# 3

# Practical Selection of Components

## INTRODUCTION

There are many aspects of selecting components that somehow end up being mentioned only in vendors' application notes and hard-to-find books. It's not enough to know that if you're dissipating 0.7W in a resistor, you should pick a resistor that can handle 1W. You also should know, for example, that a wirewound device can handle pulse power much larger than 1W, and thus maybe doesn't need to be a 1W resistor if the 700mW is actually present for only a short time. But how do you find out *how long* a 0.5W resistor can take the 700mW pulse?

More or less every component used in power supplies has idiosyncrasies like this. The usual way of finding such information is through (sometimes sorry) experience or by having the good luck of talking with someone who has learned the hard way. Rather than advocating reliance on luck, this chapter collects together many of these topics for the following electronic items: resistors, capacitors, schottky diodes, rectifier diodes, BJTs, MOSFETs, op amps, and comparators. Hopefully, those relatively new in the field will find lots of useful items here, and even those with much experience will find it handy, if nothing else, to have all the information collected in one place. However, the author has refrained from stating the obvious, having assumed that if you're reading this book in the first place, you don't require to be told that electrolytic capacitors have a polarity and should not be put in backward.

# RESISTORS

## Values

10.2kΩ. 39.9Ω. Where did all those weird numbers come from for standard values? It turns out that the values are (approximately) logarithmically distributed across a decade (such as 1kΩ to 10kΩ). Actually, there are different numbers of resistors per decade, depending on the tolerance: for example, either 48 or 96.

There is a practical maximum to how large a resistor can be used on a PCB. The problem with very large values (although they are available, for specialized applications) is that there is current leakage on the board between any two points that are at different potentials and are close enough together physically. In severe cases, leakage can be the equivalent of a 1–10MΩ resistor. Thus, if you were to try to put a 100MΩ resistor in a circuit, it could be in parallel with this leakage, and you would end up with only the 1 or 10MΩ instead of 100MΩ. For a circuit that gets around this problem in the common case of an op amp feedback, see the section below on op amps.

> **Practical Note** Avoid using resistors bigger than 1MΩ unless special precautions are taken.

## Types of Resistor

The oldest style of resistor is carbon composition (carbon comp), which nowadays you see only in hobby stores: these resistors are much bigger than metal film resistors of the same power rating, and are actually more expensive nowadays. Metal film resistors have the same frequency response as carbon comps, so be sure to tell your purchasing agent not to buy any carbons.

Then there are wirewounds. These range in size from tiny 1W packages to mammoth 1kW rheostats (see below for a definition). These resistors are called wirewound because they really are: if you cut one up, you will find a piece of (relatively) high resistance wire, usually wound around a core in a helix pattern. If you think this sounds like the description of a solenoid, that's because it is: wirewounds have plenty of inductance; as discussed below, it's also possible, by winding equal numbers of turns in opposite directions to produce a wirewound resistor having very little inductance. Table 3.1 lists the types of resistor discussed here and mentions some applications.

**TABLE 3.1** Brief Resistor Selection Guide

| Type | Suggested Applications |
|---|---|
| Carbon Comp | Not used anymore; use metal film instead. |
| Metal film | General purpose, use for most applications. |
| Wirewound (inductive) and rheostat | Use for load resistors. |
| Wirewound (noninductive) | Use for current sensing high frequencies, such as switching waveforms. |
| Shunt | Use for current sensing high currents. |
| PCB trace | Use for current sensing when cost is more important than accuracy. |

### Tolerance

It used to be that 5% resistors were the cheapest kind, and everyone used them for everything—not so today. Today, 1% resistors are both the most available and the cheapest—if you come across a purchasing agent who disagrees, suggest that a homework session is in order: there's no reason to ever use a tolerance greater than 1% on a resistor.

For that matter, during cost reduction efforts, people seem to come around to ask if you really need that 0.1% resistor, and couldn't you use a 1%? Send these people away. Unless you're making millions of this type of supply, the cost difference for a 0.1% versus a 1% is so small that almost anything else you change on the supply will have a bigger cost impact than changing that tolerance. Besides, you probably picked 0.1% to achieve the necessary output voltage tolerance or the like anyway.

### Selecting Ratios

So far you've found out how to miff a purchasing agent. Here's a way to please one. As we'll discuss in the chapter on monitoring circuitry, frequently you don't care about the actual values of the resistors you use, you just want a voltage divider (i.e., a ratio of resistances). Picking a standard value for one of the pair will considerably reduce the number of different values in a design, which doesn't affect performance but makes the supply cheaper by enabling larger quantities to be purchased. If there's no reason to choose any particular impedance, for example, I always use 10kΩ as one part of a divider.

### Maximum Voltage

Believe it or not, resistors have a maximum voltage rating. And it's not always just set by the power dissipation; resistors can actually arc. This problem is especially severe when surface mount resistors are used, because of the close spacing between the ends. So if you're dealing with voltages above, say, 100V in a supply, you might check that any resistors attached to high voltage nodes have the necessary rating.

### Temperature Coefficient

Most resistors have pretty small temperature coefficients (say, 50–250ppm/°C), although these can be important when you're trying to do worst-case analysis (Chapter 10). Wirewounds, however, can change quite a bit when they get hot, so read the specifications carefully.

### Power Rating

Everyone knows not to put a half-watt of power into a quarter-watt resistor. But what exactly *is* a quarter-watt resistor? The military decided that to increase resistor reliability, nobody would be allowed to put more than half the rated power into a (carbon or metal film) resistor. To meet this requirement, companies that design for the military apply their own derating; for example, that no one is allowed to put more than 70% of the *military* rated power into the resistor. To aid in this, some companies produce military-style resistors (e.g., RN55 or RN60) that are already derated the 50%; that is, what is really a half-watt resistor they call a quarter-watt resistor. Completely confused? The moral of the story is just that you need to keep a sanity check on resistor ratings—does it *look like* a quarter-watt resistor? Get a standard catalog, and check up on your buyers when they select alternates, to be sure the resistors you get will fit in the PCB holes.

While we're at it, what about putting 0.25W into a quarter-watt resistor? After all, the data book says the device can handle it! And so it can. However, resistors can get *extremely* hot—wirewounds are rated to operate at 275°C! Quite aside from not wanting to accidentally touch something at this temperature, resistors also give off unpleasant smells when they get too hot, and drift in resistance value very considerably.

**Practical Note** A good practical limit is to follow the lead of the military and, whenever possible use a resistor only to half its rated steady-state power.

***Pulse Power.*** While putting 1W into a 1W wirewound is not a good idea, this limitation is only for steady state (e.g., many seconds or more). For short times, a wirewound can take much more power than its rating without failing. This is *not true* for resistors of other types! You should strictly adhere to the maximum power rating of nonwirewound resistors, although for short times it's OK to put the full rated power into one; for example, you can safely put 100mW into a 100mW nonwirewound resistor for 100ms.

**EXAMPLE**

Suppose I have a 40V one-shot pulse across a 10Ω resistor for 100ms. This is a power of $P = (40\text{V})^2/10\Omega = 160\text{W}$. Do I really need a 200W resistor? Dale has provided a guide to selecting power resistors (the table from which is reproduced as Figure 3.1). To use this table, first we calculate the energy that goes into the resistor, $E = P \times t = 160\text{W} \times 100\text{ms} = 16\text{J}$, and then the energy per ohm, $E/R = 16\text{J}/10\Omega = 1.6\text{J}/\Omega$. We now use the table's first column to find an energy per ohm larger than this: the first one is $2.46\text{J}/\Omega$. Reading across, we find the resistor value larger than 10Ω, which is 10.11Ω; and reading up, we see that this can be put into a G-10 resistor, which is a 10W type, a considerable size savings!

*Note*: Dale says that this is valid only up to pulses about 100ms long, and for standard wirewounds; longer pulses should be based on "short time overload" ratings, while noninductive wirewounds can take *four times* the pulse ratings given by this table.

***Rheostats: A What?*** Rheostat is the proper name for a variable power resistor. In the lab you'll typically see rheostats that range from maybe 100W to 1kW. As with potentiometers, they have a center tap that shorts out part of the winding. Not to belabor the obvious, this means that if you are using half the resistance of a rheostat (e.g., 50Ω on a 100Ω rheostat), you can put in only half the power, too! (If it was a 300W rheostat, you wouldn't put in more than 150W in this condition.) This restriction most often gets people in trouble when they have a constant output voltage supply, and a rheostat as the variable load. Idly twirling the rheostat's knob in such a setup is not advisable. The best solution here is to put a fixed power resistor in series with the rheostat, so that even if it were set to 0Ω, it wouldn't dissipate too much power. The math is straightforward; don't be too lazy to do it right.

**Energy–Resistance Chart**

| Energy per ohm<br>[J/ohm or<br>(W·s)/ohm] | Resistance (ohms)<br>EGS-1<br>RS-$\frac{1}{4}$<br>G-1<br>per ohm | EGS-2<br>RS-$\frac{1}{2}$<br>G-2 | EGS-3<br>RS-1A<br>G-3 | RS-1B | RH-5<br>ESS-2B<br>RS-2B<br>G-5 | RS-2C<br>RS-2C | RH/PH-10<br>G-6<br>G-8 | RH-25<br>EGS-10<br>RS-5<br>RS-5-69<br>G-10 | RS-7<br>G-12 | RH-50<br>ESS-10<br>RS-10<br>RS-10-38<br>G-15 |
|---|---|---|---|---|---|---|---|---|---|---|
| $13.9 \times 10^{-6}$ | 3480 | 4920 | 10.4K | 15K | 24.5K | 32.3K | 47.1K | 90.90K | 154K | 265K |
| $20.3 \times 10^{-6}$ | 2589 | 3659 | 7580 | 11.4K | 18.69K | 24.19K | 31.79K | 69.40K | 1114.9K | 197K |
| $28.7 \times 10^{-6}$ | 1999 | 2829 | 5840 | 7960 | 14.19K | 18.29K | 26.99K | 51.70K | 8K | 152K |
| $39.5 \times 10^{-6}$ | 1549 | 2189 | 4630 | 6190 | 10.89K | 13.69K | 20.69K | 40.40K | 68.59K | 111K |
| $53.1 \times 10^{-6}$ | 1239 | 1749 | 3630 | 5280 | 8600 | 11.39K | 16.69K | 31.40K | 54.39K | 93.50K |
| $70.0 \times 10^{-6}$ | | 1414 | 2920 | 4280 | 6980 | 9250 | 13.59K | 25.90K | 44.19K | 75.50K |
| $90.6 \times 10^{-6}$ | 1000 | 1149 | 2740 | 3510 | 6550 | 7560 | 11.09K | 24.50K | 36.79K | 71.50K |
| $145 \times 10^{-6}$ | 670 | 947 | 1960 | 2870 | 4650 | 6260 | 8910 | 17.30K | 29.50K | 50.60K |
| $221 \times 10^{-6}$ | 492 | 684 | 1420 | 2060 | 3370 | 4560 | 6570 | 12.70K | 20.59K | 37.40K |
| $324 \times 10^{-6}$ | 355 | 502 | 1040 | 1510 | 2460 | 3270 | 4820 | 9220 | 15.69K | 26.90K |
| $460 \times 10^{-6}$ | 272 | 384 | 792 | 1160 | 1860 | 2480 | 3640 | 7000 | 11.89K | 20.40K |
| $632 \times 10^{-6}$ | 206 | 291 | 615 | 909 | 1340 | 1920 | 2840 | 5460 | 9240 | 15.70K |
| $850 \times 10^{-6}$ | 167 | 236 | 487 | 713 | 1150 | 1530 | 2260 | 4310 | 7320 | 12.40K |
| $1.12 \times 10^{-3}$ | 131 | 186 | 393 | 572 | 935 | 1201 | 1800 | 3850 | 5900 | 10.0K |
| $2.07 \times 10^{-3}$ | 96.3 | 136 | 283 | 415 | 571 | 910 | 1250 | 2840 | 4260 | 7540 |
| $3.54 \times 10^{-3}$ | 65.1 | 92.0 | 192 | 255 | 454 | 601 | 875 | 1690 | 2870 | 4920 |
| $5.67 \times 10^{-3}$ | 45.7 | 64.5 | 134 | 196 | 313 | 424 | 617 | 1160 | 2030 | 3460 |
| $8.65 \times 10^{-3}$ | 33.2 | 47.0 | 97.7 | 142 | 227 | 307 | 444 | 843 | 1470 | 2510 |
| $12.7 \times 10^{-3}$ | 23.8 | 33.6 | 71.1 | 103 | 168 | 222 | 310 | 622 | 1073 | 1840 |
| $20.4 \times 10^{-3}$ | 17.9 | 25.3 | 51.8 | 75.8 | 122 | 163 | 237 | 447 | 777 | 1340 |
| $33.2 \times 10^{-3}$ | 12.2 | 17.2 | 36.1 | 52.8 | 85.5 | 113 | 165 | 320 | 544 | 932 |
| $56.7 \times 10^{-3}$ | 8.22 | 11.6 | 24.2 | 34.6 | 57.8 | 76.3 | 111 | 215 | 364 | 618 |
| $55.3 \times 10^{-3}$ | 6.06 | 8.566 | 16.9 | 25.6 | 42.1 | 55.5 | 70.3 | 156 | 263 | 451 |
| $90 \times 10^{-3}$ | 4.47 | 6.32 | 12.3 | 19.4 | 31.6 | 40.5 | 51.0 | 116 | 201 | 343 |
| 0.153 | 2.98 | 4.07 | 8.52 | 13.1 | 21.1 | 27.9 | 40.8 | 78.5 | 133 | 229 |
| 0.245 | 2.18 | 3.09 | 6.28 | 9.19 | 14.8 | 19.6 | 28.6 | 55.4 | 95.0 | 160 |
| 0.374 | 1.50 | 2.13 | 4.57 | 6.49 | 10.8 | 14.2 | 21.0 | 40.2 | 68.2 | 117 |
| 0.589 | 1.12 | 1.59 | 3.27 | 4.89 | 7.86 | 10.3 | 14.9 | 22.0 | 49.0 | 84.1 |
| 0.9443 | 0.780 | 1.10 | 2.31 | 3.13 | 5.46 | 7.22 | 10.6 | 20.0 | 34.4 | 59.3 |
| 1.52 | 0.542 | 0.773 | 1.61 | 2.35 | 3.80 | 5.13 | 7.40 | 14.1 | 24.2 | 41.5 |
| 2.46 | 0.383 | 0.538 | 1.13 | 1.67 | 2.69 | 3.56 | 5.47 | 10.11 | 17.2 | 29.4 |
| 3.76 | 0.271 | 0.394 | 0.829 | 1.22 | 1.99 | 2.61 | 3.81 | 7.36 | 12.4 | 21.4 |
| 5.98 | | | 0.591 | 0.861 | 1.41 | 1.84 | 2.15 | 5.24 | 8.87 | 15.1 |
| 9.77 | | | | | | | | | | 11.29 |
| 6.57 | 0.178 | 0.280 | 0.423 | 0.644 | 0.999 | 1.36 | 2.00 | 3.52 | 5.49 | 7.09 |
| 16.6 | 0.105 | 0.121 | 0.210 | 0.366 | | | | | | |

**Figure 3.1** Dale's guide to derating power wirewound resistors. (From Ref. 1, p. 5.)

| | Resistance (ohms) | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|
| Energy per ohm [J/ohm or (W·s)/ohm] | EGS-1 RS-G-1 per ohm | EGS-2 RS-$\frac{1}{2}$ G-2 | EGS-3 RS-1A G-3 | RS-1B | RH-5 ESS-2B RS-2B G-5 | RS-2C RS-2C | RH/PH-10 G-6 G-8 | RH-25 EGS-10 RS-5 RS-5-69 G-10 | RS-7 G-12 | RH-50 ESS-10 RS-10 RS-10-38 G-15 |
| 20.9 | | | | | 0.529 | 0.784 | 1.04 | | | |
| 8.04 | | | | | | | | 2.01 | 3.19 | 5.50 |
| 20.9 | | | 0.170 | 0.259 | 0.370 | 0.499 | 0.675 | 1.27 | 1.46 | 3.69 |
| 33.2 | | | | | | | | 0.639 | 0.984 | 1.86 |
| 42.2 | | | 0.081 | 0.114 | | | | | | |
| 83.8 | | | | | 0.189 | 0.259 | 0.344 | | | |
| 25.1 | | | | | | | | 0.484 | 0.697 | 1.45 |
| 67.8 | | | | | 0.099 | 0.129 | 0.179 | 0.329 | 0.514 | 0.949 |
| 169 | | | | | 0.063 | 0.079 | 0.100 | 0.209 | 0.319 | 0.579 |
| 335 | | | | | | | | 0.139 | 0.209 | 0.399 |

**Figure 3.1** (*Continued*)

## Noninductive Wirewound Resistors

As mentioned, resistors have inductance too; it's not usually of any concern unless you want to use one for a current sensor, and decide that the power dissipation calls for a wirewound device. Because they are wound with wire, resistors of this type can have inductance so large that at typical power supply switching frequency, the inductive reactance is larger than the resistance, which then gives nonsense readings of current.

There is an alternative here: some manufacturers produce a special type of wirewound resistor that has very low inductance (although not zero) by specially winding the wire—of course, these resistors cost a bit more.

## Shunts

When you go to really high current (and still don't want to use a current transformer, perhaps because there is DC current involved) you will want to use a shunt, which is a big piece of metal with an almost zero temperature coefficient (manganin) attached to heavy-duty brass terminal blocks. Shunts come in any size you could ever want: the author once had a 1500A job that could have had another life as a boat anchor. However, in addition to resistance, shunts also have inductance, which is very limiting again. As an example, consider a 100A shunt that produces 100mV at full current (100mV and 50mV at full current are the two standard types). It obviously has a resistance of $100\text{mV}/100\text{A} = 1\text{m}\Omega$. But the metal itself is about an inch long, which corresponds to about 20nH. So the transfer function of this device has a zero at a frequency of $f = 1\text{m}\Omega/(2\pi \times 20\text{nH}) = 8\text{kHz}$. All you can do here directly is get a shunt that produces a higher voltage (this increases $R$) or one that consists of multiple stacked pieces of metal, which reduces $L$. Chapter 7, on monitoring, shows a technique for differential amplification of the signal from a shunt that removes the effect of its inductance.

### Using a Trace as a Resistor

Of course a trace is just a piece of copper, so it has some resistance, too. Sometimes, you don't need very accurate current sensing, perhaps for a converter's overcurrent limit. A trace might work just fine in such a case: it's there anyway, so there are no additional losses, and it doesn't cost anything. Of course, the resistance is only as accurate as the trace is cut, and copper has a temperature coefficient of 0.4%/°C.

> **Practical Note** The resistance of a trace is approximately given by the formula:
>
> $$R = 0.5\text{m}\Omega \frac{\text{length}}{\text{width}} \text{ (1 oz. copper)}$$
>
> at room temperature. Two-ounce copper is half this, etc.

## CAPACITORS AND THEIR USAGE

There are quite a number of different capacitors used in power supplies, and each type has its own idiosyncrasies. It's really not possible to use only one type of capacitor; different kinds must be used in different applications if you're going to have a successful design. We'll cover some of the main points, paralleling the selection guide presented in Table 3.2.

### Types of Capacitors

One common type of capacitor is the electrolytic capacitor used for the input or output of a supply. There are a variety of choices available. The most common (and cheapest) is the aluminum electrolytic. (You'll find that some people mean "aluminum electrolytic" when they just say "electrolytic.") There are also tantalum electrolytic capacitors, which are available in solid and wet varieties. The aluminums are available in the widest variety of values and voltages and can have gigantic values (millifarads, and hundreds of volts) but

**TABLE 3.2** Brief Capacitor Selection Guide

| Type | Suggested Applications |
|---|---|
| Aluminum electrolytic | Use when large capacitance is needed and size is unimportant, such as input and output capacitors on a converter. |
| Tantalum electrolytic | Use for moderately large capacitance, such as input and output capacitors on a converter. |
| Ceramic | Use for timing and signal applications. |
| Multilayer ceramic | Use for lowest ESR (e.g., in parallel with an electrolytic at the input or output of a converter). |
| Plastic | Use for high $dV/dt$, such as in quasi-resonant converters. |

they are correspondingly gigantic in size. Tantalums have substantially better high frequency performance than aluminum, but cost more and are limited to about 100V and a few hundred microfarads. Nowadays, the best choice for a medium-power power supply may be to have an aluminum as the input capacitor for a supply, and a tantalum chip as the output. (Chips of course have much smaller capacitance and voltage than discretes.)

Then there are ceramic capacitors. These are used for timing and bypass. The ordinary variety come in a range from a few picofarads to 1μF. But also on the verge of affordability is the MLC (multilayer ceramic) variety, which has extremely low ESR and much larger values available, up to a few hundred microfarads.

Let's also mention plastic capacitors, particularly polypropylene, which are used in circuits that have very high $dV/dt$ values (but see below) such as in quasi-resonant converters.

## Standard Values

Not at all like resistors, there are only a few standard values for capacitors, (1.0, 1.2, 1.5, 1.8, 2.2, 2.7, 3.3, 4.7, and 6.8, with an occasional 5.6 and 8.2). So when you're calculating a time constant or a loop compensation value, select one of these values and adjust your resistors to get the values needed—it'll be a lot cheaper than trying to synthesize that 347pF cap.

There are some practical limits to how small a capacitor you can usefully use, just as we found for maximum value resistors, and for the same reason. Again, two surfaces in close proximity form a capacitor, and for very tiny discrete capacitors, the parallel capacitance that is formed can swamp out the value you're trying to use. So again,

> **Practical Note** Avoid using capacitors smaller than 22pF unless special precautions are taken.

## Tolerance

Capacitor initial tolerances are typically ±20%, and can be substantially worse for electrolytics. You need to look very carefully at any electrolytic, to verify that it's going to be OK in production. Examine the tolerance over the temperature range even more carefully: some types lose 80% of their capacitance at −40°C!

## ESR and Power Dissipation

Modern manufacturers of electrolytics specify ESR (equivalent series resistance) of their caps, and you should try to use only those that specify it at a high frequency, such as 100kHz.

**Practical Note** *You will have no idea what the ESR of a cap is at 100kHz from data given at 120Hz.* This ESR, in addition to being a function of frequency, also depends on temperature. At −25°C, the ESR can be almost triple its value at 25°C! To come close to predicting capacitor ESR, you need data at least within an order of magnitude of your intended operating frequency.

**EXAMPLE**

I have an output ripple current of $1A_{pp}$ at 100kHz, and I need an output voltage ripple of $50mV_{pp}$. First off, I have a charge that could be as large as $1A \times (1/100kHz) = 10\mu C$, so even ignoring ESR, I need a capacitance of $C = Q/V = 10\mu C/50mV = 200\mu F$. Let's assume, then, that I'm going to use at least two 100μF electrolytics. Typically, a capacitor this size may have an ESR of something like 100mΩ at room temperature. To get down to 50mV, I need an ESR of $50mV/1A(=50m\Omega)$, which is the two caps in parallel. *But*, at −25°C, the caps have more like 300mΩ ESR each, so I actually need six caps. With six caps, then, the ripple due to ESR is 50mV at temperature, and the ripple due to capacitance is only about 17mV; since resistance and capacitance are out of phase, the total ripple will be about $I_{total} = [(50mV)^2 + (17mV)^2]^{1/2} = 53mV$. Clearly, when you're designing a bulk filter, ESR can often be more important than total capacitance.

## Aging

Although the matter of aging is easy to overlook, those specs that say "life 1000 hours" really mean something on electrolytics. If you are going to run a supply at elevated temperatures or for many years, you need to pick electrolytics that are designated as 2000h types at least, or better, 5000h. What happens is that as you approach the age rating, the capacitance goes down, and your ripple goes up until the supply ceases to meet spec. This is not an old wives' tale, either. You may not be about to wait around a year to see how bad it gets, but accelerated life tests quickly show up the differences between capacitors.

Fortunately, though, the life of the capacitor doubles with every 10°C drop in temperature, so a type rated 2000h at 85°C will last $2000h \times 2^6 = 128{,}000h = 16$ years at 25°C average temperature.

**Practical Note** Make sure you use the *average* temperature the supply sees over its lifetime for this calculation, not the maximum temperature the supply will see, nor the rated temperature—otherwise you'll find there are no caps available that will meet spec over life!

## *dV/dt*

A different type of usage of capacitor that is growing more common is the use of a metallized plastic cap for a quasi-resonant converter. In this application, there can be substantial dissipation in the ESR of the cap, and this is in fact the limiting factor on the capacitor size. Whereas electrolytics frequently are rated with a ripple current (which is basically determined by the ESR $I^2R$ loss and the thermal characteristics of the package), plastic caps have the equivalent rating in terms of $dV/dt$ [since charge $Q = C \times V$, current $I = dQ/dt = C(dV/dt)$]. To be sure that your cap is adequately rated requires measurement in-circuit. Whether you measure the current through the cap or its $dV/dt$ depends on the circuit configuration—you may need a large-bandwidth differential amplifier to measure accurately $dV/dt$, but you need a loop to measure current, which can introduce unwanted inductance. In any case, make sure you get a capacitor rated for the $dV/dt$ you're applying. Otherwise the capacitor can self-destruct!

### Putting Caps in Series

If I can't get the voltage rating needed for a cap, how about putting two (or more) capacitors in series? Remember that two capacitors in series form a (reactive) voltage divider, and so if one is smaller than the other, it will carry a greater percentage of the total voltage. This sort of design is not really recommended, but if you need it, try putting a resistor in parallel with each cap, as shown in Figure 3.2. This will tend to balance the voltages.

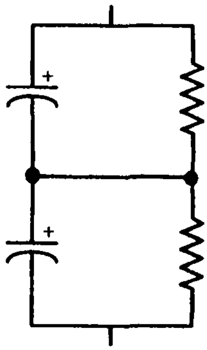


**Figure 3.2** Practical method for placing capacitors in series.

## SCHOTTKY DIODES

Schottkys are great as output rectifiers, because they have a low forward voltage and no reverse recovery time, right? Although it's true that they have no reverse recovery time *as such*, they often do have substantial *capacitance* from anode to cathode. This capacitance has to be charged and discharged every time the voltage across the schottky changes (it's largest when the schottky has almost no voltage across it). Current flowing into this capacitance looks a whole lot like reverse recovery current of an ordinary rectifier. So depending on your circuit, there can be times when it's less lossy to use an ultrafast rectifier than a schottky.

You might also take note that the anode–cathode capacitance, although low-$Q$, can still resonate with stray inductances in the circuit—this property is used intentionally in some resonant designs. So it may be necessary to add a snubber across the schottky, dissipating even more power.

> **Practical Note** Schottkys are very leaky at high temperatures and as the applied reverse voltage increases toward its rating. This leakage can look like a short on a forward's secondary, and indeed the leakage current is the main reason that no one uses germanium rectifiers today. So as a practical limit, you shouldn't try to use a schottky at more than about three-quarters of its rated reverse voltage, nor with a die temperature above about 110°C.

Given this tip, what about using something like a 100V schottky? Look carefully at the specs—as of 1996, the high voltage schottky tended to have a forward voltage comparable to that of a regular rectifier, so you might not be buying much with such a device.

## RECTIFIER DIODES

Someone has just announced to you, the design engineer, that the output of that 12V rail is going to need 1.6A, not 1A. Rather than try to get a new part, and worrying about whether it will fit in the old holes, we'll try to just parallel up two of the old 1A diodes, OK? After all, our buddy John says he did it many years ago, and it worked out. Bad idea! As diodes get hotter, their forward voltage decreases, so the one that is conducting the most current at the start will get hotter, have a lower $V_f$, and conduct yet more, and so on, until it tries to conduct the whole current and fails—positive feedback, remember? So although it is possible to parallel rectifiers by very careful thermal management (i.e. by ensuring that there is minimal thermal resistance between them), *in practice*, these schemes never work out very well.

> **Practical Note** Bite the bullet and get a single diode that can handle the whole current.

Although you can get single diodes of almost any size, it's worth noting that MOSFETs *do* share current, because as *their* temperature goes up, their resistance goes up too. Thus a FET carrying more than its fair share of current will have a higher drop than a parallel device, and will thus correct itself—a negative feedback. This is one of the attractive features of synchronous rectifiers.

### Reverse Recovery

We've mentioned that schottky diodes don't have a reverse recovery time; all other diodes, however, do. That is, after a diode has been conducting current in the forward direction, it will be able to conduct current in the opposite direction (yes, from cathode to anode) for a short time afterwards, and this time is called the reverse recovery time. Figure 3.3 illustrates this anomaly, which clearly would be very bad for converter efficiency and

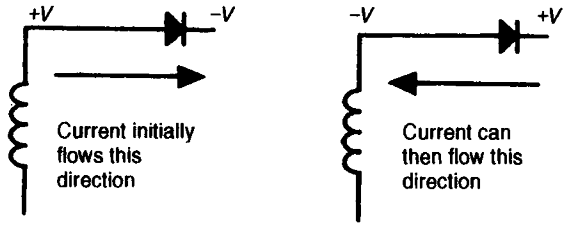


**Figure 3.3** After current flows from anode to cathode, applying a reverse voltage to the diode can cause current to flow from cathode to anode.

must be avoided. There are different grades of diode (fast, ultrafast, etc.), depending on speed of recovery.

> **Practical Note** Converters are almost always going to use either ultrafast diodes or schottkys in their output stages.

Not mentioned in the practical note was synchronous rectification. The reason is that MOSFET body diodes usually have very slow reverse recovery, often about 1 µs. They are thus not suitable for rectification, and this is why a schottky is usually paralleled with a MOSFET synchronous rectifier: the schottky takes almost all the current during the time the MOSFET is off, which means that the body diode doesn't have to reverse recover.

### Is Faster Better?

This certainly *seems* like a generic rule, for after all, a diode that recovers faster will have lower losses. In the case of a rectifier used in an off-line bridge, however, it is not a good idea to use an ultrafast rectifier. The problem is that the fast recovery time also generates fast edges: read EMI. So for this particular case, your best bet is to use that old-fashioned regular bridge rectifier with a recovery time of 5–10µs. After all, it recovers only 120 times a second, so who cares if it's a little slow?

## TRANSISTORS: BJTs

### Pulse Current

If you're using power bipolars at all, you are presumably aware that they take substantially more care and attention than is required in designing with MOSFETs. Let's talk about some performance aspects of BJTs that are often not mentioned in data sheets. First, many small-signal BJTs, and power bipolars not designed specifically for switching, tell you their maximum DC collector current but don't give any curves or numbers relating to pulse currents.

> **Practical Note** When the manufacturer doesn't (or won't) give a pulse rating for a bipolar, a reasonable guess is that the device can take twice the rated DC current for a pulse. If this were based on the fusing current of the bond wire, you would think it would depend on the time; in fact, the limit seems rather to be set by localized current hogging. You'll be safest not to exceed the 2× limit even with short current pulses.

## How Much Beta Can I Get?

The beta of a BJT (not referring to darlingtons, now) depends on all sorts of parameters—collector current, aging, temperature, not to mention initial tolerance. If you figure up all these parameters together, you may find that your bipolar has almost no gain left at all!

> **Practical Note** If you want to make a safe design, that is, one you don't have to sweat over in worst-case analysis, assume that your BJTs have a minimum beta of 10, regardless of what the data sheets seem to say.

## Don't Forget Collector Leakage Current

And don't forget that this current, too, increases with the "double for each 10°C" rule. I recently saw a design that used a 4.7MΩ pull-up on the collector of a bipolar. It seemed to work in the lab, but as soon as it saw any sort of temperature rise above ambient, the collector voltage went to zero! In practice, you had better plan on a leakage of up to 1mA, depending on the size of the device.

## Emitter–Base Zenering—Is It Bad?

Another limit on BJT performance is the emitter–base voltage $V_{eb}$, that is, how much negative voltage can be applied to the base of a bipolar with respect to its emitter. The manufacturer will usually say this limit is 5V or 6V. But what really happens if this value is exceeded? The base–emitter junction is a diode, and if you apply enough voltage to it, it will zener. (You can test this in the lab with a current-limiting resistor.) You might actually want to do this in a converter, because turning off a BJT involves sweeping out the current from the base region; the more negative the voltage on the base, the faster you can turn off the transistor. The limit, of course, is that the fastest possible turnoff occurs when the base–emitter is zenering.

> **Practical Note** You can apply any voltage emitter to base you care to, as long as you don't exceed the power rating of the die; that is, the real limitation on zenering is on the product $V_{eb} \times I_{eb}$. In practice this means ensuring that there is some sort of base resistor, to limit the current. Manufacturers typically refuse to guarantee this as a specification, but that's what they will tell you privately.

## Fast Turnoff

Better than sweeping the base charge out fast is ensuring that there isn't too much base charge in the first place. Unfortunately, the easiest tactic, running the transistor at nearly its actual beta, conflicts with the assumption that your transistor is going to have a minimum beta of 10, much lower than the typical beta. If you need to have fast turnoff, it may be worthwhile to try a Baker clamp, which is about the best you can do, although it dissipates some extra power.

The Baker clamp (Figure 3.4) works as follows. When the transistor is on, the base is one diode drop above the emitter, and so the driving source is two diode drops above the emitter. The extra diode from the driving source to the collector then assures that the collector is approximately one diode drop above the emitter, which is to say, the BJT is almost, but not quite, saturated. It can thus be turned off quickly; whether this fast turnoff decreases the circuit losses enough to compensate for the increased losses due to the increased collector–emitter voltage depends on the particulars of a given design. (Observe that these statements are only approximations, since the currents and the $V_f$ values of the various diodes differ.)

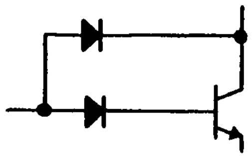

**Figure 3.4** A Baker clamp prevents deep saturation, speeding turnoff.

# TRANSISTORS: MOSFETs

## Don't Confuse JFETs and MOSFETs

When everyone used bipolars and MOSFETs were new, it was fairly common to get the terms mixed up. Just to be sure: JFETs are small-signal devices with high on-resistance frequently used in rf-type work; MOSFETs, and specifically power MOSFETs, are what are used nowadays for power applications.

## p-Channel and n-Channel

Most designs use n-channel MOSFETs, and if there's no indication otherwise, you can simply assume that all MOSFETs in a design are n-channel. One of the reasons p-channels are less popular than n-channels is that p-channels have higher on-resistance for the same voltage and die area, making them more expensive.

Still p-channels have a certain utility: they are turned off when their gate–source voltage is below a threshold (similarly to n-channels), but they are on when the gate voltage is below the source voltage, that is, negative. So in practice, they are used by attaching the source of the p-channel to a voltage (e.g., 5V) and letting the gate be at 5V to turn it off, or pulling it to ground to turn it on. The advantage is that whereas turning on an n-channel would require a voltage *higher* than 5V (such as 12V), the p-channel needs no extra supply, being turned on simply by pulling its gate to ground.

## Bidirectional Conduction

Although MOSFETs are used routinely in synchronous rectification, perhaps it should be mentioned that they can conduct current in both direction, drain-to-source as well as source-to-drain, not counting the body diode. Applying a voltage from gate to source (for n-channel FETs) enables them to conduct bidirectionally. In synchronous rectification, this "reverse direction" conduction is explicitly used to short out the body diode, since the product of the current and the $R_{DS,on}$ of the FET is much smaller than the $V_f$ of the body diode.

## Calculating Losses: Conduction Loss

There are three sources of power dissipation in switching applications of power MOSFETs; the first to be discussed is conduction loss. When a MOSFET is fully on, it has a resistance from drain to source, and this dissipates power based on how much current goes through it, $P = I^2 R_{DS,on}$. However, you need to be aware that this resistance goes up with temperature [typically $R(T) = R(25°C) \times 1.007 \exp(T - 25°C)$]; so to find the actual junction temperature, you have to calculate the total power dissipated, figure out what temperature this causes (by multiplying by the thermal resistance), then recalculate the power based on the new temperature's resistance, and so on iteratively until the calculation has converged.

> **Practical Note:** A single iteration of this calculation is almost always good enough, because of limited knowledge of the actual thermal resistance. If it doesn't converge after one iteration you're probably dissipating more power than the device can take!

While on the subject of $R_{DS,on}$, you might take note that "logic-level" FETs are a little bit of a cheat. While their gate–threshold voltage is indeed lower than for a regular FET, their on-resistance is also higher than it would be if the gate were driven to a normal level. Typical logic-level FETs may have twice the $R_{DS,on}$ at 4.5V $V_{GS}$ as at 10V.

## Calculating Losses: Gate Charge Loss

A second source of loss, though not lost in the MOSFET, is due to the MOSFET having a rather substantial equivalent gate capacitor. (The losses are in whatever devices and resistors drive the gate.) Although the capacitance is actually a highly nonlinear function of gate voltage, many modern data sheets give a total gate charge $Q_g$, required to bring the gate voltage to a certain level $V$. Power lost by driving this charge into the gate at the switching frequency $f_s$ is then $P = Q_g V f_s$. Note that there is no factor of 0.5.

> **Practical Note** If the gate voltage you actually drive the gate to differs from the one specified in the data sheet, it is probably a reasonable approximation to multiply the specified gate charge by the ratio of the two voltages; this works best if the voltage you are using is higher than that in the data sheet. (For the cognoscenti, the limiting factor in the approximation is how much charge is required to charge the Miller capacitance.)

## Calculating Losses: Switching Loss

The third source of loss in switching MOSFETs, and the second that dissipates into the MOSFET, is switching loss. Whenever a (nonresonant) transistor turns off or on, it simultaneously has both voltage on it and current through it, resulting in power dissipation.

> **Practical Note** An estimate of the switching losses can be had by assuming that the voltage is a linear function of time while the current is constant, in which case the power lost is $P = I_{pk} V_{pk} t_s f_s / 2$ for discontinuous conduction mode converters, and double this for continuous conduction mode, where $t_s$ is the transition time of the MOSFET drain-to-source voltage from on-to-off (and from off-to-on for continuous conduction converters); this is why driving the gate harder results in lower switching losses.

In summary, total losses associated with switching power MOSFETs are due to conduction loss, gate charge loss, and switching loss, of which only the first and last are dissipated in the FET. You can get a pretty good idea of the losses in the transistors by doing these calculations. Then by using the thermal resistance of the package, you should know whether the FET will be cool or hot or very hot; if it's not pretty close to what you calculate, there's something wrong!

### The Need for Gate Resistors

You (hopefully) always put a gate resistor in series with the gate of a MOSFET. But if you have two FETs in parallel, can you still use just one resistor, maybe of half the value?

> **Practical Note** You need an individual gate resistor for each MOSFET, regardless of whether the devices are in parallel, and even if they have some other current-limiting part, such as a bead, in series. The reason is that MOSFETs have both capacitance (gate–source) and inductance (in the leads). This potentially forms an underdamped resonant tank, and paralleled MOSFETs have been observed to oscillate at 100MHz! If you're using a digital oscilloscope and don't know to look for these oscillations, you may not even see them, but they can be lossy, and of course they wreak havoc in EMI. The gate resistor acts to limit the current the source has to source or sink to the gate, but its real significance is to damp the oscillations.

### Maximum Gate Voltage

One last thing to avoid. On occasion people get the bright idea that to reduce switching loss, they're going to <u>really</u> drive that gate and use a 40V source or the like to ensure that it charges past the gate threshold voltage very fast. Don't even think about it. You end up having to stick a clamping zener diode on the gate to prevent it from exceeding its maximum voltage rating (always 20V nowadays), and that then throws away more power than you could possibly have hoped to save. The right solution is to get a gate driver with lower output impedance. In bare die form, power MOSFETs have been turned on in 10ns with the right driver!

# OP AMPS

This section talks about the main parameters that can affect whether you get the anticipated operation out of your op amp design: offsets, limits on achievable gains, gain–bandwidth, phase shifts, and slew rates. Regardless of your application, you need to be familiar with these nonidealities.

## Offsets: Input Offset Voltage

What's the story with offsets? Let's try to untangle how to use these specifications on op amps, and also give tips as to when they're important.

Consider Figure 3.5, the schematic for a noninverting amplifier with a gain of 10. (To make the discussion easy to follow, the input is grounded, but the effects of the offsets would be exactly the same if instead a nonzero voltage were used as the input.) Since its input signal is ground, we might naively expect its output to be zero volts also. But now consider that the LM2902 has a typical input offset voltage $V_{os}$ of 2mV. (Plus or minus is always implied, if not explicitly stated.) What this means is that even with no input, the noninverting terminal will see (something between plus and minus) 2mV. Of course, the same thing applies at the inverting terminal when the op amp is used as an inverter. This 2mV gets amplified just like an ordinary, desired signal, and so at the output will appear as (something between plus and minus) 20mV. This signal is additive, so if we had applied 100mV at the noninverting terminal, at the output we would have, instead of exactly $100\text{mV} \times 10 = 1\text{V}$, somewhere between $(100\text{mV} - 2\text{mV}) \times 10 = 980\text{mV}$ and $(100\text{mV} + 2\text{mV}) \times 10 = 1.02\text{V}$. This value is clearly independent of the absolute values of the resistors used; it is dependent only on the gain. Thus, input offset voltage is important whenever small signals are to be measured and/or when high gain is needed.

## Offsets: Input Offset Current

The same schematic (Figure 3.5) explains about input offset current, which is very similar to input offset voltage. Since the inputs to an op amp are not of infinite impedance, applying a voltage to them causes them to draw (or source) some current. The LM2902 has a typical $I_{io}$ of 5nA. This means that the noninverting (or inverting) terminal will see (something between plus and minus) 5nA being pulled from the voltage source *different from the current being pulled by the other terminal.* In the case of the schematic, we are

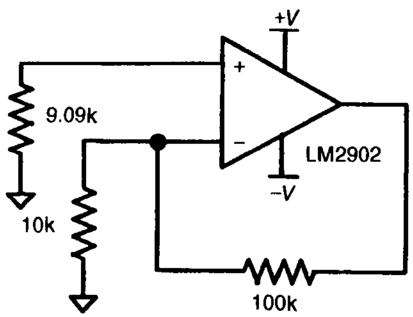


**Figure 3.5** Op amp circuit for discussion of offset errors.

pulling 5nA from ground through a 9.09kΩ resistor, and so the noninverting terminal sees a voltage of $V = (5\text{nA}) \times (9.09\text{k}\Omega) = 45\mu\text{V}$ (again, this could just as well be $-45\mu\text{V}$). This then is amplified by the gain of 10, resulting in 450μV at the output terminal; this is in addition to the offset caused by the input offset voltage.

In this example, the input offset voltage was much more important for errors than the input offset current; but since the error due to the current is determined by the absolute values of the resistances used, in addition to the gain, it is clear that for large source resistances, the offset current becomes more important than the offset voltage.

### Offsets: Input Bias Current

Now for input bias current, which is current pulled by both the inverting and noninverting terminals by the same amount. (It may help to think of input bias current as common mode current, and input offset current as normal mode current.) The LM2902 has a typical $I_b$ of 90nA. The op amp schematic of Figure 3.5 has the same input resistance for both terminals ($100\text{k}\Omega \| 10\text{k}\Omega = 9.09\text{k}\Omega$), so the effect of pulling equal amounts of current from both is none. Suppose, however, that instead of 9.09kΩ to ground, the noninverting terminal had 19.09kΩ to ground. Then there is a difference of 10kΩ in the input resistances, and this results in an offset of $V = 90\text{nA} \times 10\text{k}\Omega = 900\mu\text{V}$, which gets multiplied by the gain to give an output error of 9mV, comparable in size with the error due to the offset voltage. This is the reason for trying to use the same input resistance for both terminals, even when they're virtually grounded.

### What to Do About Offsets

To sum up, the output error of an op amp due to offsets is calculable as

$$V = [V_{os} + (I_{os} \times R) + (I_b \times \Delta R)]\text{gain}$$

where $R$ is the average of the two input resistances, and $\Delta R$ is the difference of the two. Since gain is determined by operating needs, minimizing the error must involve three actions:

1. Ensuring that the resistor values being used are limited to the smallest feasible values; this limits the effects of $I_{os}$ but also results in a need for greater currents from the sources being used to drive the signals.
2. Ensuring that resistor values to the terminals are matched eliminates the effects of $I_b$.
3. Minimizing $V_{os}$, which can be accomplished only through selection of the proper type of op amp.

Unfortunately, a low $V_{os}$ is also invariably accompanied by a higher operating current of the op amp, a lower op amp bandwidth, or both. There are thus engineering trade-offs to be made in selecting the proper op amp for an application.

### Limits on Large Resistances

Sometimes you want a large gain from an op amp and you might try something like the bad example shown in Figure 3.6.

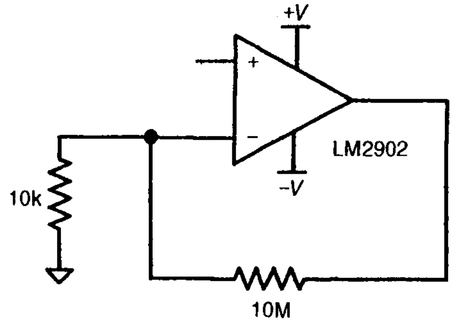


**Figure 3.6** A circuit that won't work very well.

Let's assume that the op amp has adequate gain–bandwidth for your purpose (probably this isn't true; see below)—are you really going to get a gain of 1000? Probably not. The trouble is not with the op amp, or with any of the components—as explained in the section about resistors, it's rather with the PCB on which you mount them. For various reasons, the leakage around a resistor may exceed the amount of current being supplied through the 10MΩ resistor, effectively shunting it with a lower value.

> **Practical Note** It is usually not effective to use a resistor larger than 1MΩ on a normal PCB, at least without taking special precautions. If you must have this gigantic gain, and can't reduce the 10kΩ source resistor to a 1kΩ, try instead the circuit shown in Figure 3.7.

The circuit of Figure 3.7 works as follows. Suppose you have 10mV at the noninverting input. Then the op amp forces the inverting input to have 10mV also (ignoring offsets in this calculation). With 10mV across 10kΩ, there must be a current flowing of 1μA. This current has to come from point *A* through the 90kΩ, so there must be a drop across that resistor of $1\mu A \times 90k\Omega = 90mV$, which added to the voltage at the inverting terminal means that node *A* has to be at $10mV + 90mV = 100mV$. Now 100mV at node *A* means that there must be a current of 100μA flowing into the 1kΩ resistor. This current (plus the 1μA flowing into the 90kΩ) must come from the output through the

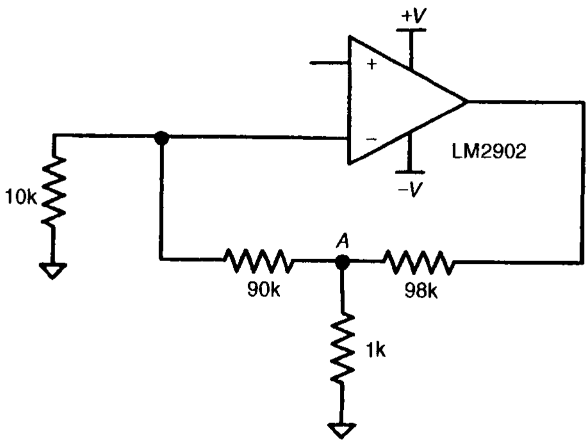


**Figure 3.7** Practical circuit to get a gain of 1000.

$98\text{k}\Omega$, which is therefore dropping $98\text{k}\Omega \times 101\mu\text{A} = 9.9\text{V}$, and the output voltage is this plus the voltage on node $A$, for a total of $9.9\text{V} + 100\text{mV} = 10\text{V}$, a gain of $10\text{V}/10\text{mV} = 1000$. No resistor in this circuit is larger than $100\text{k}\Omega$.

## Gain Bandwidth

Suppose I use an op amp to construct a gain-of-10 amplifier. Suppose next that I use it to amplify a sine wave (ignoring slew rate for the moment, see below), and I keep increasing the frequency of this sine wave. At some frequency, the op amp itself will run out of gain and the output of the amplifier will stop being 10 times larger than the input. If I increase the frequency further, at some point the output will have the same amplitude as the input. This frequency is independent of the external components used to set the gain, and is called the gain bandwidth of the op amp.

One of the places you have to watch out for this parameter occurs when you use an error amplifier in a power supply. For example, the result of a calculation on closing the control loop, discussed in detail in Chapter 6, might be that you need a gain of 300 at a frequency of 20kHz. Well, a gain of 300 isn't so bad, and of course most op amps will work well at 20kHz; unfortunately, the two parameters together imply that the op amp must have a gain bandwidth of $300 \times 20\text{kHz} = 6\text{MHz}$, which may be beyond many of the error amps included in typical PWM ICs. This problem becomes quite noticeable as converter bandwidths reach into the tens of kilohertz. The symptom of having inadequate bandwidth in an error amp may be something like an instability in the converter, even though you have correctly compensated the loop.

## Phase Shift

Even beyond gain bandwidth limitations of common op amps, there is an additional problem: as the sine wave frequency injected into the op amp in the thought experiment in the preceding section is increased, the output sine wave becomes more and more phase-shifted from the original. In the case of an op amp that is being used as the error amplifier in a converter, this translates to extra (unexpected!) phase shift in the loop, reducing phase margin. This too can cause a loop to be unstable even though apparently correctly compensated.

This is unpleasant enough, but worse yet, very few manufacturers give even typical numbers or curves for phase shift as a function of frequency, to say nothing of trying to do a worst case. It turns out that the phase shift is very dependent on the internal construction of the op amp; it is *not the case* that op amps with a higher gain bandwidth product necessarily have less phase shift at a given frequency than those with lower gain bandwidth! In fact, the only practical method for deciding whether a given op amp is going to give excess phase shift for a particular application is to measure it; for example, configure the op amp as a unity gain follower, and use a network analyzer, as described in Chapter 4.

## Slew Rate

The last limitation of op amps we consider is the speed with which they can change from one output voltage to another. In the description of the gain-of-10 amplifier in the discussion of gain bandwidth product, it was assumed that the input signal was tiny.

Suppose instead that the input was $1V_{pp}$; then the output would have to be $10V_{pp}$. If, for example, the frequency of the sine wave was 200kHz, then in one quarter of the period, $(\frac{1}{4}) \times (1/200\text{kHz}) = 1.25\mu s$, the output has to go from 0V to maximum, a change of 10V; this implies that the op amp needs a slew rate of at least $10V/1.25\mu s = 8V/\mu s$, to use the common units. Many common op amps, particularly low power devices, can't slew this fast.

When is this important? One place, again, is in high bandwidth converters. As discussed in Chapter 6, it's not enough for a converter to be small-signal stable, it also has to have adequate response to a transient. When a transient occurs, the output voltage of the error amp has to change levels. If the device happens not to have the slew rate needed to do this, you will be left puzzling about why your fast converter is so slow.

In summary, then, gain bandwidth product and phase shift for an op amp used as an error amp are related to the small-signal performance of the converter; slew rate is related to the large-signal, transient performance.

## COMPARATORS

### Hysteresis

The same offsets and biases discussed for op amps apply in exactly the same way for comparators. Comparators are unique however, in that their outputs can be expected to be either high or low, not anything in between (No, *don't* try to use that spare op amp as a comparator, or vice versa! Distinct parts are built for a reason.) Actually, since comparators are real devices, on occasion they oscillate between these two states, sometimes at surprisingly high frequencies. The usual reason for this behavior is that the comparators don't have hysteresis. This can cause all sorts of problems that take time to debug.

**Practical Note** Always use hysteresis on comparators unless they are intended to run a latch: that is, if the first time the comparator trips, it is intended to cause something from which there is no recovery.

**EXAMPLE**

For small amounts of hysteresis, you can easily guess hysteresis values. For the circuit of Figure 3.8, since $1k\Omega/100k\Omega = 0.01$, the amount of hysteresis will be about 1% of the reference voltage.

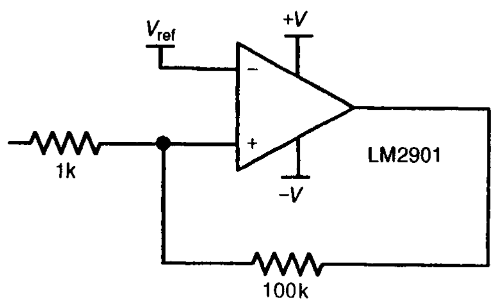


**Figure 3.8** It's easy to guess the hysteresis of this comparator.

### Output Saturation Voltage

One other unique aspect of comparators is that when they go low, they often don't go to 0V. Inspection of the data sheet of the common LM139 shows that its output is only guaranteed to be 0.7V if it is sinking 6mA. So when designing hysteresis, make sure to check how much current the output is intended to be sinking; if it's more than about 1mA, you need to include the saturation voltage in determining the hysteresis resistor values.

Saturation voltage also is important in driving an NPN transistor from the output of the comparator; at 0.7V a "low" probably will suffice to turn on a base–emitter junction and have the transistor on, so you can't use the comparator to directly drive a bipolar! For this situation, you need a blocking diode and a pull-down base resistor. Figure 3.9 shows an arrangement that will work even in worst case (the worst case of this circuit is analyzed in Chapter 10). When the comparator pulls low, even to only 700mV, the diode is off, keeping the transistor off. The pull-down on the base is needed because the base would otherwise be floating when the diode is reverse-biased, and the transistor might be partially on through leakage currents.

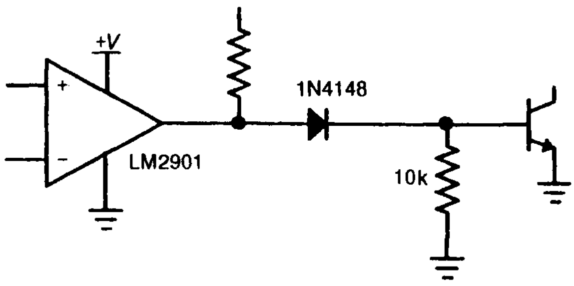


**Figure 3.9** How a comparator should be configured to drive a BJT.

## REFERENCES

1. *Pulse Handling Capability of Wirewound Resistors.* Dale Electronics, Columbus NE.