.

Similarly, you can create an output of $2V_{\text{in}}$, by arranging things so that C_1 charges as before, but then gets hooked in series with V_{in} during the second (transfer) half-cycle

[56] A good reference is M.D. Seeman & S.R. Sanders, "Analysis and optimization of switched-capacitor DC–DC converters," *IEEE Trans. Power Electron.* **23** (2) pp. 841–851 (2008).

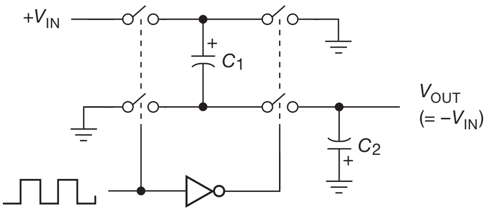

Figure 9.56. Charge-pump voltage inverter. An oscillator operates the switch pairs in alternation: the left-hand switches charge "flying capacitor" C_1 to a voltage of $V_{\rm IN}$; the right-hand switches then apply that voltage, with reversed polarity, to the output storage capacitor C_2.

(Figure 9.57). The LT1026 and MAX680 conveniently integrate a positive doubler and an inverting doubler in one package: Figure 9.58 shows the simple circuitry required to generate an unregulated split supply from a single +5 V input.

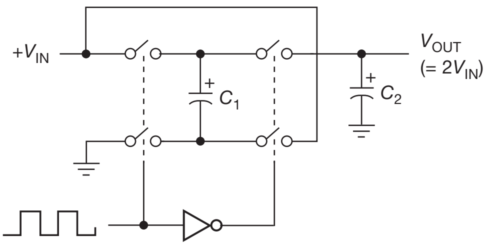

Figure 9.57. Charge-pump voltage doubler. Here the voltage on the flying capacitor, charged to $V_{\rm IN}$, is added to the input voltage to generate an output voltage of twice $V_{\rm IN}$.

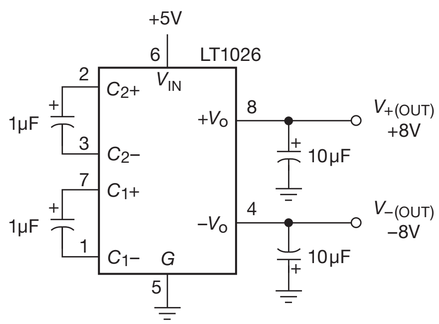

Figure 9.58. Generating a pair of unregulated ±8 V outputs from a single +5 V input.

A. Limitations of inductorless converters

This charge-pump technique is simple and efficient, and requires few parts and no inductors. However, the output is not regulated, and it drops significantly under load (Figure 9.59). Also, in common with other switching power

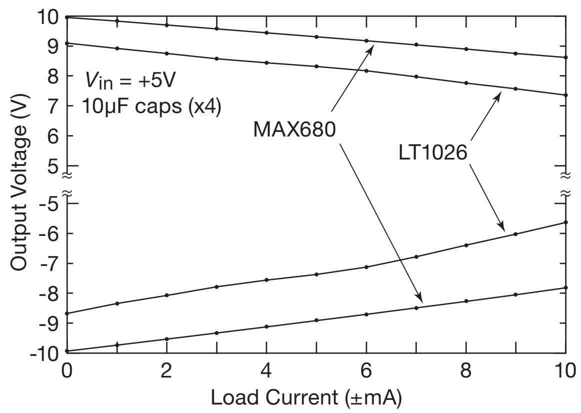

Figure 9.59. The output voltage of a charge-pump converter drops significantly under load, as seen here with measured data for the circuit of Figure 9.58, with either bipolar (LT1026) or CMOS (MAX680) devices. MOSFET switches have no voltage drop at zero current, where $V_{\rm out}$ is accurately equal to twice $V_{\rm in}$.

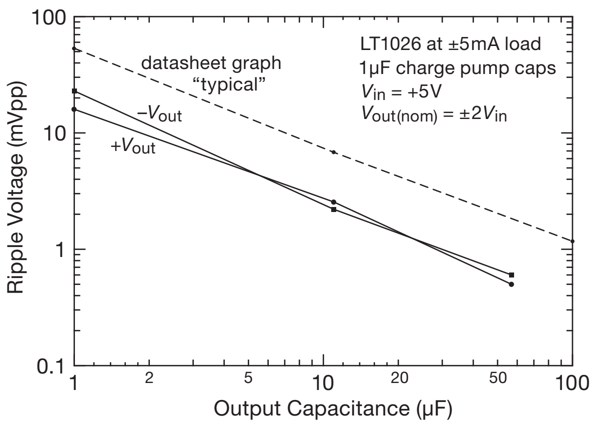

Figure 9.60. Reducing ripple with a larger output capacitor: measured peak-to-peak ripple voltage for the LT1026 doubler–inverter.

conversion techniques, the switching operation produces output ripple, which however can be reduced by using larger output capacitors (Figure 9.60), or by appending a low-dropout linear regulator (see below).[57] Furthermore, like most CMOS devices, charge pumps have a limited

[57] The ripple voltage is given approximately by $V_{\rm ripple}(pp) = I_{\rm out}/2f_{\rm osc}C_{\rm out} + 2I_{\rm out}\cdot{\rm ESR}$. The first term is just $I = C\,dV/dt$, and the second term adds the effect of the capacitor's finite equivalent series resistance.

Table 9.4 Selected Charge-pump Converters[a]

Part #	Package: DIP	SOIC	SOT23	MSOP etc	Config	V_{in} (V)	V_{out} (V)	R_{out} typ (Ω)	@ V_{in} (V)	I_{out}[z] (mA)	f_{osc} typ (kHz)	I_q typ (mA)	@ V_{in} (V)	Comments
unregulated														
LTC3261	-	-	-	•	inv	4.5-32	track	35	12	50[e]	50-500[p]	7	15	HV
TC962	•	•	-	-	inv, x2, x0.5	3-18	track	32	15	80	12 or 24[p]	0.5	15	improved 7660/2
LTC1144	•	•	-	-	inv, x0.5	2-18	track	56	15	50	10 or 100[p]	1.1[m]	15	HV ver of 1044/7660/62
ICL7662[o]	•	•	-	-	inv, x0.5	9-20 or 4.5-11	track	55	15	50	10	0.15	12.5	Maxim, orig Intersil
TC1044[k]	•	•	-	-	inv, x0.5	3-12 or 1.5-3.5	track	55	15	60	10 or 45[p]	0.15	12.5	improved 7660; see (k)
LM2681	-	-	•	-	x2, x0.5	2.5-5.5[g] or 1.8-11[h]	track	15	5	30	160	0.06	5	
ICL7660[k]	•	•	-	-	inv, x0.5	3-10 or 1.5-3.5	track	30	10	40	10 or 35[p]	0.08	5	***classic***, 5 manuf, see (k)
LT1026	•	•	-	-	inv & x2	4-10	track	b	-	20	-	15	15	
MAX680	•	•	-	-	pos & neg x2	2-6	track	100[c]	5	5	8	1	5	MAX864 for 200kHz
MAX864	-	-	-	•	pos & neg x2	1.8-6	track	40[c]	5	15[d]	7-185[p]	0.6-12[p]	5	
LM828	-	-	•	-	inv	1.8-5.5	track	20	5	25	12	0.04	5	
LM2767	-	-	•	-	x2	1.8-5.5	track	20	5	25	11	0.04	5	
TPS6040x	-	-	•	-	inv	1.6-5.5	track	10	3	60	20-250[i,q]	0.06-0.4[q]	5	f_{osc} variable ('60400)
MAX660	•	•	-	-	inv, x2	1.5-5.5	track	6.5	5	100	10 or 80[p]	0.12	5	
regulated[r]														
LTC3260	-	-	-	•	dual LDO	4.5-32	1.2-32 & -1.2 to -32	0.03	12	50	200	4	15	HV dual reg split supply
LT1054	•	•	-	-	inv	3.5-15	$-V_{in}$, or adj reg	10	-	100	25	3	15	
ADP3605		•	-	•	inv	3-6	−3.0, or −3 to −6	0.3	5	120	250	3	5	
ST662		•	-	-	reg 12V	4.5-5.5	12	0.8	5	50	400[i]	0.1	5	flash mem prog supply
MAX889		•	-	-	reg adj -Vout	2.7-5.5	−2.5 to $-V_{in}$	0.05	5	200	500-2000[x]	6	5	
MAX682		•	-	•	reg 5V	2.7-5.5	+5V	<1	3	250	20-3000[i,p]	7.5	3.6	
REG710-vv	-	-	•	-	buck-boost	1.8-5.5	2.5, 2.7,…,5.5[vv]	2	n	30	1000[i]	0.07	3.3	auto switch buck/boost
MAX1595-vv	-	-	-	•	buck-boost	1.8-5.5	3.3 or 5.0[vv]	1	3	125	1000[i]	0.23	3	auto switch buck/boost
TPS6024x	-	-	-	•	buck-boost	1.8-5.5	2.7, 3, 3.3, 5[u]	0.7	3.3	40	160	0.25	3	low noise[v]
LTC1517-5	-	-	•	-	reg 5V	2.7-5	5.0	1	3	50	800	0.006	all	micropower reg 5V
LTC3200	-	-	•	•	reg 5V	2.7-4.5	+5V	0.4	3.6	100	2000[i]	3.5	3.6	
LTC1682	-	•	-	•	LDO, adj Vout	1.8-4.4	2.5-5.5	0.2	3	50	550	0.15	3	x2 to LDO; low noise[w]
LTC1502-3.3	-	•	-	•	reg 3.3V	0.9-1.8	3.3	<0.2	1	20[s]	500[i]	0.04	1	single-cell to reg +3.3V
TPS6031x	-	-	-	•	x2 & reg 3[y]	0.9-1.8	3.0, 3.3[u]	4[w], 0.3	1	50	700[i]	0.03	1.5	single-cell to reg +3V[y]
NJU7670	•	•	-	-	neg x3 & LDO	−2.6 to −6	−8 to −18	5	−5	20	2.5	0.08	-5	neg V_{in}, tripler plus LDO

Notes: (a) all are inductorless, and require several external caps; "regulated" types include either internal linear LDO post-regulator, or regulation via control of switching; sorted within categories by decreasing maximum V_{in}. (b) bipolar, see datasheet for typical V_{out}. (c) with other output unloaded. (d) both outputs loaded. (e) at max f_{osc}. (g) in x2 mode. (h) in x0.5 mode. (i) high f_{osc} allows small capacitors. (k) LV pin for low V_{in} range; many mfgs, prefixes LMC, NJU, TC, TL; see also MAX/LTC/TC1044, 1144, and TC962. (m) maximum. (n) at $V_{in}=V_{out}/2+0.8V$. (o) or Si7661. (p) freq pin selectable or adjustable. (q) last digit of p/n sets f_{osc}, except TPS60400, where f_{osc} varies cleverly with V_{in} and I_{out}. (r) unreg outputs also available on most; unless marked "LDO," all regulate via control of switching. (s) at V_{in}=1.2V. (u) last digit of p/n sets V_{out}. (v) V_n = 170μVrms in BW = 20Hz-10MHz. (vv) suffix selects V_{out}. (w) V_n = 60μVrms in BW = 10Hz-100KHz, 600μVpp for 10Hz-2.5MHz. (x) suffix sets f_{osc}. (y) or reg +3.3V; unreg x2 output also provided. (z) maximum usable.

supply-voltage range: the original charge-pump IC (the Intersil ICL7660) allows V_{in} to range from +1.5 V to +12 V; and although some successor devices (e.g., the LTC1144) extend this range to as much as +18 V, the trend is toward lower-voltage devices with greater output current and with other features.[58] Finally, unlike *inductive* switching converters (discussed next), which can generate any output voltage you want, the flying-capacitor voltage converter can generate only small 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 split-supply op-amp or a serial-port chip (see Chapters 14 and 15) on a circuit board that has only +5 volts available. Table 9.4 lists a selection of charge-pump voltage converters, illustrating a range of capabilities (voltage, regulation, output current, and so on).

[58] For example, more than half of Maxim's offerings are limited to +5.5 V input; and of the 67 offerings from Texas Instruments, only 7 can operate above +5.5 V input (and 28 of them are limited to +3.6 V or less). The story is similar for Linear Technology's 62 charge-pump converter offerings.

B. Variations

There are interesting and useful flying-capacitor variations, many of which are listed in Table 9.4, which is organized into unregulated and regulated varieties (each sorted by maximum input voltage). The unregulated types represent variations on the original ICL7660, including its similarly named successors (from TI, NJR, Maxim, Microchip, etc.) and pin-compatible upgrades ('7662, '1044, '1144); such multiply-sourced jellybean parts are widely available and inexpensive. More recent parts, for example the low-voltage TPS6040x, offer flexibility in switching frequency, and generally lower output resistance. Operation at higher frequency reduces output ripple (e.g., 35 mV at 20 kHz, but 15 mV at 250 kHz for the TPS6040x series), but it increases quiescent current (which goes from 65 μA to 425 μA in this example).[59]

The regulated types, like the LT1054 from LTC (with a maximum output current of 100 mA), include an internal voltage reference and error amplifier, so you can connect feedback to regulate the output voltage; the internal circuitry accommodates this by adjusting the switching control. Other converters regulate the output by including an internal low-dropout linear regulator, for greatly reduced output ripple (at the expense of some additional voltage drop); examples are the LTC1550 and 1682 series, with less than 1 mV peak-to-peak output ripple. Note that most of the "regulated" types let you use them as unregulated converters, if you wish.

There are also converters that *reduce* the input voltage by a rational fraction, e.g., by a factor of 1/2 or 2/3 (see if you can figure out how that is done!). At the other end, there are converters that are voltage quadruplers, for example the LTC1502, which generates a regulated +3.3 V at 10 mA from an input of 0.9–1.8 V (e.g., to power digital logic from a single alkaline cell).[60] And there are converters that can provide up to 500 mA of output current. Some charge-pump converters include internal capacitors, if you want to be especially lazy; but the selection is limited, and the price is high.

Finally, there is the LTC1043 uncommitted flying-capacitor building block, with which you can 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 datasheet has eight pages of similarly clever applications.

Then there are integrated circuits that include charge pumps to power their primary functions:

(a) Many RS-232/485 driver–receiver chips are available with integral $\pm$10 V charge-pump supplies, to run from a single +5 V or +3.3 V supply. An example of the latter is the MAX3232E from Maxim (the originator of the MAX232, now widely second sourced), which can run from a single supply between +3 V and +5.5 V.

(b) Some op-amps use integral charge pumps to generate a voltage beyond the supply rail, so their inputs can operate rail-to-rail while maintaining a conventional high-performance architecture (see §4.6.3B); examples are the OPA369, LTC1152, and MAX1462-4.

(c) Charge pumps are used in many MOSFET "high-side drivers" (like the HIP4080 series from Intersil) and in fully integrated power MOSFETs (like the PROFET series of "smart highside high-current power switches" from Infineon); these generate the necessary above-the-rail gate bias for an *n*-channel MOSFET operating as a follower up at the positive rail.[61]

(d) Some complex digital logic devices (processors, memory) require elevated voltages, which they generate on-chip with charge pumps. The manufacturers are modest, and you don't even hear about these things.

9.6.4 Converters with inductors: the basic non-isolated topologies

The term *switching converter* (or *switchmode converter*[62]) is generally understood to mean a converter that uses some arrangement of inductors and/or transformers, in combination with transistor switches (usually MOSFETs, but also

[59] You can reduce ripple by using much larger output capacitors (with low ESR, to minimize the effect of current spikes), or, perhaps better, an output filter stage.

[60] Sadly, there are no charge-pump converters that take a single-cell input (0.9 V at end of life) up to +5 V (that would require at least a $\times$6 voltage conversion), although you could accomplish that task in two steps by cascading, say, a TPS60310 (0.9–3.3 V) with a TPS60241 (3.3–5 V). That would require two ICs and seven capacitors. But, happily, such a task is easily done with a boost-mode *inductive* switching converter (§9.6.6). For example, TI's TPS61222 comes in a tiny 6-pin SC-70 package, requires only a single external 4.7 μH inductor (plus input and output bypass caps), and delivers +5 V at 50 mA with 0.9 V input (an alkaline cell's end-of-life voltage). It costs less than \$2 in single-piece quantities. Another approach to powering from a single alkaline cell is to use a charge-pump converter to generate +3.3 V, which then powers one or more inductive switching converters to generate the full set of voltages you need; the ENABLE input to the charge-pump converter can then be used to turn power on or off.

[61] For further detail see §3.5.3 and Figures 3.96 and 3.106.

[62] Switchmode power supplies are referred to as SMPSs, thus phrases like "SMPS technology."

IGBTs[63] for high voltages), to carry out efficient dc-to-dc conversion. A common characteristic of all such converters is this: in the first portion of each switching cycle the source of input power is used to increase the current (and therefore the energy) in an inductor; that energy then flows to the output during the second portion of the switching cycle. Switchmode power conversion is a major and vital area of electronics, and these converters are used in just about every electronic device.

There are literally hundreds of switchmode circuit variations, but they can be pared down to a few fundamental topologies. In this section we describe the three basic *non-isolated* designs – step-down, step-up, and invert – shown in Figure 9.61. After that we look at isolated converter designs; then we conclude with a look at the use of isolated converters fed from the ac powerline. Tables of selected switchmode converters (Tables 9.5a,b on pages 653 and 654) and controllers (Table 9.6 on page 658) appear later.

Along with the basic power-conversion *topologies* (which describe the circuitry that performs the voltage conversion itself), there is the important topic of *regulation*. Just as with linear voltage regulators, a sample of the output voltage is compared with a voltage reference in an *error amplifier*. Here, however, the error signal is used to adjust some parameter of the switching conversion, most often the pulse width; this is known as *pulse width modulation* (PWM).[64]

As we'll see, the pulse-width modulator circuits themselves fall into two categories (voltage mode and current mode), with important consequences in terms of response time, noise, stability, and other parameters. And, to introduce a bit of further complication, any of these switchmode circuit combinations may operate in a mode with the inductor's current dropping fully to zero by the end of each switching cycle, or in a mode in which the inductor's current never drops to zero. These modes of operation are known as *discontinuous-conduction mode* (DCM) and *continuous-conduction mode* (CCM), respectively, and they have major effects on feedback stability, ripple, efficiency, and other operating parameters of a switchmode regulator. We describe the basics of PWM with a few examples; but we will touch only lightly on the more advanced topics of voltage- versus current-mode PWM, and on loop compensation.

[63] Insulated-gate bipolar transistors, §3.5.7A.

[64] In some switchmode converters the regulation is done instead by varying the pulse *frequency*.

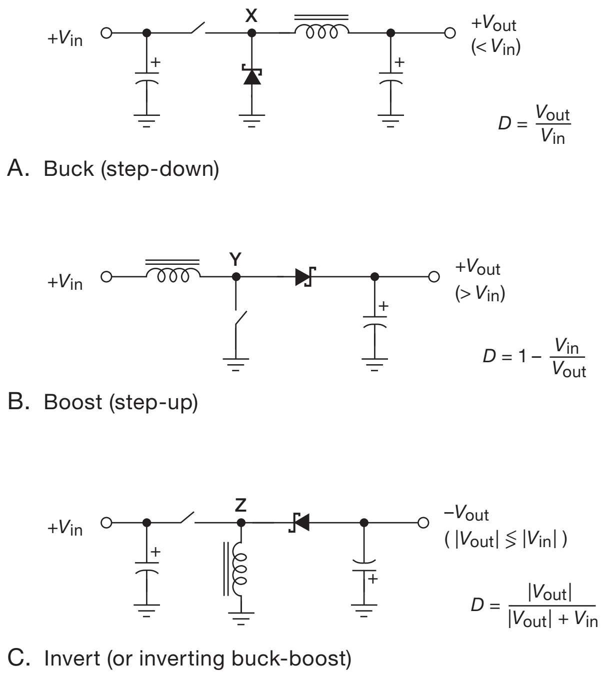

Figure 9.61. The basic nonisolated switching converters. The switch is usually a MOSFET. Schottky diodes are commonly used for the rectifiers, as shown; however, a MOSFET can be used as an efficient synchronously switched "active rectifier."

9.6.5 Step-down (buck) converter

Figure 9.61A shows the basic step-down (or "buck") switching circuit, with feedback omitted for simplicity. When the 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" (or "freewheeling diode") now conducting to complete the circuit. The inductor now finds a fixed voltage $V_{out}-V_{diode}$ across it, causing its current to decrease linearly. The output capacitor acts as an energy "flywheel," smoothing the inevitable sawtooth ripple (the larger the capacitor, the smaller the ripple voltage). Figure 9.62 shows the corresponding voltage and current waveforms, assuming ideal components. 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

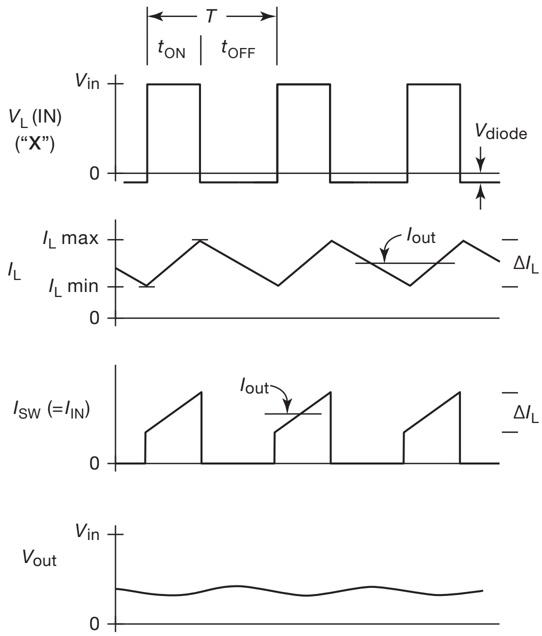

Figure 9.62. Buck converter operation. Inductor current ramps up during switch ON, and ramps down during switch OFF. The output voltage equals the input voltage times the duty cycle ($D \equiv t_{on}/T$). In the case of continuous inductor current (CCM; as shown here) the output current is equal to the average inductor current.

repetition rate (with constant pulse width) from an error amplifier that compares the output voltage with a reference.[65]

For all three circuits of Figure 9.61 the voltage drop across the catch diode wastes energy, reducing the conversion efficiency. Schottky diodes (as shown) are often used to mitigate this, but the best solution is to add a second switch across or in place of the diode. This is called *synchronous switching*; see the "synchronous" column in Tables 9.5a,b and 9.6.

Output voltage What is the output voltage? In the steady state the average voltage across an inductor must be zero, because otherwise its current is continually growing (according to $V=LdI/dt$).[66] So, ignoring voltage drops in the diode and switch, this requires that $(V_{in}-V_{out})\,t_{on} = V_{out}t_{off}$, or

$$V_{out} = DV_{in}, \qquad (9.3)$$

where the "duty cycle" (or "duty ratio") D is the fraction of the time the switch is ON, $D=t_{on}/T$, and T is the switching period ($T=t_{on}+t_{off}$).

You can think about this in another way: the *LC* output network is a lowpass filter, to which is applied a chopped dc input whose average voltage is just DV_{in}. So, after smoothing, you get that average voltage as the filtered output. Note that, assuming ideal components, the output voltage from a buck converter running at fixed duty cycle D from a fixed input voltage is intrinsically regulated: a change in load current does not change the output voltage; it merely causes the inductor's triangular current waveform to shift up or down, such that the average inductor current equals the output current. (This assumes continuous inductor current, or CCM, as we discuss below.)

Input current What is the input current? If we assume ideal components, the converter is lossless (100% efficiency), so the input power must equal the output power. Equating these, the average input current is $I_{in}=I_{out}(V_{out}/V_{in})$.[67]

Critical output current We've been assuming continuous inductor conduction in the waveforms of Figure 9.62, and also in deducing that the output voltage is simply the input voltage times the switch duty cycle. Look again at the graph of inductor current: its average current must equal the output current, but its peak-to-peak variation (call it ΔI_L) is completely determined by other factors (namely V_{in}, V_{out}, T, and L); so there is a *minimum output current* for which the inductor stays in conduction, namely when $I_{out} = \frac{1}{2}\Delta I_L$.[68] For output currents less than this critical load current, the inductor current reaches zero before the end of each cycle; the converter is then operating in discontinuous conduction mode, for which the output voltage would no longer remain stable at fixed duty cycle, but would depend on load current. Of greater importance, operating in DCM has a major effect on loop stability and regulation. For this reason many switching regulators have a minimum output current, in order to operate in CCM.[69] As the following expressions show, the minimum load current for CCM is reduced by increasing the inductance, increasing the switching frequency, or both.

[65] There is also hysteretic control, in which both pulse width and switching frequency may vary.

[66] Engineers like to say that the *volt–time product* (or the *volt–second product*) must average to zero.

[67] In real converters the efficiency is reduced by losses in the inductors, capacitors, switches, and diodes. It's a complicated subject.

[68] Operation at this current is called *critical conduction mode*.

[69] At load currents less than the minimum current for CCM they may enter other modes of operation, including "burst mode."

A. Buck converter equations (continuous-conduction mode)

From the preceding discussion and waveforms it is not terribly difficult to figure out that the ideal buck converter (Figure 9.61A), operating in continuous conduction mode, obeys these equations:

$$\langle I_{\text{in}}\rangle = I_{\text{out}}\frac{V_{\text{out}}}{V_{\text{in}}} = DI_{\text{out}}, \tag{9.3a}$$

$$\Delta I_{\text{in}} = I_{\text{out}}, \tag{9.3b}$$

$$V_{\text{out}} = V_{\text{in}}\frac{t_{\text{on}}}{T} = DV_{\text{in}}, \tag{9.3c}$$

$$D = \frac{V_{\text{out}}}{V_{\text{in}}}, \tag{9.3d}$$

$$\begin{aligned} I_{\text{out(min)}} &= \frac{T}{2L}V_{\text{out}}\left(1-\frac{V_{\text{out}}}{V_{\text{in}}}\right) \\ &= \frac{T}{2L}V_{\text{out}}(1-D), \end{aligned} \tag{9.3e}$$

$$\Delta I_{\text{C(out)}} = \frac{T}{L}V_{\text{out}}(1-D), \tag{9.3f}$$

$$I_{\text{L(pk)}} = I_{\text{out}} + \frac{T}{2L}V_{\text{out}}(1-D), \tag{9.3g}$$

$$L_{\text{min}} = \frac{T}{2}\frac{V_{\text{out}}}{I_{\text{out}}}(1-D), \tag{9.3h}$$

where $\langle I_{\text{in}}\rangle$ represents the time-averaged value of input current, and ΔI_{in} and $\Delta I_{\text{C(out)}}$ are the approximate peak-to-peak ripple currents at input and output (important for capacitor selection[70]). The first equation holds regardless of mode (CCM or DCM). The expressions for minimum inductance and minimum output current represent the critical values to maintain CCM; for these expressions use the minimum output current and the maximum value of V_{in}, respectively.

Exercise 9.8. Take the challenge: derive these equations (and be sure to tell us if we got them wrong). *Hint*: for $I_{\text{out(min)}}$ and L_{min} use the fact that the output current I_{out} equals half the peak-to-peak inductor current variation ΔI_{L}, at the threshold of CCM, as easily seen from the I_{L} waveform in Figure 9.62.

B. Buck converter example – I

Let's do a buck regulator design, using a very simple (and inexpensive) controller chip, the MC34063 (Figure 9.63). This controller dates back to the 1980s and costs about \$0.50. In spite of its ancient heritage, the MC34063 is quite popular for undemanding applications, because of its low price and simple design criteria; this 8-pin part is manufactured by a half-dozen companies, and is supplied in the usual package styles (DIP, SOIC, SOP). It includes an oscillator, error amplifier and voltage reference, current-limit comparator, and a Darlington output pair with access to both collector and emitter. Its operation is unsophisticated: it does not use the more usual PWM (in which the switch conduction time during each cycle is varied continuously, as in Figure 9.72). Instead, switch conduction cycles are enabled as long as the voltage at the feedback (FB) input is less than the +1.25 V internal reference; otherwise they are inhibited. You can think of this as a crude form of PWM, in which the modulation consists of turning on the switch for a full cycle, then skipping enough cycles to approximate the needed ratio of switch ON/OFF.[71] This feedback regulation scheme is known as *hysteretic* control.

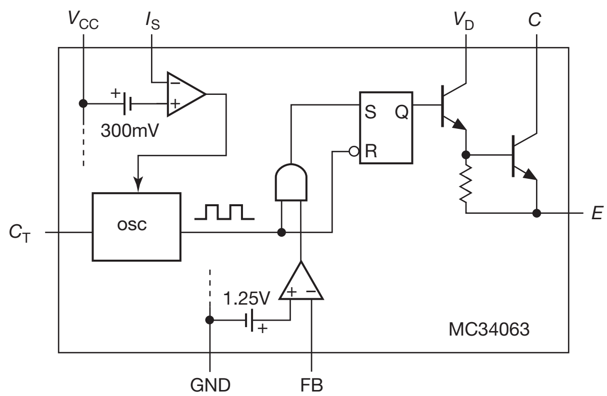

Figure 9.63. A popular \$0.50 switching converter. The external connections to both collector and emitter of the 1.5 A switch make it easy to implement buck, boost, or inverting converters.

For our design let's assume a +15 V input, and produce a +5 V regulated output for load currents up to 500 mA. Figure 9.64 shows the circuit. The design is straightforward:

1. Choose an operating frequency: we picked 50 kHz, half the chip's recommended maximum. For that frequency the datasheet specifies $C_{\text{T}} = 470\,\text{pF}$. The oscillator runs with a ratio $t_{\text{on}}/t_{\text{off}} = 6$, so the switch conduction time is $t_{\text{on}} = 17\,\mu\text{s}$.
2. Calculate the inductor value so the converter operates

[70] Note that capacitor datasheets specify maximum allowed *rms* ripple current, rather than peak-to-peak. Be sure to allow a large safety margin in this parameter when selecting input and output capacitors for power conversion.

[71] This is analogous to "bang–bang" feedback control, as contrasted with proportional (or PID) control in which the feedback signal operates in a continual manner.

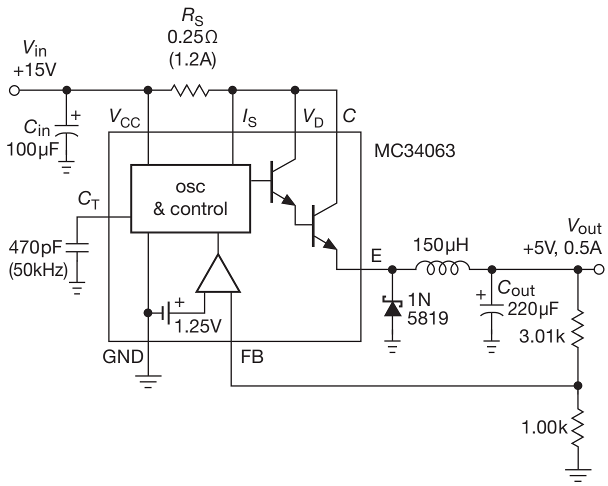

Figure 9.64. Step-down regulator using the MC34063. In contrast to proportional PWM, the chip's simple bang–bang control eliminates the need for feedback compensation components. But performance suffers.

in DCM,[72] assuming onset of CCM at minimum input voltage and maximum load current: at onset of CCM, the output current is half the peak inductor current, so, using $V = LdI/dt$ (and assuming a 1 V drop in the Darlington switch), we get $L = (V_{in} - V_{sw} - V_{out})t_{on}/2I_{out} = 153\mu$H. We'll use a standard value of 150 μH.

3. Calculate the value of sense resistor R_S to limit the peak current I_{pk} to somewhat greater than the expected 1 A, but no greater than the chip's 1.5 A rating: $R_S = 300\,\text{mV}/I_{lim} = 0.25\Omega$ (for a 1.2 A current limit).[73]

4. Choose an output capacitor value to keep the ripple voltage below some acceptable value. You can estimate the ripple by calculating the capacitor's voltage rise during one cycle of switch conduction (during which its current goes from 0 to I_{pk}), which gives a value $\Delta V = I_{pk}t_{on}/2C_{out}$. So an output capacitor of 220 μF results in a peak-to-peak ripple voltage of ~40 mV.[74]

Several comments. (a) This simple design will work, but the performance will be far from ideal. In particular, the crude bang–bang control, combined with discontinuous-conduction operation, produces lots of output ripple, and even audible noise, caused by its intermittent pulsing. (b) The Darlington output connection prevents saturation in the output stage, with some loss of efficiency; this could be remedied by connecting the driver collector line (V_D) to the input supply, through a current limiting resistor of the order of 200 Ω. (c) The internal switch is limited to 1.5 A peak current, which is inadequate for output currents greater than 0.75 A; this can be remedied with an external transistor switch, for example a *pnp* transistor or *p*-channel MOSFET (for this buck configuration). The main attractiveness here is the combination of very low cost, and lack of worries about feedback stability and compensation. You'll see this part used in relaxed applications such as cellphone chargers and the like.[75]

C. Buck converter example – II

Fortunately, there are very nice integrated switchers that implement proportional PWM and, furthermore, make it really easy to do a circuit design (many are listed in Tables 9.5a,b, discussed later). For example, National Semiconductor (part of Texas Instruments) has a series of "Simple Switcher™" ICs, individually configured for buck, boost, or invert topologies, that include all the necessary feedback loop compensation components on-chip.[76] They cover a voltage range up to 40 V or more, with currents to 5 A, and have built-in current limit, thermal limit, voltage reference, fixed-frequency oscillator, and (in some versions) features such as soft-start (see §9.6.8G), frequency synchronization, and shutdown. Best of all, they make it dead simple to design a converter either by following the step-by-step recipes in the datasheets or by using free web-based design tools: you get the component values (including recommended component manufacturers' part numbers) and performance data.

Figure 9.65 shows such a design, in this case converting a 14 V input (from an automobile battery) to a +3.3 V output that can supply up to 5 A (to power digital logic). We followed the datasheet's recipe to get the component values and part numbers shown. With these components the efficiency is 80% and the output ripple is less than 1% of V_{out} (~30 mV).

The LM2677 we used (and other "simple switcher" successors) follow on from the original LM2574,75,76 series

[72] That is, the inductor current ramps completely to zero during each switch cycle.

[73] If you find that the expected peak current is greater than the chip's limit, you will have to append an outboard transistor, or (better) use a different chip.

[74] The actual ripple voltage will be higher because of the capacitor's ESR, an effect that can also be estimated.

[75] Those who are struggling with an under-performing circuit based on an MC34063A should consider the NCP3063, a drop-in upgrade that operates to 150 kHz. This allows you to reduce the inductor size and deliver higher output currents.

[76] See for example the block diagram in the LM2677's datasheet, and associated patents for the active inductor (US patent 5,514,947) and active capacitor (US patent 5,382,918).

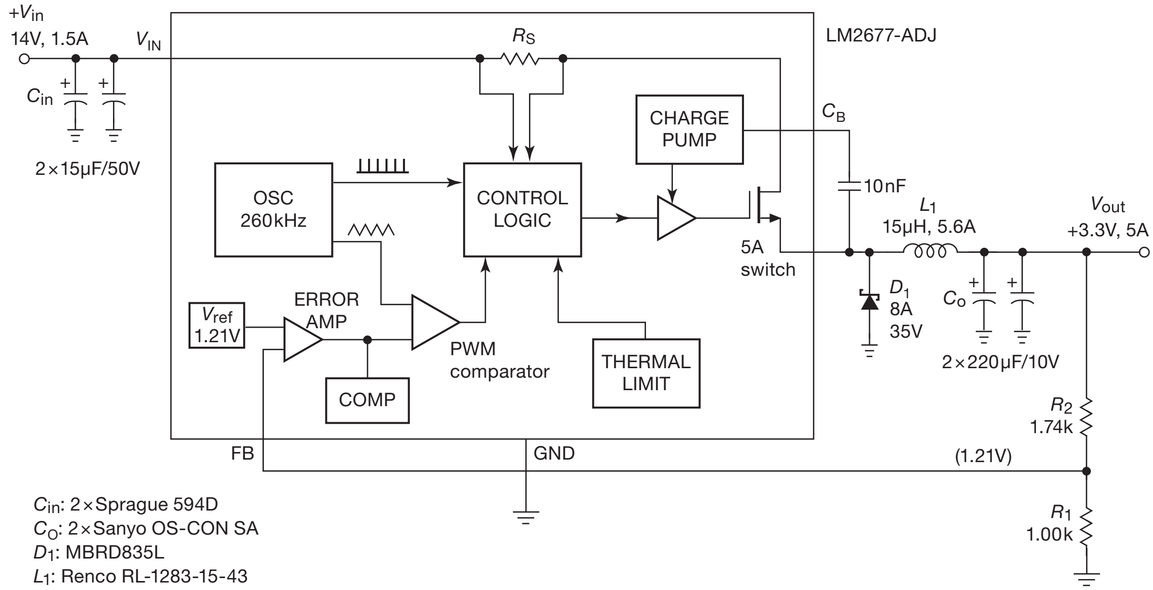

Figure 9.65. Step-down regulator using the LM2677 "Simple Switcher" (complete with elegant built-in compensation). We followed the datasheet's design recipe to get the component values and recommended part numbers shown.

(0.5 A, 1 A, and 3 A, respectively), which run at 52 kHz and which are widely popular "jellybean" parts – they are inexpensive and available from many manufacturers.[77] The LM2677 is a member of the improved LM2670 family, running at 260kHz, with output-current ratings to 5 A; it requires one additional capacitor (C_B in the figure) to drive the 5 A low-drop MOSFET.

Several comments:

(a) This converter provides ten times the output current of the previous design (Figure 9.64), and with significantly improved performance in terms of regulation, ripple, and transient response. That comes at a cost (literally), namely an IC that costs ten times as much (about \$5, versus \$0.50).[78]

(b) The good efficiency is due in part to the use of an *n*-channel MOSFET whose gate is driven from a voltage higher than V_{in}, thanks to an internal charge pump; that's the purpose of the boost capacitor C_B.

(c) Note the use of paralleled capacitors at the input and output. You see this often in switchmode converters, where it's important to keep ESR and ESL (equivalent series inductance) low: that reduces the voltage ripple caused by ripple current, and also keeps the capacitors within their ripple-current ratings.[79]

(d) For a standard output voltage like the +3.3 V here, you can save two resistors by selecting a fixed-voltage version (LM2677-3.3); but the adjustable version (LM2677-ADJ) lets you choose your output voltage, and you don't have to keep multiple versions in stock in your laboratory.

(e) Note that the input current is a lot less than the output current, representing a power-conversion efficiency of 80%; this is a major advantage over a linear regulator.

(f) Fixed efficiency means that if you increase the input voltage, the input current goes *down*: that's a negative resistance! This creates some amusing complications – for example you can get oscillation when the input is filtered with an *LC* network, a problem that applies to ac powerline input converters as well.

Exercise 9.9. What is the maximum theoretical efficiency of a linear (series pass) regulator, when used to generate regulated +3.3 V from a +14 V input?

Exercise 9.10. 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?

[77] And ON semiconductor has introduced the compatible NCV2576 family, low-cost parts rated specifically for the automotive market.

[78] Power converter ICs vary over an enormous price range; the approximate prices listed in this chapter's tables can provide some guidance in their selection.

[79] It also assists in creating a desirably low physical profile.

9.6.6 Step-up (boost) converter

Unlike linear regulators, switching converters can produce output voltages higher than their input. The basic non-isolated step-up (or "boost") configuration was shown in Figure 9.61B (repeated here as 9.66, and seen earlier, in Figure 9.55, in comparison with the linear regulator). During switch conduction (point Y near ground) the inductor current ramps up; when the switch is turned off, the voltage at point Y 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.

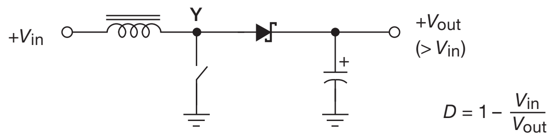

Figure 9.66. Basic boost (or "step-up") topology (non-isolated).

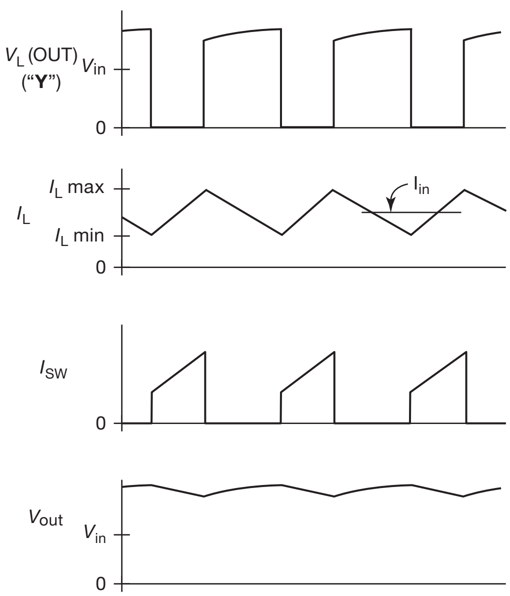

Figure 9.67. Boost converter operation. Inductor current ramps up during switch ON, and ramps down during switch OFF. The output voltage equals the input voltage divided by the fraction of the time the switch is OFF. In the case of continuous inductor current (CCM, as shown here) the input current is equal to the average inductor current.

A. Boost converter equations (continuous-conduction mode)

Figure 9.67 shows relevant voltage and current waveforms, assuming ideal components. As with the buck converter, it is not terribly difficult to figure out that the boost converter (Figure 9.61B), operating in continuous conduction mode, obeys these equations:

$$\langle I_{\rm in}\rangle = I_{\rm out}\frac{V_{\rm out}}{V_{\rm in}} = \frac{I_{\rm out}}{1-D}, \tag{9.4a}$$

$$\Delta I_{\rm in} = \frac{T}{L}V_{\rm in}D, \tag{9.4b}$$

$$V_{\rm out} = V_{\rm in}\frac{T}{t_{\rm off}} = \frac{V_{\rm in}}{1-D}, \tag{9.4c}$$

$$D = 1-\frac{V_{\rm in}}{V_{\rm out}}, \tag{9.4d}$$

$$\begin{aligned} I_{\rm out(min)} &= \frac{T}{2L}\left(\frac{V_{\rm in}}{V_{\rm out}}\right)^2(V_{\rm out}-V_{\rm in}), \\ &= \frac{T}{2L}V_{\rm out}D(1-D)^2, \end{aligned} \tag{9.4e}$$

$$\Delta I_{\rm C(out)} = \frac{I_{\rm out}}{1-D}, \tag{9.4f}$$

$$I_{\rm L(pk)} = \frac{I_{\rm out}}{1-D} + \frac{T}{2L}V_{\rm in}D, \tag{9.4g}$$

$$L_{\rm min} = \frac{T}{2I_{\rm out}}\left(\frac{V_{\rm in}}{V_{\rm out}}\right)^2(V_{\rm out}-V_{\rm in}). \tag{9.4h}$$

The first equation holds regardless of mode (CCM or DCM). The expressions for minimum inductance and minimum output current represent the critical values to maintain CCM; for these expressions use the maximum value of $V_{\rm in}$ and (for $L_{\rm min}$) the minimum output current.

Exercise 9.11. Continuing the challenge: derive these equations. *Hint*: for $I_{\rm out(min)}$ and $L_{\rm min}$ use the fact that, at the threshold of CCM, the input current $I_{\rm in}$ equals half the peak-to-peak inductor current variation $\Delta I_{\rm L}$, as easily seen from the $I_{\rm L}$ waveform in Figure 9.67.

Exercise 9.12. Why can't the step-up circuit be used as a step-down regulator?

The design procedures for step-up (and inverting) converters are analogous to those for the buck converter, and so we will resist the temptation to display actual circuit examples.

9.6.7 Inverting converter

The inverting circuit (also known as an "inverting buck–boost," or "negative buck–boost") was shown in Figure 9.61C (repeated here as 9.68). During switch conduction, a linearly increasing current flows from the input into the inductor (point *Z*) to ground. To maintain the current when the switch is open, the inductor pulls point *Z* negative, as much as needed to maintain continuous current flow. Now, however, that current is flowing into the inductor from the filter capacitor (and load). 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.

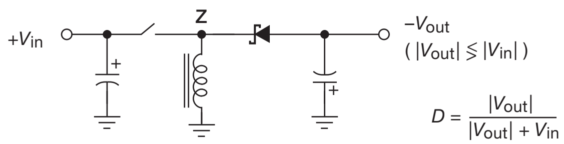

Figure 9.68. Basic inverting (or "inverting buck-boost") topology (non-isolated).

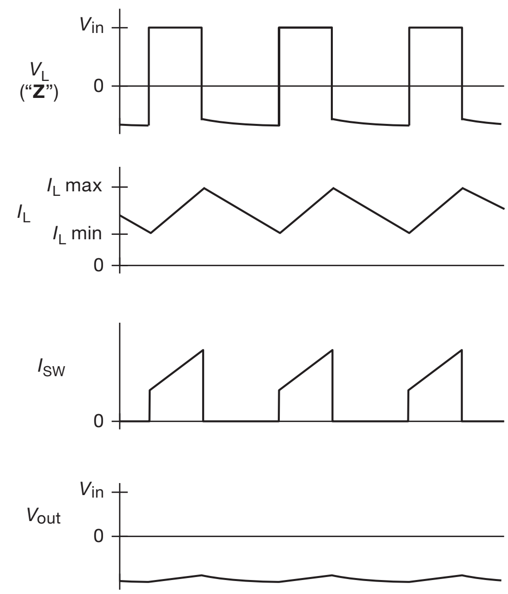

Figure 9.69. Inverting converter operation. Inductor current ramps up during switch ON, and ramps down during switch OFF. The output voltage is inverted in polarity, with a magnitude equal to the input voltage times the ratio of switch t_{on}/t_{off} (for CCM, as shown here).

A. Inverting converter equations (continuous-conduction mode)

Figure 9.69 shows the relevant voltage and current waveforms of the inverting regulator, once again assuming ideal components. With more than a bit of struggle you can figure out that the inverting converter (Figure 9.61C), operating in continuous-conduction mode, obeys these equations:

$$\langle I_{in}\rangle = I_{out}\frac{V_{out}}{V_{in}} = -I_{out}\frac{D}{1-D}, \tag{9.5a}$$

$$\Delta I_{in} = \frac{\langle I_{in}\rangle}{D}, \tag{9.5b}$$

$$V_{out} = -V_{in}\frac{t_{on}}{t_{off}} = -V_{in}\frac{D}{1-D}, \tag{9.5c}$$

$$D = \frac{|V_{out}|}{|V_{out}|+V_{in}}, \tag{9.5d}$$

$$\begin{aligned} I_{out(min)} &= \frac{T}{2L}V_{out}\left(\frac{V_{in}}{V_{in}+|V_{out}|}\right)^2 \\ &= \frac{T}{2L}V_{out}(1-D)^2, \end{aligned} \tag{9.5e}$$

$$\Delta I_{C(out)} = \frac{I_{out}}{1-D}, \tag{9.5f}$$

$$I_{L(pk)} = \frac{I_{out}}{1-D} + \frac{T}{2L}V_{in}D, \tag{9.5g}$$

$$L_{min} = \frac{T}{2}\frac{V_{out}}{I_{out}}\left(\frac{V_{in}}{V_{in}+|V_{out}|}\right)^2. \tag{9.5h}$$

As with the buck and boost converters, the first equation holds regardless of mode (CCM or DCM). The expressions for minimum inductance and minimum output current represent the critical values to maintain CCM; for these expressions use the maximum value of V_{in} and (for L_{min}) the minimum output current. In these equations we've used the absolute value symbol ($|V_{out}|$) in the two places where the reader, unmindful of the opposite polarity of input and output voltage, could go seriously off the rails.[80]

Exercise 9.13. The final (and trickiest[81]) challenge: derive these equations. *Hint*: for $I_{out(min)}$ and L_{min} use the fact that, at the threshold of CCM, the average inductor current $\langle I_L\rangle$ equals half

[80] Readers who feel insulted by such lack of trust should replace "$+|V_{out}|$" with "$-V_{out}$." They can argue, with some justification, that their signed equation correctly describes also an inverting converter that produces a positive output from a negative input rail.

[81] Dare we confess? It flummoxed more than a few of us before we got it right.

the peak-to-peak inductor current variation ΔI_L. Now figure out how $\langle I_L \rangle$ is related to I_{in} (or to I_{out}), and take it from there.

9.6.8 Comments on the non-isolated converters

This is a good place to pause, before moving on to the transformer-isolated switching converters, to discuss and review some issues common to these converters.

A. Large-voltage ratios

The ratio of output to input voltage in the basic non-isolated converters depends on the duty cycle ($D = t_{on}/T$), as given in the formulas above. For modest ratios that works fine. But to generate a large ratio, for example a buck converter converting a +48 V input to a +1.5 V output, you wind up with undesirably short pulse widths (hence greater transistor stress, in the form of high peak voltages and currents, and lower efficiency). A better solution is to take advantage of a transformer, whose turns ratio provides an additional voltage transformation. We'll see soon how this is done, in the analogous isolated converter topologies (buck converter → forward converter; inverting converter → flyback converter).

B. Current discontinuity and ripple

The three basic converters behave quite differently in terms of input- and output-current pulsation. In particular, assuming the preferred continuous-conduction mode, the buck converter has continuous current being supplied to the output storage capacitor, but pulsed input current from the $+V_{in}$ supply; the boost converter has pulsed output current, but continuous input current; and the inverting converter has pulsed current at both input and output. Pulsed (discontinuous) currents are generally undesirable at high power levels because they require larger-value storage capacitors, with lower ESR/ESL, for comparable performance. There are some interesting converter topologies (discussed presently, §9.6.8H) that address these problems; in particular, the Ćuk converter (Figure 9.70) boasts continuity of current at both input and output.

C. Regulation: voltage mode and current mode

We've talked little about the details of feedback and voltage regulation in switchmode converters, though the examples above illustrate two approaches: the simple bang–bang pulse-skipping scheme of the MC34063-style regulator (Figure 9.64); and the more commonly used proportional PWM scheme implemented in Figure 9.65. In fact, PWM control can be done in two ways, known as *voltage mode* and *current mode*: in voltage-mode PWM, the error signal is compared with the internal oscillator's sawtooth (or triangular) waveform to set the switch-ON duration. By contrast, in current-mode PWM the switch's current, ramping according to $V = L\,dI/dt$, replaces the sawtooth, and is compared with the error signal to terminate the switch's ON state, as shown below in Figure 9.71. We'll go into a bit more detail in §9.6.9.

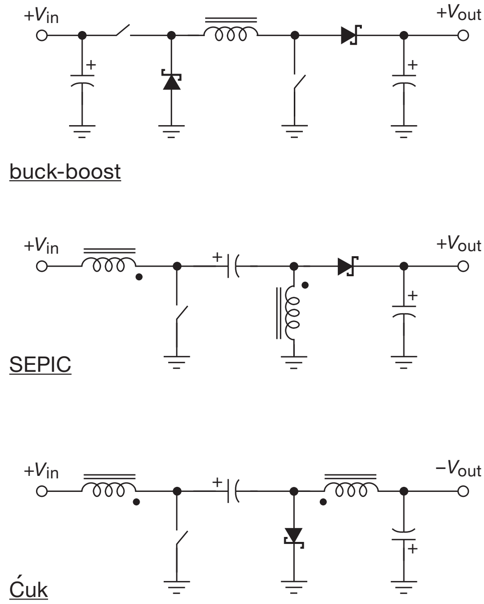

Figure 9.70. Converters allowing overlap of input and output voltage range. Both switches are operated together in the buck–boost (or "non-inverting buck–boost") configuration (A). The SEPIC (B) and Ćuk (C) configurations each use a single switch, but two (optionally coupled) inductors. The Ćuk "boost–buck" is inverting.

D. Low-noise switchers

Switchers are noisy! Figure 9.53, which compared linear and switching 5 V power converters, shows several characteristics of this undesirable "feature": first, there is plenty of noise at the switching frequency, which typically falls in the 20 kHz–1 MHz range; second, the switching frequency may vary,[82] causing interference over a range of frequencies; and, third, (and most distressingly) the switching signals can be nearly impossible to eliminate, propagating both as radiated signals and through ground currents.

[82] This is often done intentionally, in order to meet regulatory standards on interference (EMI) by "spreading" the emitted switching signals over a range of frequencies (see Figures 9.53 and 9.54). Although there is some rationale for resorting to this measure when other options are exhausted, we're not wild about this practice, which paradoxically encourages sloppy design that emits *more* total radiated power. As NASA engineer Eric Berger remarked, "When I first heard about this practice, I was appalled. The radiated energy is not reduced, just the peaks in the frequency domain are. This is like getting rid of a cow pie by stomping on it."

Figure 9.53 illustrates this latter point well: the switching noise can be heavily bypassed *at one point*, as in Figure 9.53B; but just put your 'scope probe a few inches away (Figure 9.53C and D) and *they're back!*

This problem is widely recognized, and there are various approaches to cleaning up switcher noise. At a simple level, a low-dropout regulator at the output helps considerably, as does a simple *LC* output filter. A more sophisticated approach is to use converter topologies that avoid current pulsations at the input and output (for example the Ćuk converter, §9.6.8H), or that exploit the resonant properties of inductance and capacitance so that the switches are brought into conduction at moments when the voltage across them is near zero ("zero-voltage switching," ZVS), and are opened when the current is near zero ("zero-current switching," ZCS). Finally, some converters (typified by the LT1533, LT1534, LT1738, and LT3439) incorporate circuitry to limit the switching transistor's voltage and current slew rates, which reduces both radiated and ground-conducted switching noise.

When thinking about switching converter noise, keep in mind that it emerges in multiple ways, namely:
(a) ripple impressed *across* the dc output terminals, at the switching frequency, typically of the order of 10–100 mV peak-to-peak;
(b) *common-mode* ripple on the dc output (which you can think of as ground-line ripple current), which causes the kind of mischief seen in Figure 9.53C;
(c) ripple, again at the switching frequency, impressed onto the *input* supply;
(d) *radiated* noise, at the switching frequency and its harmonics, from switched currents in the inductors and leads.

You can get into plenty of trouble with switching supplies in a circuit that has low-level signals (say 100 μV 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.

E. Inductance tradeoffs

There's some flexibility in the choice of inductance. Usually you want to run PWM converters (but not bang–bang converters like the MC34063 in our first example) in continuous-conduction mode, which sets a minimum inductance for a given switching frequency and value of minimum load current. A larger inductor lowers the minimum load current, reduces the ripple current for a given load current, and improves the efficiency; but a larger inductor also reduces the maximum load current, degrades the transient response,[83] and adds physical size to the converter. It's a tradeoff.

F. Feedback stability

Switching converters require considerably more care in the design of the frequency-compensation network than, say, an op-amp circuit. At least three factors contribute to this: the output *LC* network produces a "2-pole" lagging phase shift (ultimately reaching 180°), which requires a compensating "zero"; the load's characteristics (additional bypass capacitance, nonlinearities, etc.) affect the loop characteristics; and the converter's gain and phase versus frequency characteristics change abruptly if the converter enters discontinuous-conduction mode. And, to add a bit more complexity into an already-complex situation, there are important differences between voltage-mode and current-mode converters: for example, the latter, which are better behaved in terms of *LC*-network phase shifts, exhibit a "subharmonic instability" when operated at switch duty cycles greater than 50% (this is addressed by a technique called *slope compensation*).

The easiest approach for the casual user is to choose converters with built-in compensation (for example, the Simple Switcher series, as in Figure 9.65), or converters that provide complete recipes for reliable external compensation. Regardless, the circuit designer (you!) should be sure to *test* the stuff you've designed.[84]

G. Soft start

When input voltage is initially applied to any voltage-regulator circuit, feedback will attempt to bring the output to the target voltage. In the case of a switching converter, the effect is to command maximum duty cycle from the switch, cycle after cycle. This generates a large inrush current (from charging the output capacitor), but, worse, it can cause the output voltage to overshoot, with potentially damaging effects on the load. Worse still, the magnetic core of the inductor (or transformer) may saturate (reaching maximum flux density), whereupon the inductance drops precipitously, causing the switch current to spike. Core saturation is a major cause of component failure; you don't want it.

These problems are most severe in converters that run

[83] Transient speed is a major reason to use low inductance values in switching converters that power microprocessors, where you see the concept of *critical inductance*, i.e., an inductance small enough to handle the load step transients.

[84] When testing for stability, don't forget about the negative-resistance input characteristic of switching converters; be sure to test with whatever input filters you plan to use.

from the ac powerline, where the transformerless input stage (diode bridge and storage capacitor) causes additional inrush current, and where that input power source can deliver plenty of peak current. Many switching controller chips therefore incorporate "soft-start" circuitry, which constrains the switch duty cycle to ramp up gradually upon initial startup; these are indicated in the "soft start" column of Tables 9.5a,b and 9.6.

H. Buck–boost topologies

For the buck converter, V_{out} must be less than V_{in}, and for the boost converter V_{out} must be greater than V_{in}, required in both cases to reset the inductor current. Sometimes you'd like a converter that permits the input voltage to vary around both sides of the output voltage (for example in a battery-operated device with 2.5 V digital logic, powered by two AA cells, which begins life with 3 V input, and ends at about 1.8 V; or an automotive application, powered from a 12 V car battery, supplying 13.8 V running, but as little as 8 V starting and as much as 40 V in "load dump").

Although the inverting (buck–boost) converter (Figure 9.61C) allows the output voltage to be larger or smaller than the input, its polarity is reversed. Figure 9.70 shows three interesting configurations that allow overlap of the input- and output-voltage ranges. The first one is particularly easy to understand: both switches are operated simultaneously for a time t_{on}, applying V_{in} across the inductor; during t_{off} the inductor's current flows through the diode pair to the output. The output voltage, from the inductor's required volt–time equality (and ignoring voltage drops in the switch and diodes), is then simply $V_{\text{out}} = (t_{\text{on}}/t_{\text{off}})V_{\text{in}}$. Typical examples of buck–boost converter ICs are the LTC3534 (internal MOSFET switches) and the LTC3789 (external MOSFET switches); both use synchronous MOSFET switches in place of Schottky diodes, i.e., four MOSFETs in all. For other converters with synchronous switching see the "synchronous" column in Tables 9.5a,b on pages 653 and 654 and 9.6 on page 658.

The SEPIC (single-ended primary-inductance converter) and Ćuk[85] converters have the advantage of requiring only a single controllable switch. And the Ćuk converter has the remarkable property of producing *zero* output ripple current when the inductors are coupled (wound on the same core). This latter property was discovered accidentally, but is now part of the vocabulary of switchmode practitioners, who call it "the zero-ripple phenomenon." And while we're praising the Ćuk, it's worth noting that both input- and output-current waveforms are continuous, unlike the buck, boost, inverting, SEPIC, or buck–boost.

[85] Invented by Slobodan Ćuk (pronounced "chook") in 1976.

9.6.9 Voltage mode and current mode

There are two approaches to implementing pulse-width modulation, as we mentioned earlier in §9.6.8C; look at Figure 9.71.

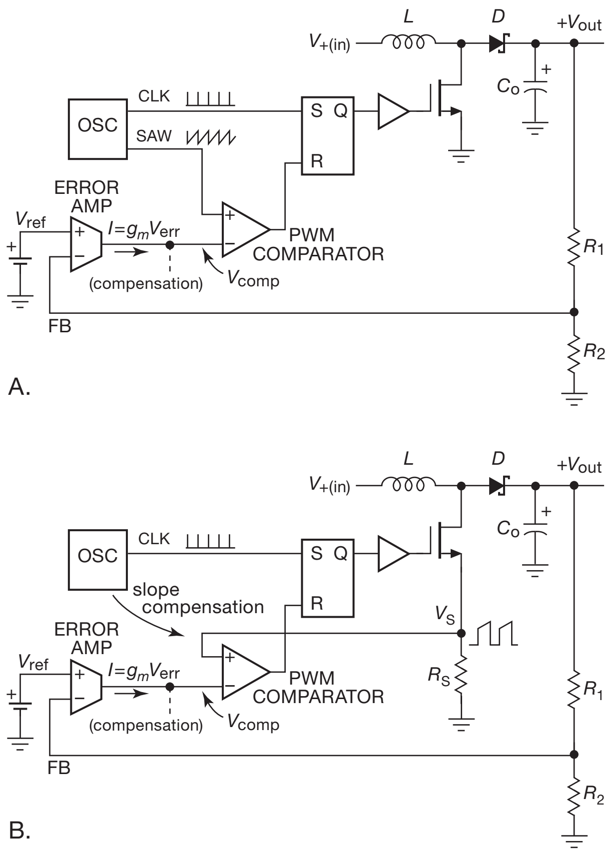

Figure 9.71. Pulse-width modulation in switchmode regulators. (A) Voltage-mode PWM compares the integrated error signal ($V_{\text{err}}{=}V_{\text{ref}}{-}\text{FB}$) with the oscillator's sawtooth whereas (B) current-mode PWM substitutes the switch's ramping current waveform.

At the top level, both methods compare the output voltage with an internal voltage reference to generate an error signal. That is, both methods are *voltage* regulators (don't confuse "current mode" with current *regulator*). The difference is in the way the error signal is used to adjust the pulse width: in *voltage-mode* PWM, the error signal is compared with the internal oscillator's sawtooth waveform to control the switch's ON duration.[86] In *current-mode* PWM, by

[86] Typically by using a *pulse* output from the oscillator to start the conduction cycle, and the output of the PWM comparator (which compares the error signal with the same oscillator's sawtooth) to end the conduction cycle, as shown in Figures 9.71A and 9.72.

contrast, the ramping current in the inductor replaces the sawtooth, with the internal oscillator used to *initiate* each conduction cycle[87] (Figures 9.71B and 9.72). Tables 9.5a,b and 9.6 indicate whether the SMPS IC employs a voltage-mode or current-mode control loop.

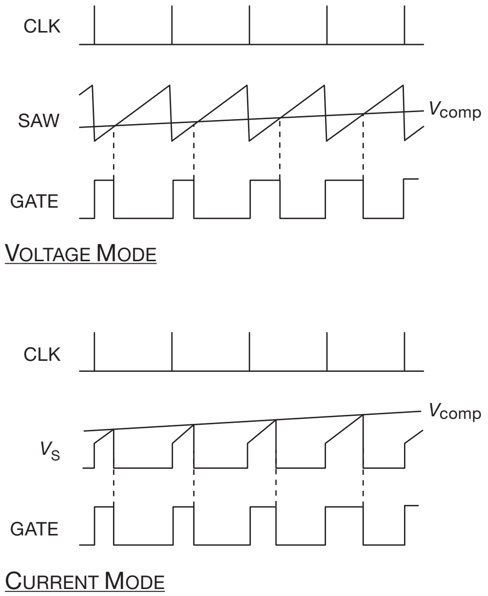

Figure 9.72. Waveforms in voltage-mode and current-mode PWM.

How to choose? Before comparing their relative merits, we offer this sensible advice: select the switching regulator chip that has the features you want (in terms of voltage and current ratings, ease of design, price and availability, component count, etc.), and don't worry about how the chip designers did their job.

Now for the comparison.

A. Voltage mode

This has been the traditional form of PWM. Its advantages include
(a) the simplicity of analyzing a single feedback path,
(b) low output impedance from the power stage, and
(c) good noise margins (because of the internally generated ramp).

Its disadvantages include
(a) the need for careful loop compensation (because of the 2-pole *LC* output filter),[88]
(b) slow loop response (especially in response to input changes), and
(c) the need for separate current-limiting circuitry for the switch transistor(s).

B. Current mode

Current-mode control became popular beginning in the 1980s when its benefits became apparent. They include
(a) rapid response to input changes,
(b) inherent pulse-by-pulse current limiting of the switch current,
(c) improved phase margin in the outer voltage-feedback loop (because the power stage's output, being current-like, effectively removes the inductor's phase shift; i.e., one pole instead of two in the feedback loop), and
(d) the ability to parallel the outputs of several identical converters.

The disadvantages of current-mode control include
(a) the greater difficulty of analyzing two nested feedback loops (mitigated by widely separating their characteristic frequencies),
(b) intrinsically higher output impedance of the power stage (the output is more affected by load changes because the fast loop tends toward a constant-current output),
(c) susceptibility to noise, particularly at low load, and to resonances (because the PWM depends on the current-derived ramp),
(d) premature termination of switch's ON-state caused by the leading-edge current spike (from parasitic capacitances and diode recovery effects), and
(e) instabilities and subharmonic resonances at high duty cycle.

Clever fixes Circuit designers are clever, and they've figured out some nice tricks to address the problems of each method. The slow response of voltage-mode controllers to input changes can be fixed by adding an input feedforward signal to the sawtooth ramp, and the slow loop response can be alleviated by running at a higher switching frequency. For current-mode control the bag of tricks includes leading-edge blanking (to ignore the switch-ON current spike), and "slope compensation" (to restore stability at high duty cycle).

Choice of control mode: both are viable In contemporary practice both modes are viable, and plenty of controller ICs are available using either technique. As a general statement, voltage-mode converters are favored
(a) in noisy applications, or in applications with light load

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87 And to generate the "slope-compensation" ramp signal.

88 As the LT3435 datasheet succinctly puts it, "A voltage fed system will have low phase shift up to the resonant frequency of the inductor and output capacitor, then an abrupt 180° shift will occur. The current fed system will have 90° phase shift at a much lower frequency, but will not have the additional 90° shift until well beyond the *LC* resonant frequency. This makes it much easier to frequency compensate the feedback loop and also gives much quicker transient response."

Table 9.5a Voltage-mode Integrated Switching Regulators[a]

Part #	Packages: TO220, DPAK	DIP	SOIC, MSOP	SOT23	smaller	Fixed-V versions[b]	Internal comp	Soft start	Burst mode, etc.	Shutdown	UVLO	Control mode[c]	Switch type[e]	Sync switching	V_{supply} min[o] (V)	V_{supply} max (V)	I_Q typ (mA)	V_{fb} typ (V)	V_{out} min (V)	V_{out} max (V)	f_{switch} (kHz)	I_{sw} max (A)	# external parts[g]	Comments
Buck																								
TPS62200	-	-	-	5	-	7	•	•	q	•	•	P	M	•	2.5	6	0.02	0.50	0.7	5.5	1000[t]	0.3	1,3	-
LT1934	-	-	-	6	•	-	•	-	•	•	•	H	B	-	3.2	34	0.012	1.25	1.25	28	~300	0.35	7	-
NCP1522B	-	-	5	-	•	-	•	•	q	•	•	P	M	•	2.7	5.5	0.05	0.90	0.9	3.3	3000[t]	1.2	4	-
CS51413	-	-	8	-	•	-	•	•	-	•	•	V2	B	-	4.5	40	4	1.27			520[t,p]	1.6	7	1
L4976	-	8	16	-	-	-	-	-	-	-	-	P	M	-	8	55	2.5	3.3	3.3	40	to 300	2	7	-
LM2574[h]	-	8	14	-	-	4	•	-	-	•	s	P	B	-	3.5	40,60	5	1.23	1.23	37,57	52[t]	1.0	2,4	-
LM2575[h]	5	-	-	-	-	4	•	-	-	•	s	P	B	-	3.4	40,60	5	1.23	1.23	37,57	52[t]	2.2	2,4	-
LM2576[h]	5	-	-	-	-	4	•	-	-	•	s	P	B	-	3.5	40,60	5	1.23	1.23	37,57	52[t]	5.8	2,4	2
NCP3125	-	-	8	-	-	-	-	•	-	•	•	P	M	•	4.5	13	4	0.80	0.8		350[t]	4 d	9	3
LT1074[h]	5	-	-	-	-	1	-	-	q	•	•	P	B	-	8	40,60	8.5	2.21	2.5	30[k]	100	5	6	4
LM2677	7	-	-	-	-	3	•	-	-	•	s	P	M	-	8	40	4.2	1.21	1.21	37	260	7	3,5	5
LMZ12010	11	-	-	-	-	-	•	•	-	•	•	P	M	•	4.3	20[f]	3	0.80	0.8	6	360[t]	10	3	-
Boost, Flyback, etc.																								
NCP1400A	-	-	-	5	-	9[n]	•	•	q	•	s	P	M	-	0.8[y]	5.5	0.03	-	1.9	5	180	0.1	2	6,8
NCP1423	-	-	-	-	10	-	•	•	•	•		P	M	•	0.8	6	0.01	0.50	1.8	3.3	to 600	0.4	4	7,8
L6920DC	-	-	8	-	-	2	•	-	-	•	•	T	M	•	0.8	5.5	0.01	1.23	1.8	5.5	1000	0.5	2,4	8
TPS61070	-	-	-	6	-	-	•	•	•	•	•	P	M	•	1.1[x]	5.5	0.02	0.50	1.8	5.5	1200	0.6	3	8
TPS61030	-	-	16	-	•	2	•	•	•	•	•	P	M	•	1.8	5.5	0.02	0.50	1.8	5.5	600	4.5	1,3	7,8

Notes: (a) all have integrated power switch(es), current-sensing, and (in some cases) loop compensation; listed in order of increasing switch current. (b) number of fixed voltages available; all except NCP1400A have adjustable versions. (c) H=hysteretic mode; P=PWM fixed frequency; T=min t_{off}, max t_{on}; V2=ONsemi "V^2" control. (d) adjustable current limit. (e) B=BJT; M=MOSFET. (f) see LMZ23608 for V_{in} to 36V. (g) typical number of external parts (not counting bypass caps); two numbers indicate fixed/adjustable. (h) 60V for HV suffix. (m) adjustable current limit. (n) no adjustable version. (o) restart threshold. (p) CS51411 for 260kHz. (q) reduced freq or pulse skipping at low load. (s) parts with SHDN can have UVLO added with an ext circuit. (t) typical. (u) plus I_{sw}/50, etc., when the switch is ON (a power-dissipation issue if used with high V_{supply}). (v) plus BJT switch drive current, on BOOST pin, taken from low-voltage buck output. (x) runs down to 0.9 volts. (y) runs down to 0.3 volts. (z) runs down to 0.5 volts.

Comments: 1: pin compatible with LTC1375. **2:** many second sources. **3:** NCP3126 and 3127 for lower current. **4:** negative V_{out} to –35V (see datasheet); V_{in} comp; LT1076 for 2A. **5:** featured in text. **6:** NCP1402 for 200mA. **7:** 96% effy, low-batty comp. **8:** single-cell stepup.

conditions, or
(b) where multiple outputs are derived from a common power stage (that is, in converters that use a transformer with multiple secondary windings).

Current-mode controllers are favored
(a) where fast response to input transients and ripple is important,
(b) where it is desirable to parallel multiple power supplies (e.g., for redundancy),
(c) where you want to avoid the complexities in designing a proper pole-zero loop compensation network, and
(d) in applications where fast pulse-by-pulse current limiting is important for reliability.[89] Tables 9.5a and 9.5b list selected "integrated" switching regulators, i.e., with *internal* power switch(es). See also Table 9.6 on page 658 for switching regulators that drive external MOSFETS, Table 3.4 (MOSFETS, page 188), and Table 3.8 (drivers, page 218).

9.6.10 Converters with transformers: the basic designs

The non-isolated switching converters of the previous sections can be modified to incorporate a transformer within

[89] Evidently SMPS integrated circuit designers (and presumably their larger customers) prefer current-mode control over voltage-mode, as reflected in the shortened length of Table 9.5a compared with Table 9.5b, and by the paucity of voltage-mode controllers in the "control mode" column of Table 9.6.

Table 9.5b Selected Current-mode Integrated Switching Regulators[a]

Part #	Packages: TO220	DPAK, D2PAK	DIP	SOIC, TSSOP	SOT23, SC70	smaller	Fixed-voltage versions	Internal comp	Slope comp	Soft start	Burst mode etc	UVLO	OVP	Control mode[c]	DMOS switch	Synchronous	V_{supply} min[o] (V)	V_{supply} max (V)	I_Q typ (mA)	V_{FB} typ (V)	V_{out} max (V)	f_{switch} min-max (kHz)	I_{sw} max (A)	# parts[g]	Comments
Buck																									
LTC1174[f]	-	-	8	8	-	-	2	•	-	-	•	•	-	P	•	-	4	13	0.45	1.25		200	0.3,0.6	2	1
LT1776	-	-	8	8	-	-	-	-	-	-	e	e	-	P	-	-	7.4	40	3.2	1.24		200	0.7	7	2
LT1933	-	-	-	-	6	6	-	•	•	•	p	•	-	P	-	-	3.6	36	1.6[v]	1.25		500[t]	1	6	-
ADP3050	-	-	-	8	-	-	2	-	•	-	-	-	-	P	-	-	3.6	30	0.7[v]	1.30		200	1.25	6/8	3
ADP2300	-	-	-	-	6	-	-	•	•	•	•	•	•	P	•	-	3	20	0.64	0.80		700[r]	1.2	5	4
ADP2108	-	-	-	-	5	5	11	•	-	•	•	•	-	P	•	•	2.3	5.5	0.02	fixed	3.3	3000[t]	1.3	1	5
LT1376	-	-	8	8	-	-	2	-	-	-	-	•	-	P	-	-	2.4	25	3.6[v]	2.42		500[t]	1.5	5/7	6
LMR12010	-	-	-	-	6	-	-	•	•	•	-	•	•	P	•	-	3	20	1.5	0.80	17	3000[t]	1.7	6	7
LT3500	-	-	-	-	-	12	-	-	•	•	-	•	-	P	-	-	3	36	2.5[v]	0.80		250-2000	2.8	12	8
NCP3170B	-	-	-	8	-	-	-	-	•	•	•	•	•	P	•	•	4.5	18	1.8	0.80	V_{in}	1000	3	5	9
A8498	-	-	-	8p	-	-	-	•	-	•	p	•	-	O	•	-	8	50	0.9	0.80	24	30-700	3.5	5	10
LT3435	-	-	-	16	-	-	-	-	•	•	•	•	-	P	-	-	3.3	60	3.3[v]	1.25		500[t,q]	4	12	11
LT1765	-	-	-	8	-	-	4	-	•	e	-	e	-	P	-	-	3	25	1.0[v]	1.20		1250	4	-	11
LT3690	-	-	-	-	-	16	-	-	•	•	•	•	-	P	w	•	3.9	36,60	0.1[v]	0.80	20	140-1500	5	8	12
Boost, Flyback, etc.																									
TPS61220	-	-	-	-	6	-	2	•	-	-	p	•	•	H	•	•	0.7	5.5	0.01	0.50	5.5	1000	0.2	1/3	13
TPS61040	-	-	-	-	5	-	-	•	-	•	•	•	-	M	•	-	1.8	6	0.03	1.233	30	1000	0.4	5	14
LT1613	-	-	-	-	5	-	-	•	•	•	p	-	-	P	-	-	1.1	10	3[u]	1.23	34[d]	1400[t]	0.8	4	15
LMR64010	-	-	-	-	5	-	-	•	•	-	-	-	-	P	•	•	2.7	14	2.1	1.23	40[d]	1600	1	5	16
LT1930A	-	-	-	-	5	-	-	•	•	-	-	-	-	P	-	-	2.6	16	5.5[u]	1.26	34[d]	2200[t]	1.2	5	-
ADP1612	-	-	-	8	-	-	-	-	-	•	•	•	-	P	•	-	1.7	5.5	4	1.235	20	650,1300	1.4	7	17
LTC3401	-	-	-	-	-	10	-	-	•	-	•	•	-	P	•	•	0.9	5.5	0.44	1.25	6	50-3000	1.6	7	18
LT1172[h]	5	5	8	16	-	-	-	-	-	e	-	-	-	P	-	-	3	40,60	6[u]	1.244	65,75[h]	100[t]	1.25	6	-
LT1171[h]	5	5	-	-	-	-	-	-	-	e	-	-	-	P	-	-	3	40,60	6[u]	1.244	65,75[h]	100[t]	2.5	6	-
LT1170[h]	5	5	-	-	-	-	-	-	-	e	-	-	-	P	-	-	3	40,60	6[u]	1.244	65,75[h]	100[t]	5	6	19
LT1534	-	-	-	16	-	-	-	-	•	•	-	•	-	P	-	-	2.7	23	12[u]	1.25	30[d]	20-250	2	12	20
LM2577[b]	5	5	16	-	-	-	2	-	•	•	-	•	-	P	-	-	3.5	40	7.5[u]	1.23	60[d]	52[t]	3	4/6	21
LM2586	7	7	-	-	-	-	3	-	•	•	-	•	-	P	-	-	4	40	11[u]	1.23	60[d]	200	3[k]	4/6	22
TPS61175	-	-	-	14p	-	-	-	-	•	•	•	•	-	-	•	-	2.7	18	3.5[m]	1.23	40[d]	200-2200	3.8	8	-
TPS55340	-	-	-	-	-	16	-	-	•	•	p	•	-	P	•	-	2.9	32	1.7	1.23	38[d]	100-1200	5	8	-
Push-pull																									
LT1533	-	-	-	16	-	-	-	-	-	•	-	-	-	-	-	-	2.7	23	12	1.25[n]	x	20-250[r]	1	13	23

Notes: (a) listed by increasing switch current; all have integrated switch, current-sensing, and in some cases loop compensation; all have shutdown capability except LM2577; all have thermal shutdown. (b) no power shutdown function; also UC2577. (c) H=hysteretic curr mode; M=Fixed peak current, with a minimum off time; O=var freq fixed off time; P=PWM fixed freq. (d) non-isolated boost higher voltages with a transformer. (e) with external parts. (f) suffix HV for 18V version. (g) typical number of external parts (not counting bypass caps); two numbers indicates fixed/adjustable. (h) suffix HV for 60V version. (k) 5A for LM2587. (m) maximum. (n) also negative, –2.5V. (o) restart threshold. (p) reduced freq or pulse skipping at low load. (r) reduced frequency during low V_{out}. (s) parts with SHDN can have UVLO added with an external circuit. (t) typical. (u) plus $I_{sw}/50$, etc., when the switch is ON (a power-dissipation issue if used with high V_{supply}). (v) plus BJT switch drive current, on BOOST pin, taken from low-voltage buck output. (w) low side. (x) transformer output.

Comments: **1:** invert OK, especially +5V to –5V converter. **2:** 60V transients OK. **3:** 60V OK for 100ms; 3.3V, 5V, and adj versions. **4:** ADP2301 for 1.4MHz. **5:** just add external inductor; 11 fixed voltages, from 1.0V to 3.3V. **6:** 5V and ADJ, see LT1507 for 3.3V. **7:** "simple switcher" nano. **8:** buck plus LDO, ext sync to 2.5MHz. **9:** power-good output; 500kHz for "A" version. **10:** adj OFF time. **11:** 100μA no-load I_Q. **12:** 80μA no-load I_Q; transients OK to 60V. **13:** boost single-cell to 1.8V–5.5V out; 3.3V, 5V, and adj versions. **14:** good for LED constant current drive. **15:** single-cell boost or flyback. **16:** "simple switcher" nano. **17:** boost from single Li-ion cell. **18:** operates down to 0.5V input; 40μA in burst mode. **19:** can regulate output using transformer's primary voltage (no feedback resistors required). **20:** low-noise, slew-rate control. **21:** 12V, 15V, and ADJ versions. **22:** 3.3V, 5V, 12V, and ADJ versions. **23:** programmable slew rate, very quiet.

the switching circuitry. This serves three important purposes: (a) it provides galvanic isolation, which is essential for converters that are powered from the ac line; (b) even if isolation is not needed, the transformer's turns ratio gives you an intrinsic voltage conversion, so that you can produce large step-up or step-down ratios while staying in a favorable range of switching duty cycle; and (c) you can wind multiple secondaries, to produce multiple output voltages; that's how those ubiquitous power supplies in computers generate outputs of +3.3 V, +5 V, +12 V, and −12 V, all at the same time.

Note that these are not the heavy and ugly laminated-core transformers that you use for the 60 Hz ac powerline: because they run at switching frequencies of hundreds to thousands of kilohertz, they do not require a large magnetizing inductance (the inductance of a winding, with all other windings open-circuited), and so they can be wound on small ferrite (or iron powder) cores. Another way to understand the small physical size of the energy-storage devices in switchmode converters – that is, the inductors, transformers, and capacitors – is this: for a given power output, the amount of energy passing through these devices in each transfer can be much less if those transfers are taking place at a much higher rate. And less stored energy ($\frac{1}{2}LI^2$, $\frac{1}{2}CV^2$) means a smaller physical package.[90]

9.6.11 The flyback converter

The *flyback* converter (Figure 9.73A) is the analog of the inverting non-isolated converter. As with the previous non-isolated converters, the switch is cycled at some switching frequency f (period $T = 1/f$), with feedback (not shown) controlling the duty cycle $D = t_{\text{on}}/T$ to maintain regulated output voltage. As with the previous converters, the pulse-width modulation can be arranged as voltage mode or current mode; and the secondary current can be either discontinuous (DCM) or continuous (CCM) from each cycle to the next, depending on load current.

What is new is the transformer, which in the flyback converter topology acts simply as an inductor with a tightly coupled secondary winding. During the switch-ON portion of the cycle, the current in the primary winding ramps up according to $V_{\text{in}} = L_{\text{pri}}\, dI_{\text{pri}}/dt$, flowing into the "dotted" terminal; during that time the output diode is reverse biased because of the positive voltage on the dotted terminals of both windings.

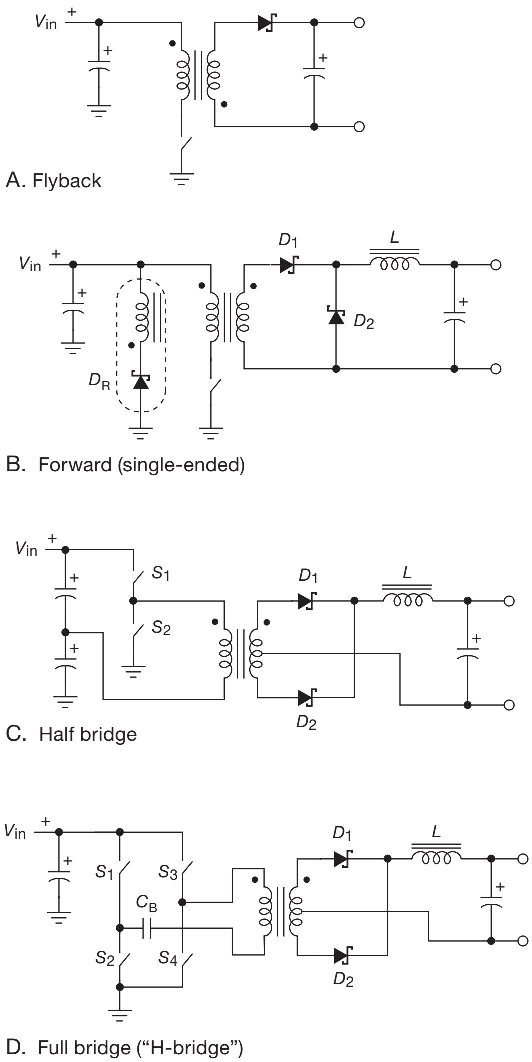

Figure 9.73. Isolated switching converters. The flyback converter (A) uses an energy-storage inductor with a secondary winding, whereas the forward and bridge converters (B–D) each use a true transformer with no energy storage (and thus require an output energy-storage inductor). The diode D_R and tertiary winding in the forward converter is one of several ways to reset the core in this single-ended design. The dc blocking capacitor C_B in the H-bridge prevents flux imbalance and consequent core saturation; for the half-bridge the series pair of capacitors serves the same function, while acting also as the input storage capacitor.

During this phase the input energy is going entirely into the magnetic field of the transformer's core. It gets its chance to go somewhere else when the switch turns OFF: unlike the situation with a single inductor, with *coupled* inductors the requirement of continuity of inductor current

[90] For the particular case of the flyback converter, discussed next, you can think of the transformer as formed by a second winding on the already-small inductor used for energy storage in the non-isolated inverting (buck–boost) converter.

is satisfied if the current continues to flow in *any* of the windings. In this case the switch-ON current, flowing into the dotted terminal, transfers itself to a similarly directed current in the secondary, but multiplied by the turns ratio $N \equiv N_{\text{pri}}/N_{\text{sec}}$. That current flows to the output (and storage capacitor), ramping down according to $V_{\text{out}} = L_{\text{sec}}\, dI_{\text{sec}}/dt$. From equality of inductor volt-seconds, the output voltage is simply

$$V_{\text{out}} = V_{\text{in}} \frac{N_{\text{sec}}}{N_{\text{pri}}} \frac{t_{\text{on}}}{t_{\text{off}}} = V_{\text{in}} \frac{N_{\text{sec}}}{N_{\text{pri}}} \frac{D}{1-D} \quad \text{(in CCM)}. \tag{9.6}$$

And, as usual, efficiency is high, so power is (approximately) conserved:

$$I_{\text{in}} = I_{\text{out}} \frac{V_{\text{out}}}{V_{\text{in}}}. \tag{9.7}$$

You can wind additional secondaries, each with its diode and storage capacitor, to create multiple output voltages (as set by the turns ratios). And, because the output windings are isolated, you can as easily generate negative outputs. Having chosen one of the outputs for regulating feedback, however, the others will not be as tightly regulated. The term "cross regulation" is used to specify the output-voltage dependencies.

A. Comments on flyback converters

Power level Flyback converters have full pulsations of input and output current. For this reason they are generally used for low- to medium-power applications (up to ∼200 W). For higher power you usually see designs using the *forward* converter, or, for really high power, *bridge* converters.

The transformer is an inductor The input energy each cycle is first stored in the transformer core (during switch-ON), then transferred to the output (during the switch-OFF). So the transformer design must provide the correct "magnetizing inductance" (acting as an inductor), as well as the correct turns ratio (acting as a transformer). This is quite different from the situation with the forward converter and the bridge converters, below, where the transformer is "just a transformer." We won't go into further detail about transformer design here, simply noting that the design of the "magnetics" is an important part of switching converter designs in general, and flybacks in particular. You have to worry about issues such as core cross-section, permeability, saturation, and deliberate "gapping" (in general, energy-storage inductors are gapped, whereas pure transformers are not). Extremely helpful resources for design are found in IC datasheets and design software (usually available at no charge from the manufacturer) that provide specifics about the choice of magnetics. We explore this important topic further in §*9x.4*.

Snubbers With ideal components, the primary current would transfer completely to the secondary when the switch turns OFF, and you wouldn't have to worry about bad things happening on the dangling drain terminal of the switch. In reality the incomplete coupling between primary and secondary creates a series "leakage inductance," whose craving for current continuity generates a positive voltage spike at the switch, even though the secondary is clamped by the load. This is not good. The usual cure is to include a *snubber network*, consisting of an *RC* across the winding, or, better, a "*DRC*" network of a diode in series with a parallel *RC*.[91]

Regulation Flyback converters can be regulated with conventional PWM, either voltage mode or current mode, with a free-running oscillator calling the shots. Alternatively, you will see inexpensive designs in which the transformer itself becomes part of a *blocking oscillator*, thereby saving a few components. We cracked open some samples of low-power (5–15 W) "wall warts" and found, well, just about *nothing* inside! We reverse engineered them to look at the circuit tricks (Figure 9.74). They seem to work just fine.

Off-line converters This final circuit (Figure 9.74) is an example of a power converter that *requires* galvanic isolation. The transformer provides isolation for the power flow; in addition, the feedback signal from the dc output must be isolated as well on its way back to the primary side. This can be done with an optocoupler, as here, or with an additional small pulse transformer. We discuss these offline converters briefly in §9.7, and in Chapter *9x* we discuss high-efficiency ("green") power supplies, including a graph comparing the performance of this 5 W supply (whose standby power is 200 mW) with others.

9.6.12 Forward converters

The *single-ended forward converter* (Figure 9.73B) is the transformer-isolated version of the buck converter. It is helpful to refer back to the basic buck circuit (Figure 9.61A), to see how it goes. The transformer converts

[91] Leakage inductance values are typically ∼1% of the magnetizing inductance. You can reduce leakage inductance greatly by splitting one of the windings (say primary) into two, with the other (secondary) sandwiched in between. And *bifilar* windings (wind primary and secondary as a pair of wires together) can reduce the leakage inductance to a low value. However these techniques increase inter-winding capacitance, and bifilar windings suffer from poor voltage insulation ratings.

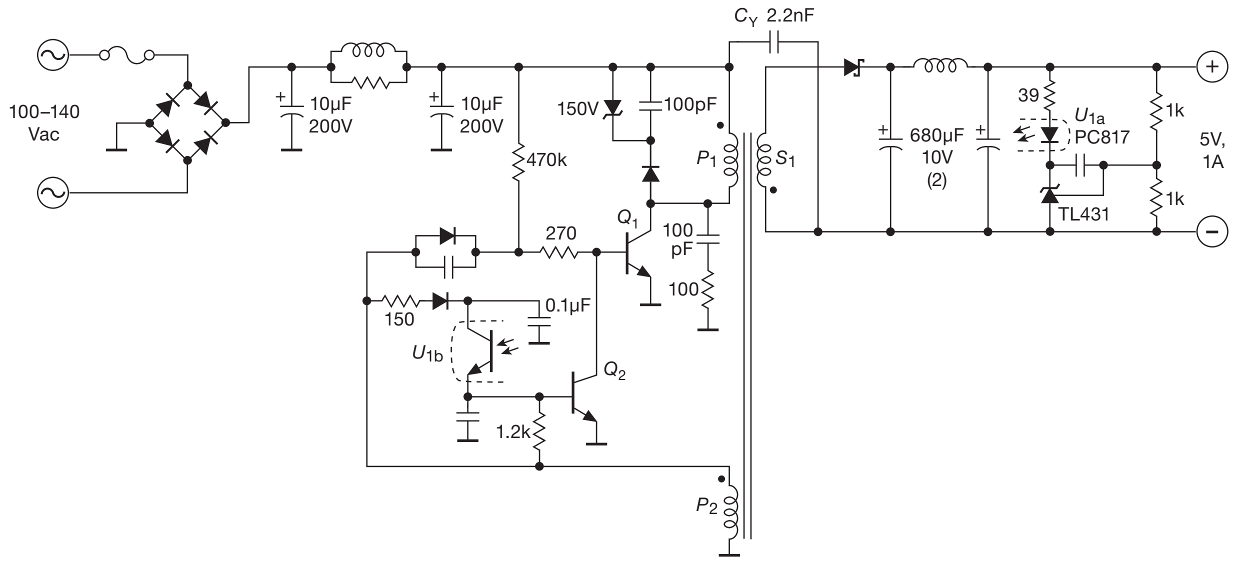

Figure 9.74. An inexpensive 5 W flyback converter, powered from 115 Vac line voltage, that uses a self-excited "blocking oscillator." Winding P2 provides positive feedback to sustain oscillation. The output voltage is sensed and compared with the TL431 shunt regulator, fed back via the optocoupler U_1 to adjust the conduction cycle.

input voltage V_{in}, during primary switch conduction, to a secondary voltage $(N_{sec}/N_{pri})V_{in}$. That transformed voltage pulse drives a buck converter circuit, consisting of catch diode D_2, inductor L, and output storage capacitor. The extra diode D_1 is needed to prevent reverse current into the secondary when the switch is OFF. Note that here, in contrast to the flyback converter, the transformer is "just a transformer": inductor L provides the energy storage, as with the basic buck circuit. The transformer does not need to store energy, because the secondary circuit conducts at the same time as the primary (energy goes "forward"), as you can see from the polarity marking.

Analogous to the buck converter, (eq'ns 9.3a–9.3h), the output voltage is simply

$$V_{out} = V_{in}\frac{N_{sec}}{N_{pri}}\frac{t_{on}}{T} = D\frac{N_{sec}}{N_{pri}}V_{in} \quad \text{(in CCM).} \qquad (9.8)$$

Resetting the core In contrast to the flyback circuit, there's an additional winding in Figure 9.73B, which is needed to *reset* the transformer's core.[92] That is because the volt-second product[93] applied to the transformer must average zero (i.e., no average dc input) in order to prevent a continual buildup of magnetic field; but the input switch alone always applies voltage in one direction only. The tertiary winding fixes this by applying voltage in the opposite direction during the switch-OFF portion of the cycle (when diode D_R conducts, from continuity of current in the winding as the magnetic field collapses).[94]

Additional comments (a) As with the flyback, and indeed with any transformer-coupled converter, the forward converter allows multiple independent secondaries, each with its inductor, storage capacitor, and pair of diodes. Regulating feedback then holds one output particularly stable. (b) The transformer isolates the output in a forward converter, if you happen to need isolation (as in a powerline-input converter); in that case you must galvanically isolate the feedback signal as well, typically with an optocoupler (as in the block diagram of Figure 9.48, or the detailed diagrams of Figures 9.74 and 9.83). On the other hand, if you do not need isolation you can have a common ground reference, and bring the error signal back to the PWM control circuit directly.

92 Reset is inherent in the flyback, but not in the single-ended forward converter, as will become evident.

93 Sometimes call "volt-time integral."

94 There are clever circuits that reset the core without requiring a tertiary winding: one method uses a pair of primary switches, one at each end of the winding, in collaboration with a pair of diodes, to reverse the voltage across the single primary (see if you can invent the circuit!). Another method uses instead a second switch to connect a small capacitor across the primary during main switch-OFF; this clever method is known as "active clamp reset," and was devised independently by Carsten, Polykarpov, and Vinciarelli. It has the virtue of *reversing* the magnetic field in the transformer core, providing better performance by allowing double the normal flux excursion.

Table 9.6 External-switch Controllers[a]

Part #	Packages: DIP	SOIC	MSOP, T/SSOP	SOT23	smaller	Control mode[aa]	Slope comp[s]	Soft start	Burst mode etc	SHDN	LEB[oo]	OVP[x]	Control mode[c]	Synch switching	V_{supply} min[o] (V)	V_{supply} max (V)	V_{out} or duty cycle max (V, %)	I_Q[v] typ (mA)	V_{FB} *V*ref (V)	f_{switch} min-max (kHz)	Drive I_{out}[d] typ (A)	Driver V_{out} (high)[k]	Ext switch	R_{sense}?	# Parts[pp]	Comments
Buck																										
LTC3863	-	-	-	-	•	I	•	•	•	•	-	•	P	•	3.5	60	–150V	0.8	0.8	50-850	0.5	8	1 P	Y	10	29
ADP1864	-	-	-	6	-	I	•	•	-	•	•	•	P	-	3.2	14	100%	0.24	0.8	580	0.6	V_{in}	1 P	Y	9	1
TLE6389	-	14	-	-	-	I	•	-	•	•	•	•	P[q]	-	5	60	100%	0.12	1.25	360	1	7	1 P	Y	9	2
ADP1872,73	-	-	10	-	-	I	-	•	•	•	-	-	V	•	2.8	20	84%	1.1	0.6	1000[f3]	1	5	2 N	N	10	-
NCV8852	-	8		-	-	I	•	•	-	•	•	-	P	-	3.4	36	100%	3	0.8	170-500	0.2	8	1 P	Y	9	3
LTC1735	-	16	16	-	-	I	•	•	•	e	•	•	P	•	4	36	6V,99%	0.45	0.8	200-550	0.6	5.2	2 N	Y	17	-
LM5116	-	-	20	-	-	I	•	•	-	•	•	-	P	•	5	100	80[u]	5	1.22	50-1000	2	7.4	2 N	Y	18	-
LTC3810	-	-	28	-	-	I	-	•	•	•	-	-	V	•	6	100	250ns	3	0.8	50-1000	2	10	2 N	N	16	-
LTC3824	-	-	10	-	-	I	x	•	•	•	•	•	P	-	4	60	100%	0.8	0.8	200-600	2.5	10	1 P	Y	13	4
LTC3830	-	8		-	-	V	na	•	•	•	-	-	P	•	2.4	9	90%	0.7	1.265	80-550	1.5[i]	V_{CC}	2 N	N	7,10	5
LTC3703	-	-	16	-	-	V	na	•	•	•	-	-	P	•	9	100	93%	1.7	0.80	100-600	2[i]	V_{CC}	2 N	N	16	-
NCP3030A	-	8	-	-	-	V	na	•	-	-	-	•	P	•	4.7	28	84%	~10	0.80	1200	1[i]	7.5	2 N	N	13	-
Buck - Boost (Vin from above to below Vout)																										
LTC3780	-	-	24	-	32	I	•	•	•	•	•	•	M	•	4	36	30V	2.4	0.80	200-400	0.6	6	4 N	Y	13	6
Boost, Flyback, etc.																										
UC384x	8	8	-	-	-	I	e	-	-	-	-	-	P	-	9, 18[ex]	30	50,100%	11	2.5	500 m	0.5	V_C	1 N	Y	20-30	7
MIC38HC4x	8	8	-	-	-	I	e	-	-	-	-	-	P	-	9, 16[ex]	20	50,100%	4	2.5	500 m	1	V_C	1 N	Y	20-30	8
ISL684x	8	8	8	-	8	I	e	-	-	-	-	-	P	-	9, 15[ex]	20	50,100%	4	2.5	2000 m	1	V_C	1 N	Y	20-30	8
UCC38C4x	8	8	8	-	-	I	e	-	-	-	-	-	P	-	9, 16[ex]	18	50,100%	2.3	2.5	1000 m	1	V_C	1 N	Y	20-30	8
UCC380x	8	8	8	-	8	I	e	•	-	-	•	-	P	-	5, 14[ex]	30	50,100%	0.5	2.5[h]	1000 m	1	V_C	1 N	Y	20-30	9
TPS40210,11	-	-	10	-	10	I	•	•	-	•	-	-	P	-	4.5	52	80%	1.5	0.26[r]	35-1000	0.4	8	1 N	Y	15	10
LTC3803	-	-	-	6	-	I	•	•	-	-	•	-	P	y	8.7	clamp	80%	0.24	0.80	200 t	0.7	V_{CC}	1 N	Y	8	11
MAX668, 69	-	-	10	-	-	I	•	•	•	•	-	-	P	-	1.8	28	90%	0.22	1.25	100-500	1	5.0	1 N	Y	10	12
LM3478	-	-	8	-	-	I	•	•	-	•	•	•	P	-	3	40	100%	3	1.25	100-1000	1	7	1 N	Y	9	13
LM5020	-	-	10	-	10	I	•	•	-	-	V	-	P	-	8	15	80%	2	1.25	50-1000	1	7.7	1 N	Y	many	14
LTC1872B	-	-	-	6	-	I	•	-	•	•	•	•	P	-	2.5	9.8	100%	0.27	0.80	550	1	V_{in}	1 N	Y	8	15
LM3481	-	-	10	-	-	I	•	•	•	•	•	•	P	-	3	48	85%	3.7	1.28	100-1000	1	5.8	1 N	Y	14	-
MAX15004	-	-	16	-	-	I	•	-	-	•	-	•	P	-	-	40	50,80%	2	1.23	15-1000	1	7.4	1 N	Y	20	-
ADP1621	-	-	10	-	-	I	•	•	•	•	•	-	P[q]	-	2.9	5.5	95%	1.8	1.215	100-1500	2	V_{in}	1 N	N[w]	10	16
LTC1871	-	-	10	-	-	I	-	-	•	•	-	•	P	-	2.5	36	92%	0.55	1.23	50-1000	2	5.2	1 N	N	13	-
NCP1450A	-	-	-	5	-	V	na	•	•	•		e	P	-	0.9	6	80%	0.14	f	180	0.05	V_{in}	1 N	N	3	17
Offline Flyback																										
FAN6300	8	8	-	-	-	I	-	•	•	-	•	•	Q	-	17[o]	25	70%	4.5	p	100[z]	0.15	18	1 N	Y	20	18,28
NCP1252	8	8	-	-	-	I	•	•	•	•	•	-	P	-	8	28	80%	1.4	p	50-500	0.5	15	1 N	Y	20	19,20
NCP1237,38	-	7	-	-	-	I	•	•	•	-	•	w	P[q]	-	13[o]	28	80%	2.5	p	65[f3]	1	13.5	1 N	Y	20	20,28
L5991	16	16	-	-	-	I	e	•	•	•	•	j	P	-	9, 16[o]	20	50,100%	7	2.5	40-2000	1	V_C	1 N	Y	30	21
Push-Pull, Forward, Half-Bridge, etc.																										
MC34025	16	16	-	-	-	I	•	•	-	-	-	-	P	-	9.6	30	t, 45%	25	5.10	5-1000	0.33	V_C	2 N	Y	many	22
LM5041	-	16	-	-	16	I	-	•	-	•	•	-	P	-	9	15	t, 50%	3	0.75	1000	1.5	V_C	4 N	Y	many	23,14
TL594	16	16	-	-	-	V	na	e	-	-	-	e	P	-	7	40	t, 45%	9	5.0[w]	1 - 300	0.2	b	2	-	9, 12	24
SG3525	16	16	-	-	-	V	na	•	-	-	-	-	P	-	8	40	t, 49%	14	5.10	0.1 - 400	0.2	V_C	2		many	24,25
LM5035	-	20	-	-	24	V	na	•	-	•	-	•	P	•	8	105	t, 50%	4	5.0	100-1000	1.25	V_{CC}	2,4 N	Y	many	26,14
NCP1395A	16	16	-	-	-	V	na	•	•	•	-	-	R	•	11	20	t, 50%	2.3	2.5	50-1000	ext	-	2 N	N	22	27

Notes: (a) all require external power switches (see listings in **Table 3.4**); all have undervoltage lockout (UVLO) and internal voltage references; listed within groups in approximate order of increasing drive current. (aa) I - current mode, V - voltage mode, P - fixed peak current, M - multiple modes. (b) uncommitted BJT output, sinks 200mA, 40V max. (c) P=PWM fixed freq; Q=quasi-res; R - resonant; V=var freq fixed width; (d) peak driver current, for controllers. (e) ext parts. (ex) lower voltage for x=3 or 5, higher voltage for x=2 or 4. (f) fixed only. (f3) three switching-frequency options. (g) unused footnote. (h) 2V for x=3 or 5. (i) adjustable current limit. (j) 25V zener clamp for V_{CC}. (k) to V_{CC} or voltage shown, whichever is less. (m) maximum. (n) nominal. (o) turn-on threshold. (oo) even with LEB (leading-edge blanking) an *RC* filter or at least a 100pF capacitor is often recommended. (p) ref pin is current-sourcing. (pp) [same note as integrated tables]. (q) reduced freq or pulse skipping at low load. (r) 0.7V for the '11. (s) helps stabilize the control loop against sub-harmonic oscillations. (t) transformer output. (u) a minimum off time (450ns) limits the duty cycle. (v) may not include dynamic gate-charge currents, etc. (w) for V_{out} below 30V, above 30V a current-sense resistor is required. (x) OVP = line over-voltage protection. (y) synchronous possible with low-voltage non-isolated flyback transfomer. (z) finds resonant frequency.

Comments: **1:** LTC1772, LTC3801 second-source. **2:** fixed 5V version available. **3:** automotive. **4:** hi-side sense. **5:** LTC3832 goes down to 0.6V. **6:** single inductor, foldback current limit. **7:** jellybean. **8:** improved UC384x. **9:** UC384x with LEB, SS, low I_Q. **10:** impressive 52V, LED drive. **11:** use with flyback xfmr. **12:** to 1.8V, slope-comp, soft-start, expensive. **13:** to 1MHz, advanced. **14:** HV pin, to 100V for startup. **15:** SOT23, low power, cute. **16:** can boost inputs as low as 1V. **17:** fixed voltage versions only, five choices 1.9V to 5.0V. **18:** quasi-resonant. **19:** inexpensive, ATX power supplies etc. **20:** freq dither. **21:** 25V zener clamp for V_{CC}. **22:** legacy, inexpensive, second sourced. **23:** programmable gap/overlap. **24:** legacy, inexpensive, flexible. **25:** also UC3525 etc. **26:** feed-forward ramp. **27:** resonant, use with FET driver IC. **28:** HV pin, to 500V for startup. **29:** optimized for inverting, V_{out} from –0.4V to –150V or more.

(c) As with all switchmode converters, snubber networks are needed to tame the voltage spikes caused by parasitic inductances (including particularly transformer leakage inductance).
(d) As with other converter types, PWM control can be either voltage mode or current mode. An alternative is to use pulse *frequency* modulation (PFM), with approximately constant pulse width, to take advantage of resonant behavior (thus avoiding "hard switching" by allowing the resonant ringing to charge and discharge parasitic capacitances, and thereby come closer to the ideal of zero-voltage/zero-current switching). (e) Single-ended forward converters are popular in the medium-power range (∼25–250 W).

9.6.13 Bridge converters

The last two transformer-isolated converters in Figure 9.73 are the *half-bridge* and *full-bridge* (H-bridge) converters. As with the single-ended forward converter, the transformer acts simply to effect voltage transformation and isolation; the secondary circuit's inductor does the energy storage, serving the same purpose as it does in the basic buck converter or single-ended forward converter. In fact, you can think of the bridge converters approximately as "double-ended forward converters." In both bridge circuits the capacitor(s) on the input side allow the voltage at the undotted end of the transformer primary to move up or down as needed to achieve zero average dc current, preventing transformer core saturation.

To understand the half-bridge converter, imagine first that switches S_1 and S_2 are operated alternately, with 50% duty cycle and with no gap or overlap. The voltage at the junction of input capacitors will float to half the dc input voltage, so what you've got is a center-tapped full-wave rectifier circuit, driven by a square wave. Power is transferred forward during both halves of each cycle, and the output voltage (ignoring diode drops) is just

$$V_{\text{out}} = V_{\text{in}} \frac{N_{\text{sec}}}{4N_{\text{pri}}}, \tag{9.9}$$

where the factor of 4 arises from the factor of $\frac{1}{2}$ for the applied input voltage and the same factor from the output center-tap. The operation of the full-bridge converter is similar, but its four switches enable it to apply the full dc input voltage across the primary during each half-cycle, so the 4 is replaced by 2 in the denominator.

Regulation With the switches operating in opposition, at 50% duty cycle, the output voltage is fixed by the turns ratio and the input voltage. To provide regulation you need to operate each switch for less than a half-cycle (Figure 9.75), with a conduction gap ("dead time") whose length is adjusted according to the error signal. You can think of each half-cycle as a forward converter, of duty cycle $D{=}t_{\text{on}}/(t_{\text{on}}{+}t_{\text{off}})$, causing the converter to produce an output voltage (assuming CCM) of

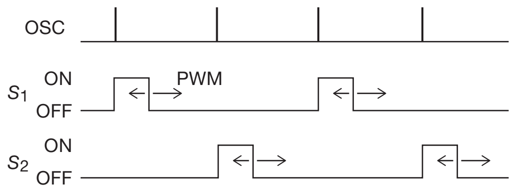

Figure 9.75. Pulse-width modulation in the half-bridge switching converter. The internal oscillator initiates switch conduction on alternate cycles, with feedback providing regulation by ending each switch's conduction according to the error signal.

$$V_{\text{out}} = D V_{\text{in}} \frac{N_{\text{sec}}}{4N_{\text{pri}}}. \tag{9.10}$$

Bridge converters are favored for high-power conversion (∼100 W and above), because they make efficient use of the magnetics by conducting during both halves of each cycle, and they cycle the magnetic flux symmetrically. They also subject the switches to half the voltage stress of a single-ended converter. By adding another pair of switches, you can convert it to a full-bridge (or H-bridge), in which the full dc input voltage is applied across the primary each half-cycle. (See the comments below, however, about flux balance.) The full-bridge configuration additionally allows another form of regulation, called "phase-shift control," in which a 50% duty cycle is maintained in each switch pair, but the relative phase of one pair is shifted relative to the other, to effectively produce a variable duty cycle.[95]

Additional comments (a) As with the single-ended forward converter, it is essential to maintain zero average voltage (or volt-time integral) across the transformer's primary. Otherwise the magnetic flux will grow, reaching destructive saturation. The H-bridge in Figure 9.73D includes a blocking capacitor C_{B} in series with the primary for this purpose; the pair of input capacitors serves the same function for the half-bridge (Figure 9.73C). That capacitor can be quite large, and it has to endure large ripple currents; so it would be nice to eliminate it, for example by connecting the bottom of the winding to a fixed voltage of $V_{\text{in}}/2$ (which is available automatically in an offline voltage-doubling input bridge). That configuration is known as "push-pull."

[95] Some phase-shift controller ICs we like are the UCC3895 from TI and the LTC3722 from Linear Technology.

However, without the blocking capacitor it is easy to violate the flux-balance condition. One solution is the use of current-mode control, in which cycle-by-cycle (or, more accurately, half-cycle by half-cycle) current limiting prevents saturation. In any case, be aware that flux imbalance in bridge converters is really bad news.
(b) In bridge converters the power switches are connected in series across the dc input supply. If there is conduction overlap, large currents can flow from rail to rail; this is known as "shoot-through" current. What you need to know is that you don't want it! In fact, turn-off delays in MOSFETs, and more seriously in BJTs, require that the control signals provide a short time gap to avoid shoot-through.
(c) Once again, snubbers are needed to tame inductive spikes.
(d) Full-bridge converters are favored for high-power converters, to 5 kW or more.
(e) At high load currents the output filter inductor has a continuous current flowing through it. During primary conduction cycles this is, of course, supplied either by D_1 or D_2, by normal transformer action. But what happens during primary *non*-conduction (the gaps in Figure 9.75)? Interestingly, the continuing inductor current flows through *both* D_1 and D_2, forcing the transformer secondary to act like a short-circuit (even though its primary is open), because equal diode currents flow in the same direction out of both ends of the center-tapped winding.

9.7 Ac-line-powered ("offline") switching converters

With the exception of Figures 9.48B and 9.74, all the switching converters and regulators we've seen so far are *dc-to-dc* converters. In many situations that's exactly what you want – for battery-operated equipment, or for creating additional voltages within an instrument that has existing dc power.[96]

Apart from battery-powered devices, however, you need to convert incoming powerline ac to the necessary regulated dc voltages. You could, of course, begin with an unregulated low-voltage dc supply of the sort in Figure 9.49, followed by a switching regulator. But the better approach is to eliminate the bulky 60 Hz step-down transformer by running an isolated switching converter directly from the rectified (unregulated) and filtered ac power, as shown earlier in Figure 9.48.[97]

Two immediate comments. (a) The dc input voltage will be approximately 160 volts[98] (for 115 Vac 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 powerline, so it's essential to use a switching converter with an isolated power stage (forward, flyback, or bridge), and with isolated feedback (via an optocoupler or transformer).

9.7.1 The ac-to-dc input stage

A. Dual-voltage configurations

Figure 9.76 shows two common input-stage configurations. The simple bridge rectifier of Figure 9.76A is perfectly OK

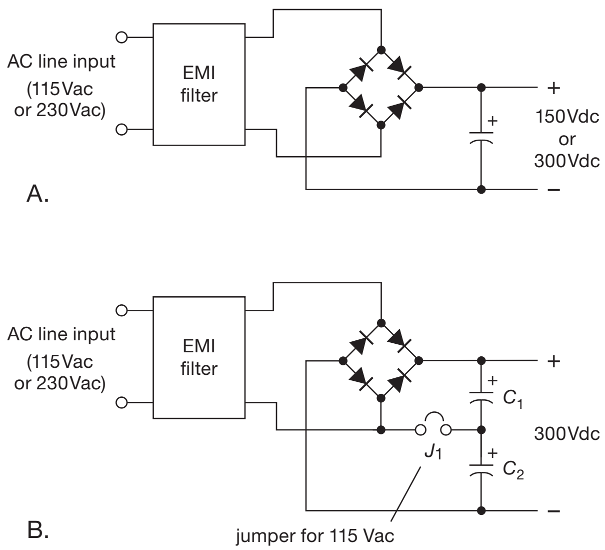

Figure 9.76. Switching power supplies run from the ac powerline (offline converters) use directly rectified dc to power an isolated converter. The jumper in the lower circuit selects bridge or voltage doubler configurations, so that either line voltage produces the same ∼300 V dc output.

[96] A common application is within a computer, where the processor may require something like 1.0 V at 100 A (!). That's a lot of current to be running around a printed circuit board! What is done, instead, is to bring a higher "bus" voltage (usually +12 V) into the vicinity of the processor, where it powers a half-dozen or so 12 V-to-1.0 V buck converters that surround the power-hungry chip and that run in multiple phases to reduce ripple. This is called "point-of-load" power conversion. The benefit, of course, is the lower current in the bus, about 8 A in this example, combined with tight voltage regulation at the load itself.

[97] A story to prove us wrong: we routinely disassemble all sorts of commercial electronic gadgets, just to see how the other half lives. Imagine our surprise, then, when we cracked open a cellphone charger and found... a tiny ac power transformer, bridge rectifier, and low-voltage storage capacitor, followed by an MC34063 switching converter! Goes to show you.

[98] And, more commonly, 320 volts; see below.

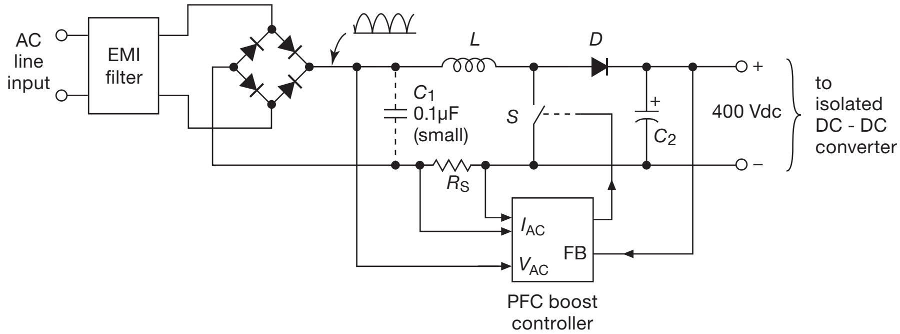

Figure 9.77. The direct rectifier circuits of Figure 9.76 create undesirable current pulses each half-cycle (low power factor). This is remedied with a power factor correction front end, consisting of a boost converter running from the (unfiltered) full-wave rectified line-voltage waveform, controlled by a special PFC chip that operates the switch to maintain the input current approximately proportional to the input voltage.

for devices intended for either 115 Vac or 230 Vac use, in which the switching converter that follows is designed for either ∼150 Vdc or ∼300 Vdc input, respectively. If you need a supply that can be switched to run on either input voltage, use the nice trick shown in Figure 9.76B: it's a simple full-wave bridge for 230 Vac input, but with the jumper connected it becomes a voltage doubler for 115 Vac input, thus generating ∼300 Vdc on either continent. (The other popular approach is to design the switching converter to accommodate a wide dc input range; most low-power chargers for consumer devices like laptop computers and cameras work this way. Check the label, though, before you plug in to 230 Vac power. And don't expect more power-hungry electronic devices to work automatically on "universal" power; they usually have a recessed slide switch that is the jumper in Figure 9.76B.)

B. Inrush current

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; even a tiny "wall-wart" can draw 25 A or more of instantaneous current when first plugged in. 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 (10 Ω) series resistor a fraction of a second after the supply is turned on. The series inductance provided by an input noise filter helps somewhat, as well. But a very nice solution comes in the form of an input power-factor correction circuit, discussed next.

C. Power-factor correction

The pulsed current waveform of rectified ac, as seen for example in Figure 9.51, is undesirable because it produces larger resistive (I^2R) losses compared with the ideal of a sinusoidal current waveform that is in phase with the voltage. (This is why it's easy to make the mistake of choosing too small a fuse rating, as discussed earlier in §9.5.1B.) Another way to say it is that a pulsed current waveform has a low *power factor*, which is defined as the power delivered divided by the product $V_{\rm rms} \times I_{\rm rms}$. Power factor made its first appearance in Chapter 1 in connection with reactive circuits, in which the phase-shifted (but still sinusoidal) current created a power factor equal to the cosine of the phase difference between the ac voltage and current. Here the problem is not phase, it's the high rms/average ratio of the pulsed-current magnitudes.

The solution is to make the power supply's input look like a passive resistor, by devising a circuit that forces the input current waveform to be proportional to the input voltage over the ac cycle. That is known as a power-factor correction (PFC) circuit, and it is connected between the full-wave rectified ac input (but with the usual storage capacitor omitted) and the actual dc–dc converter, as shown in Figure 9.77. It consists of a non-isolated boost converter, operating at the usual high switching frequency, with the switching duty cycle continually adjusted to keep the sensed input current ($I_{\rm ac}$) proportional to the instantaneous ac input voltage ($V_{\rm ac}$) over the ac cycles. At the same time, it regulates its dc output to a voltage somewhat greater than the peak ac input, usually +400 V. This dc output then powers an isolated dc–dc converter to produce the final regulated voltages.

Power-factor correction is becoming standard in most moderate-to-high-power offline switching power supplies (>100 W, say), and is required by various regulatory standards. It is quite effective, as can be seen in Figure 9.78, where we dusted off a vintage desktop computer and compared its input current waveform with that of a contemporary unit running at the same time and from the same wall outlet.

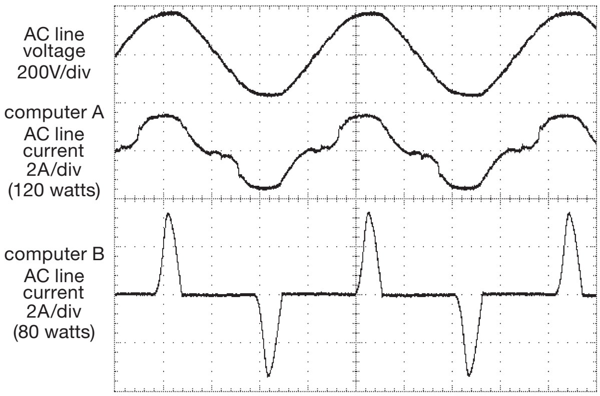

Figure 9.78. A tale of two computers. Computer A has a PFC-input power supply, causing its input current to track the input voltage. The power supply in computer B, built ten years earlier, lacks PFC; its input bridge rectifier charges the storage capacitor with short-duration current surges. Horizontal scale: 4 ms/div.

9.7.2 The dc-to-dc converter

There are some extra issues to contend with in the design of offline converters.

A. High voltage

Whether power-factor corrected or not, the dc supply to the converter–regulator will be at a substantial voltage, typically 150 V or 300 V, or somewhat higher if PFC is used. The converter itself provides the isolation, typically using one of the transformer configurations of Figure 9.73. The switch must withstand the peak voltages, which can be significantly larger than the dc supply. For example, in the forward converter with 1:1 tertiary reset winding (Figure 9.73B) the MOSFET drain swings to twice V_{in} during reset; and for the flyback the drain flies up to $V_{in} \cdot T/t_{off}$. Note also that these peak voltages assume ideal transformer behavior; leakage inductance and other non-ideal circuit realities further exacerbate the situation.

B. Switching losses

High-voltage MOSFETs do not have the extremely low R_{on} of their lower-voltage brethren. For high-voltage MOSFETs of a given die size, R_{on} increases at least quadratically with voltage rating (see Tables 3.4 and 3.5). So designers have to worry about the *conduction loss* during the conduction portion of the cycle, namely $I_D^2 R_{on}$. You can, of course, reduce conduction losses by choosing a larger MOSFET, with reduced R_{on}.[99] But larger transistors have higher capacitances, which contribute to *dynamic losses*, which become increasingly important when switching high voltages: imagine, for example, a forward converter in continuous-conduction mode; when the switch is turned ON, it must bring its drain (and attached load) from $+2V_{in}$ to ground. But there is energy stored in the switch's drain capacitance, as well as the parasitic capacitance of the transformer's winding, to the tune of $E = \frac{1}{2}CV^2$, which is squandered as heat each switching cycle. Multiply that by the switching frequency, and you get $P_{diss} = 2fCV_{in}^2$. It goes up quadratically with operating voltage, and it can be substantial: an offline forward converter, running from +300 V rectified line voltage, switching at 150 kHz, and using a 750 V MOSFET with drain (and load) capacitance of 100 pF would be dissipating 3 W from this dynamic switching loss alone.[100]

There are clever ways to circumvent some of these problems. For example, inductances can be exploited to cause the drain voltage to swing close to ground (ideally, zero-voltage switching) before the switch is activated; this is called "soft switching," and is desirable for reducing both $\frac{1}{2}CV^2$ switching losses and the component stress caused by hard switching. And the $V_D I_D$ switching loss during transitions can be minimized by driving the gate hard (to reduce switching time), and by exploiting reactances to bring about zero-current switching. These problems are not insurmountable; but they keep the designer busy, dealing with tradeoffs of switch size, transformer design, switching frequency, and techniques for soft switching. This kind of circuit design is not for the casual electronics tinkerer, nor for the faint of heart.

C. Secondary-side feedback

Because the output is deliberately isolated from the hazardous powerline input, the feedback signal has to cross

[99] Or, for high-enough voltages, use an IGBT instead; see §3.5.7.

[100] A second kind of dynamic switching loss occurs during the ramp-up and ramp-down of switch voltage, during which the instantaneous transistor power dissipation is the product of drain voltage and drain current. This is basically a dynamic conduction loss associated with switching *transitions*, to be distinguished both from the *static* conduction loss during the switch's ON state, and from the dynamic "hard-switching" losses associated with charging and discharging parasitic capacitances.

back over the same isolation barrier. The configuration in Figure 9.74 is typical: a voltage reference and error amplifier (here implemented with a simple shunt regulator) drives the LED of an optocoupler at the output, with the isolated phototransistor providing guidance to the switch control (usually PWM) on the drive side. A lesser-used alternative is a pulse transformer, driven from a "secondary-side controller" circuit. A third alternative, if a high degree of output regulation is not needed, is to regulate the output of an auxiliary winding that is not on the "output" side (for example, a winding like P2 in Figure 9.74); because it returns to the input-side common, no isolation of its feedback signal is needed. This is called *primary-side* regulation. Typically you'll get something like ±5% output regulation (over a load-current variation from 10% to 100% of rated current), compared with ±0.5% or better with secondary-side feedback.

D. The isolation barrier

Transformers and optocouplers provide galvanic isolation. Simple enough, it would seem. But, as with life itself, there's usually plenty of nuance lurking below the surface (and, as will become evident, *along* the surface as well).

There are two mechanisms by which an isolation barrier can be breached:

(a) High voltages can create a spark directly across an air gap (or through an insulating sheet); this kind of breakdown is called "arcing" (or "arc-over"), so you have to ensure a minimum *clearance* distance, defined as the shortest distance in air between a pair of conductors.

(b) A conductive path can develop on the surface of insulating material that separates a pair of conductors; this kind of breakdown is called "tracking,"[101] best prevented by ensuring a minimum *creepage* distance, defined as the shortest distance along the surface of insulating material between two conductors; see Figure 9.79. As will become evident, creepage is generally the greater worry (compared with clearance) in high-voltage circuit layouts.

It's bad news when there's breakdown of an isolation barrier; it will likely cause damage or destruction to downstream powered electronics. Worse yet, there's human safety – an electronic device whose isolation from the ac line power is lost can kill you. For these reasons there are guidelines and strict standards that govern the design of isolation barriers (codified by IEC, UL, DIN/VDE, etc.). Publications like IEC 60950 and IEC 60335 include extensive tables of minimum clearance and creepage, and web-

[101] A colorful term that describes well the little carbonized tracks you tend to find in postmortem forensics of a high-voltage device that has failed.

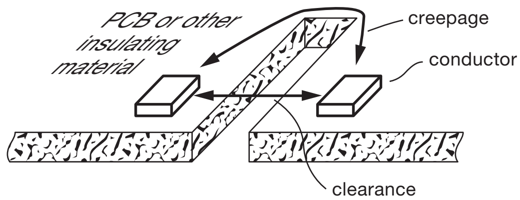

Figure 9.79. Two paths for breaching an isolation barrier: rapid arcing across the airgap (defined by the *clearance* distance), and conductive "tracking" along a path on the surface of the insulating material (defined by the *creepage* distance).

sites like www.creepage.com have delightful online calculators to keep your designs reliable and safe.

Generally speaking, clearances of 2 mm or so, and creepage distances of 4–8 mm or so, are appropriate for 120 Vac powered converters. However, there are additional variables that affect the required spacings. An example is "pollution degree" (referring to the presence of conductive dust, water, etc.); and there is the overall category of intended insulation (ranging from the merely "functional" to the strictest safety level of "reinforced"). Another factor is the intended application: for example, there are separate safety standards for products intended for household use (IEC 60335), and there are particularly strict standards for medical devices (IEC 60601). A full discussion of the subject is well beyond the scope of this book. The following treatment aims to alert the reader to the seriousness of high-voltage isolation, and some of the techniques that are used to deal with it.

The variables: insulation type, voltage, material group, pollution degree

These are the parameters you use with the tables or calculators.

Insulation type The overall level of required effectiveness, in five steps (functional, basic, supplementary, double, reinforced).

Voltage Arc-over in air or through an insulating sheet is rapid, so it's the *peak* voltage (or peak transient) that matters. By contrast, the deterioration or contamination that causes conductive creepage is slower, so you use rms or dc voltages when consulting the tables.

Material group This refers to the susceptibility of the particular insulating material to surface breakdown; the groups are called I, II, and III, going from least to most susceptible. Some standards prefer analogous parameters called "comparative tracking index" (CTI) and "performance level categories" (PLCs).

Pollution degree A curious term, which refers to the quality of air: degree 1 is clean and dry air; degree 2 is the normal home or office environment; degree 3 is nasty, with conductive dust, condensing moisture, and the like – basically it applies to service in heavy industrial or farming environments.

Increasing the creepage distance

If you've got a compact design, such that there's insufficient space to provide adequate creepage distances, you can use various measures. You'll frequently see gaps or slots cut through a printed circuit board, as in the offline switcher of Figure 9.80. You can also provide a protruding barrier to lengthen the surface-clinging path, a technique used in high-voltage optocouplers, transformer windings, and the like (see next paragraph). A conformal insulating coating applied over a populated circuit board is a particularly effective technique (but it must not delaminate, or it can be worse than no coating at all). Related techniques for individual components involve potting or molding.

Creepage considerations in component packaging and design

Components that bridge the isolation barrier, such as transformers and optocouplers, must be designed and packaged with appropriate clearances and creepage distances, both in the external leads and in the internal insulation. An example is the isolation-straddling Y-capacitor, with one foot on each side. As the photograph of Figure 9.81 shows, the leads of the disc-geometry Y-capacitor are oriented at right

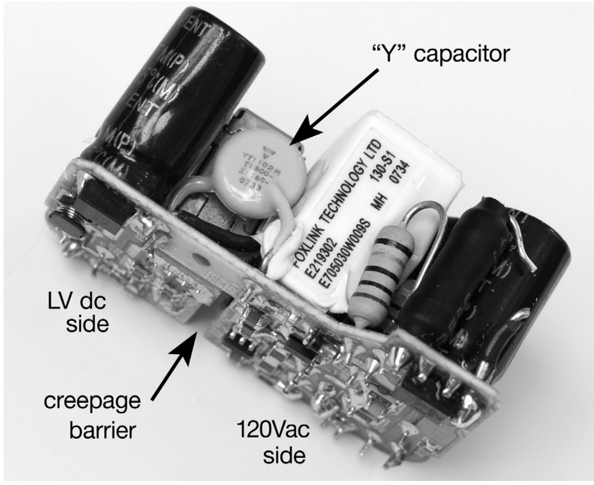

Figure 9.81. This edgewise view of the same converter reveals that the Y-capacitor's widely spaced leads preserve the 8 mm minimum creepage; by contrast, the converter's minimum *clearance* is just 1.5 mm.

angles and coated with a continuation of the same conformal insulation that covers the capacitor body. Components housed in DIP-style cases can achieve greater separation of input and output sections by omitting intermediate pins[102] (thus a "DIP-8" that's missing pins 2,3,6, and 7). An example of a fully specified high-voltage part comes from Avago, whose datasheet for an optocoupler (ACNV260E) includes an abundance of clearance and creepage specifications: both "external" and "internal" clearances (13 mm and 2 mm, respectively), and likewise for creepage distances (13 mm and 4.6 mm, described as "measured from input terminals to output terminals, shortest distance path along body" and "along internal cavity," respectively).

The leads of the switching transformer must similarly maintain adequate spacing and creepage distance. Equally important, the inter-winding insulation and winding geometry must create both appropriate insulation (by a sufficient number of layers of insulating tape, etc.) and also appropriate creepage standoff. To meet the creepage requirements, the windings may be arranged side-by-side (rather than coaxial), and separated with an insulating sheet that extends outward beyond the windings. This is good for creepage, but bad for the magnetic design, as it increases the leakage inductance. With a magnetically preferable coaxial geometry, the creepage distance can be extended by

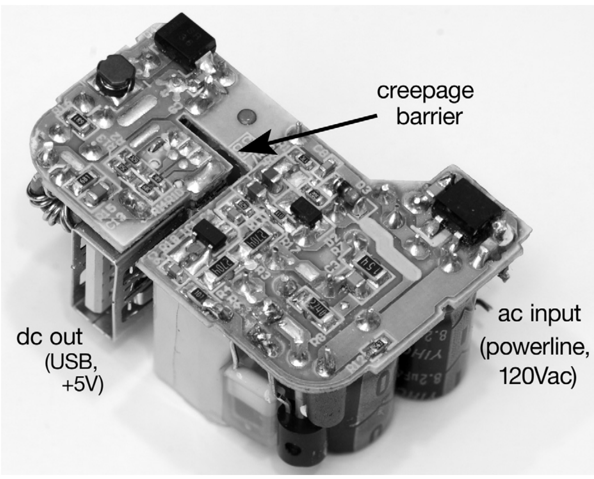

Figure 9.80. The designers of this switching converter included an L-shaped slot in the circuit board, greatly lengthening the creepage distance from the powerline circuitry to the isolated 5 V output.

[102] See for example the datasheets for the Vishay CNY64 coupler, the ON Semiconductor NCP1207 PWM controller, or the Power Integrations LNK-403 driver.

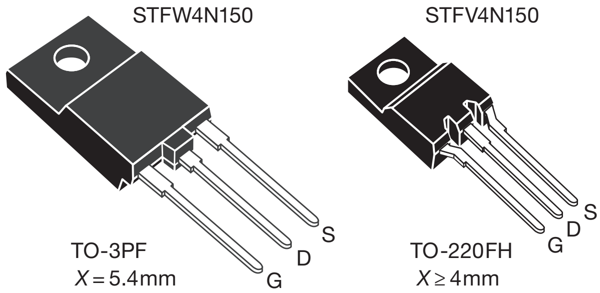

Figure 9.82. These 1500 V MOSFET packages employ shaped and grooved insulation to lengthen the creepage path length. (Adapted with permission of STMicroelectronics)

allowing the inter-winding tape to extend beyond the windings, or to wrap back around the outer winding.

Creepage effects are present whenever you deal with high voltage, whether or not an isolation barrier is involved. An example is shown in Figure 9.82, illustrating the pin configuration of two package styles of a 1500 V MOSFET. For the larger TO-3PF package (5.4 mm lead spacing) an extension of the plastic package material around the drain lead provides adequate creepage distance; for the smaller TO-220FH package (2.5 mm lead spacing) there's a grooved structure and offset lead geometry.

9.8 A real-world switcher example

To convey the additional complexity involved in a production-model line-powered switching power supply, we disassembled a commercial single-output regulated switching supply[103] (Astrodyne model OFM-1501: 85–265 Vac input, 5 Vdc @ 0–3 A output), another in our series of "Designs by the Masters," revealing the circuit of Figure 9.83.

9.8.1 Switchers: top-level view

Let's take a walk through the circuit to see how a line-powered switcher copes with real-world problems. The basic topology is precisely that of the switching converter in Figure 9.48, implemented with flyback power conversion (Figure 9.73A); there are, however, a few additional components! Let's take it first at the broad-brush level, circling back later to delight in the refinements.

At this very basic level it goes like this: the line-powered bridge rectifier D_1 charges the 47 μF storage capacitor[104] (rated at 400 Vdc, to accommodate the 265 Vac maximum input), providing the unregulated high-voltage dc input (+160 Vdc or +320 Vdc, for 115 Vac or 230 Vac input, respectively) to the high side of the 70-turn primary winding of T_1. The low side of the winding is switched to input common (the $\perp$ symbol) at fixed frequency (but with variable pulse width) by the PWM switchmode controller chip U_1, according to feedback current at its FB terminal. On the secondary side the 3-turn paralleled secondaries are rectified by Schottky diode D_5, with "flyback" polarity configuration (i.e., nonconducting during the primary ON period). The rectified output is filtered by the four low-voltage storage capacitors (totaling 2260 μF), creating the isolated 5 Vdc output. This supply uses secondary-side regulation, comparing a fraction (50% nominal) of V_{out} with U_2's internal +2.50 V reference, turning on the LED emitter of optocoupler U_3 when the output reaches its nominal 5 Vdc. This couples to phototransistor $U_{3\text{b}}$, varying the feedback current into switchmode controller U_1, thus varying the ON pulse width to maintain regulated +5 Vdc output.

At this point we've accounted for perhaps a third of the components in Figure 9.83. The rest are needed to cope with issues such as (a) auxiliary power for the controller chip; (b) powerline filtering, mostly of *outgoing* switching noise; (c) protection (fusing, reverse polarity); (d) feedback loop compensation; and (e) switching transient snubbing and damping. And, although not obvious from the schematic, but most essential to the design – the choice of transformer parameters: core size and "gapping," turns ratios, and magnetizing inductance[105] L_{M}.

Before looking into those details, though, let's see how the basic converter works. We'll be able to figure out things like the voltage and current waveforms, peak voltages and currents, and the duty cycle as a function of input voltage and output current.

9.8.2 Switchers: basic operation

The control chip operates at a fixed frequency f_{osc} of 100 kHz, adjusting its primary switch conduction duty cycle ($D = t_{\text{on}}/T$) according to voltage feedback. We've drawn ideal waveforms for one cycle (duration $T = 1/f_{\text{osc}}$) in Figure 9.84. These are what you might expect in the

[103] Pictured in the northeast corner of Figure 9.1.

[104] The input storage capacitor is often called the *bulk capacitor*.

[105] The conventional symbols for magnetizing inductance and leakage inductance are L_m and L_l, respectively. But the lower-case L subscript can be hard on the eyes, especially in a footnote. In the interest of readability, therefore, we've adopted small upper-case subscripts: L_{M} and L_{L} throughout.

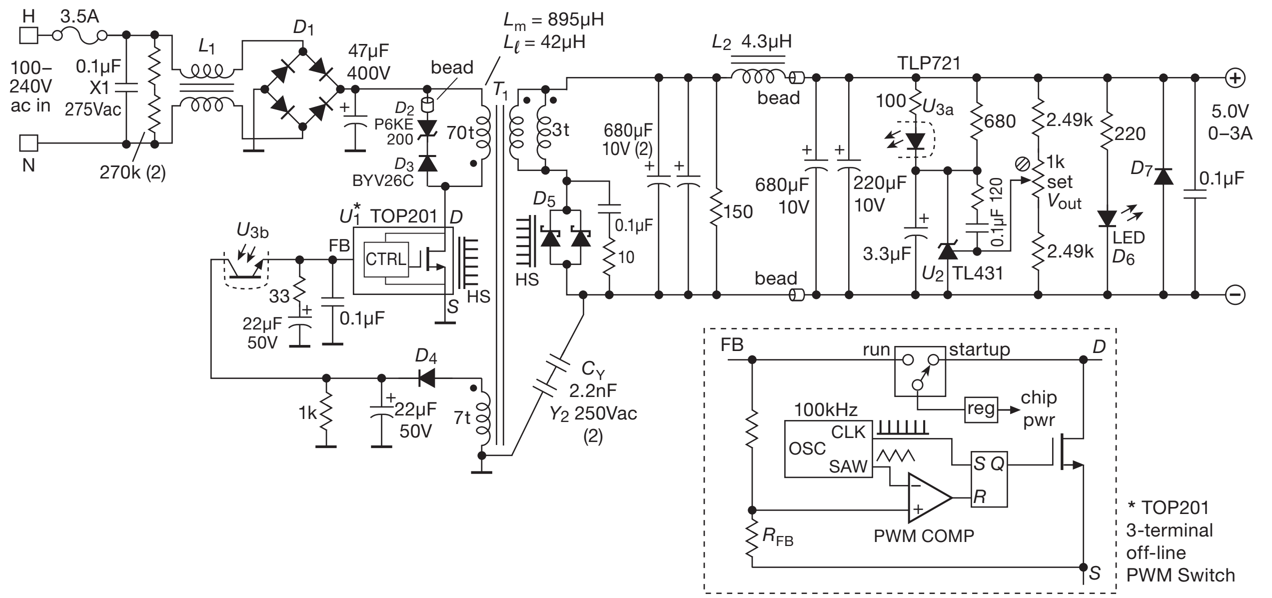

Figure 9.83. "Real-world" line-powered switching power supply. The circuit is relatively uncomplicated, thanks to its low power rating (15 W), and to the elegant 3-terminal switchmode controller U_1 from Power Integrations (with on-chip high-voltage power MOSFET). This is the open-frame "15W ac/dc switcher" shown in Figure 9.1.

absence of parasitic effects such as leakage inductance and switch capacitance.

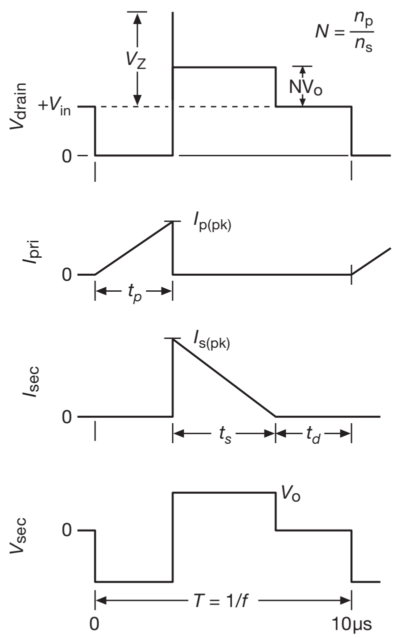

Figure 9.84. Ideal waveforms for an isolated flyback switching supply, operating in discontinuous-conduction mode.

A. The waveforms

We'll do the calculations shortly, but look first at the waveforms. (We've assumed the converter is operating in discontinuous-conduction mode, which will be borne out when we do the numbers.) During switch conduction the drain voltage is held at ground, putting $+V_{in}$ across the transformer primary and causing a ramp-up of primary current, according to $V_{in} = L_M \cdot dI_{pri}/dt$, where L_M is the primary "magnetizing inductance" (the inductance seen across the primary, with all other windings disconnected). That current ramps up to a peak value I_p, at which time there is a stored energy of $E = \frac{1}{2}L_M I_p^2$ in the transformer's core. When the switch turns off, the persistent inductive current transfers to the secondary winding, delivering that stored energy E to the output as the secondary current ramps down to zero, according to $V_{out}=L_{M(sec)} \cdot dI_{sec}/dt=(1/N^2)L_M \cdot dI_{sec}/dt$ (where $L_{M(sec)}$ is the magnetizing inductance seen at the secondary[106]). For the rest of the cycle there is no transformer current flowing.

The voltage waveforms are instructive. When the primary switch is turned off, at time t_p, the drain voltage rises

[106] Most of the time it's the magnetizing inductance seen at the *primary* that matters, for which we simply use L_M; in the few situations where we refer to the magnetizing inductance seen at the *secondary*, we add (sec) to the subscript: $L_{M(sec)}$.

well beyond the input supply voltage V_{in}: that is because the inductor tries to continue sourcing current into the drain terminal. The voltage would soar, but the secondary circuit goes into conduction instead (notice the polarity of "dotted" windings in Figure 9.83), clamping its output to V_{out}, which reflects back to the primary via the turns ratio N (shorthand for N_p/N_s). The brief spike shown in the figure is caused by some primary inductance[107] that is not coupled to the secondary, and therefore not clamped. This terrifying voltage spike is ultimately clamped by the zener clamp D_2 seen in the schematic (more on this later). When the secondary current has ramped down to zero, the voltage drop across both windings goes to zero; so the drain terminal sits at $+V_{in}$, and the voltage across the secondary winding goes to zero. Note that the latter is negative during primary switch conduction; it's a requirement that the "volt-time integral" (or "volt-second product") across any inductor average to zero, otherwise the current would rise without bound. That holds true for the primary also.

B. The calculations

Let us assume for simplicity that the converter is running at full load (5 V, 3 A) with nominal input voltage (115 Vrms or 160 Vdc).[108] We will calculate the switch duty cycle $D=t_p/T$, the secondary conduction duty cycle t_s/T, and the peak currents $I_{p(pk)}$ and $I_{s(pk)}$. It's easiest to take these in reverse order, doing the calculations from a simple energy standpoint.

The parameters We measure the magnetizing inductance seen at the primary to be L_M=895 μH, and the number of turns of primary and secondary to be N_p=70t and N_s=3t. From this we get the turns ratio $N=N_p/N_s=23.3$, which sets the voltage and current transformation ratios. Finally, from the turns ratio we get the magnetizing inductance as seen at the secondary side: $L_{M(sec)}=L_M/N^2=1.65\,\mu$H (impedances scale as N^2; see Chapter *1x*). A final parameter that we will use later is the measured primary leakage inductance $L_L=42\,\mu$H.

Peak currents The output circuit is delivering 15 W to the load; but, taking account of rectifier drop ($\sim$0.5 V) and the combined resistive losses in the secondary winding and filter inductor L_2 (10 mΩ), the transformer secondary is delivering an average power of approximately 6 V$\times$3 A, or 18 W. So, at a switching frequency of f_s=100 kHz, the transformer must deliver an energy increment of $E=P/f_s=180\,\mu$J during each switch cycle.

The rest is easy: we equate E to the magnetic energy in the core's magnetizing inductance, as seen at the secondary (because that's where it emerges). That is, $E=\frac{1}{2}L_{M(sec)}I^2_{s(pk)}$, from which we get $I_{s(pk)}=14.8$ A. Dividing by the turns ratio (N=23.3), we find that the peak primary current is $I_{p(pk)}=0.64$ A.

Conduction timing The primary switch stays on for a duration that ramps its current up to this peak current. That is, $t_p=L_M I_{p(pk)}/V_{in(dc)}=3.6\,\mu$s. The secondary conduction commences when the primary switch turns off, and continues for the time duration t_s needed to ramp its current down from $I_{s(pk)}$ to zero: $t_s=L_{M(sec)}I_{s(pk)}/V_{sec}=4.1\,\mu$s. Note that the successive conduction of primary and secondary totals 7.7 μs, which is less than the cycle time of 10 μs; that is, the converter is running in discontinuous conduction mode, as we assumed at the outset (and drew in Figure 9.84). There is a "dead time" of about 2.3 μs before the next switch conduction.

C. Comparison with reality

How well did we do with this basic model? To find out, we measured voltage and current waveforms of this converter, at nominal input voltage and full output load. They are shown in Figure 9.85. The good news is that the timing and peak currents are in very good agreement with our

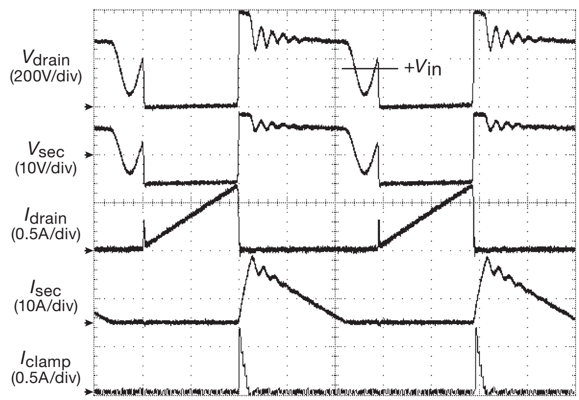

Figure 9.85. Measured waveforms for the switcher of Figure 9.83, running at full load (5 V, 3 A) and nominal input voltage (115 Vrms; V_{in}=160 Vdc). The arrows mark the location of zero voltage and current for each trace. Horizontal scale: 2 μs/div.

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107 This is in fact the infamous "leakage inductance" L_L. As with magnetizing inductance, we use the unadorned L_L to refer to leakage inductance seen at the primary winding; for secondary leakage inductance we add (sec) to the subscript: $L_{L(sec)}$.

108 Of course, a full design analysis must consider operation at the extremes, in particular at minimum input with maximum load (hence maximum duty cycle), and for the full range of output current with maximum input.

calculations. The bad news is that there are some real-world "features" that are absent from our basic waveforms of Figure 9.84. Most prominent are

(a) a substantial drain voltage spike at turn-off, followed by
(b) some fast ringing on both windings during secondary conduction, and
(c) slower ringing during the dead time at the end of the cycle.

Visible also is

(d) a drain current spike at turn-on.

These are caused by non-ideal behavior of the MOSFET switch and the transformer, as we'll discuss soon; but, to put some names onto them, these effects are due to

(a) primary leakage inductance,
(b) resonance of drain (and other) capacitances with primary leakage inductance,
(c) resonance of drain (and other) capacitances with primary magnetizing inductance, and
(d) "hard switching" of the voltage across the drain and other capacitances.

9.8.3 Switchers: looking more closely

Let's go back and fill in the missing pieces. In the real world you cannot ignore important effects such as the voltage and current transients that we saw in Figure 9.85, and numerous other details that account for all the components you see in the circuit diagram.

A. Input filtering

Beginning at the input, we find the obligatory fuse, and then an across-the-line "X" capacitor (§9.5.1D and following) and a series-coupled inductor pair, together forming an EMI and transient filter. It's always a good idea, of course, to clean up the ac power entering an instrument; here, however, filtering is additionally needed to keep RF hash generated *inside* the power supply from radiating *out* through the powerline.[109] This is not merely an act of altruism; there are regulatory standards governing permissible levels of radiated and conducted EMI.[110] The pair of 270k resistors discharges the X capacitor's residual voltage when the unit is unplugged.

B. Voltage range, inrush current, PFC

Note that this low-power (15 W) supply operates directly from a wide input voltage range (3:1), without a dual-voltage range switch in the manner of Figure 9.76B. Such wide-range operation is particularly convenient in chargers and power bricks for consumer electronics. It does, however, impose constraints on the design, because the converter must operate over a wide range of switch conduction duty cycle, and because the components must be sized for the wider range of peak voltages and currents. Absent, also, are any circuit elements to limit the inrush current during initial charging of the line-side storage capacitor. That's permissible in a small supply like this; but even with the relatively small 47 μF storage capacitor the specified typical inrush current is a hefty 20 A at 100 Vac input (and twice that for 200 Vac). Note also the absence of a PFC frontend; it's common practice to omit PFC in small supplies, but PFC is usually found in supplies of 50 W or more, at least in part from regulatory pressures. Note, by the way, that a PFC front-end reduces peak inrush current.

C. Auxiliary supply

Moving to the right, we see the interesting configuration of the "auxiliary supply," needed to power the internal circuits of the regulator–controller chip with low-voltage, low-power dc. An unattractive possibility would be 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 D_4) on T_1, thus saving a separate transformer. That's what's been done here, with the 7-turn winding, which generates a nominal +12 V output.

Sharp-eyed readers will have noticed a flaw in this scheme: the circuit cannot start itself, because the auxiliary dc is present only if the supply is already running! This turns out to be an old problem,[111] solved with a "kick-

[109] The important filter parameter here is not the converter's basic switching frequency, but rather the parasitic RF ringing frequency. If the latter is 2.5 MHz, for example, a lowpass filter with 250 kHz cutoff will attenuate the RFI by approximately $(f_{\rm RFI}/f_{\rm LPF})^2$, or 100×. With the 100 nF "X1" capacitor shown, the series inductance of the common-mode choke (its transformer leakage inductance) need be only $L=1/(2\pi f_{\rm LPF})^2C_{\rm X}=4\,\mu$H. Higher frequencies will be attenuated more, up to the frequency at which the PCB's wiring inductance and the choke's winding self-capacitance take over.

[110] In the US, electronic equipment must meet FCC Class A (for industrial settings) and Class B (more stringent, for residential settings) limits; in Europe the analogous standards are set by VDE.

[111] For example, designers of traditional CRT-based television sets faced the same quandary, when they derived all their low-voltage dc supplies from auxiliary windings on the high-frequency horizontal drive transformer, the latter itself activated by those same supplies.

start" circuit that powers initially from the high-voltage unregulated dc, switching over to its auxiliary dc power after things are running. We'd like to show you how this is implemented in detail, but we are frustrated in that worthy goal because in this supply those functions (and others) are cleverly integrated into the TOP201 controller chip (shown in simplified block diagram form in the dashed box).[112]

D. Controller chip: bias and compensation

Moving next to the controller chip itself, we see its internal high-voltage MOSFET (drawn explicitly, for clarity), which switches the low side of the primary to input common. The switch operates at fixed 100 kHz rate, varying the duty cycle according to the feedback, in a voltage-mode regulator. The chip is packaged in a 3-pin TO-220 plastic power package, and requires a small heatsink. Think about that – a *3-pin* switching regulator! Impossible, you say: it needs pins at least for common, drain, feedback, and chip power ("bias"). Surprisingly, this clever chip does it with just three, with the feedback terminal doing double duty as a bias pin. Feedback takes the form of a current into the FB pin, with an internal voltage divider to create the voltage-feedback signal that is presented to the PWM (duty-cycle) comparator, and a linear regulator to create the (higher) internal bias voltage. The remaining components on the primary side are for loop compensation (the series *RC* and *C* shunting the FB terminal), and for clamping and damping the inductive spike at the end of the conduction cycle (the 200 V zener transient suppressor and ferrite bead).

E. Input transient clamp (snubber)

At first you might reason that no clamp is needed, because the secondary circuit clamps the flyback voltage (as transformed to the secondary side by the turns ratio) to the output voltage. That is, after all, how a flyback works: the magnetic energy added to the core during switch conduction is stored in the transformer's magnetizing inductance ($E_M = \frac{1}{2} L_M I_p^2$), and released to the secondary circuit when the switch is turned OFF. But there is also "leakage inductance" (L_L, see Chapter *1x*), an effective series inductance caused by incomplete magnetic coupling between the windings.[113] The magnetic energy stored in L_L ($E_L = \frac{1}{2} L_L I_p^2$) is *not* transferred to, nor clamped at, the secondary, which is why you need the zener clamp on the primary side. (You can think of this unclamped energy as arising from the magnetic field of the primary that is not linked by the secondary.) This energy can be substantial – we'll see just how robust a zener is needed, even for this low-power switcher, when we do the clamp calculations in the next paragraph. It's worth noting that the effects of leakage inductance loom particularly large in a line-powered supply, because the required high-voltage insulation between primary and secondary mandates that the windings be physically well separated, causing incomplete flux coupling.

Let's take a moment to understand the drain voltage spike waveform in Figure 9.85. The primary-side leakage inductance, here measured to be 42 μH, though a smallish fraction (~5%) of the magnetizing inductance of 895 μH, stores that fraction of the total energy put into the transformer during primary switch conduction, and it is not transferred to the secondary; instead, it comes back out and is dissipated in the zener clamp D_2. That's about 0.84 W, which accounts for the robust zener that the designers chose. We can estimate the time duration of the primary current ramp to zero (call it t_{clamp}), mediated by the zener clamp. Look at Figure 9.86: the leakage inductance sees a clamp voltage equal to the zener voltage minus the reflected secondary voltage, which acts to ramp the primary

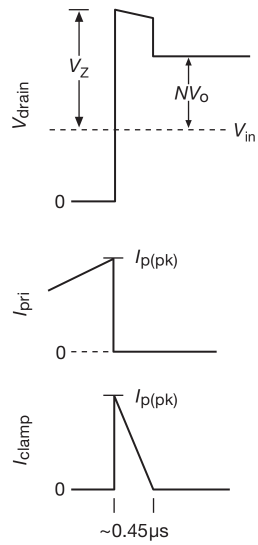


Figure 9.86. Drain-voltage spike caused by transformer leakage inductance. The zener clamp, whose voltage is higher than the reflected secondary output voltage, ramps the current to zero according to $V_Z - NV_{out} = L_L\, dI_D/dt$.

[112] Look in our second edition, where we devote six pages (pp. 361–366) to a complex offline switcher, if you want to see the gory implementation details of these and other features.

[113] Referring all inductances to the primary side, the magnetizing inductance L_M is what you measure across the primary terminals with all other windings left open-circuited, and the leakage inductance L_L is what you measure with all other windings short-circuited.

current down to zero from its starting value of $I_{p(pk)}$. So, from $V=LdI/dt$ we get $V_Z-NV_{out}=L_LI_{p(pk)}/t_{clamp}$, so $t_{clamp}=0.45\,\mu$s. This is in good agreement with the measured waveforms of Figure 9.85.

A final note on the clamp network: the zener D_2 is not a normal zener, but rather a "transient voltage suppressor" type (TVS; see discussion in Chapter *9x*), designed and specified to absorb large pulses of energy. The series diode D_3 is needed to prevent conduction during the switch-ON cycle, when the zener would conduct as a normal diode. There's an interesting problem associated with D_3, namely the fact that ordinary diodes have a "reverse recovery time" after forward conduction, which is due to charge storage effects, before they become non-conducting (this is the origin of the curious microsecond-scale spikes seen in a simple 60 Hz unregulated power supply; see §*9x.6*). For this reason D_3 in this circuit is a "fast soft-recovery" rectifier: the "fast" means that it turns off quickly (<30 ns), and the "soft" means that it does so smoothly, not abruptly. That's useful, because an abrupt current transition to nonconduction produces large inductive spikes ($V=LdI/dt$). In addition, the designers added a ferrite bead to damp and suppress such effects.

F. The transformer

In a flyback converter the primary and secondary conduction cycles do not overlap (as they do in, say, a forward converter). So all the energy that is being moved from primary to secondary must take up temporary residence in the transformer's core. That is, in a flyback converter the transformer is not "just a transformer": in addition to the usual transformer functions (voltage and current transformation by the turns ratio, and galvanic isolation), it is also an *inductor*, storing energy from the primary cycle in its magnetizing inductance to the tune of $E=\frac{1}{2}L_MI_{p(pk)}$. In fact, it's probably more accurate to think of it as "an inductor with a secondary winding." To enhance the energy storage functions, such transformers are usually designed with a deliberate gap in the magnetic material, which has the effect of raising the stored energy for a given applied volt-second product. This particular transformer is evidently gapped, because its value of A_L (the ratio of magnetizing inductance to turns squared) is low: $A_L=L_M/N_p^2=183\,\text{nH/t}^2$, compared with a value of the order of 1500 for an ungapped ferrite core of this size. (The nonconducting ferrite core is used to eliminate eddy current losses at the high operating frequency.)

As we discovered above, this converter runs in discontinuous conduction mode at nominal input voltage and full load current. In fact, it stays in DCM even at minimum input voltage (90 Vrms) and full load current, which is the combination that brings it closest to CCM. With a bit more transformer inductance it would enter CCM; presumably the design choice was based on the desire to keep it small, and also to avoid some issues associated with CCM.[114]

As we hinted earlier, the transformer's inductances are responsible for the ringing seen in the waveforms of Figure 9.85. Let's do a simple calculation of what frequencies we expect. During secondary conduction (immediately following primary switch turn-off) the primary circuit looks like a parallel LC, with leakage inductance L_L in parallel with the parasitic capacitances of the MOSFET and other components (clamp diode, primary winding). A reasonable estimate for the combined capacitances is something like 75 pF, due mostly to the transformer wiring and the clamp zener. So the parallel LC formed with the leakage inductance of 42 μH resonates at about 2.8 MHz, in good agreement with the observed ringing (∼2.5 MHz). At the completion of secondary conduction, the primary side no longer sees the leakage inductance (because the secondary is no longer clamped by the load); instead it sees the magnetizing inductance L_M of 895 μH (because the secondary is now open-circuited).[115] The new calculated resonant frequency then drops to about 615 kHz. You can see the first half-cycle of this slower resonance in the measured waveforms, centered on the +160 V dc input voltage, and interrupted by the onset of the next conduction cycle. (We later ran the converter at 25% load, which allowed three cycles of ringing at ∼600 kHz, in excellent agreement with this estimate.)

While we're on the subject of parasitic capacitances, this is a good time to note the ∼0.3 A current spikes at switch turn-on. This occurs because the switch is abruptly shorting a charged capacitor (the parallel capacitance of

114 Most notably some tendency for the output voltage to overshoot when there is an abrupt drop in load current (owing to a larger required inductance in a CCM design, perhaps influenced also by the nonzero magnetic field throughout the cycle), and also a change in the feedback loop behavior (because of the different functional dependence of output voltage versus duty cycle, and, more interestingly, the fact that in CCM the duty cycle is fixed, for a given output voltage, and is independent of load current). Given this last fact, it may seem paradoxical that regulation against changes in load current is even possible! What happens is that, once in CCM, a change in load current causes a *transient* change in duty cycle, such that the baseline (minimum) primary current moves up or down to accommodate the changed load current; having established that new baseline current, the duty cycle then returns to the fixed value appropriate to the regulated output voltage.

115 Somewhat overdamped by the reflected impedance of about 5 kΩ in series with ∼200 pF from the 10 Ω+0.1 μF secondary snubber network.

the switch itself, plus attached components). This is called “hard switching,” and is responsible for significant power losses in converters running at high switching frequencies. Here, for instance, we can estimate the power dissipated in the switch by multiplying $\frac{1}{2}CV^2$ by the switching frequency, giving $P_{\text{diss}} \approx 0.15$ W. That’s not too serious at this modest switching frequency of 100 kHz, being just 1% of the output power; but its relative contribution is greater at low load current, and in any case it contributes to switch dissipation and stress. And it becomes increasingly important as you try to increase the switching frequency (in order to reduce size). The solution is to strive for “soft switching,” in which the voltage across the switch is brought close to zero before switch activation (by exploiting reactive currents to discharge parasitic capacitances); this goal is called “zero-voltage switching” (ZVS).

G. Secondary power train

Moving to the secondary side, the rectifier is a Schottky type, which has both low forward voltage drop and zero recovery time (absence of charge storage).[116] Schottky rectifiers (also known as *hot carrier* rectifiers) are available at voltages up to ~100 V; above that you’d use a “fast-recovery” (or “fast soft-recovery,” like D_3) rectifier. Power rectifiers often come packaged as duals for applications that require two; here they’ve simply connected them in parallel. Note the heatsink: 3 A of average load current flowing through a 0.5 V (Schottky) forward drop dissipates 1.5 W, enough to merit a small heatsink. The series *RC* provides some damping and attenuation of switching transients, as do the ferrite beads. The series inductor L_2 filters ripple at the switching frequency: its reactance at the 100 kHz switching frequency is 2.7 Ω, compared with an impedance of ~0.1 Ω (dominated by series resistance) for the downstream storage capacitors.

H. Secondary regulation

This power supply uses the popular TL431 “shunt regulator,” which includes an internal voltage reference and error amplifier, and which goes into heavy conduction when the reference pin reaches 2.5 V above the ground pin. That turns on the optocoupler U_3’s LED emitter, for currents above about 2 mA (the threshold set by the 680 Ω resistor). The resistive divider and trimmer allow ±0.4 V output adjustment, and the series *RC* around the TL431 is a compensation network to prevent oscillation. The large shunt capacitor limits the loop bandwidth, and also accomplishes “soft-start” at power-on: this it does by tricking the opto-emitter into thinking that the TL431 is conducting, when in fact the LED current is coming from the ramp-up of output voltage. It’s easy to verify that an output ramp-up of 1.5 V/ms produces a sinking current of 5 mA at the cathode of the optocoupler LED, thus stretching the startup to about 3 ms, and therefore setting the secondary current needed to charge the four output storage capacitors to ~3.4 A, roughly equal to the supply’s maximum current rating.

I. Other design features

There are just a few additional goodies in this circuit. Capacitor C_Y is used to suppress conducted EMI. Because it bridges the isolation barrier, it must have appropriate “Y-capacitor” safety ratings (see §9.5.1). Rectifier D_7 protects against reverse polarity, in case some misbehaving load decides to create mayhem. The small output capacitor ensures low output impedance at high frequencies, where the large electrolytic capacitors become less effective (owing to internal inductance and ESR). And, finally, the switchmode controller itself (U_1) includes a host of nice features: internal oscillator requiring no external timing components, internal cycle-by-cycle current limit, overtemperature protection, automatic restart, internal regulator and dc source switching, and on-chip high-voltage power MOSFET, all integrated in an elegant 3-terminal configuration. Its high level of integration stole our thunder: it robbed us of the opportunity to show these important circuits explicitly!

9.8.4 The “reference design”

This is a nice power supply. We’ve bought a lot of them, and they work reliably and well. The circuit design might seem forbiddingly complicated, certainly for those inexperienced in offline switchmode supply design. In fact, we recommend strongly that the *user* of such supplies should not try to design and build them – *buy* them from the expert folks who do this for a living (see below).

But, how did those experts come up with this particular design? As it turns out, the manufacturers of interesting ICs have a great interest in making it painless to use their products. For this noble objective they provide what are known as *reference designs*, which basically consist of a complete circuit example (usually available from them as a “development board” or “evaluation board”). For the regulator chip used in this particular power supply, for example, Power Integrations (the TOP201 manufacturer) provides four example circuits, with increasing levels of regulation

[116] To deal with the high 15 A peak current in D_5, the designers selected a YG802C04 Schottky rectifier with its pair of 40 V 10 A sections paralleled (each of which specifies a forward drop of 0.53 V at 7 A), attached to its own heatsink.

stability (called "minimum parts count," "enhanced minimum parts count," "simple optocoupler feedback," and "accurate optocoupler feedback"). And in a pair of "Application Notes"[117] they provide a step-by-step recipe for these designs, complete with flowcharts, formulas, and graphs. You can hardly go wrong. The supply of Figure 9.83, in fact, closely follows the "accurate optocoupler feedback" design, differing primarily in the inclusion of soft start, ferrite beads, and reverse polarity protection. This is not to say that the design is a trivial exercise – the implementation of the transformer, the packaging and layout, and the process of testing and regulatory approval are all major challenges.

9.8.5 Wrapup: general comments on line-powered switching power supplies

- Line-powered switchers are ubiquitous, and for good reasons. 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 industrial and consumer electronics.
- Switchers are noisy! Their outputs have tens of millivolts of switching ripple; they put garbage onto the powerline; and they can even scream audibly! One cure for output ripple, if that's a problem, is to add an external high-current *LC* lowpass filter; alternatively, you can add a low-dropout linear post-regulator.[118] Some commercial converters include this feature, as well as complete shielding and extensive input filtering.
- 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 +3.3 V or +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 V output, say, changes when you vary the load on the +5 V 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!
- Line-powered switchers, like other switching converters, may have a minimum load-current requirement. If your load current could drop below the minimum, you'll have to add some resistive loading; otherwise the output may soar or oscillate.
- When working on a line-powered switcher, *watch out!* This is no empty warning – you can get yourself killed. Many components are at or above line potential and can be lethal. You can't clip the ground of your 'scope probe to the circuit without catastrophic consequences! (Use a 1:1 isolation transformer at the input, if you must go poking around.)
- Switchers usually include overvoltage "shutdown" 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.[119] For maximum safety you may want to add an autonomous outboard SCR-type crowbar.
- 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.
- Line-powered switchers are definitely complex and tricky to design reliably. Our advice is to avoid the design phase entirely, by *buying* what you need! After all, why build what you can buy?
- A switching supply, operating at roughly constant efficiency, presents a load that looks like a negative resistance (averaged over the ac wave) to the powerline that drives it. It can cause some crazy effects, including (but not limited to) oscillations, when combined with the input reactance of noise filters.

9.8.6 When to use switchers

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

[117] AN-14: "TOPSwitch Tips, Techniques, and Troubleshooting Guide"; AN-16: "TOPSwitch Flyback Design Methodology."

[118] You can get a switcher and LDO combined as a single regulator IC, for example in the Micrel "High Efficiency Low Dropout" (HELDO™) series.

[119] A personal anecdote: we smelled smoke, and found a dead 'scope in our lab one day. We opened it up, and found that the PFC's output storage capacitor (470 μF, 450 V) had failed, exuding lots of gooey stuff. No problem, we thought, we'll just replace it; especially since a new supply costs $800! Power on, looks good, go to lunch... come back, *smoke!* Turns out the PFC controller chip failed in a way that prevented both regulation and overvoltage shutdown, so the boost circuit just kept boosting, until the capacitor cried uncle.

- For *digital* systems, you usually need something like +2.5 V, +3.3 V, or +5 V, often at high current (10A or more). *Advice:* (a) Use a line-powered switcher. (b) Buy it (perhaps adding filtering, if needed).
- For analog circuits with low-level signals (small-signal amplifiers, signals less than 100 μV, etc.). *Advice:* Use a linear regulator; switchers are too noisy – they will ruin your life.[120] *Exception:* For some battery-operated circuits it may be better to use a low-power dc–dc switching converter.
- For high-power anything. *Advice:* Use a line-powered switcher. It's smaller, lighter, and cooler.
- 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 Simple Switcher 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.

9.9 Inverters and switching amplifiers

The benefits of switchmode power conversion – high efficiency and small size – can be applied to the generation of a *time-varying* output voltage. You can think of this as "dc-to-ac" conversion, as contrasted with a dc-to-dc power converter. In essence, you can imagine substituting an input *signal* for the fixed dc voltage reference of a switchmode dc regulator; the output will follow the input signal as long as the input signal bandwidth is well below the switching frequency.

Switching converters of this sort are used widely, for example to provide multiphase ac power for motor driving, or to generate the individual winding currents for microstepping motors. A variable-frequency motor driver lets you control the motor speed. Powerline-frequency dc-to-ac converters are often called *inverters*, such as those used in uninterruptible power supplies (UPSs) for computers. At higher power levels such inverters are used to generate powerline-frequency ac from the high-voltage dc that is shipped across the countryside (at dc voltages up to a *mega*volt, would you believe?). And, closer to home, switching audio amplifiers (known as "class-D" amplifiers; see §2.4.1C) are dominant in consumer electronics. In that application a passive *LC* lowpass filter smooths a rail-to-rail switching waveform (typically at frequencies of ~250 kHz or higher) whose duty cycle is modulated in accordance with the input signal. See Figure 2.73 for waveforms from a low-power class-D audio amplifier.

To get just a taste of this subfield of power electronics, take a look at Figure 9.87, where we've captured waveforms from two styles of uninterruptible 120 Vac power supplies, along with the raw 120 Vac wallplug power in our lab. You might guess that the clean middle waveform is the utility power, but in fact that waveform is the loaded output from a UPS that boasts "low-distortion sine wave." The top waveform is the wallplug power, showing rather typical levels of distortion. The 3-level waveform at the bottom is euphemistically called a "modified sinewave," and is typical of less-expensive inverters and UPSs. It isn't handsome, but it does the job: if switched to ±170 V rails 25% of the time, and to zero (or unpowered) in between, it's easy to figure out that it has the same rms voltage (120 Vrms) and peak voltage (170 Vpk) as a 120 Vrms sinewave.[121] So it delivers the same power to resistive loads, etc., and it

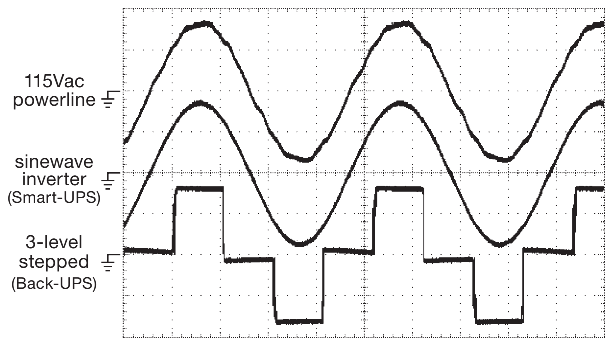


Figure 9.87. A true sinewave inverter generates a cleaner sinewave than wallplug ac. The 3-level waveform (sometimes called *modified sinewave*), though hardly a sinewave, has the same rms and peak voltages, and suffices for most loads. For these measured waveforms we loaded the APC UPSs with a 75 W incandescent lamp. Vertical: 100 V/div; horizontal: 4 ms/div.

[120] *Really!* Here's a pithy quote from James Bryant (from the Analog Devices "Rarely Asked Questions" series), in answer to the question "How can I prevent switching-mode power supply noise from devastating my circuit performance?" Answer: "With great difficulty – but it can be done." He continues: "Switching-mode power supplies are inherently the noisiest circuits imaginable. A large current from the supply is being turned on and off at high frequency with very fast dI/dt. There are inevitably large fast voltage and current transients."

[121] Just add up the squared voltages in equal time intervals, then take the square root of their average: $V_{\rm rms} = \left[(V_1^2 + V_2^2 + \cdots + V_n^2)/n\right]^{1/2}$.

charges the input side of dc power supplies or converters to the same voltage as would the 120 Vrms utility line power.

There's more to think about than simply having the same rms and peak voltage, of course. There's *distortion*: the 3-level waveform has no even harmonics, but it has strong harmonics at all odd multiples of the fundamental frequency (there are various multilevel schemes to address this problem). Then there's the worry about systems that exploit zero crossings of the incoming ac power for timing, for which the 3-level waveform (or any stepped waveform with an odd number of levels) would wreak havoc.[122] There's plenty more to think about, even limiting ourselves to the subject of multilevel inverters.[123]

This area of power electronics is a rich subject; but, sadly, life is finite, and so (barely) is the size of this book.

9.10 Voltage references

Quite apart from their use in integrated voltage regulators, 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 the best integrated regulators. Or you might want to construct a precision constant-current supply. Other applications requiring precision references (but not a precision power supply) include A/D and D/A converters, precision waveform generators, and accurate voltmeters, ohmmeters, or ammeters.

Integrated voltage references come in two styles: *2-terminal* (or *shunt*), and *3-terminal* (or *series*). Two-terminal references act like zener diodes, maintaining a constant voltage drop when current is flowing; the external circuit must provide a reasonably stable operating current. Three-terminal references (V_{in}, V_{out}, GND) act like linear voltage regulators, with internal circuitry taking care of biasing the internal reference (whether a zener diode, or something else). In Tables 9.7 and 9.8 (on pages 677 and 678, respectively) we've listed an abundance of currently available references of both types.

There are four different technologies used in currently available voltage references, all of which exploit some physical effect to maintain a well-defined and stable voltage – *zener diodes*, *bandgap references*, *JFET pinchoff references*, and *floating gate references*. They are all available as stand-alone (2-terminal or 3-terminal) components; they are also commonly incorporated as an internal voltage reference within a larger IC such as an A/D converter. Let's take them in order.

9.10.1 Zener diode

The simplest form of voltage reference is the zener diode, a 2-terminal device we introduced in §1.2.6A. 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 (being simple, inexpensive, passive 2-terminal devices), zeners lose their luster when you look a bit more closely: it is necessary to stock a selection of values, the voltage tolerance is poor except in high-priced precision zeners, they are noisy (above 7 V), and the zener voltage depends on current and temperature. As an example of the last two effects, a 27 V zener in the popular 1N5221 series of 500 mW zeners has a temperature coefficient of $+0.1\%/^\circ$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 are quiet, become very stiff against changes in current, and simultaneously achieve a nearly zero temperature coefficient. The graphs in Figures 9.88 and 9.89 illustrate the effects.[124] If you need a zener for use as a stable voltage reference only, and you don't care what voltage it is, one possibility is to use one of the compensated zener references constructed from a 5.6 V zener (approximately) in series with a forward-biased diode – if you can find one! (Read on...) The zener voltage is chosen to give a positive coefficient to cancel the diode's temperature coefficient of -2.1 mV/°C. Temperature compensation can be

[122] One solution is a 6-interval, 4-level waveform: V_{pk}, $V_{\text{pk}}/2$, $-V_{\text{pk}}/2$, $-V_{\text{pk}}$, spending twice as much time at $V_{\text{pk}}/2$ as at V_{pk}, and never dwelling at zero. This eliminates the 3rd harmonic (as well as all the even harmonics), and produces 120 Vrms if V_{pk}=170 V.

[123] A nice review is found in J. Rodriguez *et al.*, "Multilevel inverters: a survey of topologies, controls, and applications," *IEEE Trans. Indus. Electronics.*, **49**, 724–738 (2002), complete with 78 references.

[124] This peculiar behavior comes about because there are two competing mechanisms going on in zener diodes: zener effect at low voltages, with negative tempco; and avalanche breakdown at high voltages, with positive tempco.

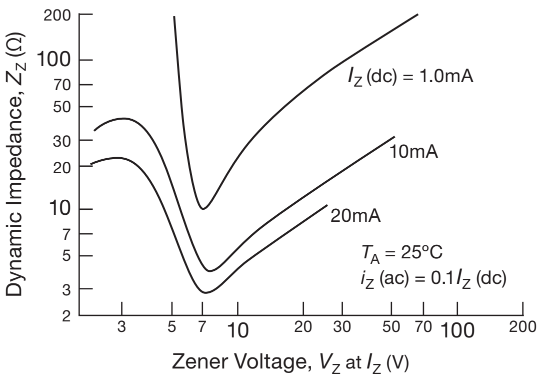


Figure 9.88. Zener diode dynamic impedance for zener diodes of various voltages. (Courtesy of Motorola, Inc.)

accomplished for other zener voltages also, for example in the 1N4057–85 series, which go from 12 V to 200 V, with tempcos of 20 ppm/°C.

Let's follow this thread – which, as we'll see, will lead us to a far better solution in the form of fully integrated voltage references (including those with a temperature-compensated zener on-chip) with superior characteristics. In fact, temperature-compensated zeners, as *discrete* devices, have become largely an extinct breed.

A. Providing operating current

A compensated zener could be used as stable voltage reference within a circuit, but it must be provided with constant current.[125] The tightly specified[126] 1N4895, for example, is specified as 6.35 V±5% at 7.5 mA, with a tempco of 5 ppm/°C (max) and incremental resistance of 10 Ω (max). So a change in bias current of 1 mA can change the reference voltage by 10 mV, three times as much as a change in temperature from 0°C to +100°C. You can, of course, rig up a separate current source circuit to bias the zener; but you can do better – Figure 9.90 shows a clever way to use the zener voltage itself to provide a constant bias current. The op-amp is here wired as a noninverting amplifier in order to generate an output of +10.0 V. That stable output is itself used to provide a precision 7.5 mA 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 nuisance problem.[127] Be sure to use an op-amp that has common-mode input range to the negative rail ("single-supply" op-amps).

There are compensated zeners available that characterize the stability of zener voltage with *time*, a specification that normally tends to get left out. The 1N4895, for example, specifies stability of better than 10 ppm/1000h. The ultimate example is probably the LTZ1000, a 7.15 V integrated zener whose datasheet specifies an astonishing long-term stability of 0.15 ppm/$\sqrt{\text{kHr}}$ (typ). This puppy includes an on-chip temperature stabilizing heater, and claims to deliver a tempco as low as 0.05 ppm/°C, if treated with respect.[128] Such zeners are not cheap: the LTZ1000 will set you back $50.

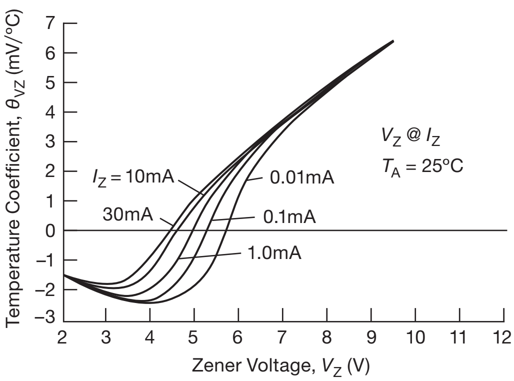


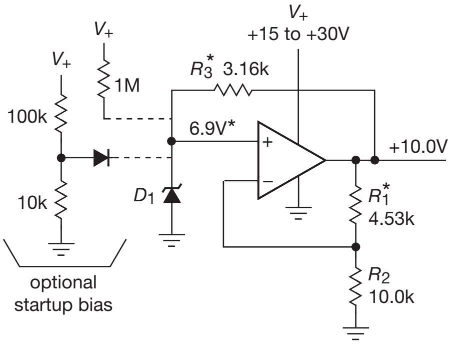
Figure 9.89. Temperature coefficient of zener diode breakdown voltage versus the voltage of the zener diode. (Courtesy of Motorola, Inc.)

[125] Most small zener diodes are specified at an operating current of 20 mA (though you can run them at lower currents). But, happily for those seeking low-current zeners, there's the 1N4678 to 1N4713 family (MMSZ4678–4713 for surface-mount SOD-123 package), specified at 50 μA.

[126] And individually tested for 1000 hours! It's available only from the manufacturer, Microsemi.

[127] With a caveat: the circuit could get stuck at zero output if the op-amp's input offset voltage is greater than the ground-saturated output voltage. This could happen with a rail-to-rail CMOS output stage, which is why we chose a BJT op-amp (whose saturation voltage is at least a few millivolts from ground). If you select a CMOS op-amp (say a precision chopper), or if you're losing sleep over the remote possibility of a stuck circuit, you can force the circuit to start correctly with either of the dotted supplements to the circuit.

[128] An example of appropriate respect is the prevention of thermal gradients: a junction of two dissimilar metals (a "thermocouple") generates a thermal EMF, about 35 μV/°C for the connection of the LTZ1000's Kovar-alloy leads to a circuit board. That's about 7 ppm per °C temperature *difference* in the two leads, which is a hundred times larger than that of the chip's zener itself!



Op-amp:			
	LT1077	2µV/°C max	$2.08
	½LM358A	15µV/°C max	$0.14

	1N5232B	LM329B	LM399A	1N4895	
V_Z	5.6	6.9	6.95	6.35	volts
I_Z	1	1	1	7.50	mA
R_3	4.42k	3.16k	3.01k	487	Ω
R_1	7.87k	4.53k	4.42k	5.76k	Ω
tempco	380	20	1	5	ppm/°C, max
drift	–	20	8	10	ppm/khr
price	0.14	1.80	9.37	RFQ	$US, qty 25
op-amp	LM358A	←	LT1077	→	

* values shown for LM329B

Figure 9.90. Stable output voltage provides a stable zener bias current over varying supply voltages V_+. The op-amp must operate to the negative rail. For a jellybean zener like the 1N5232 you can use an inexpensive op-amp like an LM358; but use a precision op-amp (e.g., the LT1077) to preserve the low tempco of a precision reference like the LM329 or LM399. Don't use a split supply here, because the output could happily settle to a negative output.

B. IC zeners

We hinted that precision-compensated zeners as *discrete* devices have largely disappeared; you can check this out for yourself by going to a site like Octopart.com, looking for once-popular parts like the 1N4895 or 1N821–29 series.

That's the bad news. The good news is that excellent compensated zeners now come in integrated form, as the internal reference within a variety of IC voltage references. Table 9.7 lists several of these, from the inexpensive (less than $1) LM329 to the spectacular LTZ1000. These include additional circuitry to yield improved performance (most notably, constancy of terminal voltage with applied current), in the form of an integrated circuit; it looks electrically just like a zener, with just two terminals, although internally it includes additional active devices. Being zener-based, these devices operate around the sweet spot of 7 V, although some (like the LT1236 in the table) include internal amplifier circuitry to create a round-number 10.0 V "zener."[129]

The jellybean LM329 is worth keeping in mind when you need only a "good-enough" zener reference; it has low noise, a zener voltage of 6.9 V, and in its best version it has a temperature coefficient of 10 ppm/°C (max), when provided with a constant current of 1 mA. Where better performance is needed, consider the LT1236A or the thermally regulated (on-chip heater) LM399A, the latter with an admirable 1 ppm/°C worst-case tempco!

When thinking about 2-terminal zener references, don't overlook the other voltage reference technologies that are available as 2-terminal (shunt) devices (see Table 9.7 on the facing page). From the outside they behave just like zener diodes, but they use other tricks (e.g., a V_{BE} drop) internally to create their stable reference voltage. Among other benefits, such devices come in desirable low voltages (1.25 V and 2.5 V are common), and some can operate at currents as low as 1 μA. Read on!

And remember always not to limit yourself to *2-terminal* references – there are excellent 3-terminal references, both zener based and otherwise. A fine example is the LT1027B, a zener-based 5.0 V reference with excellent tempco (2 ppm/°C, max) and low noise (3 μVpp typ, 0.1–10 Hz). A nice feature of most of the integrated references (both 2-terminal and 3-terminal) is the convenient output voltages they provide: instead of having to deal with something like V_{out}=6.95 V±4% (the voltage specification of the excellent LM399 temperature stabilized 2-terminal zener reference), you get precise round-number output voltages like 1.25 V, 2.50 V, 5.0 V, and 10.0 V, factory trimmed to an accuracy as good as ±0.02% (see Tables 9.7 and 9.8).[130]

[129] 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; this is the technology that is used to achieve the very low noise of parts like the LT1236 and LTZ1000 references.

[130] Also powers-of-2 voltages (2.048 V, 4.096 V) to set round-number LSB steps in ADCs and DACs.

Table 9.7 Shunt (2-terminal) Voltage References[a]

Part #	Packages: TO-92	DIP	SOIC	SOT-23	SC70	other	Voltages: 1.235	2.048	2.50	3.0	4.096	5.00	other	Accy max (%)	Adjust version: trim pin	avail	V_{max}	Zener current: min[q] (µA)	max (mA)	Noise: 0.1-10Hz typ (µVpp)	noise density[x] (nV/√Hz)	Load cap C_L	Tempco typ (ppm/°C)	Tempco max (ppm/°C)	R_{out} typ[r] (Ω)	Price qty 25 ($US)	Comments
shunt / feedback references																											
TL431A	•	8	-	3,5	-	8	-	-	•	-	-	-	-	1	-	•	36	1000	100	10^{o}	20	6uF^{g}	6^{f}	16^{f}	0.22^{b,u}	0.32	1,2
LMV431B	•	-	-	3,5	-	-	w	-	-	-	-	-	-	0.5	-	•	30	80	20	7^{o}	195	2nF	4^{f}	12^{f}	0.25^{b,u}	0.85	1,3
TLV431B^{p}	•	-	8	3,5	6	-	w	-	-	-	-	-	-	0.5	-	•	16^{v}	100	20	15^{o}	220	20uF	6^{f}	20^{f}	0.25^{b,u}	0.67	1,4
bandgap references																											
LM4431	-	-	-	3	-	-	-	-	•	-	-	-	-	2	-	-	-	100	15	-	170	N	30	-	1	0.75	
LM336B-2.5	•	-	8	-	-	-	-	-	w	-	-	-	-	2	•	-	-	400	10	-	95	-	1.8^{f}	6^{f}	0.27^{u}	0.95	5
LM336B-5.0	•	-	8	-	-	-	-	-	-	-	-	•	-	2	•	-	-	600	10	-	95	-	4^{f}	12^{f}	0.6	0.95	5
LM336Z5	•	-	-	-	-	-	-	-	-	-	-	•	-	2	•	-	-	600	10	-	-	-	4^{f}	12^{f}	0.6	0.06	5,6
LM385B	•	-	8	-	-	20	w	-	-	-	-	-	-	1	-	•	5	11	20	-	400	N	-	150	0.4^{b}	1.38	7
LM385B-1.2	•	-	8	-	-	8	•	-	-	-	-	-	-	1	-	-	-	15	20	-	490	N	20	-	0.4	0.54	-
LM385B-2.5	•	8	8	-	-	-	-	-	•	-	-	-	-	1.5	-	-	-	18	20	-	480	N	20	-	0.4	0.54	-
LT1034	•	-	-	-	-	-	w	-	•	-	-	-	•	1.2	-	-	-	30^{g}	20	-	24	-	20	40	0.5	3.33	8
ADR510	-	-	-	3	-	-	1.00V	-	-	-	-	-	-	0.35	•	-	-	100	10	4	-	N	-	70	0.3	1.42	9
LT1004	•	-	8	-	-	-	•	-	•	-	-	-	-	0.8	-	-	-	20^{g}	20	-	260	N^{k}	20	-	0.2	1.74	10
LT1004 (TI)	-	-	8	-	-	-	•	-	•	-	-	-	-	0.3	-	-	-	10^{h}	20	-	310	N^{k}	20	-	0.2	0.94	10,11
MAX6006A	-	-	8	3	-	-	w	-	-	-	-	-	-	0.2	-	-	-	1.0^{h}	2	30	high	R	30	-	1.5	1.56	12,13
LT1009	•	-	8	-	-	8	-	-	•	-	-	-	-	0.2	•	-	-	400	10	-	48	-	15	25	0.2	1.74	14
LT1029A	•	-	-	-	-	-	-	-	-	-	-	•	-	0.2	•	-	-	700	10	-	quiet	N^{s}	8	20	0.2	3.34	-
LT1029	•	-	-	-	-	-	-	-	-	-	-	•	-	1.0	•	-	-	700	10	-	quiet	N^{s}	12	34		2.07	15
LM4040A	•	-	-	3	5	-	-	•	•	•	•	•	2	0.1	-	-	-	75	15	-	165	N^{s}	15g	-	0.3^{g}	2.32	16
ADR5041B	-	-	-	3	3	-	-	•	•	•	•	•	-	0.1	-	-	-	50	15	3.2	600	N	10	75	0.2^{u}	0.86	17
LM4041A	•	-	-	3	5	-	w	-	-	-	-	-	-	0.1	-	•	15	60	12	-	165	N^{s}	15h	-	0.5^{h}	1.55	16
LM4050A	-	-	-	3	-	-	-	•	•		•	•	2	0.1	-	-	-	60	15	-	180	N^{s}	15g	50^{g}	0.3^{g}	2.43	-
AD1580B	-	-	-	3	3	-	w	-	-	-	-	-	-	0.1	-	-	-	50	20	5	160	N^{k}	-	50	0.4	1.56	-
MAX6138	-	-	-	-	3	-	w	•	•	•	•	•	•	0.1	-	-	-	65	15	20	325	N^{s}	4	25	0.3	2.08	-
LT1634A	•	8	8	-	-	-	w	-	•	-	•	•	-	0.05	-	-	-	8	20	15^{g}	-	N^{s}	4	10	0.15^{g,u}	3.13	19
LT1389A	-	-	8	-	-	-	w	-	•	-	•	•	-	0.05	-	-	-	0.7	2	25	noisy	N^{s}	4	10	0.25	7.43	20
buried zeners																											
LM329	•	-	-	-	-	-	-	-	-	-	-		6.9	5	-	-	-	600	15	-	11	-	50	100	1^{u}	0.79	21
LT1236A-10	-	8	8	-	-	-	-	-	-	-	-		10.0	0.05	•	-	-	1700	20	6	13	-	2	5	0.5^{b,u}	5.09	22
LM399AH	-	TO-46 metal					-	-	-	-	-		6.95	4	-	-	-	500	10	-	13	-	0.3	1	0.5	9.37	23
LTZ1000	-	TO-5 metal					-	-	-	-	-		7.15	4	-	-	-	1000 s	5	1.2	5.5	-	0.05	-	-	55.00	23,24
zener diodes																											
1N4370A	DO-35						2.4 V							5	-	-	-	20mA spec		-	-	-	−600	-	30	2.11	25
1N752A	DO-35						5.6 V							5	-	-	-	20mA spec		-	-	-	300	-	11	1.56	26
1N4895	DO-7						6.35 V							5	-	-	-	7.5mA spec		-	-	-	5e	-	10	na	27
1N3157	DO-7						8.4 V							5	-	-	-	10mA spec		-	-	-	10c	-	15^{m}	na	28

Notes: (a) generally listing best accuracy grade.; sorted by type and increasing accuracy. (b) wired as a zener. (c) at 10mA. (C_L) load capacitor — R: >10nF required; N: not required but allowed, or recommended for transient-loads; µF = min required if more than a small cap is added, see datasheet; blank = no comment. The ac output impedance rises with frequency and will resonate with the load capacitor's reactance. A small resistor (22 to 100Ω, etc.) can isolate the capacitor and lower the resonance Q. (d) 5-10mA. (e) at I_Z=7.5mA. (f) ΔV (mV) over temp. (g) for the 2.5V version (the 1.2V version is generally less). (h) for the 1.225 version, or V_{ref} for the adj version. (k) an RC is suggested, e.g. 22Ω. (m) min or max. (n) nominal. (na) not available. (o) of the 1.24V ref, gained up to V_{clamp}. (p) also TLVH431A. (q) minimum operating current (maximum, i.e., worst-case); often higher for higher fixed voltages. (r) usually at 1mA, but not current dependent. (s) see datasheet. (t) typical. (u) spec'd over operating range. (v) 6V for TI's TLV431, 16V for Onsemi TLV431 or TI TLVH431. (w) see datasheet for exact value, chosen for minimum tempco. (x) scaled to 1.0V output; multiply listed value by V_{out}.

Comments: **1:** two resistors set V_{clamp}. **2:** I_{ref}=4µA max. **3:** I_{ref}=0.5µA max; complementary to LM385-adj. **4:** I_{ref}=0.5µA max; TLV432 is alternate pinout. **5:** LM336 has voltage-trim pin. **6:** multiple-source jellybean. **7:** -BX version is 30ppm/°C; I_{ref}=15nA. **8:** dual: bandgap and 7V zener (1.6%, 40ppm/°C typ, 90Ω), common neg terminal. **9:** lowest V_{ref} shunt ref. **10:** 1.235V is 0.3% tol, 2.45V is 0.8% tol. **11:** TI's -CDR suffix costs $0.25 (qty 25). **12:** nanopower, min I_Z=1µA; 40ms turn-on settling time with 1.2µA bias and 10nF cap. **13:** MAX6007, 08, 09 for other voltages. **14:** LM336 upgrade. **15:** non -A version. **16:** -B, -C, -D suffix looser tolerance. **17:** -A suffix 0.2% tolerance. **19:** -C suffix for looser tolerance. **20:** nanopower. **21:** -A suffix is 5ppm/°C typ, 10ppm/°C max. **22:** series ref used in shunt mode. **23:** on-chip heater; lowest guaranteed tempco. **24:** factory purchase. **25:** low-voltage zeners are poor! **26:** optimum zener voltage. **27:** tested 1k hours; "reference zener," spec'd at 7.5mA only. **28:** temp comp zener reference.

Table 9.8 Series (3-terminal) Voltage References[a]

Part #	Reference[c]	Pkgs[b]					Available Voltages[d] (V)											ADJ version	Accy	Tempco		Supply		I_{out}		Trim pin	Filter pin	Shutdown pin	sense +	Noise		Regulation	C_{in}[z]	C_{out}[z]		Price	Comments
		DIP	SOIC	SOT-23	TO-92	other	1.024	1.25	1.8	2.048	2.5	3.0	3.3	4.096	5.0	10.0	other		max (%)	typ (ppm/°C)	max (ppm/°C)	range min-max (V)	I_Q typ (μA)	source (mA)	sink (mA)					0.1Hz-10Hz typ (μVpp)	10Hz-10kHz typ (μVrms)	max (ppm/V)	min (μF)	min (μF)	max (μF)	qty 25 ($US)	
LP2950A	B	-	-	-	•	P	-	-	-	-	-	•	•	-	●	-	-	-	0.5	20	100	5.4-30	100	100	-	-	-	-	-	-	430[r]	0.1%[q]	1	1	-	0.66	1
LP2951	B	8	8	-	•	S	-	-	-	-	-	●	•	-	•	-	-	•	0.5	20	120	3.35-30	100	100	-	•	•	•	•	-	430[r,s]	0.1%[q]	1	1	-	0.56	2
LT6650	B	-	-	5	-	-	-	-	-	-	-	-	-	-	-	-	0.4	j	0.75	30	-	1.4-18	5.6	0.2	0.2	-	-	-	•	20	23[e]	900	0.1	1	-	1.53[k]	3
LM4128A-2.5	B	-	-	5	-	-	-	-	•	•	●	•	•	•	-	-	-	-	0.1	-	75	2.9-5.5	60	20	-	-	-	-	-	275	-	50 t	0.1[w]	-	10	1.80	4
AD680J	B	8	8	-	•	-	-	-	-	-	●	-	-	-	-	-	-	-	0.4	10	25	4.5-36	200	10	-	-	-	-	-	8	-	16	0.1	-	0.05	3.20	5
REF43G	B	8	8	-	-	-	-	-	-	-	●	-	-	-	-	-	-	-	0.2	10	25	4.5-40	340	20	1.2	•	-	-	-	-	7[e]	2	0.1	-	10	6.66	6,7
ISL21010-25	B	-	-	3	-	-	•	•	-	•	●	•	•	•	-	-	1.5	-	0.2	15	50	2.6-5.5	46	25	1	-	-	-	-	67	37[e]	100	-	0.1	10	0.70[k]	-
LM4120A-2.5	B	-	-	5	-	-	-	-	•	•	●	•	•	•	•	-	-	-	0.2	14	50	2.7-12	160	5	5	-	-	•	-	20	36[p]	80	-	0.022	0.047	2.37	8
LM4125A-2.5	B	-	-	5	-	-	-	-	-	•	●	-	-	•	-	-	-	-	0.2	14	50	2.7-6	160	5	5	-	-	-	-	20	36[p]	80	-	0.022	0.1	1.75	8
LM4132A-2.5	B	-	-	5	-	-	-	-	•	•	●	•	•	•	-	-	-	-	0.05	-	10	2.9-5.5	60	20	-	-	-	•	-	240	-	50 t	0.1[w]	-	10	2.60	4
ISL60002B-25	F	-	-	3	-	-	•	•	•	•	●	•	•	-	-	-	1.2	-	0.04	-	20	2.7-5.5	0.35	7	7	-	-	-	-	30	-	100	-	-	0.001[v]	5.86	4
REF3125	B	-	-	3	-	-	-	•	-	•	●	•	•	•	-	-	-	-	0.2	5	15	2.55-6	100	10	10	-	-	-	-	33	48	25	0.47	-	-	2.07	7
REF3225	B	-	-	6	-	-	-	•	-	•	●	•	•	•	-	-	-	-	0.2	4	7	2.55-6	100	10	10	-	-	•	•	33	48	65	0.47	1	10	3.60	7
AD584K	B	8	-	-	-	M	-	-	-	-	●	-	-	-	•	•	7.5	-	0.14	-	15	4.5-30	750	10	5	•	•	•	•	50	40	50	-	-	-	10.61	11
X60003B-50	F	-	-	3	-	-	-	-	-	-	-	-	-	●	•	-	-	-	0.02	-	10	5.1-9	0.5	10	10	-	-	-	-	30	-	60	-	-	0.001[v]	2.68[h]	7
LT6656A-2.5	B	-	-	6	-	T	-	•	-	•	●	•	•	•	•	-	-	-	0.05	5	10	2.9-18	0.85	5	0.1	-	-	-	-	30	80[e]	25	0.1	1	-	6.42	4,13
REF5025	B	-	8	-	-	S	-	-	-	•	●	-	-	•	•	-	-	-	0.05	4	6.5	5.2-18	800	10	10	•	•	-	-	7.5	-	1	1	1	50	2.97	6
ADR3425	B	-	-	6	-	-	-	-	-	-	●	•	-	•	•	-	-	-	0.1	2.5	8	2.7-5.5	85[m]	10	3	-	-	•	•	18	42	50	0.1	0.1	-	2.34	-
LT1021B-5	Z	8	8	-	-	-	-	-	-	-	-	-	-	-	●	•	7.0	-	0.1	2	5	7.2-36	800	10	10	•	-	-	-	3	2.2[e]	12	-	-	-	7.00	7,14
LT6654A-2.5	B	-	-	6	-	T	-	•	-	•	●	•	•	•	•	-	-	-	0.05	3	10	2.7-36	350	10	10	-	-	-	-	1.5	2[e]	5	0.1	1	-	5.66	4,7,15
LT1019A-2.5	B	8	8	-	-	-	-	-	-	-	●	-	-	-	•	•	4.5	-	0.05	3	5	4-36	650	10	10	•	-	-	-	6.3	6.3[e]	3	-	y	y	7.50	6,14
LTC6652A-2.5	B	-	-	-	-	S	-	•	-	•	●	•	•	•	•	-	-	-	0.05	2	5	2.8-12	350	5	5	-	-	•	-	5	7.5[e]	50	-	-	-	4.80	4
ISL21090B-25	B	-	8	-	-	S	-	•	-	-	●	-	-	-	•	•	7.0	-	0.02	-	7	3.7-36	930	20	10	•	-	-	-	1.9	1.6[e]	18	-	0.1	10	7.56	7
LT1236A-5	Z	8	8	-	-	-	-	-	-	-	-	-	-	-	●	•	-	-	0.05	2	5	7.2-36	800	10	10	•	-	-	-	3	2.2[e]	12	-	-	-	5.09	7
MAX6033A-25	B	-	-	6	-	-	-	-	-	-	●	-	•	-	-	-	-	-	0.04	1.5	7	2.7-12.6	75[m]	15	-	-	-	-	•	16	12[e]	10	-	0.1	100	6.32	7
ADR02B	B	-	8	5	-	-	-	-	-	-	•	•	-	-	●	•	-	-	0.06	1	3	7-36	650	10	-	•	-	-	-	10	-	30	-	-	-	5.00	4
LM4140A-2.5	B	-	8	-	-	-	•	•	-	•	●	-	-	•	-	-	-	-	0.1	-	3	2.7-5.5	230	8	0.01	-	-	•	-	2.2	-	200	-	1[x]	4.7[x]	2.87	4,7
ADR441B	J	-	8	-	-	S	-	-	-	•	●	•	-	•	•	-	-	-	0.04	1	3	3-18	3000	10	5	•	-	-	-	1.2[g]	-	20	0.1	0.1	-	6.53	9,10
ISL21009B25	F	-	8	-	-	S	-	•	-	•	●	•	-	-	-	-	-	-	0.02	-	3	3.5-16.5	95	7	7	•	-	-	-	4.5	2.2[e]	60	-	-	0.001[v]	6.11	4,7,12
ADR421B	J	-	8	-	-	-	-	-	-	•	●	•	-	-	•	-	-	-	0.04	1	3	5-15	390	10	-	•	-	-	-	1.8	-	35	-	-	-	7.76	4,9
LT1027B	Z	8	8	-	-	-	-	-	-	-	-	-	-	-	●	-	-	-	0.02	1	2	8-40	2200	15	10	•	•	-	-	3	2[e]	6	-	-	-	6.17	4,7
LTC6655B-2.5	B	-	-	-	-	S	-	•	-	•	●	•	•	•	•	-	-	j	0.025	1	2	3-13	5000	5	5	-	-	-	•	0.6	1.7[e]	25	0.1	2.7	100	8.16	7,9,16
ADR4525B	B	-	8	-	-	-	-	-	-	•	●	•	•	•	•	-	-	-	0.02	-	2	3-15	700	10	10	-	-	-	-	1.25[g]	-	10	0.1	1[u]	100	5.85	9,17
MAX6325C	Z	8	8	-	-	-	-	-	-	-	●	-	-	•	•	-	-	-	0.04	0.5	1	8-36	1800	15	15	•	•	-	-	1.5	1.3[e]	5	-	-	100	13.40	4,7,18

Notes: (a) sorted approximately by tolerance, tempco and 0.1-10Hz noise; generally listing best accuracy grade. (b) other packages: M - TO-99 metal can; P - DPAK power pkg; S - small (micro8, MSOP); T - tiny (DFN, LCC). (c) B: bandgap; F: floating gate; J: JFET pinchoff; Z: zener. (d) tabulated data corresponds to the voltage choice indicated by a large bullet. (e) 10Hz-1kHz. (f) in LCC pkg. (g) for 2.5V version. (h) qty 3k. (j) adjustable via external resistors. (k) qty 1k. (m) min or max. (n) nominal. (o) 15nV/√Hz with C_{NR}=1uF. (p) peak-to-peak. (q) over V_{in} range. (r) 10Hz to 100kHz. (s) 100μV with 10nF filter cap. (t) typical. (u) 0.1μF for V_{out}≥3V. (v) up to 10uF with recommended pole-zero network. (w) a minimum of 0.1uF or C_{out}, whichever is larger. (x) ESR must fall in min-max range, see datasheet. (y) no min or max for all except the 2.5V version, which may oscillate with 400pF<C_{out}<2μF when sinking between 1mA and 6mA. (z) a "-" means no minimum or maximum; in all cases consult datasheet for details.

Comments: **1:** inexpensive LDO reg/ref. **2:** inexpensive LDO reg/ref with dropout flag; can add ext BJT. **3:** lowest V_{ref}; op-amp FB input for V_{out} from 0.4V to rail. **4:** other suffixes for relaxed tempco and accy. **5:** -A grade for 0.2% accy; temp output (in 8-pin pkgs). **6:** temperature output. **7:** load regulation 20ppm/mA or better. **8:** no suffix for relaxed tempco and accy. **9:** low noise, low tempco. **10:** a favorite. **11:** pin selectable V_{out}; ext resistors for variable V_{out}. **12:** ISL21007 for V_{in}=2.7-5.5V and I_Q=75μA. **13:** 10mV no-load dropout. **14:** can use in shunt mode. **15:** low noise, wide supply. **16:** very low noise, low tempco. **17:** low noise, low tempco. **18:** lowest noise and tempco.

Well, you say, I could do that with the circuit of Figure 9.90, which lets me set the dc output voltage via the ratio of R_1/R_2. Sure. But hold on – standard metal-film resistors come in 1% precision, with tempcos in the range of ± 50 ppm/°C. You *can* get fixed resistors and arrays with tempcos down in the ± 1 ppm/°C territory (see §5.6), but you'll pay a dear price, and the resistance selection is sparse. And don't forget that you've still got to trim the gain to reach the precise round-number output voltage. Use a trimpot? Not a good idea, because the tempco will suffer, and you will have to worry about resistance stability (wiper resistance, mechanical stability, etc.). You're likely to conclude that an on-chip factory-trimmed resistive divider (matched tempcos, thus very low tempco of gain) is the way to go. And it is.

9.10.2 Bandgap (V_{BE}) reference

This method exploits the ~0.6 V base–emitter voltage drop of a transistor operating at constant collector current (it should properly be called a V_{BE} reference), as given by the Ebers–Moll equation. Because that voltage has a negative temperature coefficient, the technique 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.

Figure 9.91 shows how it works. We start with a current mirror with two transistors operating at different emitter current densities (typically a ratio of 10:1). Using the Ebers–Moll equation, it is easy to show that I_{Q2} has a positive temperature coefficient, because 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 Figure 2.62). You may wonder where we get the constant current to program the mirror. 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} (here Q_3's V_{BE}). 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.[131] It turns out that zero temperature coefficient occurs when the total voltage equals the silicon bandgap voltage (extrapolated to absolute zero), about 1.22 V. The circuit in the box is the reference. Its own output is used (via R_3) to create the constant mirror programming current we initially assumed.

The classic bandgap reference requires three transistors, two for ΔV_{BE} and the third to add a V_{BE}. However, Widlar and Dobkin cleverly created a two-transistor version, first used in the LM317, see Figure 9.13.

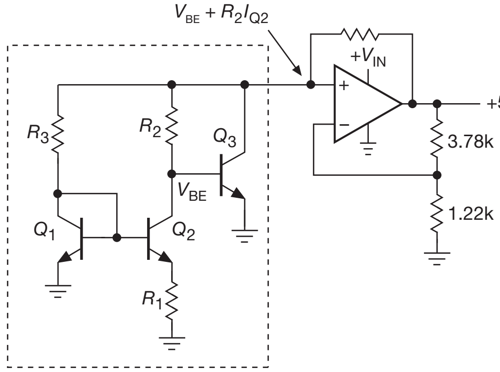


Figure 9.91. Classic V_{BE} bandgap voltage reference. The transistor pair Q_1Q_2 is a ratio current mirror, typically $I_{Q1} = 10I_{Q2}$; that ratio puts 60 mV across R_1, which sets the PTAT current I_{Q2}.

IC bandgap references

An example of an IC bandgap reference is the inexpensive (about \$0.50) 2-terminal LM385-1.2, with a nominal operating voltage of 1.235 V, $\pm 1\%$ (the companion LM385-2.5 uses internal circuitry to generate 2.50V), usable down to 10 μA. That's much less than you normally run zeners at, making these references excellent for micropower equipment.[132] The low reference voltage (1.235 V) is often more convenient than the approximately 5 V minimum usable voltage for zeners (you can get zeners rated at voltages as low as 1.8 V, but they are pretty awful, with very soft knees). The best grade of LM385 guarantees 30 ppm/°C maximum tempco and has a typical dynamic impedance of 1 Ω at 100 μA. Compare this with the equivalent figures for a 1N4370 2.4 V zener diode: tempco = 800 ppm/°C (typ), dynamic impedance ≈3000 Ω at 100 μA, at which the "zener voltage" (specified as 2.4 V at 20 mA) is down to 1.1 V! 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

[131] The complete expression for V_{ref} is therefore $V_{ref}=V_{BE3} + (V_{BE1}-V_{BE2})R_2/R_1$.

[132] But note that low-current references tend to be noisy: the LM385-2.5 (20 μA min) running at 100 μA has a noise voltage density of 800 nV/$\sqrt{\text{Hz}}$, compared with 120 nV/$\sqrt{\text{Hz}}$ for the analogous LT1009 or LM336-2.5 references (400 μA min) running at 1 mA. And, if you're willing to squander more operating current, the LTC6655 low-noise 3-terminal bandgap reference (with a quiescent current of 5 mA) has an output noise density of just 50 nV/$\sqrt{\text{Hz}}$, and, impressively, with a $1/f$ noise corner below 10 Hz.

2-terminal LT1634A (2.5 V or 5 V, 10 ppm/°C max, about \$6), or the 3-terminal AD586 (5 V, 2 ppm/°C max, about \$9).

One other interesting bandgap-based voltage reference is the extremely popular TL431. It is an inexpensive (less than \$0.10 in large quantities) 2-terminal shunt regulator–reference, but with a third terminal to set the voltage. You hook it up as shown in Figure 9.92. The "zener" turns on when the control voltage reaches 2.50 V; the device draws only a few microamps from the control terminal, and gives a typical tempco of output voltage of 10 ppm/°C. The circuit values shown give a zener voltage of 10.0 V, for example. This versatile device comes in TO-92, mini-DIP, and a half-dozen surface-mount packages, and can handle currents to 100 mA and voltages to 36 V. Its low-voltage and low-power (80 μA min) cousins, the TLV431 and TLVH431, work the same way, but with a 1.25 V internal bandgap reference and limited output voltage and current.[133] Both types come in accuracy grades of ±2%, ±1%, and ±0.5%.

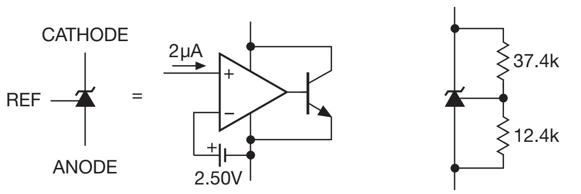


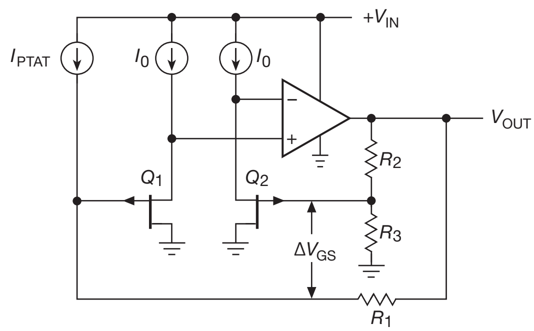
Figure 9.92. TL431 adjustable shunt regulator–reference. The resistive divider in the application circuit on the right sets the "zener" voltage to 10.0 V.

Bandgap temperature sensors

The predictable V_{BE} variation with temperature can be exploited to make a temperature-measuring IC. In Figure 9.91, for example, the difference in V_{BE}'s of $(kT/q)\log_e r$ implies that the current through Q_2 (and also Q_1) is proportional to absolute temperature (PTAT). The circuit can be rearranged (the Brokaw circuit) to produce simultaneously both an output voltage proportional to temperature, and a (fixed) bandgap voltage reference of 1.25 V. This is the case with a number of bandgap references, for example the AD680, a 2.50 V reference with an additional TEMP pin whose output voltage is 2.0 mV/K (thus 596 mV at 25°C). If you want only a temperature sensor, and don't need the bandgap reference, you can get nice stand-alone temperature sensors, for example the LM35, a 3-terminal sensor with 10 mV/°C output (0 V at 0°C), or the LM61, whose output is offset by +600 mV so it can measure from −30°C to +100°C. The LM61 costs a half dollar, compared with \$3 for the 8-pin multifunction AD680.

9.10.3 JFET pinch-off (V_P) reference

This recent technique is analogous to the V_{BE}-based bandgap reference, but uses instead the gate-source voltages of a pair of JFETs. A single JFET operating at fixed drain current has a wicked tempco of V_{GS}, but this can be circumvented by cleverly using a JFET pair. Figure 9.93 shows the configuration used in the ADR400-series of "XFET" voltage references from Analog Devices. The JFET pair Q_1Q_2 have identical geometry and run at equal drain currents; but their different channel doping produces a gate-voltage *difference* of ∼0.5 V that is quite stable, with a relatively small tempco of −120 ppm/°C. That's much smaller than the tempco of a V_{BE} drop (roughly −3000 ppm/°C), and requires only a small dose of positive tempco correction, here applied by the voltage drop across R_1.



$$V_{OUT} = (\Delta V_{GS} - R_1 I_{PTAT})\left(1 + \frac{R_3}{R_2}\right)$$

Figure 9.93. JFET voltage reference. The asymmetrically doped JFET pair, running at the same drain current, generates a difference voltage ΔV_{GS} between the gates. The relatively small tempco is compensated by a current derived from a bandgap-type reference (not shown).

The result is a voltage reference with excellent tempco (e.g., 3 ppm/°C or 10 ppm/°C for the two grades in the ADR400-series). An important benefit of this technique is its unusually low noise (1.2 μVpp for the 2.5 V part[134]).

[133] 6 V and 15 mA for the TLV431; 18 V and 80 mA for the TLVH431. The adjustable LM385 works similarly, but with an operating current range of 10 μA–20 mA, and voltages to 5.3 V.

[134] Better specified as 0.5 ppm(pp), because the noise voltage scales linearly with output voltage.

Bandgap references cannot match this sort of noise performance, because the process of compensating for their large intrinsic temperature coefficient introduces the lion's share of their output noise.

9.10.4 Floating-gate reference

This most recent entry into the voltage reference sweepstakes is, well, bizarre. If you were challenged to come up with an idea that would most likely fail, you might invent the "floating-gate array" (FGA) reference. Intersil did that, but they made it succeed! The idea is to put some charge onto the buried and well-insulated gate of a MOSFET, during manufacturing, which puts it at some voltage (thinking of it as a capacitor); the MOSFET then acts as a voltage follower (or op-amp input) to create a stable output voltage.

The stability over time depends, of course, on the tiny capacitor not losing or gaining any charge. That's a tall order – you'd like it to remain stable to perhaps 100 ppm over several years, over the full operating temperature range. A gate capacitance of 100 pF charged to 1 V, for example, would require that gate leakage be no more than 10^{-22} A; that's about two electrons per hour!

Somehow the folks at Intersil have made it work. They also dealt with stability over temperature, with several tricks: one method uses capacitors of different construction to cancel the already-small tempco of approximately 20 ppm/°C; another method uses capacitors of one type only, cancelling the small residual tempco by adding a voltage of known tempco (as in the bandgap and JFET references).

The results are impressive: the ISL21009 series claim long-term stabilities of order 10 ppm per square-root-kilohour; tempcos of 3 ppm/°C, 5 ppm/°C, and 10 ppm/°C (max) for the three grades;[135] noise of 4.5 μVpp; and very low supply current of 0.1 mA (typ). They come in preset voltages of 1.250 V, 2.500 V, 4.096 V, and 5.000 V, each available in several grades of accuracy and tempco.

9.10.5 Three-terminal precision references

As we remarked earlier, these clever techniques make possible voltage references of remarkable temperature stability (down to 1 ppm/°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 40 ppm/°C. There are two methods used to make references of the highest stability.

[135] But see the paragraph and footnote about ionizing radiation on page 684.

A. Temperature-stabilized references

A good way of 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. In this way the circuit can deliver equivalent performance with a greatly relaxed temperature coefficient, because 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 10 min or more). These problems are greatly reduced if the thermal stabilization is done at the chip level by integrating a heater circuit (with sensor) onto the IC itself. This approach was pioneered in the 1960s by Fairchild with the μA726 and μA727 temperature-stabilized differential pair and preamp, respectively.

This technique is used in the LM399 and LTZ1000 references, which claim stabilized tempcos below 1 ppm/°C (max). Users should be aware that the subsequent op-amp circuitry, including gain-setting resistors, may degrade performance considerably unless extreme care is used in design. In particular, low-drift precision op-amps and matched-tempco resistor arrays are essential. These aspects of precision circuit design are discussed in Chapter 5.

B. Precision unheated references

Clever chip design has made possible unheated references of nearly comparable stability. For example, the MAX6325 series from Maxim have tempcos of 1 ppm/°C (max), with no heater power or warm-up delays. Furthermore, they exhibit low noise and long-term drift. Their chief drawback is the difficulty in getting them! These high-stability references (LTZ1000, LM399, and MAX6325) all use buried zeners.

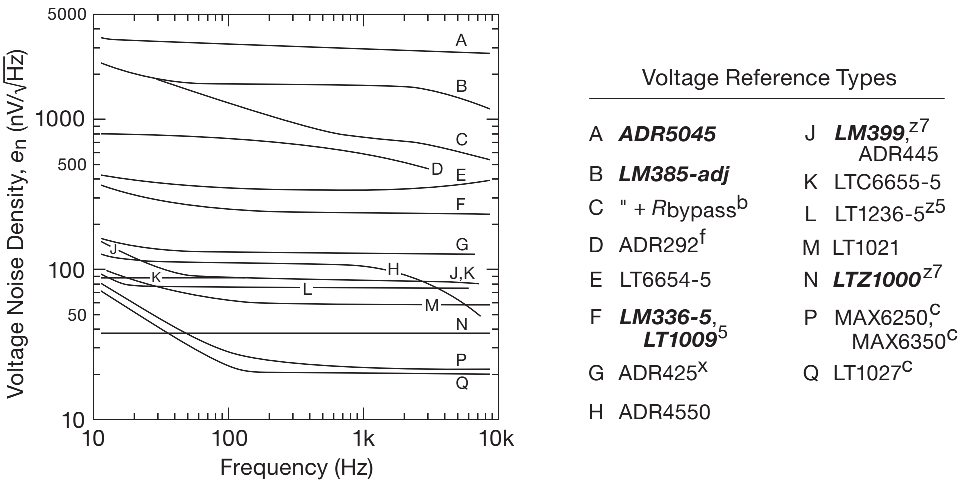


Figure 9.94. Noise density (e_n) versus frequency for a selection of voltage references. All are for 5 V output, except as noted; **boldface** part numbers are shunt type (2-terminal), the rest are series type (3-terminal). The LTZ1000 is operated at 4 mA. *Notes*: (5) 2.5 V reference, curve shown is 2× datasheet's e_n plot; (b) upper resistor bypassed; (c) with 1 μF noise reduction cap; (f) 4.096 V part; (x) "XFET" reference; (z5) buried zener, 5 V buffered output; (z7) 7 V buried zener.

9.10.6 Voltage reference noise

We mentioned briefly the business of *noise*, in connection with low-power references on page 679. You can always add filtering to suppress power supply or reference noise at higher frequencies (see the discussion of the *capacitance multiplier* in §8.15.1), but there's no substitute for a quiet reference at low frequencies, where the noise properties of the reference set a lower limit on output noise. The listings in Table 9.7 on page 677 and 9.8 on page 678 include datasheet values for integrated low-frequency noise (0.1–10 Hz, in units of μVpp), as well as rms noise voltage at somewhat higher frequencies (usually 10 Hz–10 kHz). In Figure 9.94 we've plotted noise-density curves (e_n, in units of nV/$\sqrt{\text{Hz}}$) for those references whose datasheets are considerate enough to provide such information. It's often useful to normalize the specified noise voltage values by the reference's voltage, to get a fair comparison among competing devices.

You can add lowpass filtering to reduce the noise from a voltage reference. Some references bring out an internal node on a "filter" pin (or "bypass," or "noise-reduction" pin) that you can bypass to ground; Table 9.3 indicates those in the "filter pin" column. Often the datasheets for such parts include numeric or graphical information to guide you in the choice of filter capacitor.[136]

Another technique you can use is the addition of an external lowpass filter, with an op-amp follower. Figure 9.95 shows this simple scheme, with an interesting twist: the basic lowpass filter is R_2C_2, with a time constant of 2.2 seconds (3 dB rolloff at 0.07 Hz). But why on earth is it perched atop C_1?! That's done in order to eliminate C_2's leakage current (which would produce an accuracy-damaging voltage drop across R_2) by bootstrapping the low side of C_2 – cute! The inclusion of R_1C_1 affects the rolloff and settling waveform, putting the 3 dB rolloff at 0.24 Hz and producing an 8% overshoot with a lengthened settling time (30 seconds) to 0.1%.[137] For this application you need a pretty good op-amp: input bias current low enough to prevent error from $\Delta V = I_B R_2$, and noise voltage low enough to add only insignificantly to the filtered reference's output. The OP-97E or LT1012A are similar and do the job well (or you could use a smaller R and larger C, permitting larger op-amp input current).

Generally speaking, a reference that runs at very low current will exhibit more noise, a trend evident in

[136] One of the more interesting pieces of such guidance appears in the LTC1844 datasheet, which cautions "Additionally, some ceramic capacitors have a piezoelectric response. A piezoelectric device generates voltage across its terminals due to mechanical stress, similar to the way a piezoelectric accelerometer or microphone works. For a ceramic capacitor the stress can be induced by vibrations in the system or thermal transients. The resulting voltages produced can cause appreciable amounts of noise, especially when a ceramic capacitor is used for noise bypassing." See the discussion in §*1x.3*.

[137] That is, with $R_1 = R_2$ and $C_1 = C_2$ the 3 dB rolloff frequency becomes $f_{-3\text{dB}} \approx 3.3/2\pi RC$, and the settling time to 0.1% (if you care) is approximately $\tau \approx 14RC$. Put a diode across R_2 to shorten the turn-on time.

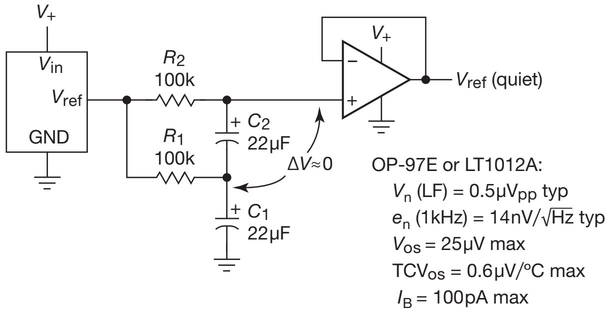


Figure 9.95. External lowpass filter with dc bootstrap quiets any voltage reference, while suppressing error from capacitor leakage currents. Use a quiet op-amp follower with low input current.

Table 9.8. This is easy to understand in the case of a bandgap (V_{BE}) reference, because BJT voltage noise decreases as the square root of collector current (see §8.3). You might conclude from this that a given shunt (2-terminal) IC reference would be quieter when biased at higher currents; but you would be wrong – a shunt reference runs its internal bandgap circuit at a current close to the part's "minimum operating current" (with corresponding noise voltage), and so running the reference at a higher current doesn't help.

9.10.7 Voltage references: additional Comments

As should be evident from Tables 9.7 and 9.8, there are lots of things to think about when selecting a voltage reference. Here are some bits of advice, to help the bewildered circuit designer (that's *you*).

Accuracy and drift There's *initial* accuracy, of course, often with a choice of grades designed by a suffix (-A, -B, etc.), with prices to match. But parts *age*, and a well-specified part will include a "long-term drift" figure (usually parts-per-million per thousand hours, or, perhaps more properly,[138] per $\sqrt{\text{kHr}}$), and sometimes a "thermal hysteresis" specification (the voltage offset after thermal cycling over the part's operating temperature range). Taking the LTC6655B (best grade) as an example, the initial accuracy is ±0.025%, the tempco is 1 ppm/°C (typ) and 2 ppm/°C (max), the long-term drift is 60 ppm/$\sqrt{\text{kHr}}$ (typ), and the thermal hysteresis is 35 ppm (typ) for thermal cycling between −40°C and +85°C. From these figures it's clear that initial accuracy is only a part of the story.

A caution about "temperature coefficient": most often we've used the description in terms of *slope*, i.e., ppm/°C (or μV/°C, etc.), giving perhaps a typical and a maximum (worst-case) value. But you'll sometimes see the description in terms of maximum deviation over the part's temperature range; an example is the LM385, where the specified worst-case *average* tempco of 150 ppm/°C is described in a footnote: "The average temperature coefficient is defined as the maximum deviation of reference voltage at all measured temperatures from T_{MIN} to T_{MAX}, divided by $T_{MAX} - T_{MIN}$." A maximum tempco so defined is guaranteed to be smaller than the maximum value of "slope" tempco (i.e., the maximum value of $\Delta V/\Delta T$ over the same temperature range), as you can convince yourself by drawing some wiggly curves. Figure 9.96 shows an example, adapted from the datasheet for the ADR4520–50 series of 3-terminal precision references.

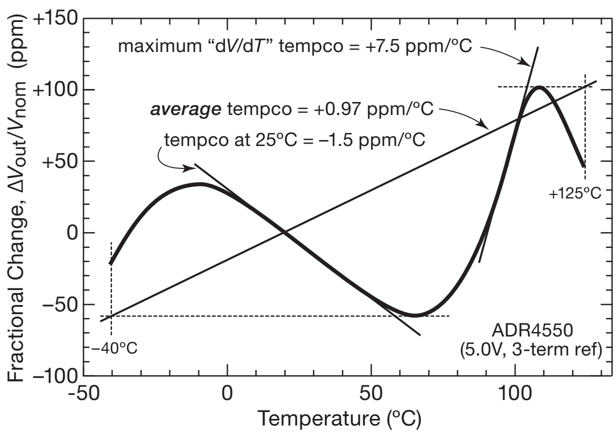


Figure 9.96. Three ways of defining the temperature coefficient of a voltage reference, illustrated with the ADR4550's serpentine datasheet curve. The datasheet's specified tempco is stated as 2 ppm/°C (max) over the full temperature range.

Some references include a "trim" terminal, which sounds like a great idea. But, as with op-amp offset trim configurations, it often provides *too much* trim range! You can try to remedy this by reconfiguring the trim network

[138] From the LTC6655 datasheet: "Long-term stability typically has a logarithmic characteristic and therefore, changes after 1000 hours tend to be much smaller than before that time. Total drift in the second thousand hours is normally less than one-third that of the first thousand hours with a continuing trend toward reduced drift with time. Long-term stability is also affected by differential stresses between the IC and the board material created during board assembly." Another bit of advice from LTC: "Significant improvement in long-term drift can be realized by preconditioning the IC with a 100–200 hour, 125°C burn in." A typical long-term drift spec is 60 ppm/$\sqrt{\text{kHr}}$. The datasheet for the REF5025 is instructive: it shows lower drift for the MSOP-8 package than an SO-8, 50 versus 90 ppm/$\sqrt{\text{kHr}}$, and it further shows 50 ppm/$\sqrt{\text{kHr}}$ for the first 1000 hours and 5 ppm/$\sqrt{\text{kHr}}$ for 1000 to 2000 hours. Better improvement than one gets from aging wine!

to supply less current. But a caution: some parts require that the external trim circuitry exhibit a specified temperature coefficient. Perhaps you're better off, as with op-amps, simply by choosing a reference with a tighter spec; a reference of better initial precision usually provides better tempco as well.

Self-heating Voltage references are happiest when they are only lightly loaded. If a reference IC is used to power a load, on-chip heating produces thermal gradients that can seriously degrade the part's accuracy and drift. For such applications it's best to buffer the output with an op-amp. Most good op-amps have lower noise and offset voltages than the voltage reference itself (you can do the calculation!), so it does not degrade the reference voltage. Quite the opposite, in fact, considering the degrading effect of substantial load current with an unbuffered reference – that's the whole point. And even middle-of-the-road op-amps have far lower tempcos of offset voltage than most voltage references (but use a precision op-amp for a precision reference, as we did in Figure 9.90).

An op-amp buffer also provides an ideal opportunity to add an *RC* noise filter; see §9.10.6, with its unusual filter configuration (Figure 9.95).

External influences As the footnote on the previous page suggests, you can seriously degrade a precision reference's accuracy by physically stressing the package; the stability is also compromised by the gradual infiltration of humidity through the plastic package. Sometimes you'll see improved specs for hermetically sealed package versions: the LT1236LS8 is packaged in hermetic LCC, and offers an improved drift specification relative to the plastic LT1236 version. And the most stable references are offered exclusively in hermetic metal packages to circumvent these problems: for example the LM399 (thermally stabilized buried zener reference) comes only in a TO-46 metal can; it specifies an excellent long-term drift specification of 8 ppm/$\sqrt{\text{kHr}}$ (typ). Pretty good – but handily outdone by the LTZ1000 "Ultra Precision" shunt reference (also in a metal-hermetic package), with a spectacular 0.3 ppm/$\sqrt{\text{kHr}}$ (typ)!

A recent contribution to the gallery of badnesses is the exposure of floating-gate references (§9.10.4) to ionizing radiation, most seriously in the form of airport luggage inspection x-ray machines or PCB post-assembly x-ray inspection. According to Intersil's Application Note,[139] by actual experiment at US airports the voltage shift for nine samples of a 5.0 V floating-gate reference (ISL21009) after six passes through carry-on x-ray machines averaged 25 ppm (negative); this is small compared with the initial accuracy of ±100 ppm, but in the same ballpark as the specified long-term (1000 hour) drift and hysteresis specs (both 50 ppm, typ).

Line and load regulation You've got to worry about regulation against input-voltage variations ("line regulation") for a voltage reference that is powered from unregulated dc, for example in a battery-powered application. For such use, a reference isn't suitable that boasts a worst-case tempco of 3 ppm/°C, but whose output varies 200 ppm per volt of input change (these are the specs of an entry in Table 9.8), though that reference would be fine if powered from a reasonably regulated dc rail. For this reason we've listed the worst-case line regulation for the references in Table 9.8 – these vary over nearly a 1000:1 range!

Load regulation matters also if you are using the reference as a voltage regulator, i.e., to power a load that draws a few milliamps, perhaps with load-current variation. But we discourage such use of a precision reference, because it produces on-chip heating and drifts; for that reason (and lack of space) we've not listed load regulation specifications in the tables.

9.11 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.[140]

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 Artesyn, Astec, Astrodyne, Acopian, Ault, Condor, CUI, Elpac, Globtek, Lambda, Omron, Panasonic, Phihong, Power-One, V-Infinity, and literally hundreds more. They offer

139 Intersil App Note #1533, 23Feb2010, *X-Ray Effects on Intersil FGA References*, which explains "The floating gate capacitor is susceptible to radiation degradation from various particles and photons in excessive doses, as the electrons generated in the silicon dioxide are collected in the storage cell. Normal radiation from cosmic rays or radon which exist in small amounts on earth will not cause the FGA reference voltage to drift appreciably for over 100 years. Artificial sources of radiation such as X-ray machines are capable of high enough doses to cause output voltage shift. Note that Flash memory devices are also susceptible to X-ray radiation degradation, although to a lesser degree as they are not precision analog devices."

140 "Dude, this is a league game."

Figure 9.97. 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.)

both switching and linear supplies, and they come in several basic packages (Figure 9.97).

- "Board-mount" supplies: these are relatively small packages, no more than a few inches on a side, with stiff wire leads on the bottom so you can mount them directly on a circuit board. Both ac–dc supplies and dc–dc converters come in this style, and they may be of "potted" or open construction. You can get linear or switchers, and they come with single or multiple output voltages. A typical PC-mounting open-case ac–dc triple-output switching supply provides +5 V at 2 A and ±12 V at 0.2 A and costs about $30 in small quantities. Linear board-mount supplies fall in the 1 W to 10 W range, switchers in the 15 W to 50 W range. In the dc–dc category (which are always switching converters) you can get isolated or non-isolated converters.[141] These are commonly used to generate additional needed voltages (e.g., ±15 V from +5 V), as we've seen in this chapter. But another important use is point-of-load (POL) conversion, for example to create +1.0 V at 75 A right at the chip's pins, to power a high-performance microprocessor. POL converters come

[141] Common package styles include the "full-brick," "half-brick," and "quarter-brick" configurations (4.6″×2.2″, 2.3″×2.2″, and 1.45″×2.3″, respectively), originated by Vicor; these span the range of 50–500 W, and may include an aluminum baseplate for heatsinking.

in regulated and unregulated versions, the latter with a fixed reduction ratio from a regulated dc input.[142]

- "Chassis-mount" supplies: these are larger power supplies, intended to be fastened to the inside of a larger instrument. They come in both "open frame" and "enclosed-case" styles; the former have all the components on display, whereas the latter (for example, the "ATX" power supplies you find in a desktop or server computer) are wrapped in a perforated metal box. They are available in an enormous variety of voltages, both single and multiple outputs. Chassis-mount linear supplies fall in the 10–200 W range, switchers in the 20–1500 W range.
- "External adapters": these are the familiar black "wall-wart" and desktop ("desk-wart"?) that come with small consumer electronic gadgets, and which are widely available from dozens of manufacturers. They actually come in three varieties, namely (a) step-down ac transformer only, (b) unregulated dc supply, and (c) complete regulated dc supply; the latter can be either linear or switcher. Some of the switching units allow a full 95 to 252 Vac input range, useful for traveling instruments.
- "DIN-rail mount" supplies: a popular way of mounting some kinds of industrial electronics (relays, circuit breakers, surge protectors, connectors, terminal blocks, and the like) is the European-originated DIN rail, which consists of a length of formed metal rail, 35 mm in width. Rail mounting makes it easy to assemble electrical equipment in industrial settings, and you can get a variety of switching supplies in this style.

9.12 Energy storage: batteries and capacitors

No chapter dealing with regulators and power conversion would be complete without a discussion of portable power. That usually means batteries (replaceable or rechargeable), sometimes assisted by energy-storage capacitors. Contemporary life is awash in portable electronic devices, which have driven the development and availability of improved batteries and capacitors. In this section we provide an introduction to battery choices and properties, and to the use of capacitors for energy storage. Because this chapter is already, well, *huge*, we defer further discussion of the care and feeding of batteries to Chapter *9x* in the advanced volume.[143]

As we remarked in the previous edition, the (now out-of-print) Duracell "Comprehensive Battery Guide" listed 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 (lithium-ion, nickel metal-hydride, and sealed lead-acid gel-type), by comparison, are designed to be recharged, typically from 200 to 1000 times. You usually make your choice among battery types 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 11.2 V mercury battery? They're better forgotten...)

[142] Although *unregulated* POLs might be expected to deliver poor "regulation," in fact they can surprise you: for example, Vicor's VTM series of high-efficiency fixed-ratio "current multipliers" (i.e., voltage step-down) include a unit rated at 130 A and 40:1 stepdown (thus 1.0 Vdc output from 40 Vdc input) with a worst-case output impedance of 0.00094 Ω at 100°C (thus < 0.1 V output change for a 100 A current step). Not bad. And you can always add feedback to the regulated supply that delivers the +40 Vdc input, to further stiffen the output for load change variations within the loop bandwidth. Moreover, these converters operate at 1.2 MHz switching frequency, so the residual ripple is at 2.4 MHz, conveniently bypassed with relatively small filter capacitors. In fact, the datasheet shows pretty good output-voltage waveforms with 0–130 A load steps, exhibiting essentially no overshoot even when the output is not filtered or bypassed with any external capacitor at all.

[143] See also the expansive discussion in Chapter 14 of the previous edition of this book.

9.12.1 Battery characteristics

If you want a primary (non-rechargeable) battery, your choices are essentially alkaline ("Zn/MnO_2") or one of the lithium chemistries ("Li/MnO_2," "Li/FeS_2," or "$Li/SOCl_2$"). Lithium batteries have a higher single-cell terminal voltage (~3 V), higher energy density, flatter discharge curves (i.e., constancy of voltage as their life ebbs, see Figure 9.98), better performance at low temperatures (where alkaline batteries just fade away) and higher price. By contrast, the alkaline types (the basic grocery-store battery) are cheap and plentiful, and you can buy them inexpensively at "big-box" stores (if you don't mind getting them in packages of several dozen); they're fine for undemanding applications.

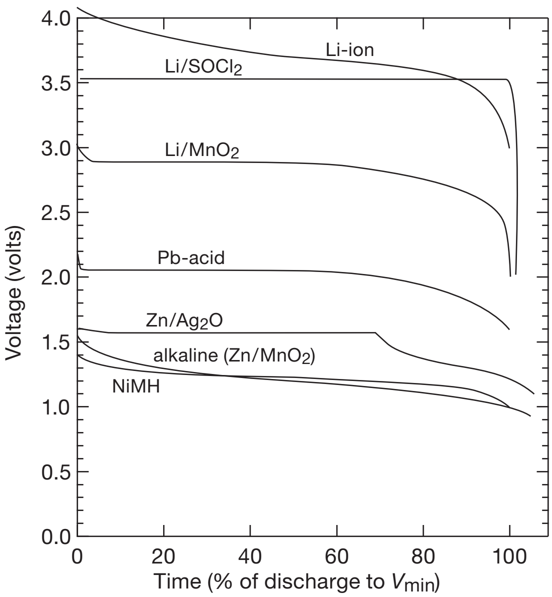


Figure 9.98. Battery discharge curves, as shown on the respective datasheets. In each case 100% discharge corresponds to the voltages as listed in the notes to Table 9.9.

Your choices for a secondary (rechargeable) battery are lithium-ion ("Li-ion"), nickel metal-hydride ("NiMH"), or lead-acid ("Pb-acid"). Li-ion batteries are lightweight, and provide the highest energy density and charge retention, but there are safety issues with lithium chemistries, and these aren't the kind of thing you buy off-the-shelf; they are the darlings of the manufacturers of smartphones, tablets, and laptop computers. Nickel metal-hydride batteries are the more common "consumer" rechargeables, and they come in standard form factors (AA, 9V sizes); early formulations had discouraging memory effects and self-discharge rates (~30%/month!), but recent versions ("LSD" – low self-discharge) are greatly improved. Lead-acid batteries are the heavy lifters, with their very low internal resistance; they are dominant in uninterruptible power supplies (UPSs) and other power-hungry devices (like boats and automobiles!); they don't come in tiny packages, but you can get them in sizes as small as D-cells.

Recharging secondary batteries can be a complicated business, particularly for fussy chemistries like Li-ion. Taking first the example of the humble lead-acid battery, a good charging method is the so-called two-step technique: 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 V 2.5 Ah lead-acid battery, typical values are I_{max}=0.5 A, V_{OC}=14.8 V, I_{OCT}=0.05 A, and V_F=14.0 V.

Although this all sounds rather complicated, it results in rapid recharge of the battery without damage. TI makes some nice ICs, for example the UC3906 and BQ24450, that have just about everything you need to do the job. They include internal voltage references that track the temperature characteristics of lead-acid cells, and they require only an external *pnp* pass transistor and four parameter-setting resistors.

Charging Li-ion batteries requires a bit more care, but once again the semiconductor industry has responded with easy-to-use single-chip solutions. Figure 9.99 shows an example of the kind of thing that is commonly seen. Here the power from a USB port (+5 V nominal, able to supply 100 mA or 500 mA) is used to charge a single-cell 4.2 V Li-ion battery; the latter's output (which is down to ~3.5 V when mostly discharged) is reduced to a stable +3.3 V logic-supply level with a linear LDO regulator. In this circuit the charger IC (U_1) takes care of the charging current and voltage profiles, and it includes safety features to sense over-voltage, short-circuit, and both chip and battery over-temperature (the latter uses the optional thermistor, which is found in many battery packs, or can be added externally). The cell temperature is also used to adjust the charging current or voltage when the temperature is outside the normal 10–45°C range, according to what's called JEITA standards. The ISET2 pin sets the input current limit as indicated; the USB protocol allows 100 mA drain initially, which can be increased to 500 mA by a negotiation through the USB data pins D− and D+ (this requires a microcontroller or other smart chip, not shown here). The LEDs indicate status (charging, input power good). The

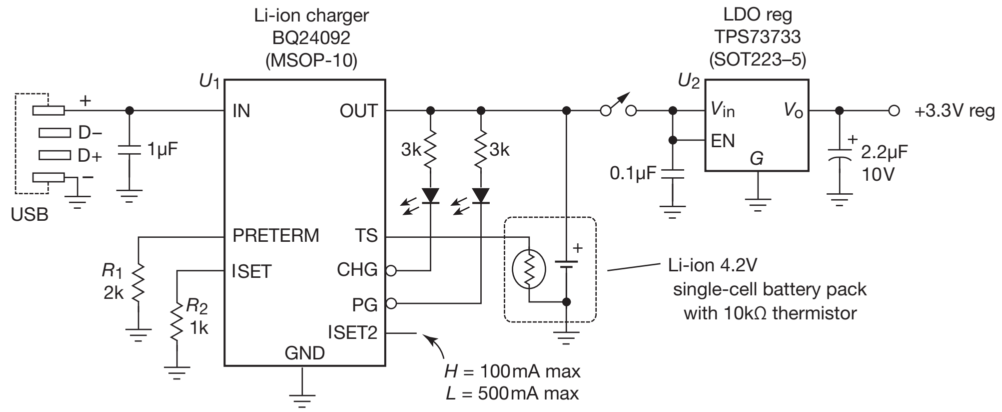


Figure 9.99. The +5 V provided by a USB port is ideal for charging a single-cell Li-ion battery; LDO regulator U_2 converts the battery's 3.5–4.2 V output to stable +3.3 V. See text for discussion of what to do with the USB's D+ and D− pins.

care and feeding of Li-ion batteries is discussed in more detail in §*9x.2*.

9.12.2 Choosing a battery

Table 9.9 lists characteristics of most of the batteries you might consider, and Figure 9.100 shows an assortment of common battery types. Here is a capsule summary of the most distinguishing characteristics of available batteries intended for use in electronic devices.

Primary Batteries (non-rechargeable)

Alkaline (Zn/MnO_2) Inexpensive; widely available (1.5V/cell AA, C, & D, and 9V pkgs); excellent shelf life; good low-temperature performance; sloping discharge.

Lithium (Li/MnO_2) High energy-density; good high-drain performance; 3V AA, C, & D, and 9V pkgs; excellent shelf life; excellent low-temperature performance; flat discharge.

Lithium (Li/FeS_2) Extraordinary shelf life (90% after 15 yrs); excellent low-temperature performance; flat discharge.

Lithium ($LiSOCl_2$) Extraordinary low-temperature performance (to −55°C); excellent shelf life; very flat discharge (but varies with I_{load}).

Silver (Zn/Ag_2O) Button cells; very flat discharge.

Zinc-air (ZnO_2) High energy-density (it breathes); flat discharge; short life after seal is removed.

Secondary Batteries (rechargeable)

Lithium-ion (Li-ion) High energy-density; popular; 3.6V/cell; flat discharge; very low self-discharge; safety issues.

Nickel metal-hydride (NiMH) Inexpensive & popular; standard packages (AA, 9V); 1.2V/cell; flat discharge; recent formulations have low self-discharge.

Lead-acid (Pb-acid) High current (low R_{int}); 2V/cell; flat discharge; moderate self-discharge.

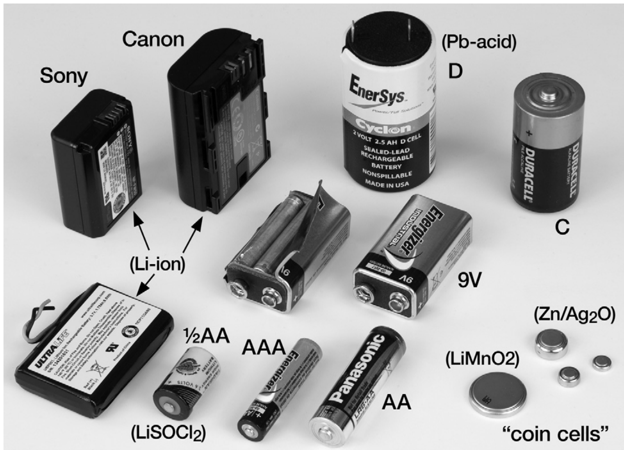


Figure 9.100. A collection of battery specimens. The Li-ion and Pb-acid types are rechargeable ("secondary"); the rest are non-rechargeable ("primary"). Those with unlabeled chemistry are alkaline. Some can opener activity revealed the 9 V innards – six itty-bitty alkaline cells.

9.12.3 Energy storage in capacitors

Batteries store energy *chemically*, either with reversible reactions (rechargeable batteries) or irreversible reactions (non-rechargeable). But batteries aren't the only way to store electrical energy: a charged capacitor stores $CV^2/2$ joules in its electric field, and a current-carrying inductor

Table 9.9 Battery Choices[a]

Chemistry	Part #	V_{nom} (V)	Discharge Capacity (mAh)	at mA	(mAh)	at mA	Size (mm)	Weight (gm)	Comments
Primary (non-rechargeable)									
9V "1604"									
carbon-zinc	122	9	320[b]	5	150[b]	25	17.5 x 12.9 x 46	37	
alkaline MnO_2	MN1604	9	550[b]	10	320[b]	100	17.5 x 12.9 x 46	45	popular
lithium MnO_2	DL1604	9	1200[b]	20	850[b]	100	17.5 x 12.9 x 46	34	
zinc-air	146X	8.4	1300[b]	10	-	-	17.5 x 12.9 x 46	34	pull tab & it breathes!
cylindrical									
D alkaline	MN1300	1.5	11500[f]	250	3700[f]	1000	34D x 61L	139	D size
D LiMnO2	U10013	3	11100[c]	250	10400[c]	2000	34D x 61L	115	D size
C alkaline	MN1400	1.5	5100[f]	250	1300[f]	1000	26D x 50L	69	C size
AA alkaline	MN1500	1.5	2800[f]	10	2400[f]	100	14.5D x 50.5L	24	AA size; popular
AA $LiFeS_2$	L91	1.5	3200[f]	25	3000[f]	500	14.5D x 50.5L	15	90% after 15y at 20°C
AA $LiSOCl_2$	ER14505	3.6	2100[c]	1	1600[c]	16	14.5D x 50.5L	18	very flat discharge
AAA alkaline	MN2400	1.5	1200[f]	10	1000[f]	100	10D x 44L	11	AAA size
CR2 $LiMnO_2$	CR2	3	850[c]	20	-	-	15.5D x 27L	11	popular
2/3A $LiMnO_2$	CR123A	3	1550[c]	20	-	-	17D x 34.2L	17	popular
2/3A Li poly	BR-2/3A	3	1200[c]	2.5	-	-	17D x 33.5L	13.5	
button									
silver	357	1.55	195[d]	0.2	-	-	11.5D x 4.8H	2.3	
zinc-air	675	1.45	600[e]	2	-	-	11.6D x 5.4H	1.9	4yrs unactivated
$LiMnO_2$	CR2032	3	225[c]	0.2	175[c]	2	20D x 3.2H	2.9	2032 size; popular
Li poly	BR2032	3	190[c]	0.2	90[c]	2	20D x 3.2H	2.5	2032 size
Secondary (rechargeable)									
cylindrical									
NiMH	HHR210AA/B	1.2	2000[f]	2000	-	-	14.5D x 50.5L	29	AA size; R_S=25mΩ; 80%/6mo
Li-ion	NCR18650	3.6	2900[h]	500	2500[h]	5000	18D x 65.2L	45	popular
Pb-acid	0810-0004	2	2500[g]	250	1900[g]	2000	34D x 61L	178	D size; R_S=5mΩ
button									
LiMn	ML2020	3	45[c]	0.12	40[c]	1	20D x 2.0H	2.2	memory backup
LiMnTi	MT621	1.5	2.5[f]	0.05	-	-	6.8D x 2.1H	0.25	memory backup
LiNb	NBL414	2	1[f]	0.004	-	-	4.8D x 1.5H	0.08	memory backup
LiV_2O_5	VL3032	3	100[c]	0.2	95[c]	1	30D x 3.2H	6.2	memory backup
rectangular									
Pb-acid	LC-R061	6	1200[k]	100	800[k]	1000	96 x 24 x 50	300	80% after 6mo; R_S=50mΩ
Pb-acid	LC-R127	12	7200[m]	500	5000[m]	5000	151 x 65 x 94	2470	80% after 6mo; R_S=40mΩ

Notes: (a) listed part numbers are representative (there are many manufacturers). (b) to 6V. [c] to 2V. (d) to 1.2V. (e) to 1.1V. (f) to 1.0V. (g) to 1.7V. (h) to 2.5V. (k) to 4.8V. (m) to 9.6V. (n) also Tadiran TL-2100.

stores $LI^2/2$ joules in its magnetic field. In quantitative terms these stored energies are dwarfed by those stored in batteries; but for some applications capacitors are just what you want. Among their other virtues, they have long lives, infinite endurance (charge/discharge cycles), the ability to be fully charged and discharged in seconds (or fractions of a second), and very high peak current capability (i.e., very low internal resistance, ESR). A storage capacitor, teamed with a conventional battery, can provide the best of both worlds: extraordinary peak power along with substantial energy storage. Furthermore, the energy density of recent "supercapacitors" is creeping up on the tail end of batteries. These points are seen nicely in a *Ragone plot* (Figure 9.101). To put numbers on it, we've collected data on some real-world representative capacitors and batteries; these are listed in Table 9.10. Capacitors excel in low ESR and high peak current (and therefore in high *power* density: W/gm or W/m^3), but batteries beat the pants off capacitors in *energy* density (Wh/gm or Wh/m^3).

Table 9.10 Energy Storage: Capacitor versus Battery[a]

Parameter	Conditions	Ultracap Maxwell K2 3000F, 2.5V	Aluminum Electrolytic Panasonic T-UP 180,000μF, 25V	Lead-acid gel Yuasa NP7-12 12V, 7Ah	Lithium-ion Saft VL34570 3.7V, 5.4Ah	Alkaline[p] Duracell MN1500 AA: 1.5V, 2Ah
generic[a]						
Wh/kg	1hr discharge	6	0.05	16	150	40
Wh/m^3	1hr discharge	7800	73	44000	360000	120000
W/kg	maximum	1000	1400	170	500	65
kW/m^3	maximum	1300	2000	500	1100	180
charge time	fast charge	30s	0.25s	1hr	1hr	-
charge cycles		10^6	(note b)	500	500	0
self discharge	25°C	7×10^5hr	0.2hr	10^4hr	10^5hr[e]	10^5hr
life	float, 25°C	10yr	25yr	5yr	?	10yr
specific to exemplar						
ESR	maximum	0.29mΩ	9mΩ	25mΩ	30mΩ[e]	120mΩ
I_{max}	cont	210A	17A	40A	11A	1A
P_{max}	cont	525W	425W	440W	35W	1.2W
weight		510g	300g	2650g	125g	24g
energy (Wh)	1hr discharge	3.0	0.015	45	18.9	1
volume (cm^3)	with terminals	390	206	950	53	8.4

(a) from mfgrs' datasheets for listed parts. (b) no wearout, limited by lifetime only. (e) estimated. (p) primary cell (non-rechargeable).

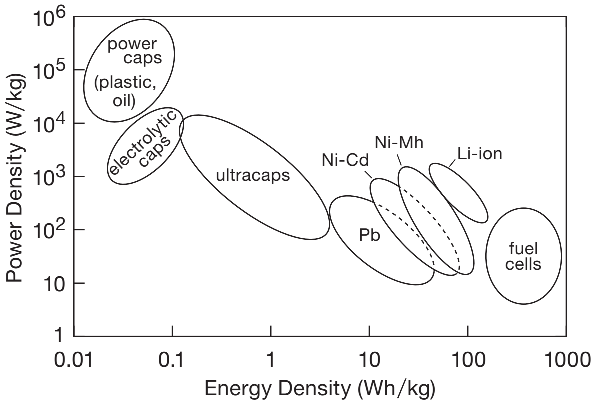


Figure 9.101. Energy-storage capacitors excel in delivering peak power, but batteries win out in energy storage, as seen in this "Ragone plot."

9.13 Additional topics in power regulation

9.13.1 Overvoltage crowbars

As we remarked in §9.1.1C, it is often a good idea to include some sort of overvoltage protection at the output of a regulated supply. Take, for instance, a +3.3 V high-current switching supply used to power a large digital system. Failure of a component in the regulation circuit (even something as simple as a resistor in the output voltage sensing divider) can cause the output voltage to soar, 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 low-voltage logic and VLSI, which operate from dc supply voltages as low as +1.0 V, and cannot tolerate an overvoltage of as much as 1 V without damage.[144] Another situation with considerable disaster potential arises when you operate something from a wide-range "bench" supply, where the unregulated input to the linear regulator may be 40 volts or more, regardless of the output voltage. We've encountered some aberrant bench supplies that soar to their full output voltage, briefly, when you switch them off. But "briefly" is all it takes to ruin your whole day!

A. Zener sensing

Figure 9.102 shows three classic crowbar circuits – (A) is simple and robust, but inflexible; (B) uses an IC trigger circuit that lets you set the trigger point more accurately; and

144 And sometimes much less: the datasheets for Xilinx's Virtex-5 and Virtex-6 FPGAs, for example, specify a core voltage of 1.0 V±5%, with an absolute maximum of 1.1 V! The Virtex-7 has the same 1.1 V limit and 1.0 V nominal core voltage, but it tightens the tolerance on the latter to ±3%; ouch!

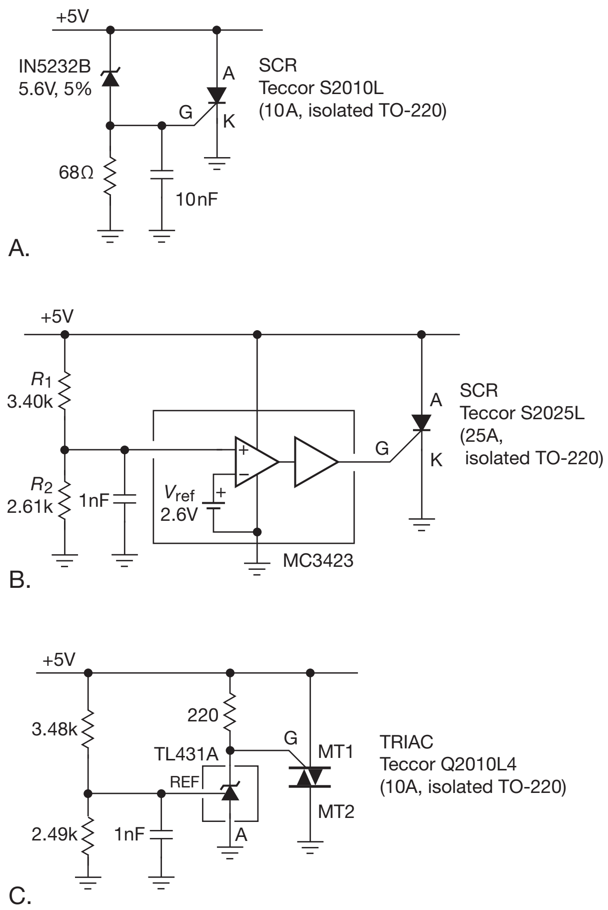


Figure 9.102. Overvoltage crowbars.

(C) uses a popular and precise 3-terminal "shunt regulator," with 1% setpoint accuracy.

In each case you hook the circuit between the regulated output terminal and ground; no additional dc supply is needed – the circuits are "powered" by the dc line they protect. For the simple circuit (Figure 9.102A) the SCR is turned on if the dc voltage exceeds the zener voltage plus a diode drop (about 6.2 V for the zener shown), and it remains in a conducting state until its anode current drops below a few milliamps. An inexpensive SCR like the S2010L can sink 10 amps continuously and withstand 100 amp surge currents; its voltage drop in the conducting state is typically 1.1 V at 10 A. The particular unit here is electrically isolated, so you can attach it directly to the metal chassis (SCRs usually connect their anode to the attachment tab, so normally you'd have to use an insulating spacer, etc.). The 68 Ω resistor is provided to generate a reasonable zener current (10 mA) at SCR turn-on, and the capacitor is added to prevent crowbar triggering on harmless short spikes.

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 5V supply for digital logic, whose typical 5% or 10% tolerance demands a crowbar voltage at least 5.5V. But that minimum is raised because of transient overshoot of the regulated supply: the voltage can jump when there's an abrupt load current change, 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 be set not less than about 6.0 V, but it cannot exceed 7.0 V 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 and 6.6 V, even using the relatively precise 5% zener indicated.

B. IC overvoltage sensing

The second circuit (Figure 9.102B) addresses these problems[145] by using a crowbar trigger IC, in this case the venerable MC3423, which has internal voltage reference (2.6 V±6%), comparators, and SCR drivers. Here we've set the external divider R_1R_2 to trigger at 6.0 V, and we've chosen a 25 A (continuous) SCR, also with an isolated mounting tab; it costs about a dollar. The MC3423 belongs to the family of so-called *power-supply supervisory* chips; the most sophisticated of these not only sense undervoltage and overvoltage, but can 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.

The third circuit (Figure 9.102C) dispenses with a supervisor IC, using instead the wildly popular[146] TL431 shunt regulator to trigger a triac (a bidirectional SCR) when the voltage presented to the reference input exceeds the internal reference voltage of 2.495 V±1%; that causes heavy

[145] And others, for example the desirability of fast gate overdrive when crowbarring large capacitive loads; see ON Semiconductor MC3423 datasheet.

[146] A quick check of DigiKey's excellent website shows approximately a half million pieces in stock, in 134 variants, from five manufacturers. They cost as little as $0.09 in large quantities.

conduction from cathode (K) to ground, triggering the triac in what's known as "third quadrant" operation.[147] This circuit can be flexibly extended to higher supply voltages (the TL431 operates to 37 V); and, with the low-voltage TLV431 variant (whose internal reference is 1.240 V) to very low trigger voltages.

The preceding circuits, like all crowbars, put an unrelenting 1 V "short circuit" across the supply when triggered by an overvoltage condition, and it can be reset only by turning off the supply. Because the SCR maintains a low voltage while conducting, there isn't much problem with the crowbar itself failing from overheating. As a result, these are reliable crowbar circuits. 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.

C. 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, because the zener will stop drawing current when the overvoltage condition disappears (unlike an SCR or triac, which have the memory of an elephant). However, a crowbar consisting of a simple power zener itself has its own problems. If the regulator fails, the crowbar has to contend with high power dissipation ($V_{\text{zener}}I_{\text{limit}}$) and may itself fail. We witnessed just such a failure in a commercial 15V 4A magnetic disk supply. When the pass transistor failed, the 16V 50W power zener found itself dissipating more than rated power, and it proceeded to fail too.

A better alternative, if you really want a power zener, is an "active zener" constructed from a small zener and a power transistor. Figure 9.103 shows two such circuits, in which a zener pulls the base or gate of a transistor into conduction, with a pull-down resistor to bring the zener current into the knee region at transistor turn-on. The TIP142 (in Figure 9.103A) is a popular Darlington bipolar power transistor, priced around $1, good for 75 W dissipation at 75°C case temperature, and with a minimum beta of 1000 at 5 A. For higher voltage and current, and where effective zener voltage accuracy is not critical, the MOSFET circuit (Figure 9.103B) is better: most MOSFETs do not have the second-breakdown-limited safe-operating areas of BJTs, and they are widely available in robust high-power versions. The circuit shown permits 130 W or 300 W dissipation at 75°C case temperature, for the IRF1407 or IRFP2907, respectively. These are "automotive" MOSFETs, rated at 75 V, and priced at $2.50 and $10, respectively. Note particularly the high peak current rating, limited only by transient thermal resistance (discussed in Chapter *9x*). One caution: a MOSFET clamp is prone to oscillation, particularly when implemented with a high-voltage (low-capacitance) part.

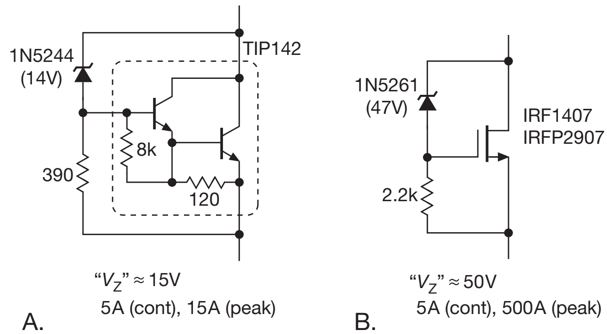


Figure 9.103. Active power zener.

D. Low-voltage clamp–crowbar

These techniques – zener crowbars, IC crowbars, and zener clamps – are generally inadequate for the low-voltage, high-current supplies used to power contemporary microprocessor systems; these may require +3.3 V (or lower) at 50–100 A: low-voltage zeners are imprecise and suffer from a soft knee, and crowbar trigger circuits like the MC3423 require too high a supply voltage (e.g., 4.5 V minimum for the 3423). Also, when an SCR triggers, it crowbars the supply until the power is cycled – not a good thing to do to a computer, particularly if the cause was a momentary (and harmless) transient.

We've wrestled with this problem, and, following the teachings of Billings,[148] have come up with a nice circuit for a low-voltage clamp–crowbar: it is adjustable, and it will operate down to 1.2 V. And (best of all) it operates in two steps – it *clamps* the transient, up to a peak current of 5–10 A; but if the transient persists, or rises above that current, it throws in the towel and fires a crowbar SCR that

[147] First and second quadrants have MT2 more positive than MT1, and trigger when the Gate is brought positive or negative with respect to MT1, respectively; third and fourth quadrants have MT2 more negative than MT1, and trigger when the Gate is brought negative or positive with respect to MT1, respectively. Quadrants two and four suffer from lower gate sensitivity.

[148] "Overvoltage clamping with SCR 'crowbar' backup," in K. Billings, *Switchmode Power Supply Handbook*, McGraw-Hill, 2nd ed. (1999).

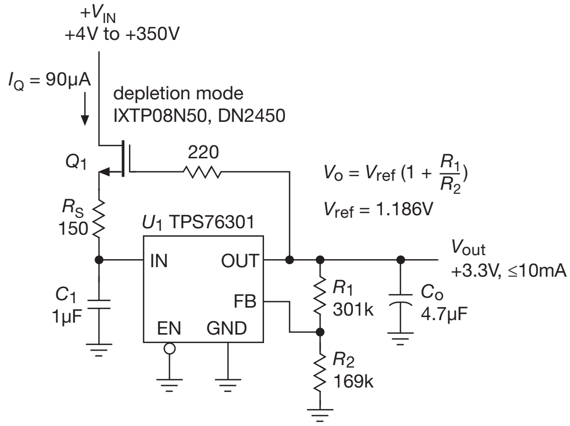
can handle 70 A continuous (1000 A peak). Because you may be dealing with a high-power system, it also has provision to switch off the ac input. *That* sure is a belt-and-suspenders solution! Because this is a bit off the beaten path, we've grouped this material with other advanced topics in Chapter *9x*.

9.13.2 Extending input-voltage range

As mentioned in §§9.3.12 and 3.6.2, linear regulators have a limited range of input voltage, typically +20 V to +30 V for BJT types, or as little as +5.5 V for CMOS types. Figure 9.104 (a completion of the block diagram in Figure 3.114) shows a nice way to extend the permissible range of V_{IN}, to as much as 1000 V. Q_1 is a high-voltage depletion-mode MOSFET (see Table 3.6), here configured as an input follower to hold U_1's V_{IN} a few volts above its regulated output. For the parts shown, V_{GS} is at least -1.5 V, a comfortable margin for any LDO;[149] and their rated maximum V_{DS} of 400 V and 500 V provides plenty of input-voltage flexibility (substitute an IXTP08N100 if you want to go to 1 kV).

A few details. (a) In this circuit we've used a small gate resistor to suppress the oscillation tendency of high voltage MOSFETs. (b) The source resistor R_S sets an output current limit of approximately V_{GS}/R_S, which is essential here because this regulator by itself is capable of output currents to 350 mA, which would cause more than 150 W dissipation in Q_1 for V_{IN}=500 V. Here we've chosen R_S for I_{lim}∼10 mA, thus 3.5 W maximum dissipation, handled easily with a modest heatsink attached to Q_1 in its TO-220 power package. (c) A power resistor could be added in Q_1's drain, to offload some of its power dissipation. (d) Low-dropout regulators specify (and require!) a minimum output capacitor C_{out} for stability (along with a specification of allowable range of its effective series resistance ESR); the value shown meets the specification for the TPS76301. (e) Some alternative choices for a fixed-voltage (+3.3 V; omit R_1 and R_2) low-power LDO are listed, with a few relevant parameters. All but the LM2936 are available also in adjustable versions, set with a resistive divider as we've done with the TPS76301 (but refer to their datasheets for V_{ref} and divider resistances).

149 The dropout voltage available from the depletion-mode FET can easily be increased (by a factor of 2× or 3× if necessary) by connecting the gate to a resistive divider between the LDO's output and the FET's source terminal. Note that this raises the minimum output load current.



Fixed 3.3V LDO choices for U_1

Part #	I_Q typ (µA)	V_{IN} max (V)	C_O min (µF)	V_{DO} max (mV)	@I_{load} (mA)	Pkgs
TPS76333	85	10	4.7	450	150	SOT-23
LP2950/1–33	75	30	2.2	600	100	TO-92, DIP, SOIC
LM2936–3.3	15	40	22	400	50	TO-92, SOT-23, SOIC
TPS71533	3.2	24	0.47	740	50	SC-70

Figure 9.104. Extending LDO input-voltage range. The listing includes some other low-power 3.3 V fixed-voltage LDO choices.

9.13.3 Foldback current limiting

In §9.1 we showed the basic current-limit circuit, which is often adequate to prevent damage to the regulator or load during a fault condition. However, 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 exceeds the maximum value of dissipation that would otherwise occur under normal load conditions. Look, for instance, at the regulator circuit of Figure 9.105, designed to deliver +15 V at currents up to 1 A. If it were equipped with simple current limiting, the pass transistor would dissipate up to 25 watts with the output shorted (+25 V input, current limit at 1 A), whereas the worst-case dissipation under normal load conditions is 10 watts (10 V drop at 1 A). And 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.

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

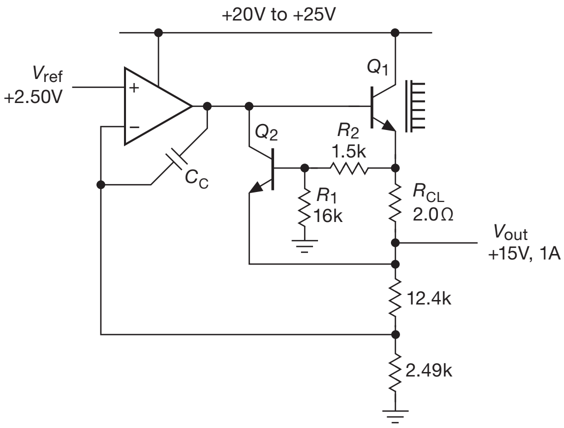


Figure 9.105. Linear regulator with foldback current limiting.

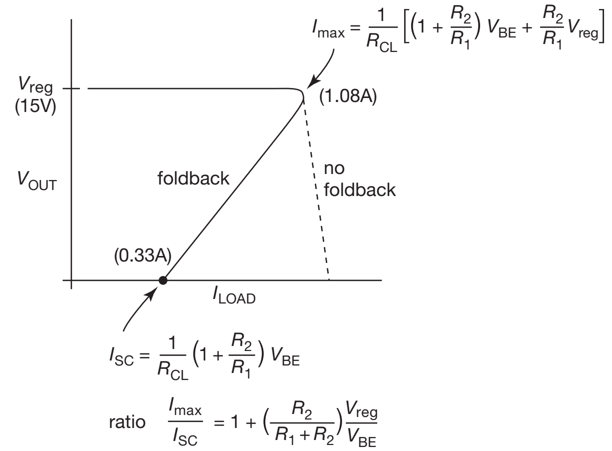


Figure 9.106. Foldback current limiting for the circuit of Figure 9.105.

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 heatsinks and transistors of higher power ratings (and safe-operating area; see §9.4.2) than necessary. Even so, it isn't a good idea to have large currents flowing into the powered circuit under fault conditions, because other components in the circuit may then be damaged. A better solution is to use *foldback* current limiting, a circuit technique that reduces the output current under short-circuit or overload conditions.

Look again at Figure 9.105. The divider at the base of the current-limiting transistor Q_2 provides the foldback. At +15 V output (the normal value) the circuit will limit at about 1 A, because Q_2's base is then at +15.55 V while its emitter is at +15 (V_{BE} is typically somewhat below the usual 0.6 V in the hot environment of power electronics). But the short-circuit current is less; with the output shorted to ground, the output current is about 0.3 A, holding Q_1's dissipation down to *less* (about 7.5 W) than in the full-load case (10 W). This is highly desirable, since excessive heatsinking 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; see Figure 9.106.

An important caution: use care in choosing the short-circuit current, because it is possible to be overzealous and design a supply that will not "start up" into certain loads. Figure 9.107 shows the situation with two common nonlinear loads: an incandescent lamp (whose resistance rises with voltage), and a linear regulator's input (which begins as an open circuit, then looks like its load resistance while operating below dropout, and finally forms a constant-current load above dropout). As a rough guide, when designing a foldback circuit, the short-circuit current limit should be set no less than one-third to one-half the maximum load current at full output voltage.

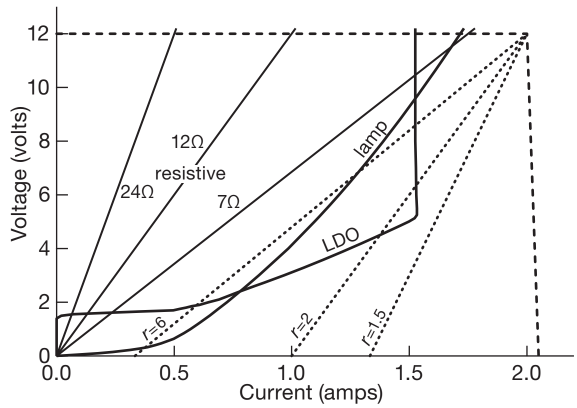


Figure 9.107. Excessive foldback may prevent startup into some loads. We measured VI curves for an automotive lamp (12 V, 21 W) and for a 5 V low-dropout regulator with 3.3 Ω load. The dashed line is normal 2 A current limiting, and the dotted lines show three values of foldback current limiting. A foldback current ratio of $r = I_{max}/I_{SC} = 6$ would fail to start up either load (it would get stuck at the lower intersection); for $r = 2$ the lamp is OK, but the LDO is not. A resistive load is never a problem.

9.13.4 Outboard pass transistor

Three-terminal linear regulators are available with 5 A or more of output current, for example the adjustable 10 A LM396. However, such high-current operation may be undesirable, because the maximum chip operating temperature for these regulators is lower than for power transistors, mandating oversized heatsinks. Also, they are expensive. An alternative solution is the use of external pass transistors, which can be added to integrated linear regulators like the 3-terminal fixed or adjustable regulators, whether of conventional or low-dropout configuration. Figure 9.108 shows the basic (but flawed!) circuit.

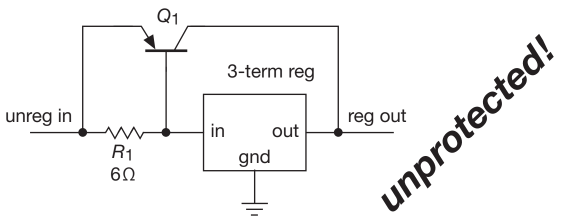


Figure 9.108. Basic three-terminal regulator with current-boosting outboard transistor. *Don't build this* – it has no current-limiting circuitry!

The circuit works normally for load currents less than 100 mA. For greater load currents, the drop across R_1 turns on Q_1, limiting the actual current through the 3-terminal regulator to about 100 mA. 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 100 mA! With this circuit the input voltage must exceed the output voltage by the dropout voltage of the regulator (e.g., 2 V for an LM317) 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 β 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 9.109 shows two methods of current limiting.

In both circuits, Q_1 is the high-current pass transistor, and its emitter-to-base resistor R_1 has been chosen to turn it on at about 100 mA load current. In the first circuit, Q_2 senses the load current via the drop across R_{SC}, cutting off Q_1's drive when the drop exceeds a diode drop. There are a couple of drawbacks to this circuit: for load currents near the current limit, the input voltage must now exceed the regulated output voltage by the dropout voltage of the 3-terminal regulator plus two diode drops. Also, the small resistor values required in Q_2's base makes it difficult to add foldback limiting.

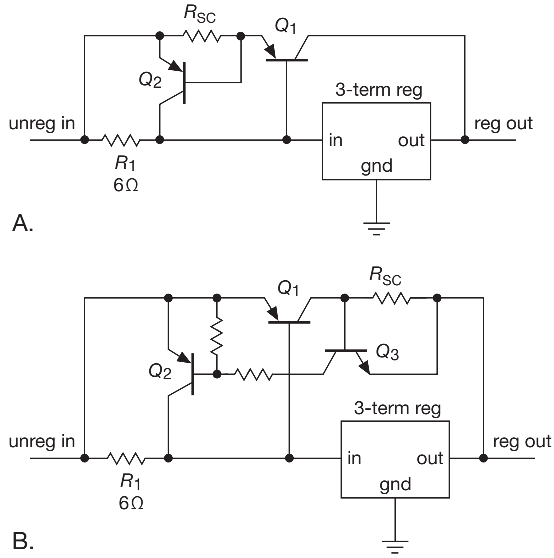


Figure 9.109. Outboard transistor booster with current limiting.

The second circuit helps solve these problems, at the expense of some additional complexity. With high-current linear 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_3's base to a divider from Q_1's collector to ground, rather than directly to Q_1's collector. Note that in either circuit Q_2 must be capable of handling the full current limit of the regulator.

A caution: With an external pass transistor you do not get the overtemperature protection that is included in nearly all integrated regulators. So you have to be careful to provide adequate heatsinking for both normal and short-circuit load conditions.

9.13.5 High-voltage regulators

Some special problems arise when you design linear regulators to deliver high voltages, and you often need to resort to some clever circuit trickery. This section presents a few such techniques.

A. Brute force: high-voltage components

Power transistors, both bipolar and MOSFET, are available with breakdown voltages to 1200 volts and higher, and they're not even very expensive. And IGBTs are available at even higher voltage ratings, up to 6000 volts.

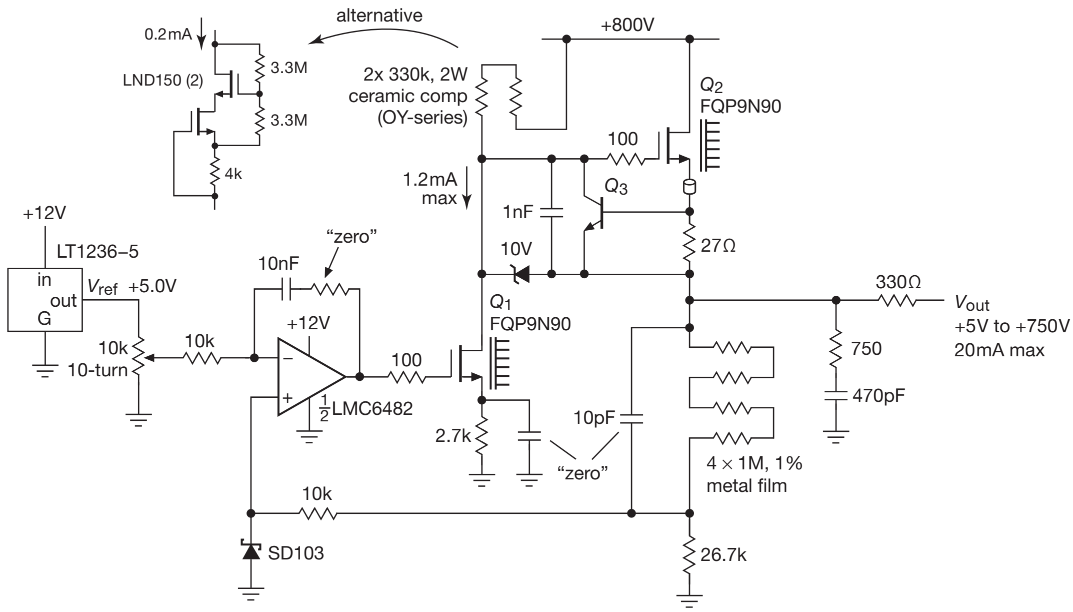


Figure 9.110. High-voltage regulated supply. The current-source pullup on Q_1's drain is a preferable alternative to the simpler resistive drain load; see §9.3.14C. See also Figure 3.111.

ON Semiconductor's MJE18004, for example, is a 5 A *npn* power transistor with conventional (V_{CEO}) collector-to-emitter breakdown of 450 V, and base back-biased breakdown (V_{CEX}) of 1000 volts; it costs less than a dollar in single quantities. And power MOSFETs are often excellent choices for high-voltage regulators owing to their excellent safe operating area (absence of thermally induced second breakdown); they are widely available with 800–1200 V ratings, and currents up to 8 A or more. For example, the FQP9N90 *n*-channel MOSFET (9 A, 900 V) from Fairchild costs about $1.75. See listings in Tables 3.4b and 3.5.

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 9.110 shows the idea, in this case a +5 V to +750 V regulated supply using an 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 a fraction of the output with a precision +5 V reference. Q_3 provides current limiting by shutting off drive to Q_2 when the drop across the 27 Ω resistor equals a V_{BE} drop. The remaining components serve more subtle, but necessary, functions: the zener 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); it also protects against forward gate breakdown, for example if the output is abruptly shorted. The Schottky diode likewise protects the op-amp's input from a negative current spike, coupled through the 10 pF capacitor.

Note the use of several resistors in series, to withstand the large voltages; the OY-series of 1 W and 2 W ceramic composition resistors from Ohmite are non-inductive, as are the precision metal-film resistors in the output sensing divider. 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). Likewise, the 330 Ω series output resistor promotes stability by decoupling capacitive loads (at the cost of degraded regulation). And Q_2's series gate resistor and source-lead ferrite bead suppress oscillations, to which high-voltage MOSFETs are particularly prone. *An important caution:* Power-supply circuits like this present a real electrical shock hazard – use care!

We can't resist an aside here: in slightly modified form (reference replaced by signal input) this circuit makes a very nice high-voltage amplifier, useful for driving crazy loads such as piezoelectric transducers; see Figure 3.111 for a simple 1 kV amplifier configured this way. For that particular application the circuit must be able both to sink and to source current into the capacitive load. Oddly enough, the circuit (called a "totem pole") acts like a "pseudo-push-pull" output, with Q_2 sourcing current and Q_1 sinking current (via the diode), as needed. See §*3x.8* for a detailed discussion about capacitive loading of MOSFET source-followers.

If a high-voltage regulator is designed to provide a fixed output only, you can use a pass transistor whose breakdown voltage is less than the output voltage. For example, you could modify this circuit to produce a fixed +500 V output, using a 400 V transistor for Q_2. But with such a circuit you must ensure that the voltage across the regulator never exceeds its ratings, even during turn-on, turn-off, and output short-circuit conditions. A few strategically placed zeners can do the job, but be sure to think about unusual fault conditions such as an abrupt upstream short-circuit (a spark, or probing fault), as well as "normal" events like an output short. It's remarkably easy to have a pretty-good (and tested) high-voltage circuit fail, abruptly (and usually with a "snap" sound), leaving precious few clues as to the cause. Learn from our hard-earned experience here: use a MOSFET rated beyond the full supply voltage.

Exercise 9.14. Add foldback current limiting to Figure 9.110.

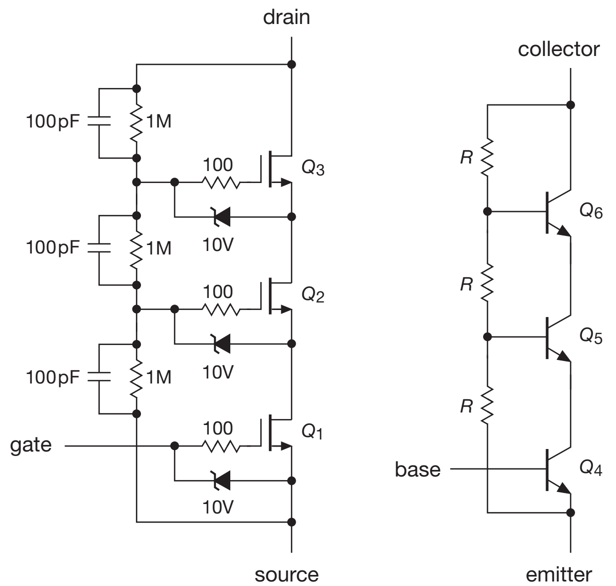


Figure 9.111. Connecting transistors in series to raise the breakdown voltage, to distribute the power dissipation, and (in power BJTs) to stay within the SOA.

B. Transistors in series

Figure 9.111 shows a trick for connecting transistors in series to increase the breakdown voltage. In the circuit on the left-hand side, the equal gate resistors distribute the dc voltage drops across the series-connected MOSFETs, and the paralleled capacitors ensure that the divider action extends to high frequencies. (The capacitors should be chosen large enough to swamp differences in transistor input capacitance, which otherwise cause unequal division, reducing overall breakdown voltage.) The zener diodes protect against gate breakdown.[150] And the 100 Ω series gate resistors help suppress oscillations, common in high-voltage MOSFETs (use some ferrite beads on the source and gate leads, if there's an oscillation sighting).

For series-connected bipolar transistors you can distribute the voltage drops with resistors alone, as shown, because the rugged base–emitter junctions are not susceptible to damage analogous to the MOSFET's oxide punch-through (in the forward direction they simply conduct a small current; and the small reverse currents from the base divider are generally benign, and they can be prevented entirely with 1N4148-style small-signal diodes connected between base and emitter). Small-signal transistors like the 300 V MPSA42 and MPSA92 (*npn* and *pnp*, respectively) and 400 V MPSA44 (available in TO-92 and surface-mount packages) are usefully extended to higher voltages this way.

Note that the series-connected string has considerably poorer saturation voltage than that of an equivalent high-voltage transistor: for three MOSFETs (as shown) the ON-voltage is $V_{\text{sat}} \approx 3V_{\text{DSon}} + 3V_{\text{GSon}}$; for the BJT circuit it's $3V_{\text{CEon}} + 3V_{\text{BE}}$.

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 1/4-watt type. A more subtle trap awaits the unwary, namely the maximum *voltage* rating of a resistor, regardless of its power-dissipation rating. For example, standard 1/4-watt axial-lead resistors are limited to 250 V, and often less for surface-mount types.[151] Another underappreciated effect is the astonishing voltage coefficients of carbon composition resistors, when run at

[150] Many designers use ordinary signal diodes, rather than zeners, assuming that there is no worry about *forward* gate breakdown, because the MOSFETs should turn on vigorously long before gate-channel breakdown. We're not so sure.

[151] Specifically, 200 V, 150 V, 75 V, 50 V, and 30 V for Vishay's CRCW thick-film resistors in sizes 1206, 0805, 0603, 0402, and 0201, respectively.

higher voltages. For example, in an actual measurement (Figure 9.112) a 1000:1 divider (10M, 10k) produced a division ratio of 775:1 (29% error!) when driven with 1 kV; note that the *power* was well within ratings. This nonohmic effect is particularly important in the output-voltage-sensing divider of high-voltage supplies and amplifiers – beware! Companies like Ohmite (Victoreen division) and Caddock make resistors in many styles designed for high-voltage applications like this. See §*1x.2* for additional measurements and discussion.

Quite apart from their use in high-voltage applications, another motivation for series connection of multiple transistors is to distribute a large power dissipation. For such power applications, where you're not dealing with high voltages, you can, of course, use a *parallel* connection. But then you have to ensure that the current is divided approximately equally among the multiple transistors. For BJTs in parallel this is usually done with individual emitter-ballasting resistors, as we saw in §2.4.4. But that scheme is problematic with MOSFETs, because they have a spread of gate-source voltages, forcing you to allow an uncomfortably large voltage drop across the source resistors. This can be addressed, though, with some cleverness; see Figure 3.117 for a nice solution.

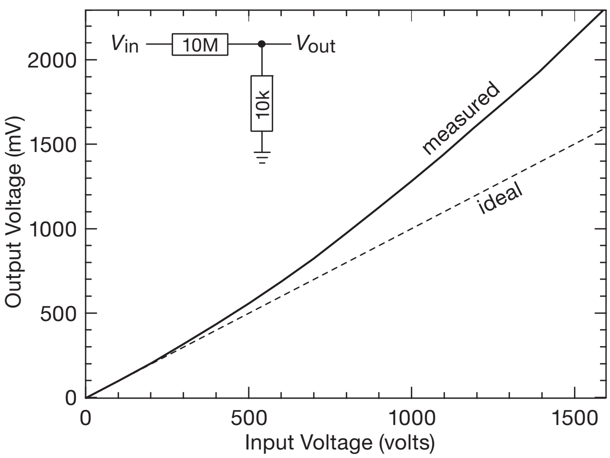


Figure 9.112. Carbon composition resistors exhibit a reduction in resistance as they approach their rated 250 V. Don't use resistors above their voltage rating!

C. Floating regulator

Another method sometimes used to extend the voltage range of integrated regulators, including the simple 3-terminal type, is to float the entire regulator above ground, for example as shown in Figure 9.113. Here the zener D_2 limits the drop across the 3-terminal regulator to just a few volts (the zener voltage minus Q_1's gate-source voltage), with outboard MOSFET Q_1 taking up the rest of the voltage drop. The LT3080 is a nice choice here, with its simple 10 μA programming current setting the output voltage. We used the trick of a pair of resistors (a "current divider," which you can think of as a current multiplier, viewed from the 10 μA source side) to raise the effective programming current to 1 mA, so we could use a 500k pot to set the voltage (rather than the awkward – and unobtainable – 50 megohms that otherwise would be needed); the boosted programming current serendipitously provides the LT3080's 0.5 mA minimum load current as well. Zener current is provided by the handy LND150 depletion-mode MOSFET from Supertex, here throttled down to 0.2 mA with a source self-biasing resistor.

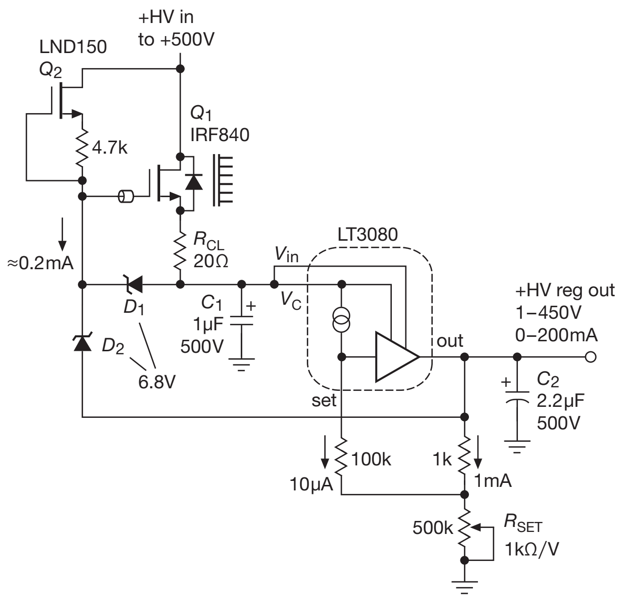


Figure 9.113. High-voltage floating three-terminal regulator.

The remaining components are easy to understand: D_1 protects Q_1's gate; the ferrite bead suppresses oscillations (you can use a 150 Ω gate resistor instead); and the LT3080 is outfitted with its minimum required input and output bypass capacitors. If there's an opportunity for the HV input to drop below the output voltage, add a diode across the LT3080 (heck, just do it anyway).

Exercise 9.15. Explore replacing Q_1 with a high-current depletion-mode MOSFET, like an IXTP3N50; see Table 3.6. Can you think of a way of using it to eliminate Q_2 and the zener diodes, despite the fact its $-V_{GS}$ may be less than the LT3080's required V_{DO}(max)=1.6 V at high currents? *Hint*: an LM385-2.5 may be useful.

Review of Chapter 9

An A-to-K summary of what we have learned in Chapter 9. This summary reviews basic principles and facts in Chapter 9, but it does not cover application circuit diagrams and practical engineering advice presented there.

¶A. Voltage Regulator Taxonomy.
Voltage regulators provide the stable dc voltages needed to power all manner of electronic circuits. The simplest (and least noisy) type is the *linear* regulator (Figure 9.2), in which the output error signal, suitably amplified and compensated, is used to control a linear "pass transistor" (BJT or MOSFET) that is in series with a higher (and perhaps unregulated) dc input voltage. Linear regulators are not power efficient, with dissipation $P_{\text{diss}}{=}I_{\text{out}}(V_{\text{in}}{-}V_{\text{out}})$, and they are not able to a produce dc output that is larger than the input, nor a dc output of reversed polarity.

The *switching* regulator (or switching *converter*, *switch-mode power supply*, SMPS, or just "switcher," §9.6) addresses these shortcomings, at the cost of some induced switching noise and greater complexity. Most switching power supplies use one or more inductors (or transformers), and one or more saturated switches (usually MOSFETs) operating at relatively high switching frequencies (50 kHz–5 MHz), to convert a dc input (which may be unregulated) to one or more stable and regulated output voltages; the latter can be lower or higher than the input voltage, or they can be of opposite polarity. The inductor(s) store and then transfer energy, in discrete switching cycles, from the input to the output, with the switch(es) controlling the conduction paths; with ideal components there would be no dissipation, and the conversion would be 100% efficient. The output error signal, suitably amplified and compensated, is used to vary either the pulse width ("PWM") or the pulse frequency ("PFM"). Switching converters can be *non-isolated* (i.e., input and output sharing a common ground, Figure 9.61), or *isolated* (e.g., when powered from the ac powerline, Figure 9.73); for each class there are dozens of topologies, see ¶D below.

A minor subclass of switching converter is the *inductorless converter* (or "charge-pump" converter; see §9.6.3), where a combination of several switches and one or more "flying" capacitors is used to create a dc output that can be a multiple of the dc input, or an output of opposite polarity (or a combination of both). For many of these, the output(s) track the dc input (i.e., unregulated), but there are also variants that regulate the output by controlling the switching cycle. See Table 9.4 and Figures 9.56 and 9.57.

¶B. The DC Input.
Regardless of the kind of converter or regulator circuit, you need to provide some form of dc input. This may be poorly regulated, as from a battery (portable equipment) or from rectified ac (line-powered equipment, Figures 9.25 and 9.48); or it may be an existing stable dc voltage already present within a circuit (e.g., Figure 9.64). For a line-powered instrument that uses a linear regulator, the "unregulated" dc input (with some ac ripple) consists of a transformer (for both galvanic isolation and voltage transformation) plus rectifier (for conversion to dc) plus bulk storage capacitor(s) (to smooth the ripple from the rectified ac). By contrast, in a line-powered (confusingly called "off-line") switcher the powerline transformer is omitted, because a transformer in the isolated switcher circuit provides galvanic isolation and it is far smaller and lighter since it operates at the much higher switching frequency; see Figure 9.48.

A diode bridge and storage capacitor converts an ac input to full-wave unregulated dc, whether from a powerline transformer or directly from the ac line. Ignoring winding resistance and inductances, the dc output voltage is approximately $V_{\text{dc}}{=}1.4V_{\text{rms}} - 2V_{\text{diode}}$, and the peak-to-peak ripple voltage is approximately[152] $\Delta V_{\text{ripple(pp)}}{\approx}I_{\text{load}}/2fC$, where C is the capacitance of the output dc storage capacitor and f is the ac input frequency (60 Hz or 50 Hz, depending on geographic and political boundaries). The ac input current is confined to relatively short pulses during the part of the waveform leading up to the positive and negative peaks (see Figures 9.51 and 9.78). This low "power factor" waveform is undesirable, because it produces excessive I^2R heating and more stressful peak currents. For this reason all but the smallest switching converters use a power-factor correction (PFC) input stage (Figure 9.77) to spread out the current waveform and thus create an input current approximately proportional to the instantaneous ac input voltage.

Line-powered instruments need a few additional components, both for safety and convenience. These include a fuse, switch, and optional line filter and transient suppressor; these are often combined in an IEC "power entry module," see Figure 9.49.

¶C. Linear Voltage Regulators.
The basic linear voltage regulator compares a sample of the dc output voltage with an internal voltage reference (see ¶G, below) in an *error amplifier* that provides negative

[152] From $I = C dV/dt$, assuming approximately constant discharge current I.

feedback to a *pass transistor*; see Figure 9.2. The dc output voltage can be greater or less than the reference (Figures 9.4 and 9.5), but it is always less than the dc input. You can think of this as a feedback power amplifier, which is prone to instability with capacitive loads, thus the compensation capacitor C_c in Figure 9.2D,E. The final circuit in that figure shows a current limiting circuit (R_{CL} and Q_2), and also an *overvoltage crowbar* (D_1 and Q_3, see ¶J below) to protect the load in the event of a regulator fault.

All the components of the original 723-type linear voltage regulator can be integrated onto a single IC (Figure 9.6), forming a "3-terminal" fixed regulator, e.g., the classic 78xx-style (where "xx" is its output voltage). These require only external bypass capacitors (Figure 9.8) to make a complete regulator. There are only a few standard voltages available, however (e.g., +3.3 V, +5 V, +15 V); so a popular variant is the 3-terminal *adjustable* regulator (e.g., the classic 317-type, see Figure 9.9), which lets you adjust the output voltage with an external resistive divider (Figure 9.10). Figures 9.14, 9.16, and 9.18 show some application hints for this very flexible regulator. Both fixed and adjustable 3-terminal regulators are also available in negative polarities (79xx and LM337, respectively), as well as in low-current variants (78Lxx and LM317L, respectively).

One drawback of these classic linear regulators is their need for an input voltage that is at least ~2 V greater than the output (its *dropout voltage*); that is needed because their pass transistor operates as an emitter follower (thus at least a V_{BE} drop) and the current-limit circuit can drop up to another V_{BE}. Two volts may not sound like a lot, but it looms large in a low-voltage regulator circuit, e.g., one with a +2.5 V dc output. To circumvent this problem, you can use a *low-dropout* (LDO) regulator, in which the pass transistor (BJT or MOSFET) is configured as a common-emitter (or common-source) amplifier, see Figure 9.20; the resulting dropout voltages are down in the tenths of a volt (Figure 9.24). LDOs are nice, but they cost more, and they are more prone to instability because their high output impedance (collector or drain) causes a lagging phase shift into the substantial load capacitance. LDOs may require significant minimum output bypass capacitance (as much as 10 μF or 47 μF), often constrained with both a minimum and maximum equivalent series resistance (ESR), e.g., 0.1 Ω min, 1 Ω max; see Table 9.3.

¶D. Switching Converter Topologies.

The basic *non-isolated* switcher topologies are the *buck* (or "step-down"), the *boost* (or "step-up"), and the *invert* (or "inverting buck–boost"); see §9.6.4 and Figure 9.61. The power train of these each uses one inductor, one switch, and one diode (or a second switch acting as an active rectifier), in addition to input and output storage capacitors. A complete converter requires additional components: an oscillator, comparator, error amplifier, drive circuits, and provisions for compensation and fault protection; see for example Figure 9.65. As with linear regulators, the semiconductor manufacturers have stepped in to provide most of the necessary components as packaged ICs, see Tables 9.5a,b and 9.6.

For the buck converter $V_{out} < V_{in}$, and for the boost converter $V_{out} > V_{in}$. The inverting converter produces an output of opposite polarity to the input, and whose voltage magnitude can be larger or smaller than the input voltage (this is true also of the remarkable Ćuk converter, §9.6.8H). The respective dc output voltages are $V_{out(buck)}=DV_{in}$, $V_{out(boost)}=V_{in}/(1-D)$, and $V_{out(invert)}=-V_{in}D/(1-D)$, where D is the switch-ON duty cycle $D=t_{on}/T$. There are also *non*-inverting buck-boost topologies that permit the output voltage range to bracket the input (i.e., able to go above or below the input). Examples are the 2-switch buck–boost (two switches, two diodes, one inductor), and the SEPIC (one switch, one diode, two inductors), see Figure 9.70. Of course, a switching converter with a transformer (whether isolated or not) provides flexibility in output polarity, as well as improved performance for large ratio voltage conversion.

Isolated switching converters use a transformer (for isolation), in addition to one or more inductors (for energy storage), see Figure 9.73. In the *flyback* converter (Figure 9.73A) the transformer acts also as the energy-storage inductor (thus no additional inductor), whereas in the *forward* converter and *bridge* converters (Figures 9.73B–D) the transformer is "just a transformer," and the diodes and inductor complete the energy storage and transfer. The respective dc output voltages are $V_{out(flyback)}=V_{in}[N_{sec}/N_{pri}][D/(1-D)]$ and $V_{out(forward)}=DV_{in}(N_{sec}/N_{pri})$. Speaking generally, flyback converters are used in low-power applications ($\lesssim$200 W), forward converters in medium-power applications (to ~500 W), and bridge converters for real power applications.

¶E. Switcher Regulation: Hysteretic, Voltage Mode, and Current Mode.

There are several ways to regulate a switching converter's dc output voltage. Simplest is *hysteretic* feedback, in which the error signal simply enables or disables successive switching cycles. It's a form of simple "bang–bang" control, with no stability issues that require a compensation network; see Figure 9.64 for a buck converter design with the popular MC34063. Proportional PWM control is better, and comes in two flavors: voltage mode and current mode. Both methods compare the output voltage with a fixed reference to regulate the output voltage, but they do it in different ways. In *voltage-mode* PWM, the output voltage error signal is compared with the internal oscillator's sawtooth waveform to control the primary switch's ON duration, whereas in *current-mode* PWM the comparison ramp is generated by the rising inductor current, with the internal oscillator used only to initiate each conduction cycle. See §9.6.9, and particularly Figures 9.71 and 9.72. In either case the controller terminates a conduction cycle if the switch exceeds a peak current, the input drops below an "undervoltage lockout" threshold, or the chip exceeds a maximum temperature. Figure 9.65 shows a simple voltage-mode PWM buck converter.

Voltage-mode and current-mode control loops both require compensation for stability, and each has its advantages and disadvantages. Current-mode controllers appear to be winning the popularity contest, owing to their better transient response, inherent switch protection (owing to pulse-by-pulse current limiting), improved outer-loop phase margin, and ability to be paralleled.

¶F. Switching Converter Miscellany.

Switching conversion is a rich subject, many details of which are well beyond the scope of this chapter (or this book). Some topics – ripple current and inductor design, core saturation and reset, magnetizing inductance and snubbing, soft start, diode recovery, CCM and DCM conduction modes, switching losses, loop compensation, burst mode, inrush current, isolation barriers, PFC, switching *amplifiers* – are treated lightly here and in Chapter *9x*. Consider this chapter's treatment of switching converters as a lengthy introduction to a specialty field that can easily consume a professional lifetime.

¶G. Voltage References.

A stable voltage reference is needed in any voltage regulator, as well as in accurate applications such as precision current sources, A/D and D/A conversion, and voltage- and current-measurement circuits. Often a good voltage reference is included in a regulator or converter IC (see for example Table 13.1), but you may want the improved performance you can get with a high-quality external reference. And, often, you need a stand-alone voltage reference for other uses in a circuit.

The simplest voltage reference is the discrete *zener diode* (§9.10.1), but most voltage references are multi-component integrated circuits that behave externally either like an extremely good zener ("2-terminal," or *shunt*; Table 9.7), or like an extremely good linear regulator ("3-terminal," or *series*; Table 9.8). Shunt references must be biased into conduction (just like a zener) by supplying current from a higher-voltage rail (use a resistor or a current source), while series references are powered by connecting their supply pin directly to the dc supply. References of either type are available in a small set of standard voltages, typically in the range of 1.25–10.0 V.

The discrete 2-terminal zener is fine for non-critical applications, but its typical accuracy of ±5% is inadequate for precision circuits. Integrated references of either kind are far better, with worst-case accuracies in the range of 0.02% to 1%, and tempcos ranging from 1 ppm/°C to 100 ppm/°C, as seen in the tables. Most integrated references are based on a circuit that temperature compensates the V_{BE} of a BJT (a so-called "bandgap reference"), generating a stable voltage of approximately 1.24 V; but others use a buried zener diode with $V_Z \approx 7$ V. The latter are generally quieter, but bandgap references can operate from low supply voltages and are widely available in voltages of 1.24 V, 2.50 V, etc. Two newer technologies with surprisingly good performance are the *JFET pinchoff reference* from ADI (the ADR400 "XFET" references), and the *floating-gate* reference from Intersil. Both exhibit very good tempco and low noise. Other important characteristics of voltage references are *regulation* (R_{out} for shunt types, PSRR for series types), minimum *load capacitance* and stability into capacitive loads, *trim* and *filter* pins, and *package* style.

¶H. Heat and Power Dissipation.

Along with power electronics comes... *heat!* You remove it with a combination of convection (air flow) and conduction (thermal contact with a heat-dissipating *heatsink*). Conductive heat flow is proportional to the temperature difference between the hot and cold sides (Newton's law of cooling), $\Delta T = P_{diss} R_\Theta$, where R_Θ is known as the "thermal resistance." For a succession of conductive joints the thermal resistances add; thus, for example, the junction temperature T_J of a power semiconductor dissipating P_{diss} watts is $T_J = T_A + P_{diss}(R_{\Theta JC} + R_{\Theta CS} + R_{\Theta SA})$, where T_A is the ambient temperature, and the successive R_Θ's represent the thermal resistances from junction to case, case to heatsink, and heatsink to ambient. Printed circuit foil patterns are often adequate for dissipation of a few watts or less (Figure 9.45); finned heatsinks or metallic chassis surfaces are used for greater heat removal, with forced airflow generally needed when the power dissipation reaches levels of 50 W or more (Figure 9.43). Semiconductor devices can withstand considerably greater power dissipation during short pulses; this is sometimes specified as a graph of *transient thermal resistance* (i.e., R_Θ versus pulse duration and duty cycle), or as elevated contours on a plot of Safe Operating Area (see ¶I).

¶I. Safe Operating Area.

A power transistor (whether BJT or MOSFET) has specified maximum values of voltage and current, and also (because of maximum allowed junction temperature) a maximum product $V_{DS}I_D$ (i.e., power dissipation) for a given case temperature; the latter is just $V_{DS}I_D \leq (T_{J(max)} - T_C)/R_{\Theta JC}$. These limits define a *safe operating area* (SOA, §9.4.2), usually shown as contours on log–log axes of current versus voltage; see for example Figure 3.95. That plot shows two further features: (a) greater dissipation is allowed for short pulses; (b) the SOA of BJTs (but not MOSFETs) is further constrained by a phenomenon known as "second breakdown."

¶J. Overvoltage Crowbars.

Some failure modes of power converters cause output overvoltage, for example a shorted pass transistor in a linear regulator, or loss of feedback control in a switcher. This is likely to damage or destroy powered circuitry. An *overvoltage crowbar* (§9.13.1) senses overvoltage and triggers an SCR to short the output. A less brute-force technique shuts down conversion when an overvoltage is sensed; these are indicated in the "OVP" column in Table 9.5b.

¶K. Current Sources.

"Regulator" usually means a stable *voltage* source; but there are many uses for a controllable *current* source (§9.3.14). Three-terminal linear regulators are easily coaxed into current-source service (§9.3.14A). There are also dedicated current-source ICs like the LM334 and REF200. JFETs make convenient 2-terminal current sources, and depletion-mode MOSFETs make excellent current sources that can operate up to voltages as high as 1 kV, see §9.3.14C. And don't forget about the humble discrete BJT current source (§§2.2.6 and 2.3.7B), or the op-amp current-source circuits (§4.2.5: Howland; op-amp + transistor).