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Chapter 7

Power Supplies

Our transceivers, amplifiers, accessories, computers and test equipment all require power to operate. This chapter illustrates the various techniques, components and systems used to provide power at the voltage and current levels our equipment needs. Topics range from basic transformers, rectifiers and filters to linear voltage regulation, switchmode power conversion, high voltage techniques and batteries. Expanded and updated by Rudy Severns, N6LF, with support from Chuck Mullett, KR6R, the material presented here builds on the earlier work of Ken Stuart, W3VVN, and other contributors. Alan Applegate, KØBG contributed the section on selecting batteries for mobile use.

7.1 The Need for Power Processing

Electronic equipment requires a source of electrical power. Sometimes the source can be as simple as a battery, but in many cases a battery may not be able to supply sufficient power and/or the voltages required with adequate regulation. For example, the voltage of a vehicle battery may vary from 14 V when fully charged down to 10 V or less when discharged. The lower end of the voltage range can compromise the performance of a transceiver, so it may be desirable to insert a device between the battery and the transceiver to maintain the transceiver input voltage at about 13 V as the battery voltage declines during use.

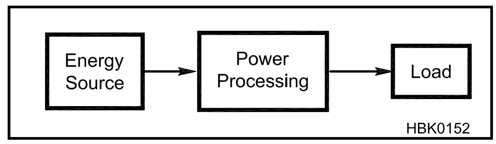


Fig 7.1 — Basic concept of power processing.

Another example is operation from commercial utility power, which may vary from 90 to 270 V ac depending on where you are in the world. In most cases the signal generating and processing equipment we use requires well-regulated dc voltages, sometimes at low voltages (3-24 V) and at other times high voltages (100-3000 V).

Fig 7.1 illustrates the concept of a power processing unit inserted between the energy

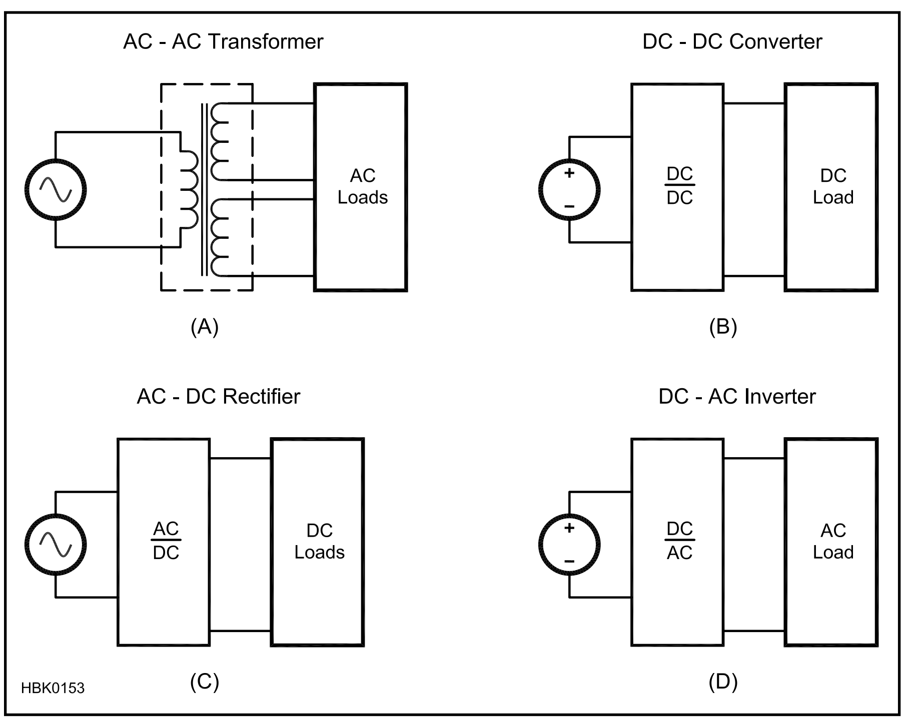


Fig 7.2 — Four power processing schemes: ac-ac, dc-dc, ac-dc and dc-ac.

source and the electronic equipment or load. The *power processor* is often referred to as the *power supply*. That's a bit misleading in that the energy "supply" actually comes from some external source (battery, utility power and so forth), which is then converted to useful forms by the power processor. Be that as it may, in practice the terms "power supply" and "power processor" are used interchangeably.

The real world is even more arbitrary. Power processors are frequently referred to as *power converters* or simply as *converters*, and we will see other terms used later in this chapter. It is usually obvious from the context of the discussion what is meant and the glossary at the end of this chapter gives some additional information.

Power conversion schemes can take the form of: ac-to-ac (usually written ac-ac), ac-dc, dc-ac and dc-dc. Examples of these schemes are given in **Fig 7.2**. Specific names may be given to each scheme: ac-dc => rectifier, dc-dc => converter and dc-ac => inverter. These are the generally recognized terms but you will see exceptions.

In the following discussion it will be assumed that the reader is familiar with the component characteristics presented in the **Electrical Fundamentals** chapter.

7.2 AC-AC Power Conversion

In most US residences, three wires are brought in from the outside electrical-service mains to the house distribution panel. In this three-wire system, one wire is neutral and should be at earth ground potential. (See the **Safety** chapter for information on electrical safety.) The neutral connection to a ground rod or electrode is usually made at the distribution panel. The voltage between the other two wires is 60-Hz ac with a potential difference of approximately 240 V RMS. Half of this voltage appears between each of these wires and the neutral, as indicated in **Fig 7.3A**. In systems of this type, the 120-V household loads are divided at the breaker panel as evenly as possible between the two sides of the power mains. Heavy appliances such as electric stoves, water heaters, central air conditioners and so forth, are designed for 240-V operation and are connected across the two ungrounded wires.

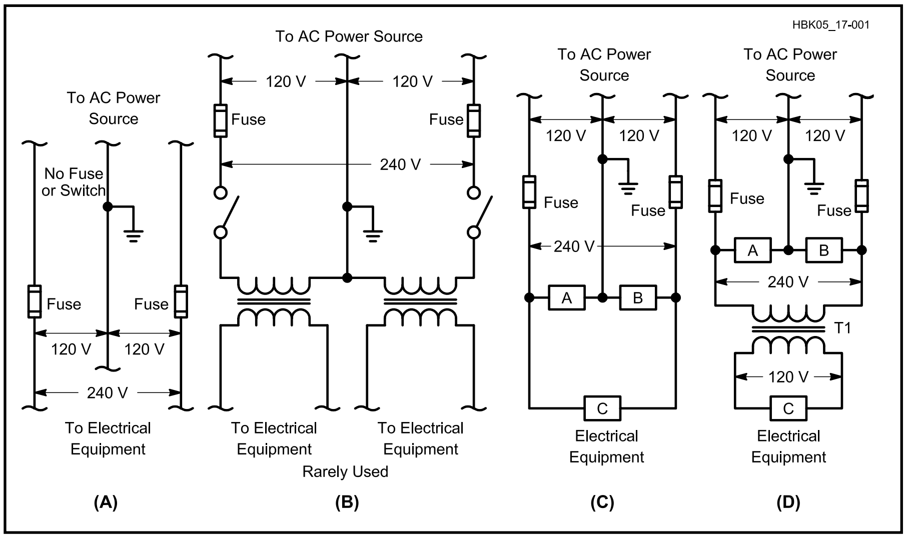


Fig 7.3 — Three-wire power-line circuits. At A, normal three-wire-line termination. No fuse should be used in the grounded (neutral) line. The ground symbol is the power company's ground, not yours! Do not connect anything other than power return wiring, including the equipment chassis, to the power neutral wire. At B, the "hot" lines each have a switch, but a switch in the neutral line would not remove voltage from either side of the line and should never be used. At C, connections for both 120 and 240-V transformers. At D, operating a 120-V plate transformer from the 240-V line to avoid light blinking. T1 is a 2:1 step-down transformer.

Both hot wires for 240 V circuits and the single hot wire for 120 V circuits should be protected by either a fuse or breaker. A fuse or breaker or any kind of switch should *never* be used in the neutral wire. Opening the neutral wire does not disconnect the equipment from an active or "hot" line, possibly creating a potential shock hazard between that line and earth ground.

Another word of caution should be given at this point. Since one side of the ac line is grounded (through the green or bare wire — the standard household wiring color code) to earth, all communications equipment should be reliably connected to the ac-line ground through a heavy ground braid or bus wire of #14 or heavier-gauge wire. This wire must be a separate conductor. You must not use the power-wiring neutral conductor for this safety ground. (A properly-wired 120-V outlet with a ground terminal uses one wire for the ac hot connection, one wire for the ac neutral connection and a third wire for the safety ground connection.) This provides a measure of safety for the operator in the event of accidental short or leakage of one side of the ac line to the chassis.

Remember that the antenna system is frequently bypassed to the chassis via an RF choke or tuned circuit, which could make the antenna electrically "live" with respect to the earth ground and create a potentially lethal shock hazard. A *Ground Fault Circuit Interrupter* (GFCI or GFI) is also desirable for safety reasons, and should be a part of the shack's electrical power wiring.

7.2.1 Fuses and Circuit Breakers

All transformer primary circuits should be fused properly and multiple secondary outputs should also be individually fused. To determine the approximate current rating of the fuse or circuit breaker on the line side of a power supply it is necessary to determine the total load power. This can be done by multiplying each current (in amperes) being drawn by the load or appliance, by the voltage at which the current is being drawn. In the case of linear regulated power supplies, this voltage has to be the voltage appearing at the output of the rectifiers before being applied to the regulator stage. Include the current drawn by bleeder resistors and voltage dividers. Also include filament power if the transformer is supplying vacuum tube filaments. The National Electrical Code (NEC) also specifies maximum fuse ratings based on the wire sizes used in the transformer and connections.

After multiplying the various voltages and currents, add the individual products. This is the total power drawn from the line by the supply. Then divide this power by the line voltage and add 10 to 30% to account for the inefficiency of the power supply itself. Use a fuse or circuit breaker with the nearest larger current rating. Remember that the charging of filter capacitors can create large surges of current when the supply is turned on. If fuse blowing or breaker tripping at turn on is a problem, use slow-blow fuses, which allow for high initial surge currents.

For low-power semiconductor circuits, use fast-blow fuses. As the name implies, such fuses open very quickly once the current exceeds the fuse rating by more than 10%.

7.3 Power Transformers

Numerous factors are considered to match a transformer to its intended use. Some of these parameters are:

1. Output voltage and current (volt-ampere rating).
2. Power source voltage and frequency.
3. Ambient temperature.
4. Duty cycle and temperature rise of the transformer at rated load.
5. Mechanical considerations like weight, shape and mounting.

7.3.1 Volt-Ampere Rating

In alternating-current equipment, the term *volt-ampere* (*VA*) is often used rather than the term watt. This is because ac components must handle reactive power as well as real power. If this is confusing, consider a capacitor connected directly across the secondary of a transformer. The capacitor appears as a reactance that permits current to flow, just as if the load were a resistor. The current is at a 90º phase angle, however. If we assume a perfect capacitor, there will be no heating of the capacitor, so no real power (watts) will be delivered by the transformer. The transformer must still be capable of supplying the voltage, and be able to handle the current required by the reactive load. The current in the transformer windings will heat the windings as a result of the I^2R losses in the winding resistances. The product of the voltage and current in the winding is referred to as "volt-amperes," since "watts" is reserved for the real, or dissipated, power in the load. The volt-ampere rating will always be equal to, or greater than, the power actually being drawn by the load.

The number of volt-amperes delivered by a transformer depends not only upon the dc load requirements, but also upon the type of dc output filter used (capacitor or choke input), and the type of rectifier used (full-wave center tap or full-wave bridge). With a capacitive-input filter, the heating effect in the secondary is higher because of the high peak-to-average current ratio. The volt-amperes handled by the transformer may be several times the power delivered to the load. The primary winding volt-amperes will be somewhat higher because of transformer losses. This point is treated in more detail in the section on ac-dc conversion. (See the **Electrical Fundamentals** chapter for more information on transformers and reactive power.)

7.3.2 Source Voltage and Frequency

A transformer operates by producing a magnetic field in its core and windings. The intensity of this field varies directly with the instantaneous voltage applied to the transformer primary winding. These variations, coupled to the secondary windings, produce the desired output voltage. Since the transformer appears to the source as an inductance in parallel with the (equivalent) load, the primary will appear as a short circuit if dc is applied to it. The unloaded inductance of the primary (also known as the *magnetizing inductance*) must be high enough so as not to draw an excess amount of input current at the design line frequency (normally 60 Hz in the US). This is achieved by providing a combination of sufficient turns on the primary and enough magnetic core material so that the core does not saturate during each half-cycle.

The voltage across a winding is directly related to the time rate of change of magnetic flux in the core. This relationship is expressed mathematically by $V = N\, d\Phi/dt$ as described in the section on Inductance in the **Electrical Fundamentals** chapter. The total flux in turn is expressed by $\Phi = A_e B$, where A_e is the cross-sectional area of the core and B is the flux density.

The maximum value for *flux density* (the magnetic field strength produced in the core) is limited to some percentage (< 80% for example) of the maximum flux density that the core material can stand without saturating, since in saturation the core becomes ineffective and causes the inductance of the primary to plummet to a very low level and input current to rise rapidly. Saturation causes high primary currents and extreme heating in the primary windings.

At a given voltage, 50 Hz ac creates more flux in an inductor or transformer core because the longer time period per half-cycle results in more flux and higher magnetizing current than the same transformer when excited by same 60-Hz voltage. For this reason, transformers and other electromagnetic equipment designed for 60-Hz systems must not be used on 50-Hz power systems unless specifically designed to handle the lower line frequency.

7.3.3 How to Evaluate an Unmarked Power Transformer

Hams who regularly visit hamfests frequently develop a junk box filled with used and unmarked transformers. Over time, transformer labels or markings on the coil wrappings may come off or be obscured. There is a good possibility that the transformer is still useable, but the problem is to determine what voltages and currents the transformer can supply. First consider the possibility that you may have an audio transformer or other impedance-matching device rather than a power transformer. If you aren't sure, don't connect it to ac power!

If the transformer has color-coded leads, you are in luck. There is a standard for transformer lead color-coding, as is given in the **Component Data and References** chapter. Where two colors are listed, the first one is the main color of the insulation; the second

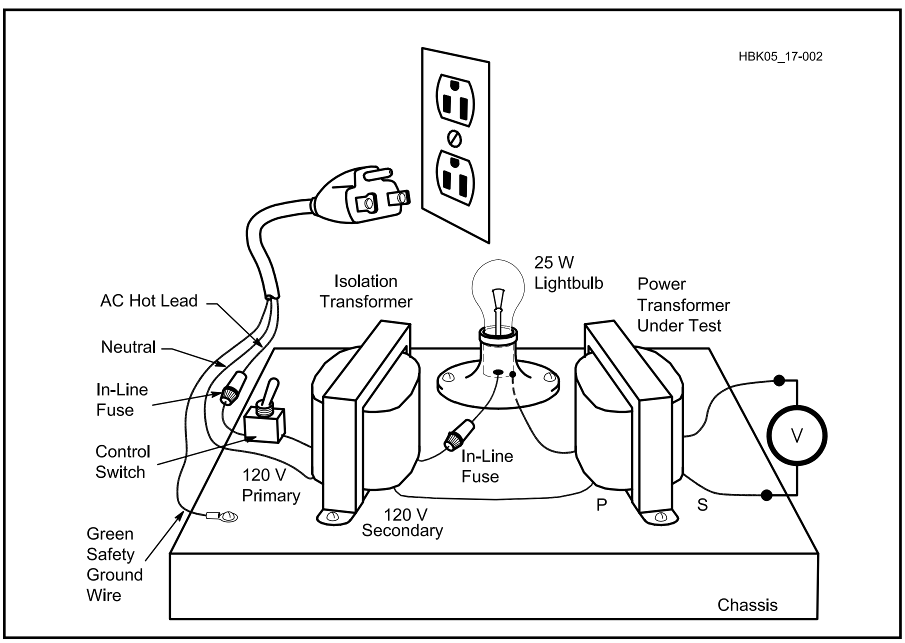


Fig 7.4 — Use a test fixture like this to test unknown transformers. Don't omit the isolation transformer, and be sure to insulate all connections before you plug into the ac mains.

is the color of the stripe.

Check the transformer windings with an ohmmeter to determine that there are no shorted (or open) windings. In particular, check for continuity between any winding and the core. If you find that a winding has been shorted to the core, do not use the transformer! The primary winding usually has a resistance higher than a filament winding and lower than a high-voltage winding.

Fig 7.4 shows that a convenient way to test the transformer is to rig a pair of test leads to an electrical plug with a 25-W household light bulb in series to limit current to safe (for the transformer) levels. For safety reasons use an isolation transformer and be sure to insulate all connections before you plug into the ac mains. Switch off the power while making or changing any connections. You can be electrocuted if the voltmeter leads or meter insulation are not rated for the transformer output voltage! If in doubt, connect the meter with the circuit turned off, then apply power while you are not in contact with the circuit. *Be careful! You are dealing with hazardous voltages!*

Connect the test leads to each winding separately. The filament/heater windings will cause the bulb to light to full brilliance because a filament winding has a very low impedance and almost all the input voltage will be across the series bulb. The high-voltage winding will cause the bulb to be extremely dim or to show no light at all because it will have a very high impedance, and the primary winding will probably cause a small glow. The bulb glows even with the secondary windings open-circuited because of the small magnetizing current in the transformer primary.

When the isolation transformer output is connected to what you think is the primary winding, measure the voltages at the low-voltage windings with an ac voltmeter. If you find voltages close to 6-V ac or 5-V ac, you know that you have identified the primary and the filament windings. Label the primary and low voltage windings.

Even with the light bulb, a transformer can be damaged by connecting ac mains power to a low-voltage or filament winding. In such a case the insulation could break down in a primary or high-voltage winding because of the high turns ratio stepping up the voltage well beyond the transformer ratings.

Connect the voltmeter to the high-voltage windings. Remember that the old TV transformers will typically supply as much as 800 V_{pk} or so across the winding, so make sure that your meter can withstand these potentials without damage and that you use the voltmeter safely.

Divide 6.3 (or 5) by the voltage you measured across the 6.3-V (or 5-V) winding in this test setup. This gives a multiplier that you can use to determine the actual no-load voltage rating of the high-voltage secondary. Simply multiply the ac voltage measured across the high-voltage winding by the multiplier.

The current rating of the windings can be determined by loading each winding with the primary connected directly (no bulb) to the ac line. Using power resistors, increase loading on each winding until its voltage drops by about 10% from the no-load figure. The current drawn by the resistors is the approximate winding load-current rating.

7.4 AC-DC Power Conversion

One of the most common power supply functions is the conversion of ac power to dc, or *rectification*. The output from the rectifier will be a combination of dc, which is the desired component, and ac *ripple* superimposed on the dc. This is an undesired but inescapable component. Since most loads cannot tolerate more than a small amount of ripple on the dc voltage, some form of filter is required. The result is that ac-dc power conversion is performed with a rectifier-filter combination as shown in **Fig 7.5**.

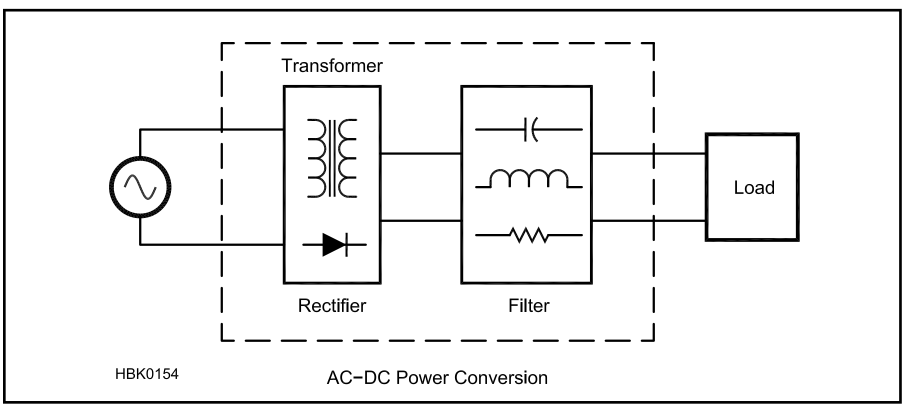


Fig 7.5 — Ac-dc power conversion with a rectifier and a filter.

As we will see in the rectifier circuit examples given in the next sections, sometimes the rectifier and filter functions will be separated into two distinct parts but very often the two will be integrated. This is particularly true for voltage and current multipliers as described in the sections on multipliers later in the chapter. Even when it appears that the rectifier and filter are separate elements, there will still be a strong interaction where the design and behavior of each part depends heavily on the other. For example the current waveforms in the rectifiers and the input source are functions of the load and filter characteristics. In turn the voltage waveform applied to the filter depends on the rectifier circuit and the input source voltages. To simplify the discussion we will treat the rectifier connections and the filters separately but always keeping in mind their interdependence.

The following rectifier-filter examples assume a conventional 60 Hz ac sine wave source, but these circuits are frequently used in switching converters at much higher frequencies and with square wave or quasi-square wave voltage and current waveforms. The component values may be different but the basic behavior will be very similar.

There are many different rectifier circuits or "connections" that may be used depending on the application. The following discussion provides an overview of some of the more common ones. The circuit diagrams use the symbol for a semiconductor diode, but the same circuits can be used with the older types of rectifiers that may be encountered in older equipment.

For each circuit we will show the voltage and current waveforms in the circuit for resistive, capacitive and inductive loads. The inductive and capacitive loads represent commonly used filters. We will be interested in the peak and average voltages as well as the RMS currents.

7.4.1 Half-Wave Rectifier

Fig 7.6 shows several examples of the half-wave rectifier circuit. It begins with a simple transformer with a resistive load (Fig 7.6A) and goes on to show how the output voltage and transformer current varies when a diode and filter elements are added.

Without the diode (Fig 7.6A) the output voltage (V_R) and current are just sine waves, and the RMS current in the transformer windings will be the same as the load (R) current.

Next, add a rectifier diode in series with the load (Fig 7.6B). During one half of the ac cycle, the rectifier conducts and there is current through the rectifier to the load. During the other half cycle, the rectifier is reverse-biased and there is no current (indicated by the broken line in Fig 7.6B) in R. The output voltage is pulsating dc, which is a combination of two components: an average dc value of $0.45 E_{RMS}$ (the voltage read by a dc voltmeter) and line-frequency ac ripple. The transformer secondary winding current is also pulsating dc. The power delivered to R is now 1/2 that for Fig 7.6A but the secondary RMS winding current in Fig 7.6B is still 0.707 times what it was in Fig 7.6A. For the same winding resistance, the winding loss, in proportion to the output power, is twice what it was in Fig 7.6A. This is an intrinsic limitation of the half-wave rectifier circuit — the RMS winding current is larger in proportion to the load power. In addition, the dc component of the secondary winding current may bias the transformer core toward saturation and increased core loss.

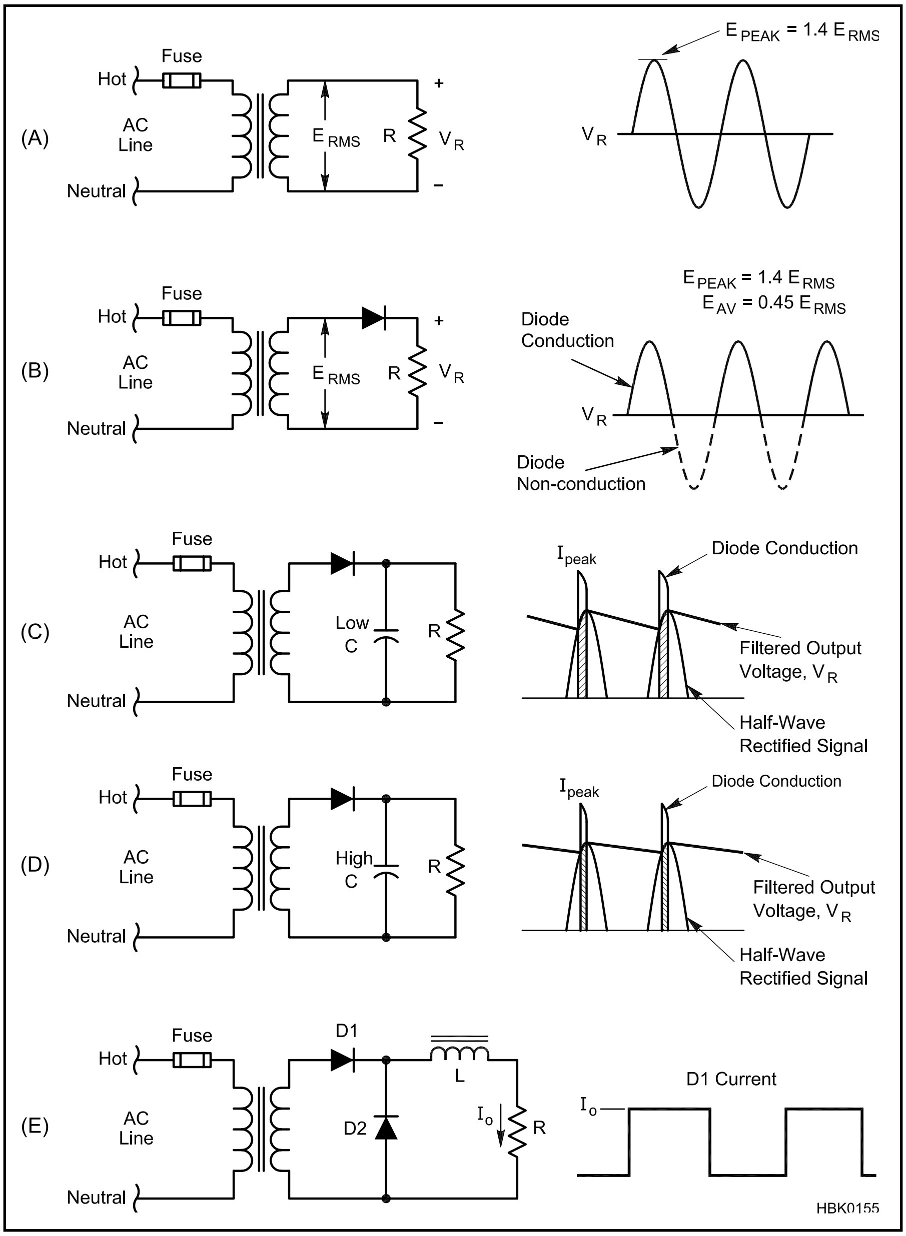


Fig 7.6 — Half-wave rectifier circuits. A illustrates the voltage waveform at the output without a rectifier. B represents the basic half-wave rectifier and the output waveform. C and D illustrate the impact of small and large filter capacitors on the output voltage and input current waveforms. E shows the effect of using an inductor filter with the half-wave rectifier. Note the addition of the shunt diode (D2) when using inductive filters with this rectifier connection.

A filter can be used to smooth out these variations and provide a higher average dc voltage from the circuit. Because the frequency of the pulses (the ripple frequency) is low (one pulse per cycle), considerable filtering is required to provide adequately smooth dc output. For this reason the circuit is usually limited to applications where the required current is small. Parts C, D and E in Fig 7.6 show some possible capacitive and inductive filters.

As shown in Fig 7.6C and D, when a capacitor is used for filtering the output dc voltage will approach

$$V_{pk} = \sqrt{2} \times E_{RMS} \qquad (1)$$

and the larger we make the filter capacitance, the smaller the ripple will be.

Unfortunately, as we make the filter capacitance larger, the diode, capacitor and transformer winding currents all become high-amplitude narrow pulses which will have a very high RMS value in proportion to the power level. These current pulses are also transmitted to the input line and inject currents at harmonics of the line frequency into the power source, which may result in interference to other equipment. Narrow high-amplitude current pulses are characteristic of capacitive-input filters in all rectifier connections when driven from voltage sources.

As shown in Fig 7.6E, it is possible to use an inductive filter instead, but a second diode (D2, sometimes called a *free-wheeling diode*) should be used. Without D2 the output voltage will get smaller as we increase the size of L to get better filtering, and the output voltage will vary greatly with load. By adding D2 we are free to make L large for small output ripple but still have reasonable voltage regulation. Currents in D1 and the winding will be approximately square waves, as indicated. This will reduce the line harmonic currents injected into the source but there will still be some.

Peak inverse voltage (*PIV*) is the maximum

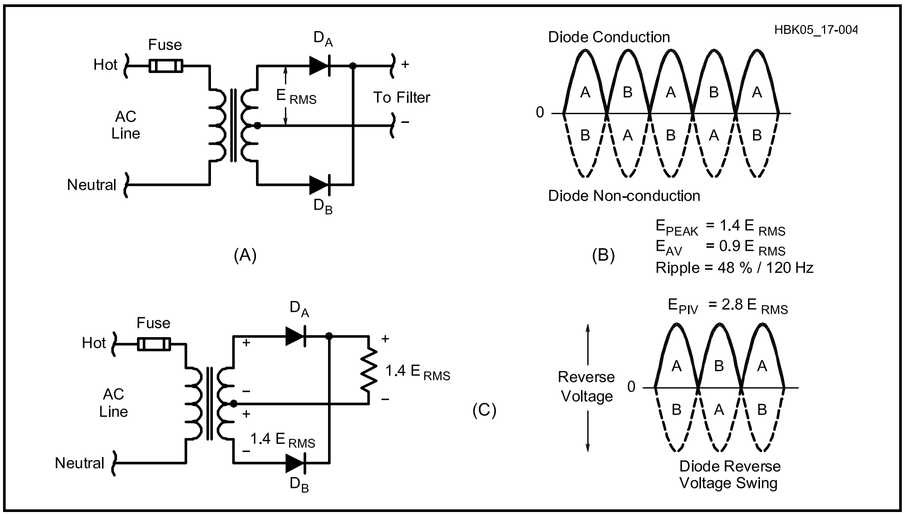


Fig 7.7 — Full-wave center-tap rectifier circuits. A illustrates the basic circuit. Diode conduction is shown at B with diodes A and B alternately conducting. The peak inverse voltage for each diode is 2.8 E_{RMS} as depicted at C.

voltage the rectifier must withstand when it isn't conducting. This varies with the load and rectifier connection. In the half-wave rectifier, with a resistive load the PIV is the peak ac voltage ($1.4 \times E_{RMS}$); with a capacitor filter and a load drawing little or no current, the PIV can rise to $2.8 \times E_{RMS}$.

7.4.2 Full-Wave Center-Tapped Rectifier

The full-wave center-tapped rectifier circuit is shown in **Fig 7.7**. It is essentially an arrangement where the outputs of two half-wave rectifiers are combined so that both halves of the ac cycle are used to deliver power to the output. A transformer with a center-tapped secondary is required.

The average output voltage of this circuit is $0.9 \times E_{RMS}$ of half the transformer secondary (the center-tap to one side); this is the maximum that can be obtained with a suitable choke-input filter. The peak output voltage is $1.4 \times E_{RMS}$ of half the transformer secondary; this is the maximum voltage that can be obtained from a capacitor-input filter.

As can be seen in Fig 7.7C, the PIV impressed on each diode is independent of the type of load at the output. This is because the peak inverse voltage condition occurs when diode D_A conducts and diode D_B is not conducting. The positive and negative voltage peaks occur at precisely the same time, a condition different from that in the half-wave circuit. As the cathodes of diodes D_A and D_B reach a positive peak (1.4 E_{RMS}), the anode of diode D_B is at a negative peak, also 1.4 E_{RMS}, but in the opposite direction. The total peak inverse voltage is therefore 2.8 E_{RMS}.

Fig 7.7C shows that the ripple frequency is twice that of the half-wave rectifier (two times the line frequency). Substantially less filtering is required because of the higher ripple frequency. Since the rectifiers work alternately, each handles half of the load current. The current rating of each rectifier need be only half the total current drawn from the supply.

The problem with dc bias in the transformer core associated with the half-wave connection is largely eliminated with this circuit and the RMS current in the primary winding will also be reduced.

7.4.3 Full-Wave Bridge Rectifier

Another commonly used rectifier circuit that does not require a center-tapped transformer is illustrated in **Fig 7.8**. In this arrangement, two rectifiers operate in series on each half of the cycle, one rectifier being in the lead supplying current to the load, the other being the current return lead. As shown in Figs 7.8A and B, when the top lead of the transformer secondary is positive with respect to the bottom lead, diodes D_A and D_C will conduct while diodes D_B and D_D are reverse-biased. On the next half cycle, when the top lead of the transformer is negative with respect to the bottom, diodes D_B and D_D will conduct while diodes D_A and D_C are reverse-biased.

The output voltage wave shape and ripple frequency are the same as for the full-wave center-tapped circuit. The average dc output voltage into a resistive load or choke-input filter is 0.9 times E_{RMS} delivered by the transformer secondary; with a capacitor filter and a light load, the maximum output voltage is 1.4 times the secondary E_{RMS} voltage.

Fig 7.8C shows the PIV to be 1.4 E_{RMS} for each diode which is half that of the full-wave center-tapped circuit for the same output voltage. When an alternate pair of diodes (such as D_A and D_C) is conducting, the other diodes are essentially connected in parallel (the conducting diodes are essentially short circuits) in a reverse-biased direction. The reverse stress is then 1.4 E_{RMS}. Each pair of diodes conducts on alternate

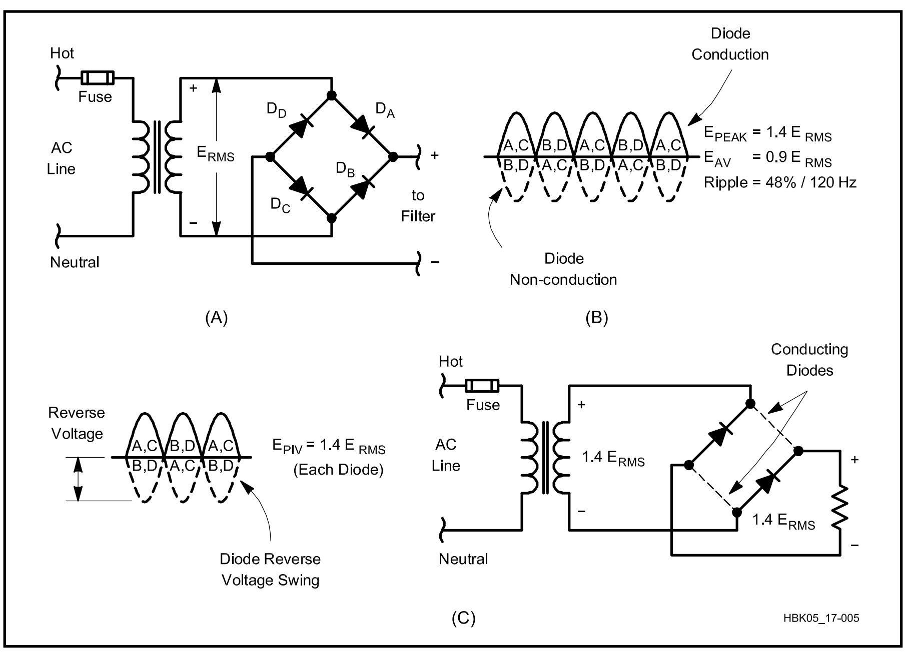


Fig 7.8 — Full-wave bridge rectifier circuits. The basic circuit is illustrated at A. Diode conduction and nonconduction times are shown at B. Diodes A and C conduct on one half of the input cycle, while diodes B and D conduct on the other. C displays the peak inverse voltage for one half cycle. Since this circuit reverse-biases two diodes essentially in parallel, 1.4 E_{RMS} is applied across each diode.

half cycles, with the full load current through each diode during its conducting half cycle. Since each diode is not conducting during the other half cycle the average diode current is one-half the total load current drawn from the supply.

Compared to the half-wave and full-wave center-tapped circuit, the full-wave bridge circuit further reduces the transformer RMS winding currents. In the case of a resistive load the winding currents are the same as when the resistive load is connected directly across the secondary. The RMS winding currents will still be higher when inductive and especially capacitive filters are used because of the pulsating nature of the diode and winding currents.

7.4.4 Comparison of Rectifier Circuits

Comparing the full-wave center-tapped and the full-wave bridge circuits, we can see that the center-tapped circuit has half the number of rectifiers as the bridge but these rectifiers have twice the PIV rating requirement of the bridge diodes. The diode current ratings are identical for the two circuits. The bridge makes better use of the transformer's secondary than the center-tapped rectifier, since the transformer's full winding supplies power during both half cycles, while each half of the center-tapped circuit's secondary provides power only during its positive half-cycle.

The full-wave center-tapped rectifier is typically used in high-current, low-voltage applications because only one diode conducts at a time. This reduces the loss associated with diode conduction. In the full-wave bridge circuit there are two diodes in series in conduction simultaneously, which leads to higher loss. The full-wave bridge circuit is typically used for higher output voltages where this is not a serious concern. The lower diode PIV and better utilization of the transformer windings makes this circuit very attractive for higher output voltages and higher powers typical of high voltage amplifier supplies.

Because of the disadvantages pointed out earlier, the half-wave circuit is rarely used in 60-Hz rectification except for bias supplies or other small loads. It does see considerable use, however, in high-frequency switching power supplies.

7.5 Voltage Multipliers

Other rectification circuits are sometimes useful, including *voltage multipliers*. These circuits function by the process of charging one or more capacitors in parallel on one half cycle of the ac waveform, and then connecting that capacitor or capacitors in series with the opposite polarity of the ac waveform on the alternate half cycle. In full-wave multipliers, this charging occurs during both half-cycles.

Voltage multipliers, particularly *voltage doublers*, find considerable use in high-voltage supplies. When a doubler is employed, the secondary winding of the power transformer need have only half the voltage that would be required for a bridge rectifier. This reduces voltage stress in the windings and decreases the transformer insulation requirements. This is not without cost, however, because the transformer-secondary *current* rating has to be correspondingly doubled for a given load current and charging of the capacitors leads to narrow high-RMS current waveforms in the transformer windings and the capacitors.

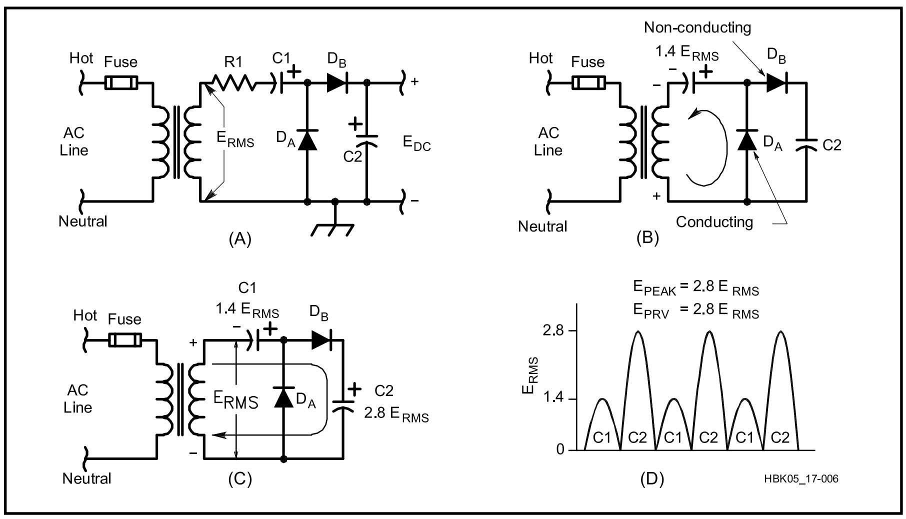


Fig 7.9 — Part A shows a half-wave voltage-doubler circuit. B displays how the first half cycle of input voltage charges C1. During the next half cycle (shown at C), capacitor C2 charges with the transformer secondary voltage plus that voltage stored in C1 from the previous half cycle. The arrows in parts B and C indicate the conventional current. D illustrates the levels to which each capacitor charges over several cycles.

7.5.1 Half-Wave Voltage Doubler

Fig 7.9 shows the circuit of a half-wave voltage doubler and illustrates the circuit operation. For clarity, assume the transformer voltage polarity at the moment the circuit is activated is that shown at Fig 7.9B. During the first negative half cycle, D_A conducts (D_B is in a nonconductive state), charging C1 to the peak rectified voltage (1.4 E_{RMS}). C1 is charged with the polarity shown in Fig 7.9B. During the positive half cycle of the secondary voltage, D_A is cut off and D_B conducts, charging capacitor C2. The amount of voltage delivered to C2 is the sum of the transformer peak secondary voltage plus the voltage stored in C1 (1.4 E_{RMS}). On the next negative half cycle, D_B is non-conducting and C2 will discharge into the load. If no load is connected across C2, the capacitors will remain charged — C1 to 1.4 E_{RMS} and C2 to 2.8 E_{RMS}. When a load is connected to the circuit output, the voltage across C2 drops during the negative half cycle and is recharged up to 2.8 E_{RMS} during the positive half cycle.

The output waveform across C2 resembles that of a half-wave rectifier circuit because C2 is pulsed once every cycle. Fig 7.9D illustrates the levels to which the two capacitors are charged throughout the cycle. In actual operation the capacitors will usually be large enough that they will discharge only partially, not all the way to zero as shown.

7.5.2 Full-Wave Voltage Doubler

Fig 7.10 shows the circuit of a full-wave voltage doubler and illustrates the circuit operation. During the positive half cycle of the transformer secondary voltage, as shown in Fig 7.10B, D_A conducts charging capacitor C1 to 1.4 E_{RMS}. D_B is not conducting at this time.

During the negative half cycle, as shown in Fig 7.10C, D_B conducts, charging capacitor C2 to 1.4 E_{RMS}, while D_A is non-conducting. The output voltage is the sum of the two capacitor voltages, which will be 2.8 E_{RMS} under no-load conditions. Fig 7.10D illustrates that each capacitor alternately receives a charge once per cycle. The effective filter capacitance is that of C1 and C2 in series, which is less than the capacitance of either C1 or C2 alone.

Resistors R1 and R2 in Fig 7.10A are used to limit the surge current through the rectifiers. Their values are based on the transformer voltage and the rectifier surge-current rating, since at the instant the power supply is turned on, the filter capacitors look like a short-circuited load. Provided the limiting resistors can withstand the surge current, their current-handling capacity is based on the maximum load current from the supply. Output voltages approaching twice the peak voltage of the transformer can be obtained with the voltage doubling circuit shown in Fig 7.10.

Fig 7.11 shows how the voltage depends upon the ratio of the series resistance to the load resistance, and the load resistance times the filter capacitance. The peak inverse voltage across each diode is 2.8 E_{RMS}. As indicated by the curves in Fig 7.11, the output voltage regulation of this doubler connection is not very good and it is not attractive for providing high voltages at high power levels.

There are better doubler connections for higher power applications, and two possibilities are shown in **Fig 7.12**. The connection in Fig 7.12A uses two bridge rectifiers in series with capacitive coupling between the ac terminals of the bridges. At the expense of more diodes, this connection will have much better output voltage regulation at higher power levels. Even better regulation can be achieved by using the connection shown in Fig 7.12B. In this example, two windings on the transformer are used. It is not essential that both windings have the same voltage, but both must be capable of providing the desired output current. In addition, the insulation of the upper winding must be adequate to accommodate the additional dc bias applied to it from the lower winding.

7.5.3 Voltage Tripler and Quadrupler

Fig 7.13A shows a voltage-tripling circuit. On one half of the ac cycle, C1 and C3 are charged to the source voltage through D1, D2 and D3. On the opposite half of the cycle, D2 conducts and C2 is charged to twice the source voltage, because it sees the transformer plus the charge in C1 as its source (D1 is cut off during this half cycle). At the same time, D3

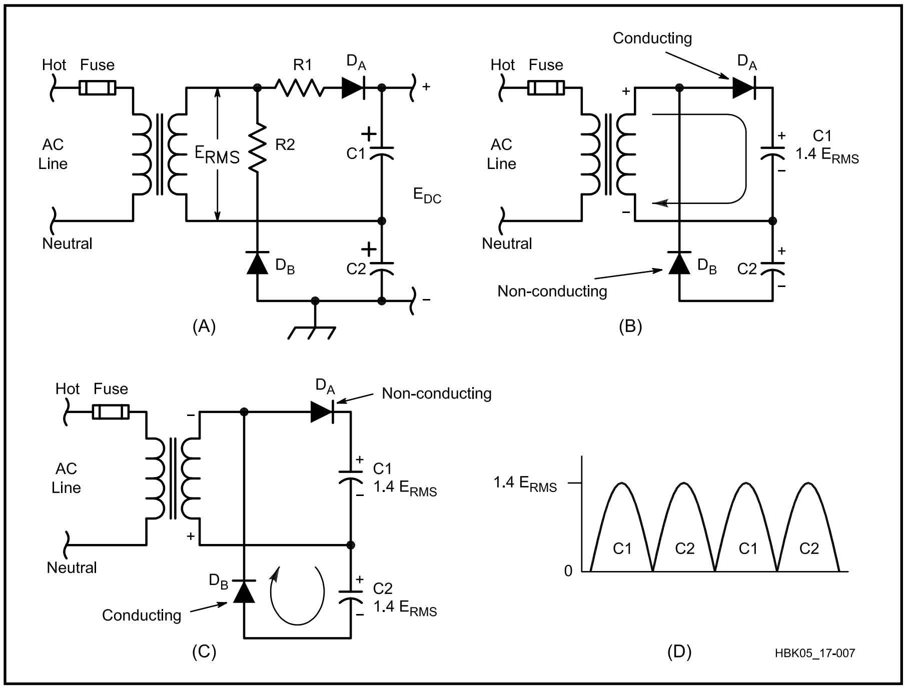


Fig 7.10 — Part A shows a full-wave voltage-doubler circuit. One-half cycle is shown at B and the next half cycle is shown at C. Each capacitor receives a charge during every input-voltage cycle. D illustrates how each capacitor is charged alternately.

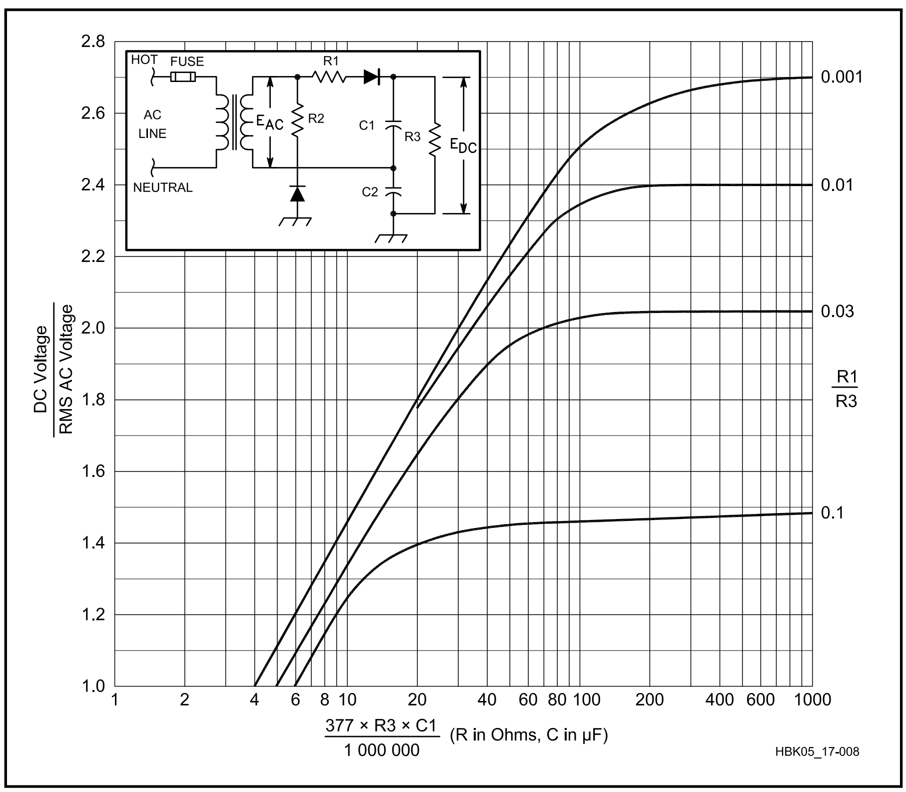
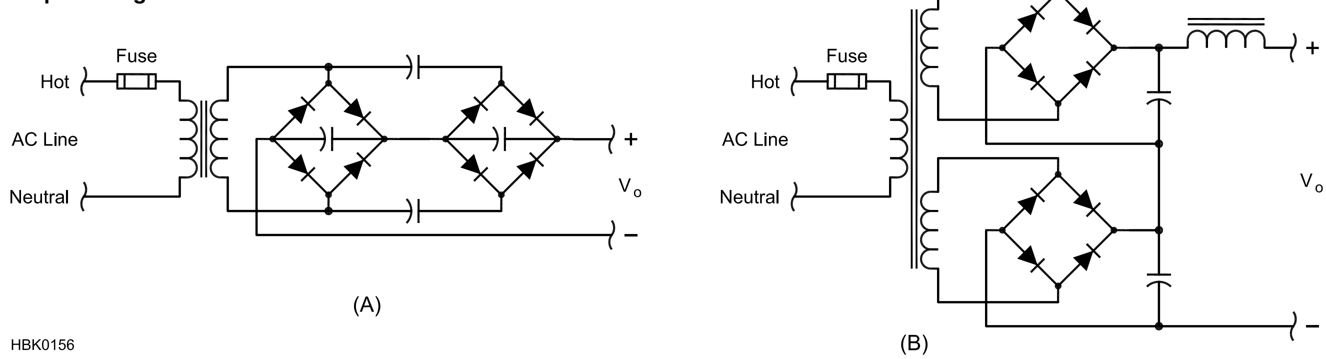


Fig 7.11 — DC output voltages from a full-wave voltage-doubler circuit as a function of the filter capacitances and load resistance. For the ratio R1 / R3 and for the R3 × C1 product, resistance is in ohms and capacitance is in microfarads. Equal resistance values for R1 and R2, and equal capacitance values for C1 and C2 are assumed. These curves are adapted from those published by Otto H. Schade in "Analysis of Rectifier Operation," *Proceedings of the I. R. E.*, July 1943.

Fig 7.12 — Voltage-doubler rectifier connections for higher power levels. A is a capacitor-coupled doubler which can be extended to more sections for a higher multiplying factor. The circuit in B uses multiple transformer windings to boost the output voltage.



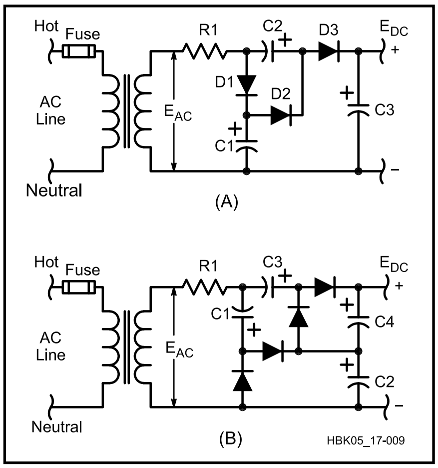


Fig 7.13 — Voltage-multiplying circuits with one side of the transformer secondary used as a common connection. A shows a voltage tripler and B shows a voltage quadrupler. Capacitances are typically 20 to 50 µF, depending on the output current demand. Capacitor dc ratings are related to E_{PEAK} (1.4 E_{RMS}):
C1 — Greater than E_{PEAK}
C2 — Greater than 2 E_{PEAK}
C3 — Greater than 3 E_{PEAK}
C4 — Greater than 2 E_{PEAK}

conducts, and with the transformer and the charge in C2 as the source, C3 is charged to three times the transformer voltage.

The voltage-quadrupling circuit of Fig 7.13B works in similar fashion. In either of the circuits of Fig 7.13, the output voltage will approach an exact multiple of the peak ac voltage when the output current drain is low and the capacitance values are large.

7.6 Current Multipliers

Just as there are voltage multiplier connections for high-voltage, low-current loads, there are current multiplier connections for low-voltage, high-current loads. An example of a current-doubler is given in **Fig 7.14A.**

To make the circuit operation easier to visualize, we can represent L1 and L2 as current sources (Fig 7.14B) which is a good approximation for steady-state operation. When terminal 1 of the secondary winding is positive with respect to terminal 2**,** diode D_A will be reverse-biased and therefore non-conducting. The current flows within the circuit are shown in Fig 7.14B. Note that all of the output current (I_o) flows through D_B but only half of I_o flows through the winding. At the cathode of D_B the current divides with half going to L2 and the other half to the transformer secondary. The output voltage will be one-half the voltage of the average winding voltage (0.45 E_{RMS}). This rectifier connection divides the voltage and multiplies the current! Because of the need for two inductors, this circuit is seldom used in line-frequency applications but it is very useful in high-frequency switching regulators with very low output voltages (<10 V) because it makes the secondary winding design easier and can improve circuit efficiency. At high frequencies, the inductors can be quite small.

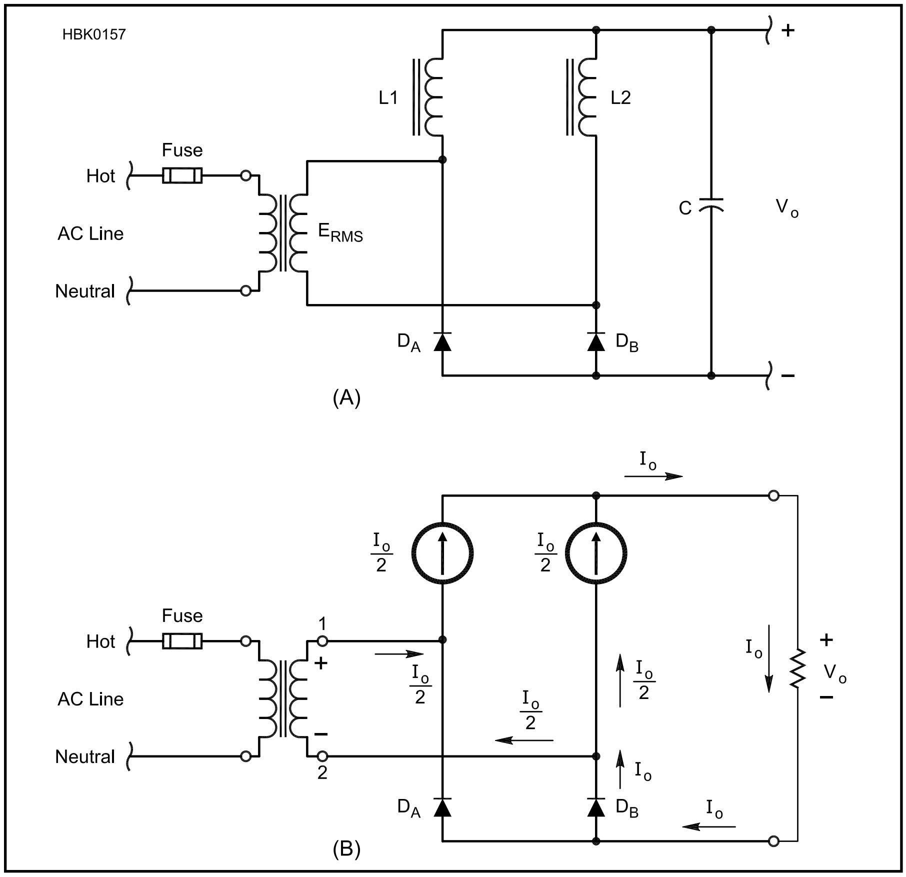


Fig 7.14 — A current doubler rectifier connection. A is the basic circuit. B illustrates current flow within the circuit.

7.7 Rectifier Types

Rectifiers have a long history beginning with mechanical rectifiers in the 1800s to today's abundant variety of semiconductor devices. While many different devices have been created for this purpose, they all have the characteristic that they block current flow in the reverse direction, withstanding substantial reverse voltage and allowing current flow in the forward direction with minimum voltage drop. The simplest rectifiers are diodes, but it is also possible to have three-terminal devices (such as a thyristor) that can be controlled to regulate the output dc in addition to providing rectification. It is also possible to use devices like MOSFETs as synchronous rectifiers with very low forward drop during conduction. This is typically done to improve efficiency for very low voltage outputs. The following is a brief description of several of the more common examples.

7.7.1 Vacuum Tube

Once the mainstay of the rectifier field, the vacuum-tube rectifier has largely been supplanted by the silicon diode, but it may be found in vintage equipment. Vacuum-tube rectifiers require filament power and are characterized by high forward voltage drop during conduction, which leads to inherently poor regulation of the output voltage. They are largely immune to ac line transients (also known as "spikes") that can destroy other rectifier types.

7.7.2 Mercury Vapor

The mercury-vapor rectifier was an improvement over the vacuum tube rectifier in that the electron stream from cathode to plate would ionize vaporized mercury in the tube and greatly reduce the forward voltage drop. Because of the lower voltage drop, the power dissipation was much lower for a given current and these tubes could carry relatively high currents. They were popular in transmitter high-voltage power supplies, some of which amateurs may still encounter.

Mercury rectifiers have to be treated with special care. When power is initially applied, the tube filament has to be turned on first to vaporize condensed mercury before high-voltage ac can be applied to the plate. This could take from one to two minutes. Also, if the tube was handled or the equipment transported, filament power would have to be applied for about a half hour to vaporize any mercury droplets that might have been shaken onto tube insulating surfaces. Mercury vapor rectifiers have mostly been replaced by silicon diodes.

7.7.3 Early Solid-State Rectifiers

Copper oxide and selenium rectifiers were the first of the solid-state rectifiers to find their way into commercial equipment. Voltage breakdown per rectifying junction was only a few volts for the copper oxide rectifiers and about 20 V for selenium. Multi-junction stacked versions of selenium rectifiers were used for higher voltage and had a relatively low forward voltage drop. Selenium rectifiers found their way into the plate supplies of test equipment and accessories that needed only a few tens of milliamperes of current at about a hundred volts. Examples include such as grid-dip meters, VTVMs and so forth.

Selenium rectifiers had a relatively low reverse resistance leading to high reverse leakage currents and were therefore inefficient. These may still be encountered in older equipment but they are usually replaced with silicon diodes. Very conveniently, a key indicator of a failed selenium rectifier is the smell of hydrogen selenide, similar to rotten eggs.

A word of caution is warranted when replacing older rectifiers such as vacuum tubes, mercury-vapor rectifiers, selenium or copper oxide. The voltage drop introduced by these rectifiers would have been included in the design of the equipment. When replaced with silicon diodes, the output voltage of the rectifier is very likely to be higher than with the original design. Care should be taken to make sure the higher voltage will not damage the filter elements or load circuits. Furthermore, mercury and selenium are toxic and should not be disposed of with regular household trash. Contact your local recycling agency for information about where to dispose of these devices.

7.7.4 Semiconductor Diodes

Rectifier diodes can be made from a number of different semiconductor materials such as germanium, silicon, silicon-carbide or gallium-arsenide, and no doubt other materials will appear in the future. The choice will depend on the application and as always cost is a factor.

Germanium diodes were the first of the solid-state semiconductor rectifiers. They have an extremely low forward voltage drop but are relatively temperature sensitive, having high reverse leakage currents at higher temperatures. They can be easily destroyed by overheating during soldering as well. Germanium diodes are no longer used as power rectifiers.

Today, silicon diodes are the primary choice for virtually all power rectifier applications. They are characterized by extremely high reverse resistance (low reverse leakage), forward drops of a volt or less and operation at junction temperatures up to 125 °C. Some multi-junction HV diodes will have forward drops of several volts, but that is still low compared to the voltage at which they are being used.

Many different types of silicon diodes are available for different applications. Silicon rectifiers fall into to two general categories: PN-junction diodes and Schottky barrier diodes (see the **Analog Basics** chapter). Schottky diodes are the usual choice for low output voltages (<20 V) where their low forward conduction drop is critical for efficiency. For higher voltages however, the high reverse leakage of Schottky diodes is not acceptable and PN-junction diodes are normally chosen.

For 50/60 Hz applications, diodes with reverse recovery times of a microsecond or even more are suitable and very economical. For switchmode converters and inverters that regularly operate at 25 kHz and higher frequencies, fast-recovery diodes are needed. These converters typically have waveform transitions of less than 1 µs within the circuit. MOSFET power transistors often have transitions of less than 100 ns.

During the switching transitions, previously conducting diodes see a reversal of current direction. This change tends to reverse-bias those diodes, and thereby put them into an open-circuit condition. Unfortunately, as explained in the **Analog Basics** chapter, solid-state rectifiers cannot be made to cease conduction instantaneously. As a result, when the opposing diodes in a bridge rectifier or full-wave rectifier become conductive at the time the converter switches states, the diodes being turned off will actually conduct in the reverse direction for a brief time. That effectively short circuits the converter for a period of time depending on the reverse recovery characteristics of the rectifiers. This characteristic can create high current transients that stress the switching transistors and lead to increased loss and electromagnetic interference. As the switching frequency increases, more of these transitions happen each second, and more power is lost because of diode cross-conduction.

These current transients and associated losses are reduced by using *fast recovery* diodes, which are specially doped diodes designed to minimize storage time. Diodes with recovery times of 50 ns or less are available.

7.7.5 Rectifier Strings or Stacks

DIODES IN SERIES

When the PIV rating of a single diode is not sufficient for the application, similar diodes may be used in series. (Two 500 PIV diodes in series will withstand 1000 PIV and so on.) There used to be a general recommendation to place a resistor across each diode in the string to equalize the PIV drops. With modern diodes, this practice is no longer necessary.

Modern silicon rectifier diodes are constructed to have an avalanche characteristic. Simply put, this means that the diffusion process is controlled so the diode will exhibit a Zener characteristic in the reverse-biased direction before destructive breakdown of the junction can occur. This provides a measure of safety for diodes in series. A diode will go into Zener conduction before it self-destructs. If other diodes in the chain have not reached their avalanche voltages, the current through the avalanched diode will be limited to the leakage current in the other diodes. This should normally be very low. For this reason, shunting resistors are generally not needed across diodes in series rectifier strings. In fact, shunt resistors can actually create problems because they can produce a low-impedance source of damaging current to any diode that may have reached avalanche potential.

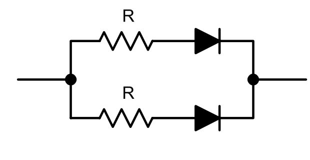


Fig 7.15 — Diodes can be connected in parallel to increase the current-handling capability of the circuit. Each diode should have a series current-equalizing resistor, with a value selected to have a drop of several tenths of a volt at the expected current.

DIODES IN PARALLEL

Diodes can be placed in parallel to increase current-handling capability. Equalizing resistors should be added as shown in **Fig 7.15**. Without the resistors, one diode may take most of the current. The resistors should be selected to have a drop of several tenths of a volt at the expected peak current. A disadvantage of this form of forced current sharing will be the increase in power loss because of the added resistors.

7.7.6 Rectifier Ratings versus Operating Stress

Power supplies designed for amateur equipment use silicon rectifiers almost exclusively. These rectifiers are available in a wide range of voltage and current ratings: PIV ratings of 600 V or more and current ratings as high as 400 A are available. At 1000 PIV, the current ratings may be several amperes. It is possible to stack several units in series for higher voltages. Stacks are available commercially that will handle peak inverse voltages up to 10 kV at a load current of 1 A or more.

7.7.7 Rectifier Protection

The discussion of rectifier circuits included the peak reverse voltage seen by the rectifiers in each circuit. You will need this information to select the voltage rating of the diodes in given application. It is normal good practice to not expose the diodes to more than 75% of their rated voltage for the worst case reverse voltage. This will probably be when operating at the highest input voltage but should also take into account transients that may occur.

The important specifications of a silicon diode are:

1. PIV — the peak inverse voltage.
2. I_0 — the average dc current rating.
3. I_{REP} — the peak repetitive forward current.
4. I_{SURGE} — a non-repetitive peak half-sine wave of 8.3 ms duration (one-half cycle of 60-Hz line frequency).
5. Switching speed or reverse recovery time.
6. Power dissipation and thermal resistance.

The first two specifications appear in most catalogs. I_{REP} and I_{SURGE} are not often specified in catalogs, but they are very important. Except in some switching regulator and capacitive filter circuits, rectifier current typically flows half the time — when it does conduct, the rectifier has to pass at least twice the average direct current. With a capacitor-input filter, the rectifier conducts much less than half the time. In this case, when it does conduct, it may pass as much as 10 to 20 times the average dc current, under certain conditions.

CURRENT INRUSH

When the supply is first turned on, the filter capacitors are discharged and act like a dead short. The result can be a very heavy current surge through the diode for at least one half-cycle and sometimes more. This current transient is called I_{SURGE}. The maximum surge current rating for a diode is usually specified for a duration of one-half cycle (at 60 Hz), or about 8.3 ms. Some form of surge protection is usually necessary to protect the diodes until the filter capacitors are nearly charged, unless the diodes used have a very high surge-current rating (several hundred amperes). If a manufacturer's data sheet is not available, an educated guess about a diode's capability can be made by using these rules of thumb for silicon diodes commonly used in Amateur Radio power supplies:

Rule 1. The maximum I_{REP} rating can be assumed to be approximately four times the maximum I_0 rating, where I_0 is the average dc current rating.

Rule 2. The maximum I_{SURGE} rating can be assumed to be approximately 12 times the maximum I_0 rating. This figure should provide a reasonable safety factor. Silicon rectifiers with 750-mA dc ratings, for example, seldom have 1-cycle surge ratings of less than 15 A; some are rated up to 35 A or more. From this you can see that the rectifier should be selected on the basis of I_{SURGE} and not on I_0 ratings.

Although you can sometimes rely on the resistance of the transformer windings to provide surge-current limiting, this is seldom adequate in high-voltage power supplies. Series resistors are often installed between the secondary and the rectifier strings or in the transformer's primary circuit, but these can be a deterrent to good voltage regulation.

One way to have good surge current limiting at turn-on without affecting voltage regulation during normal operation is to have a resistor in series with the input, along with a relay across the resistor that shorts it out after 50 ms or so. This kind of arrangement is particularly important in HV supplies.

VOLTAGE TRANSIENTS

Vacuum-tube rectifiers had little problem with voltage transients on the incoming power lines. The possibility of an internal arc was of little consequence, since the heat produced was of very short duration and had little effect on the massive plate and cathode structures.

Unfortunately, such is not the case with silicon diodes. Because of their low forward voltage drop, silicon diodes create very little heat with high forward current and therefore have tiny junction areas. However, conduction in the reverse direction beyond the normal reverse recovery time (reverse avalanching) can cause junction temperatures to rise extremely rapidly with the resulting destruction of the semiconductor junction.

To protect semiconductor rectifiers from voltage transients, special surge-absorption devices are available for connection across the incoming ac bus or transformer secondary. These devices operate in a fashion similar to a Zener diode; they conduct heavily when a specific voltage level is reached. Unlike Zener diodes, however, they have the ability to absorb very high transient energy levels without damage. With the clamping level set well above the normal operating voltage range for the rectifiers, these devices normally appear as open circuits and have no effect on the power-supply circuits. When a voltage transient occurs, however, these protection devices clamp the transient and thereby prevent destruction of the rectifiers.

Transient protectors are available in three basic varieties:

1. *Silicon Zener diodes* — large junction Zeners specifically made for this purpose and available as single junction for dc (unipolar) and back-to-back junctions for ac (bipolar). These silicon protectors are available under the trade name of TransZorb from General Semiconductor Corporation and are also made by other manufacturers. They have the best transient-suppressing characteristics of the three varieties mentioned here, but are expensive and have the least energy absorbing capability per dollar of the group.

2. *Varistors* — made of a composition metal-oxide material that breaks down at a certain voltage. Metal-oxide varistors, also known as MOVs, are cheap and easily obtained, but have a higher internal resistance, which allows a greater increase in clamped voltage than the Zener variety. Varistors can also degrade with successive transients within their rated power handling limits (this is not usually a problem in the ham shack where transients are few and replacement of the varistor is easily accomplished).

Varistors usually become short-circuited when they fail. Large energy dissipation can result in device explosion. Therefore, it is a good idea to include a fuse that limits the short-circuit current through the varistor, and to protect people and circuitry from debris.

3. *Gas tube* — similar in construction to the familiar neon bulb, but designed to limit conducting voltage rise under high transient currents. Gas tubes can usually withstand the highest transient energy levels of the group. Gas tubes suffer from an ionization time problem, however. A high voltage across the tube will not immediately cause conduction. The time required for the gas to ionize and clamp the transient is inversely proportional to the level of applied voltage in excess of the device ionization voltage. As a result, the gas tube will let a little of the transient through to the equipment before it activates.

In installations where reliable equipment operation is critical, the local power is poor and transients are a major problem, the usual practice is to use a combination of protectors. Such systems consist of a varistor or Zener protector, combined with a gas-tube device. Often there is an indicator light to warn when a surge has blown out the varistor. Operationally, the solid-state device clamps the surge immediately, with the beefy gas tube firing shortly thereafter to take most of the surge from the solid-state device.

HEAT

The junction of a diode is quite small, so it must operate at a high current density. The heat-handling capability is, therefore, quite small. Normally, this is not a prime consideration in high-voltage, low-current supplies. Use of high-current rectifiers at or near their maximum ratings (usually 2-A or larger, stud-mount rectifiers) requires some form of heat sinking. Frequently, mounting the rectifier on the main chassis — directly or with thin thermal insulating washers — will suffice.

When a rectifier is directly mounted on the heatsink it is good practice to use a thin layer of thermal grease between the diode and the heat sink to assure good heat conduction. Most modern insulating thermal washers do not require the use of grease, but the older mica and other washers may benefit from a *very thin layer* of grease. Thermal grease and heat conducting insulating washers and pads are standard products available from mail-order component sellers.

Large, high-current rectifiers often require special heat sinks to maintain a safe operating temperature. Forced-air cooling from a fan is sometimes used as a further aid. Safe case temperatures are usually given in the manufacturer's data sheets and should be observed if the maximum capabilities of the diode are to be realized. See the thermal design section in the chapter on **Electrical Fundamentals** for more information.

7.8 Power Filtering

Most loads will not tolerate the ripple (an ac component) of the pulsating dc from the rectifiers. Filters are required between the rectifier and the load to reduce the ripple to a low level. As pointed out earlier, some capacitances or inductances may be inherent in the rectifier connection, reducing the ripple amplitude. In most cases, however, additional filtering is required. The design of the filter depends to a large extent on the dc voltage output, the desired voltage regulation of the power supply and the maximum load current. Power-supply filters are low-pass devices using series inductors and/or shunt capacitors.

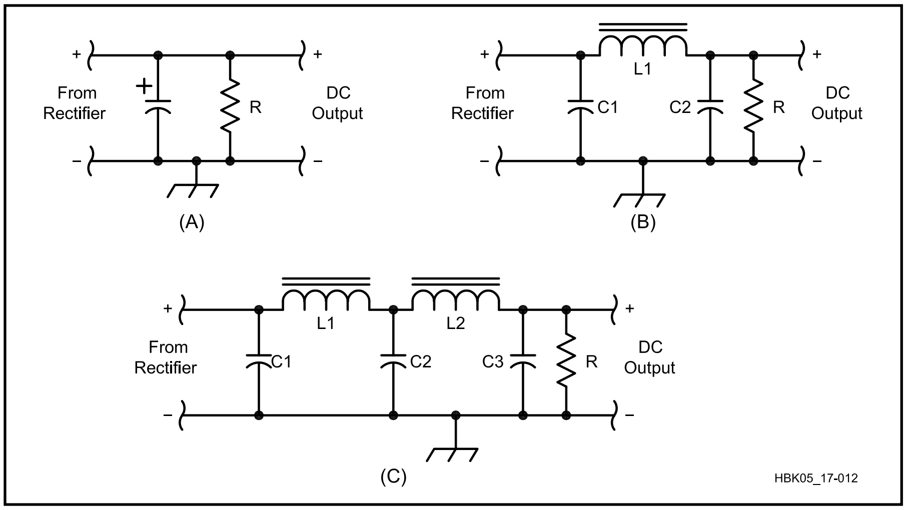


Fig 7.16 — Capacitor-input filter circuits. At A is a simple capacitor filter. B and C are single- and double-section filters, respectively.

7.8.1 Load Resistance

In discussing the performance of power-supply filters, it is sometimes convenient to characterize the load connected to the output as a resistance. This *load resistance* is equal to the output voltage divided by the total load current, including the current drawn by the bleeder resistor.

7.8.2 Voltage Regulation

In an unregulated supply, the output voltage usually decreases as more current is drawn. This happens not only because of increased voltage drops in the transformer and filter chokes, but also because the output voltage at light loads tends to soar to the peak value of the transformer voltage as a result of charging the first capacitor. Proper filter design can reduce this effect. The change in output voltage with load is called the *voltage regulation* and is expressed as a percentage.

$$\text{Percent Regulation} = \frac{(E1 - E2)}{E2} \times 100\% \qquad (2)$$

where

E1 = the no-load voltage

E2 = the full-load voltage.

A steady load, such as that represented by a receiver, speech amplifier or unkeyed stages of a transmitter, does not require good (low) regulation as long as the proper voltage is obtained under load conditions. The filter capacitors must have a voltage rating safe for the highest value to which the voltage will rise when the external load is removed.

Typically the output voltage will display a larger change with long-duration changes in load resistance than with short transient changes. The reason for this is that transient load currents are supplied from energy stored in the output capacitance. The regulation with long-term changes is often called the *static regulation*, to distinguish it from the *dynamic regulation* (transient load changes). A load that varies at a syllabic or keyed rate, as represented by some audio and RF amplifiers, usually requires good dynamic regulation (<15%) if distortion products are to be held to a low level. The dynamic regulation of a power supply can be improved by increasing the output capacitance.

When essentially constant voltage is required, regardless of current variation (for stabilizing an oscillator, for example), special voltage regulating circuits described later in this chapter are used.

7.8.3 Bleeder Resistors

A *bleeder resistor* is a resistance (R) connected across the output terminals of the power supply as shown in **Fig 7.16A**. Its functions are to discharge the filter capacitors as a safety measure when the power is turned off and to improve voltage regulation by providing a minimum load resistance. When voltage regulation is not of importance, the resistance may be as high as 100 Ω per volt of output voltage. The resistance value to be used for voltage-regulating purposes is discussed in later sections. From the consideration of safety, the power rating of bleeder resistors should be as conservative as possible — having a burned-out bleeder resistor is dangerous!

7.8.4 Ripple Frequency and Voltage

The ripple at the output of the rectifier is an alternating current superimposed on a steady direct current. From this viewpoint, the filter may be considered to consist off: 1) shunt capacitors that short circuit the ac component while not interfering with the flow of the dc component; and/or 2) series chokes that readily pass dc but will impede the ac component.

The effectiveness of the filter can be expressed in terms of percent ripple, which is the ratio of the RMS value of the ripple to the dc value in terms of percentage.

$$\text{Percent Ripple (RMS)} = \frac{E1}{E2} \times 100\% \qquad (3)$$

where

E1 = the RMS value of ripple voltage

E2 = the steady dc voltage.

Any frequency multiplier or amplifier supply in a CW transmitter should have less than 5% ripple. A linear amplifier can tolerate about 3% ripple on the plate voltage. Bias supplies for linear amplifiers should have less than 1% ripple. VFOs, speech amplifiers and receivers may require no greater than 0.01% ripple.

Ripple frequency refers to the frequency of the pulsations in the rectifier output waveform — the number of pulsations per second. The ripple frequency of half-wave rectifiers is the same as the line-supply frequency — 60 Hz with a 60-Hz supply. Since the output pulses are doubled with a full-wave rectifier, the ripple frequency is doubled — to 120 Hz with a 60-Hz supply.

The amount of filtering (values of inductance and capacitance) required to give adequate smoothing depends on the ripple frequency. More filtering is required as the ripple frequency is reduced. This is why the filters used for line-frequency rectification are much larger than those used in switchmode converters where the ripple frequency is often in the hundreds of kHz.

7.8.5 Capacitor-Input Filters

Typical capacitor-input filter systems are shown in Fig 7.16. The ripple can be reduced by making C1 larger, but that can lead to very large capacitances and high inrush currents at turn-on. Better ripple reduction will be obtained when moderate values for C1 are employed and LC sections are added as shown in Fig 7.16B and C.

INPUT VERSUS OUTPUT VOLTAGE

The average output voltage of a capacitor-input filter is generally poorly regulated with load-current variations. As shown earlier (Fig 7.6) the rectifier diodes conduct for only a small portion of the ac cycle to charge the filter capacitor to the peak value of the ac waveform. When the instantaneous voltage of the ac passes its peak, the diode ceases to conduct. This forces the capacitor to support the load current until the ac voltage on the opposing diode in the bridge or full wave rectifier is high enough to pick up the load and recharge the capacitor. For this reason, the peak diode currents are usually quite high.

Since the cyclic peak voltage of the capacitor-filter output is determined by the peak of the input ac waveform, the minimum voltage and, therefore, the ripple amplitude, is determined by the amount of voltage discharge, or "droop," occurring in the capacitor while it is discharging and supporting the load. Obviously, the higher the load current, the proportionately greater the discharge, and therefore the lower the average output.

There is an easy way to approximate the peak-to-peak ripple for a certain capacitor and load by assuming a constant load current. We can calculate the droop in the capacitor by using the relationship:

$$C \times E = I \times t \qquad (4)$$

where

C = the capacitance in microfarads

E = the voltage droop, or peak-to-peak ripple voltage

I = the load current in milliamperes

t = the length of time in ms per cycle during which the rectifiers are not conducting, during which the filter capacitor must support the load current. For 60-Hz, full-wave rectifiers, t is about 7.5 ms.

As an example, let's assume that we need to determine the peak-to-peak ripple voltage at the dc output of a full-wave rectifier/ filter combination that produces 13.8 V dc and supplies a transceiver drawing 2.0 A. The filter capacitor in the power supply is 5000 µF. Using the above relationship:

$$C \times E = I \times t \qquad (5)$$

$$5000\ \mu F \times E = 2000\ mA \times 7.5\ ms$$

$$E = \frac{2000\ mA \times 7.5\ ms}{5000\ \mu F} = 3\ V_{P\text{-}P}$$

Obviously, this is too much ripple. A capacitor value of about 20,000 µF would be better suited for this application. If a linear regulator is used after this rectifier/filter combination, then it is possible to trade off higher ripple voltage against high power dissipation in the regulator. A properly designed linear regulator can reduce the ripple amplitude to a very small value.

7.8.6 Choke-Input Filters

Choke-input filters provide the benefits of greatly improved output voltage stability over varying loads and low peak-current surges in the rectifiers. On the negative side, the output voltage will be lower than that for a capacitor-input filter.

In line-frequency power supplies, choke-input filters are less popular than they once were. This change came about in part because of the high surge current capability of silicon rectifiers, but more importantly because size, weight and cost are reduced when large filter chokes are eliminated. However, choke input filters are frequently used in high-frequency switchmode converters where the chokes will be much smaller.

As long as the inductance of the choke is large enough to maintain a continuous current over the complete cycle of the input ac waveform, the filter output voltage will be the average value of the rectified output. The average dc value of a full-wave rectified sine wave is 0.637 times its peak voltage. Since the RMS value is 0.707 times the peak, the output of the choke input filter will be (0.637 / 0.707), or 0.9 times the RMS ac voltage.

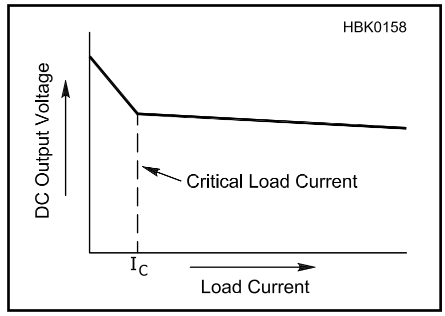


Fig 7.17 — Inductive filter output voltage regulation as a function of load current. The transition from capacitive peak charging to inductive averaging occurs at the critical load current, I_C.

As shown in **Fig 7.17**, there is a minimum or "critical" load current below which the choke does not provide the necessary filtering. For light loads, there may not be enough energy stored in the choke during the input waveform crest to allow continuous current over the full cycle. When this happens, the filter output voltage will rise as the filter assumes more and more of the characteristics of a capacitor-input filter. One purpose of the bleeder resistor is to keep the minimum load current above the critical value.

The value for the critical (or minimum) inductance for a given maximum value of load resistance in a single phase, full-wave

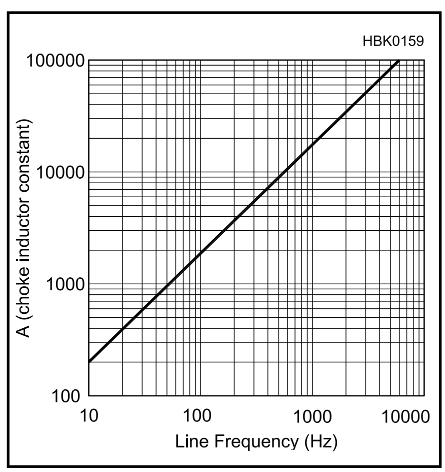


Fig 7.18 — The choke inductor constant, A, is used to solve equation 6.

rectifier with a sine wave source voltage can be approximated from:

$$L_c = \frac{R}{A} \qquad (6)$$

where

R = the maximum load resistance

A = a constant obtained from **Fig 7.18**, derived from the frequency of the input current (see Reference 1).

Low values for minimum load current (high minimum load R) can lead to large values for L_c which may not be practical. Standard filter inductors typically have a relatively constant value for L as the dc current is varied but it is possible to use a *swinging choke* instead. This is an inductor which has a high inductance at low currents and much lower inductance at high currents. Using a swinging choke will usually result in a much smaller filter choke and/or better output regulation.

7.9 Power Supply Regulation

The output of a rectifier/filter system may be usable for some electronic equipment, but for today's transceivers and accessories, further measures may be necessary to provide power sufficiently clean and stable for their needs. Voltage regulators are often used to provide this additional level of conditioning.

Rectifier/filter circuits by themselves are unable to protect the equipment from the problems associated with input-power-line fluctuations, load-current variations and residual ripple voltages. Regulators can eliminate these problems, but not without costs in circuit complexity and power-conversion efficiency.

7.9.1 Zener Diodes

A Zener diode (developed by Dr Clarence Zener) can be used to maintain the voltage applied to a circuit at a practically constant value, regardless of the voltage regulation of the power supply or variations in load current.

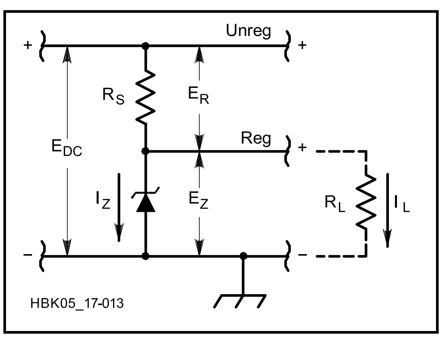


Fig 7.19 — Zener-diode voltage regulation. The voltage from a negative supply may be regulated by reversing the power-supply connections and the diode polarity.

The typical circuit is shown in **Fig 7.19**. Note that the cathode side of the diode is connected to the positive side of the supply.

Zener diodes are available in a wide variety of voltages and power ratings. The voltages range from less than 2 V to a few hundred volts, while the power ratings (power the diode can dissipate) run from less than 0.25 W to 50 W. The ability of the Zener diode to stabilize a voltage depends on the diode's conducting impedance. This can be as low as 1 Ω or less in a low-voltage, high-power diode or as high as 1000 Ω in a high-voltage, low-power diode.

The circuit in Fig 7.19 is a *shunt* regulator in that it "shunts" current through a controlling device (the Zener diode) to maintain a constant output voltage. To design a Zener shunt regulator, you must know the minimum and maximum input voltage (E_{DC}); the output voltage, which is equal to the Zener diode voltage (E_Z); and the minimum and maximum load current (I_L) through R_L. If the input voltage is variable, you must specify the maximum and minimum values for E_{DC}. As a rule of thumb, the current through the Zener should be $I_L/10$ for good regulation and must be greater than the $I_{Z(min)}$ at which the Zener diode maintains its constant voltage drop. Once these quantities are known the series resistance, R_S, can be determined:

$$R_S = \frac{E_{DC(min)} - E_Z}{1.1\, I_{L(max)}} \qquad (7)$$

The power dissipation of the Zener diode, P_D, is

$$P_{DZ} = \left[\frac{E_{DC(max)} - E_Z}{R_S} - I_{L(min)} \right] E_Z \qquad (8)$$

and of the series resistor, R_S,

$$P_{DR} = \frac{\left(E_{DC(max)} - E_Z\right)^2}{R_S} \qquad (9)$$

It is good practice to provide a five times rated power dissipation safety margin for both the series resistor and the Zener diode. This avoids heating in the Zener and the resulting drift in voltage. High-power Zener diodes (10 W dissipation or more) will require heat-sinking as discussed in the section on Managing Heat in the **Electrical Fundamentals** chapter.

7.9.2 Linear Regulators

Linear regulators come in two varieties, *series* and *shunt*, as shown in **Fig 7.20**. The shunt regulator is simply an electronic (also called "active") version of the Zener diode. For the most part, the active shunt regulator (Fig 7.20B) is rarely used since the series regulator is a superior choice for most applications.

The series regulator (Fig 7.20A) consists of a stable voltage reference, which is usually established by a Zener diode, a transistor in series between the power source and the load (called a *series pass transistor*), and an error amplifier. In critical applications a temperature-compensated reference diode would be used instead of the Zener diode.

The output voltage is sampled by the error amplifier, which compares the output (usually scaled down by a voltage divider) to the reference. If the scaled-down output voltage becomes higher than the reference voltage, the error amplifier reduces the drive current to the pass transistor, thereby allowing the

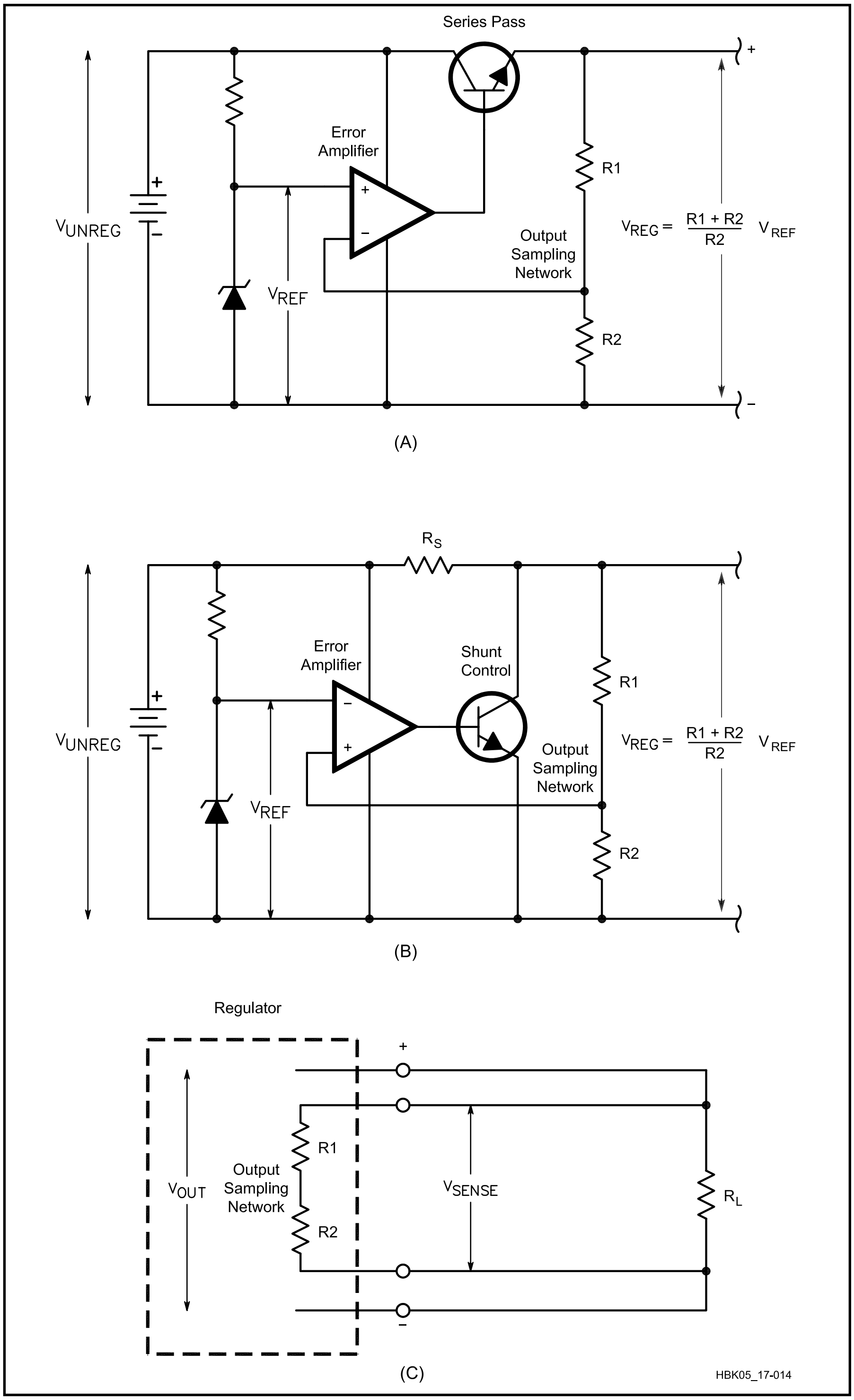


Fig 7.20 — Linear electronic voltage-regulator circuits. In these diagrams, batteries represent the unregulated input-voltage source. A transformer, rectifier and filter would serve this function in most applications. Part A shows a series regulator and Part B shows a shunt regulator. Part C shows how remote sensing overcomes poor load regulation caused by the I R drop in the connecting wires by bringing them inside the feedback loop.

output voltage to drop slightly. Conversely, if the load pulls the output voltage below the desired value, the amplifier drives the pass transistor into increased conduction.

The "stiffness" or tightness of regulation of a linear regulator depends on the gain of the error amplifier and the ratio of the output scaling resistors. In any regulator, the output is cleanest and regulation stiffest at the point where the sampling network or error amplifier is connected. If heavy load current is drawn through long leads, the voltage drop can degrade the regulation at the load. To combat this effect, the feedback connection to the error amplifier can be made directly to the load. This technique, called *remote sensing*, moves the point of best regulation to the load by bringing the connecting loads inside the feedback loop. This is shown in Fig 7.20C.

INPUT VERSUS OUTPUT VOLTAGE

In a series regulator, the pass-transistor power dissipation is directly proportional to the load current and input/output voltage differential. The series pass element can be located in either leg of the supply. Either NPN or PNP devices can be used, depending on the ground polarity of the unregulated input.

The differential between the input and output voltages is a design tradeoff. If the input voltage from the rectifiers and filter is only slightly higher than the required output voltage, there will be minimal voltage drop across the series pass transistor. A small drop results in minimal thermal dissipation and high power-supply efficiency. The supply will have less capability to provide regulated power in the event of power line brownout and other reduced line voltage conditions, however. Conversely, a higher input voltage will provide operation over a wider range of input voltage, but at the expense of increased heat dissipation.

7.9.3 Linear Regulator Pass Transistors

DARLINGTON PAIRS

A simple Zener-diode reference or IC op-amp error amplifier may not be able to source enough current to a pass transistor that must conduct heavy load current. The Darlington configuration of **Fig 7.21A** multiplies the pass-transistor beta, thereby extending the control range of the error amplifier. If the Darlington arrangement is implemented with discrete transistors, resistors across the base-emitter junctions may be necessary to prevent collector-to-base leakage currents in Q1 from being amplified and turning on the transistor pair. These resistors are contained in the envelope of a monolithic Darlington device.

When a single pass transistor is not available to handle the current required from a regulator, the current-handling capability may be increased by connecting two or more pass transistors in parallel. The circuit of Fig 7.21B shows the method of connecting these pass transistors. The resistances in the emitter leads of each transistor are necessary to equalize the currents.

TRANSISTOR RATINGS

When bipolar (NPN, PNP) power transistors are used in applications in which they are called upon to handle power on a continuous

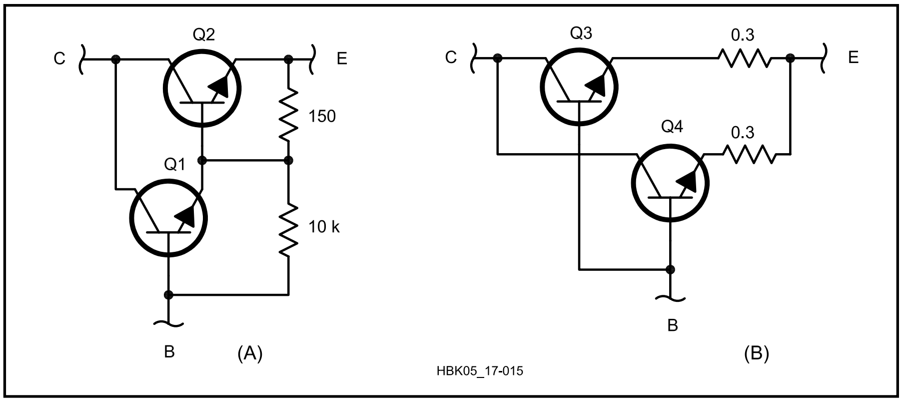


Fig 7.21 — At A, a Darlington-connected transistor pair for use as the pass element in a series-regulating circuit. At B, the method of connecting two or more transistors in parallel for high-current output. Resistances are in ohms. The circuit at A may be used for load currents from 100 mA to 5 A, and the one at B may be used for currents from 6 A to 10 A.
Q1 — Motorola MJE 340 or equivalent
Q2-Q4 — Power transistor such as 2N3055 or 2N3772

basis, rather than switching, there are four parameters that must be examined to see if any maximum limits are being exceeded. Operation of the transistor outside these limits can easily result in device failure, and these parameters must be considered during the design process.

The four limits are maximum collector current (I_C), maximum collector-emitter voltage (V_{CEO}), maximum power and *second breakdown* (I_{SB}). All four of these parameters are graphically shown on the transistor's data sheet on what is known as a *safe operating area (SOA)* graph. (see **Fig 7.22**) The first three of these limits are usually also listed prominently with the other device information, but it is often the fourth parameter — secondary breakdown — that is responsible for the "sudden death" of the power transistor after an extended operating period.

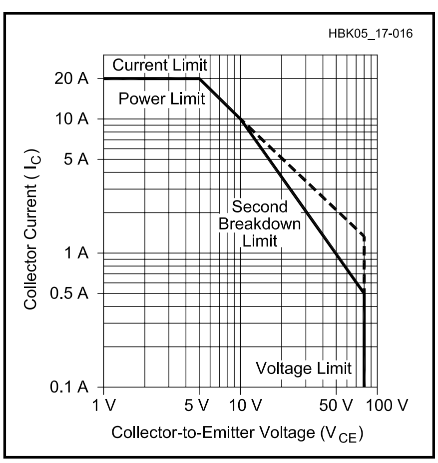


Fig 7.22 — Typical graph of the safe operating area (SOAR) of a transistor. See text for details. Safe operating conditions for specific devices may be quite different from those shown here.

The maximum current limit of the transistor ($I_{C\ MAX}$) is usually the current limit for fusing of the bond wire connected to the emitter, rather than anything pertaining to the transistor chip itself. When this limit is exceeded, the bond wire can melt and open circuit the emitter. On the operating curve, this limit is shown as a horizontal line extending out from the Y-axis and ending at the voltage point where the constant power limit begins.

The maximum collector-emitter voltage limit of the transistor ($V_{CE\ MAX}$) is the point at which the transistor junctions can no longer withstand the voltage between collector and emitter.

With increasing collector-emitter voltage drop at maximum collector current, a point is reached where the power in the transistor will cause the junction temperature to rise to a level where the device leakage current rapidly increases and begins to dominate. In this region, the product of the voltage drop and the current would be constant and represent the maximum power (P_t) rating for the transistor; that is, as the voltage drop continues to increase, the collector current must decrease to maintain the power dissipation at a constant value.

With most transistors rated for higher voltages, a point is reached on the constant power portion of the curve whereby, with further increased voltage drop, the maximum power rating is *not* constant, but decreases as the collector to emitter voltage increases. This decrease in power handling capability continues until the maximum voltage limit is reached.

This special region is known as the *forward bias second breakdown* (FBSB) area. Reduction in the transistor's power handling capability is caused by localized heating in certain small areas of the transistor junction ("hot spots"), rather than a uniform distribution of power dissipation over the entire surface of the device.

The region of operating conditions contained within these curves is called the safe operating area, or SOA. If the transistor is always operated within these limits, it should provide reliable and continuous service for a long time.

MOSFET TRANSISTORS

The bipolar junction transistor (BJT) is rapidly being replaced by the MOSFET in new power supply designs because MOSFETs are easier to drive. The N-channel MOSFET (equivalent to the NPN bipolar) is more popular than the P-channel for pass transistor applications.

There are some considerations that should be observed when using a MOSFET as a linear regulator series pass transistor. Several volts of gate drive are needed to start conduction of the device, as opposed to less than 1 V for the BJT. MOSFETs are inherently very-high-frequency devices and will readily oscillate with stray-circuit capacitances. To prevent oscillation in the transistor and surrounding circuits, it is common practice to insert a small resistor of about 100 Ω directly in series with the gate of the series-pass transistor to reduce the gate circuit Q.

OVERCURRENT PROTECTION

Damage to a pass transistor can occur when the load current exceeds the safe amount. **Fig 7.23A** illustrates a simple current-limiter circuit that will protect Q1. All of the load current is routed through R1. A voltage difference will exist across R1; the value will depend on the exact load current at a given time. When the load current exceeds a predetermined safe value, the voltage drop across R1 will forward-bias Q2 and cause it to conduct. Because Q2 is a silicon transistor, the voltage drop across R1 must exceed 0.6 V to turn Q2 on. This being the case, R1 is chosen for a value that provides a drop of 0.6 V when the maximum safe load current is drawn. In this instance, the drop will be 0.6 V when I_L reaches 0.5 A. R2 protects the base-emitter junction of Q2 from current transients, or from destruction in the event Q1 fails under short-circuit conditions.

When Q2 turns on, some of the current through R_S flows through Q2, thereby depriving Q1 of some of its base current. This action, depending upon the amount of Q1 base current at a precise moment, cuts off Q1 conduction to some degree, thus limiting the current through it.

FOLDBACK CURRENT LIMITING

Under short-circuit conditions, a constant-

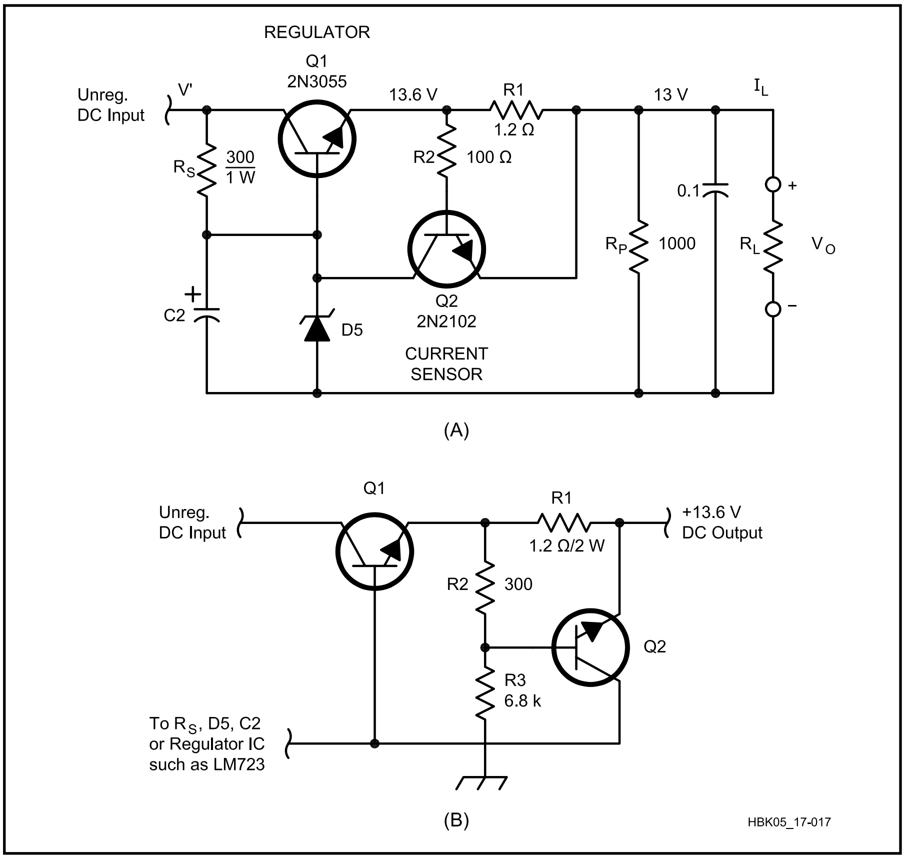


Fig 7.23 — Overload protection for a regulated supply can be implemented by addition of a current-overload-protective circuit, as shown at A. At B, the circuit has been modified to employ current-foldback limiting.

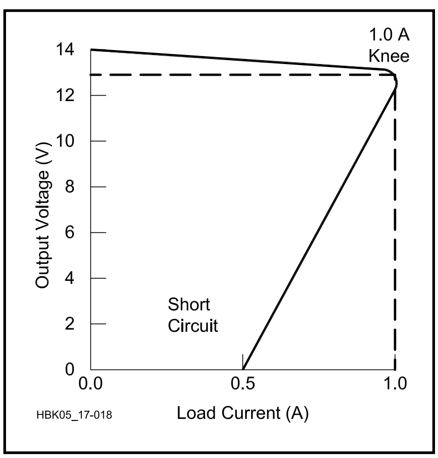


Fig 7.24 — The 1-A regulator shown in Fig 7.23B will fold back to 0.5 A under short-circuit conditions. See text.

current type current limiter must still withstand the full source voltage and limited short-circuit current simultaneously, which can impose a very high power dissipation or second breakdown stress on the series pass transistor. For example, a 12-V regulator with current limiting set for 10 A and having a source of 16 V will have a dissipation of 40 W [(16 V – 12 V) × 10 A] at the point of current limiting (knee). But its dissipation will rise to 160 W under short-circuit conditions (16 V × 10 A).

A modification of the limiter circuit can cause the regulated output current to decrease with decreasing load resistance beyond the over-current knee. With the output shorted, the output current is only a fraction of the knee current value, which protects the series-pass transistor from excessive dissipation and possible failure. Using the previous example of the 12-V, 10-A regulator, if the short-circuit current is designed to be 3 A (the knee is still 10 A), the transistor dissipation with a short circuit will be only 16 V × 3 A = 48 W.

Fig 7.23B shows how the current-limiter example given in the previous section would be modified to incorporate foldback limiting. The divider string formed by R2 and R3 provides a negative bias to the base of Q2, which prevents Q2 from turning on until this bias is overcome by the drop in R1 caused by load current. Since this hold-off bias decreases as the output voltage drops, Q2 becomes more sensitive to current through R1 with decreasing output voltage. See **Fig 7.24**.

The circuit is designed by first calculating the value of R1 for short-circuit current. For example, if 0.5 A is chosen, the value for R1 is simply 0.6 V / 0.5 A = 1.2 Ω (with the output shorted, the amount of hold-off bias supplied by R2 and R3 is very small and can be neglected). The knee current is then chosen. For this example, the selected value will be 1.0 A. The divider string is then proportioned to provide a base voltage at the knee that is just sufficient to turn on Q2 (a value of 13.6 V for 13.0 V output). With 1.0 A flowing through R1, the voltage across the divider will be 14.2 V. The voltage dropped by R2 must then be 14.2 V –13.6 V, or 0.6 V. Choosing a divider current of 2 mA, the value of R2 is then 0.6 V / 0.002 A = 300 Ω. R3 is calculated to be 13.6 V / 0.002 A = 6800 Ω.

7.9.4 Three-Terminal Voltage Regulators

The modern trend in regulators is toward the use of three-terminal devices commonly referred to as *three-terminal regulators*. Inside each regulator is a voltage reference, a high-gain error amplifier, temperature-compensated voltage sensing resistors and a pass element. Many currently available units have thermal shut-down, overvoltage protection and current foldback, making them virtually destruction-proof. It is easy to see why regulators of this sort are so popular when you consider the low price and the number of individual components they can replace.

Three-terminal regulators have connections for unregulated dc input, regulated dc output and ground, and they are available in a wide range of voltage and current ratings. Fixed-voltage regulators are available with output ratings in most common values between 5 and 28 V. Other families include devices that can be adjusted from 1.25 to 50 V.

The regulators are available in several different package styles, depending on current ratings. Low-current (100 mA) devices frequently use the plastic TO-92 and DIP-style cases. TO-220 packages are popular in the 1.5-A range, and TO-3 cases house the larger 3-A and 5-A devices. They are available in surface mount packages too.

Three-terminal regulators are available as positive or negative types. In most cases, a positive regulator is used to regulate a positive voltage and a negative regulator a negative voltage. Depending on the system ground requirements, however, each regulator type may be used to regulate the “opposite” voltage.

Fig 7.25A and B illustrate how the regulators are used in the conventional mode. Several regulators can be used with a common-input supply to deliver several voltages with a common ground. Negative regulators may be used in the same manner. If no other common supplies operate from the input

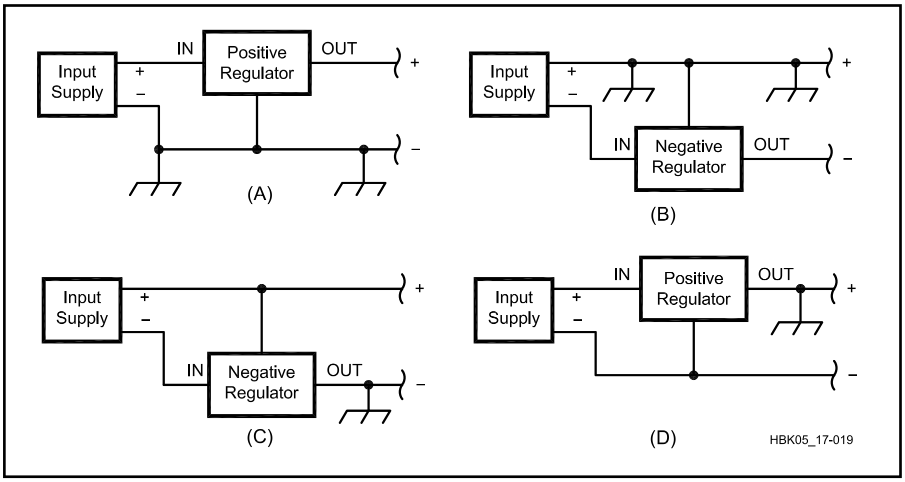


Fig 7.25 — Parts A and B illustrate the conventional manner in which three-terminal regulators are used. Parts C and D show how one polarity regulator can be used to regulate the opposite-polarity voltage.

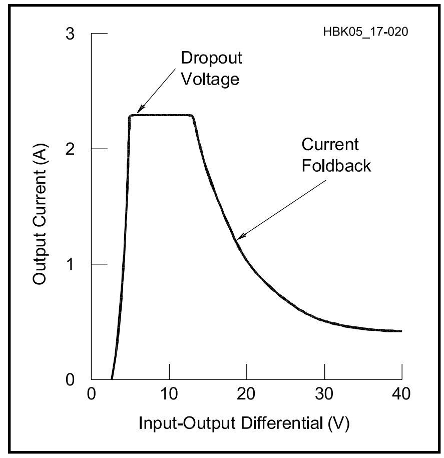


Fig 7.26 — Effects of input-output differential voltage on three-terminal regulator current.

supply to the regulator, the circuits of Fig 7.25C and D may be used to regulate positive voltages with a negative regulator and vice versa. In these configurations the input supply is floated; neither side of the input is tied to the system ground.

Manufacturers have adopted a system of family numbers to classify three-terminal regulators in terms of supply polarity, output current and regulated voltage. For example, 7805 describes a positive 5-V, 1.5-A regulator and 7905 a negative 5-V, 1.5 A unit. Depending on the manufacturer, the full part number might have a prefix such as LM, UA or MC, along with various suffixes (for example, LM7805CT or MC-7805CTG). There are many such families with widely varied ratings available from manufacturers. More information may be found in the **Component Data and References** chapter.

SPECIFYING A REGULATOR

When choosing a three-terminal regulator for a given application, the most important specifications to consider are device output voltage, output current, minimum and maximum input-to-output differential voltages, line regulation, load regulation and power dissipation. Output voltage and current requirements are determined by the load with which the supply will ultimately be used.

Input-to-output differential voltage is one of the most important three-terminal regulator specifications to consider when designing a supply. The differential value (the difference between the voltage applied to the input terminal and the voltage on the output terminal) must be within a specified range. The minimum differential value, usually about 2.5 V, is called the *dropout voltage*. If the differential value is less than the dropout voltage, no regulation will take place. Special *low dropout regulators* with lower minimum differential values are available as well. At the other end of the scale, maximum input-output differential voltage is generally about 40 V. If this differential value is exceeded, device failure may occur.

Increases in either output current or differential voltage produce proportional increases in device power consumption. By employing current foldback, as described above, some manufacturers ensure that maximum dissipation will never be exceeded in normal operation. **Fig 7.26** shows the relationship between output current, input-output differential and current limiting for a three-terminal regulator nominally rated for 1.5-A output current. Maximum output current is available with differential voltages ranging from about 2.5 V (dropout voltage) to 12 V. Above 12 V, the output current decreases, limiting the device dissipation to a safe value. If the output terminals are accidentally short circuited, the input-output differential will rise, causing current foldback, and thus preventing the power-supply components from being over stressed. This protective feature makes three-terminal regulators particularly attractive in simple power supplies.

When designing a power supply around a particular three-terminal regulator, input-output voltage characteristics of the regulator should play a major role in selecting the transformer-secondary and filter-capacitor component values. The unregulated voltage applied to the input of the three-terminal device should be higher than the dropout voltage, yet low enough that the regulator does not go into current limiting caused by an excessive differential voltage. If, for example, the regulated output voltage of the device shown in Fig 7.26 was 12 V, then unregulated input voltages of between 14.5 and 24 V would be acceptable if maximum output current is desired.

In use, all but the lowest-current regulators generally require an adequate external heat sink because they may be called on to dissipate a fair amount of power. Also, because the regulator chip contains a high-gain error amplifier, bypassing of the input and output leads is essential for stable operation.

Most manufacturers recommend bypassing the input and output directly at the leads where they protrude through the heat sink. Solid tantalum capacitors are usually recommended because of their good high-frequency capabilities.

External capacitors used with IC regulators may discharge through the IC junctions under certain circuit conditions, and high-current discharges can harm ICs. Look at the regulator data sheet to see whether protection diodes are needed, what diodes to use and how to place them in any particular application.

Adjustable Regulators

In addition to fixed-output-voltage ICs, high-current, adjustable voltage regulators are available. These ICs require little more than an external potentiometer for an adjustable output range from 5 to 24 V at up to 5 A. The unit price on these items is only a few dollars, making them ideal for test-bench power supplies. A very popular low current, adjustable output voltage three terminal regulator, the LM317, is shown in **Fig 7.27**. It develops a steady 1.25-V reference, V_{REF}, between the output and adjustment terminals. By installing R1 between these terminals, a constant current, I1, is developed, governed by the equation:

$$I1 = \frac{V_{REF}}{R1} \qquad (10)$$

Both I1 and a 100-µA error current, I2,

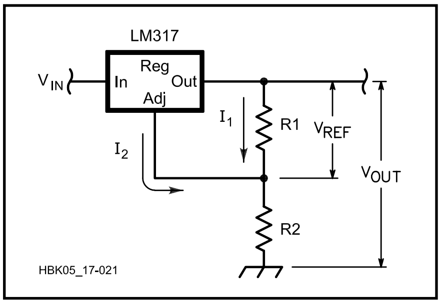


Fig 7.27 — By varying the ratio of R2 to R1 in this simple LM317 schematic diagram, a wide range of output voltages is possible. See text for details.

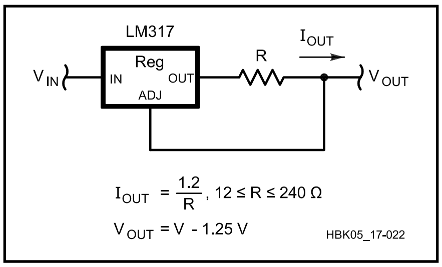


Fig 7.28 — The basic LM317 voltage regulator is converted into a constant-current source by adding only one resistor.

flow through R2, resulting in output voltage V_O. V_O can be calculated using the equation:

$$V_O = V_{REF}\left(1+\frac{R2}{R1}\right)+I2\times R2 \qquad (11)$$

Any voltage between 1.2 and 37 V may be obtained with a 40-V input by changing the ratio of R2 to R1. At lower output voltages, however, the available current will be limited by the power dissipation of the regulator.

Fig 7.28 shows one of many flexible applications for the LM317. By adding only one resistor with the regulator, the voltage regulator can be changed into a constant-current source capable of charging NiCd batteries, for example. Design equations are given in the figure. The same precautions should be taken with adjustable regulators as with the fixed-voltage units. Proper heat sinking and lead bypassing are essential for proper circuit operation.

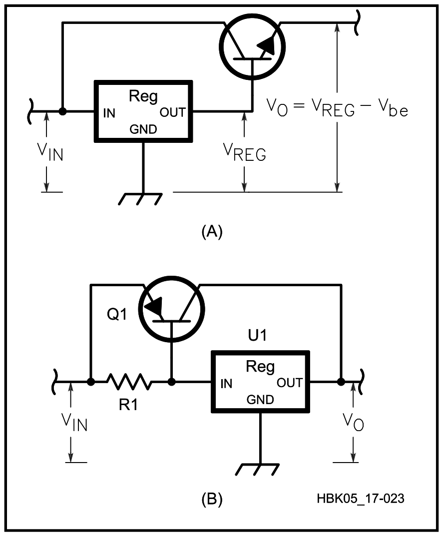


Fig 7.29 — Two methods for boosting the output-current capacity of an IC voltage regulator. Part A shows an NPN emitter follower and B shows a PNP "wraparound" configuration. Operation of these circuits is explained in the text.

INCREASING REGULATOR OUTPUT CURRENT

When the maximum output current from an IC voltage regulator is insufficient to operate the load, discrete power transistors may be connected to increase the current capability. **Fig 7.29** shows two methods for boosting the output current of a positive regulator, although the same techniques can be applied to negative regulators.

In A, an NPN transistor is connected as an emitter follower, multiplying the output current capacity by the transistor beta. The shortcoming of this approach is that the base-emitter junction is not inside the feed-back loop. The result is that the output voltage is reduced by the base-emitter drop, and the load regulation is degraded by variations in this drop.

The circuit at B has a PNP transistor "wrapped around" the regulator. The regulator draws current through the base-emitter junction, causing the transistor to conduct. R1 provides bias voltage for turning on Q1 so that U1 doesn't see the excess current. For example, a 6 Ω resistor will limit the current U1 sees to 100 mA. The IC output voltage is unchanged by the transistor because the collector is connected directly to the IC output (sense point). Any increase in output voltage is detected by the IC regulator, which shuts off its internal-pass transistor, and this stops the boost-transistor base current.

7.10 "Crowbar" Protective Circuits

Electronic components *do* fail from time to time. In a regulated power supply, the only component standing between an elevated dc source voltage and your transceiver is one transistor, or a group of transistors wired in parallel. If the transistor, or one of the transistors in the group, happens to short internally, your equipment could suffer lots of damage.

To safeguard the load equipment against possible overvoltage, some power-supply manufacturers include a circuit known as a *crowbar*. This circuit usually consists of a silicon-controlled rectifier (SCR) or thyristor connected directly across the output of the power supply, with an over-voltage-sensing trigger circuit tied to its gate. The SCR is large enough to take the full short-circuit output current of the supply, as if a crowbar was placed across the output terminals, thus the name.

In the event the output voltage exceeds the trigger set point, the SCR will fire, and the output is short circuited. The resulting high current in the power supply (shorted output in series with a series pass transistor failed short) will blow the power supply's line fuses. This is a protection for the supply as well as an indicator that something has malfunctioned internally. For these reasons, never replace blown fuses with ones that have a higher current rating.

An example of a crowbar overvoltage protection circuit can be found as a project at the end of this chapter. It provides basic design equations that can be adapted to a wide range of power supply applications.

7.11 DC-DC Switchmode Power Conversion

Very often the power source is dc, such as a battery or solar cell, or the output of an unregulated rectifier connected to an ac source. In most applications high conversion efficiency is desired, both to conserve energy from the source and to reduce heat dissipation in the converter. When high efficiency is needed, some form of switching circuit will be employed for dc-dc power conversion. Besides being more efficient, switching circuits are usually much smaller and lighter than conventional 60 Hz, transformer-rectifier circuits because they operate at much higher frequencies — from 25 to 400 kHz or even higher. Switching circuits go by many names; *switching regulators* and *switchmode converters* are just two of the more common names.

The possibility of achieving high conversion efficiency stems directly from the use of switches for the power conversion process, along with low-loss inductive and capacitive elements. An active switch is a device that is either ON or OFF and the state of the switch (ON or OFF) can be controlled with an external signal. The loss in the switch is always the product of the voltage across the switch and the current flowing through it ($P = E \times I$).

In the ON (conducting) state the voltage drop across the switch is small, and in the OFF state the current through the switch is small. In both cases the losses can be small relative to the power level of the converter. During transitions between ON and OFF states, however, there will simultaneously be both substantial voltage across the switch and significant current flowing through it. This results in power dissipation in the switch, called *switching loss*. This loss is minimized by making the switching transitions as short as possible. In this way even though the instantaneous power dissipation may be high, the average loss is low because of the small duty cycle of the transitions. Of course the more frequently the switch operates (higher switching frequency), the higher the average loss will be and this eventually limits the maximum operating frequency. A limitation on the lower end of the switching frequency range is that it needs to be above audible frequencies (>25 kHz).

This is quite different from the linear regulators discussed earlier in which there is always some voltage drop across the pass transistor (which is acting as a controlled variable resistor) while current is flowing through it. As a result the efficiency of a linear regulator can be very low, often 65% or less. Switchmode converters on the other hand will typically have efficiencies in the range of 85 to 95%.

Switchmode circuits can also generate radio-frequency interference (RFI) through VHF because of switching frequency harmonics and ringing induced by the rapid rise and fall times of voltage and current. In attempting to minimize the ON-OFF transition time, significant amounts of RF energy can be generated. To prevent RFI to sensitive receivers, careful bypassing, shielding and filtering of both input and output circuits is required. (RFI from switchmode or "switching" supplies is also discussed in the chapter on **Electromagnetic Capability**.)

There are literally hundreds of different switchmode circuits or "topologies" (see Reference 2) but we will only look at a few of those most commonly used by amateurs. Fortunately, the characteristics of the simpler circuits are to a large extent replicated in more complex circuits so that an understanding of the basic circuits provides an entry point to many other circuits.

The following discussion is only an introduction to the basics of switchmode converters. To really get a handle on designing these circuits the reader will have to do some additional reading. Fortunately a very large amount of useful information is freely available on-line. Many useful application notes are available from semiconductor manufacturers, and additional information can be found on the Web sites of filter capacitor and ferrite core manufacturers (see References 3 and 4). There are also numerous books on the subject, often available in libraries and used book stores.

Switchmode circuits can and have been implemented with many different kinds of switches, from mechanical vibrators to vacuum and gas tubes to semiconductors. Today however, semiconductors are the universal choice. For converters in the power range typical of amateur applications (a few watts through 2-3 kW) the most common choices of semiconductor switches would be either bipolar junction transistors (BJTs) or power MOSFETs. BJTs have a long history of use in switchmode applications, but to employ BJTs in a reliable and trouble-free circuit requires a relatively sophisticated understanding of them. While the MOSFET must also be used carefully, it is generally easier for a newcomer. The following circuit diagrams will show MOSFETs or generic switch symbols for the switches but keep in mind that all of the circuits can be implemented with other types of switches.

7.11.1 The Buck Converter

A schematic for a *buck converter* is shown in **Fig 7.30A**. This circuit is called a "buck" converter because the output voltage is always less than or equal to the input voltage. Power is supplied from the dc source (V_S) through the input filter (L1, C1) to the drain of the switch (Q1). The load (R) is connected across the output filter (L, C2).

The equivalent circuit when Q1 is ON is shown in Fig 7.30B (where the input filter

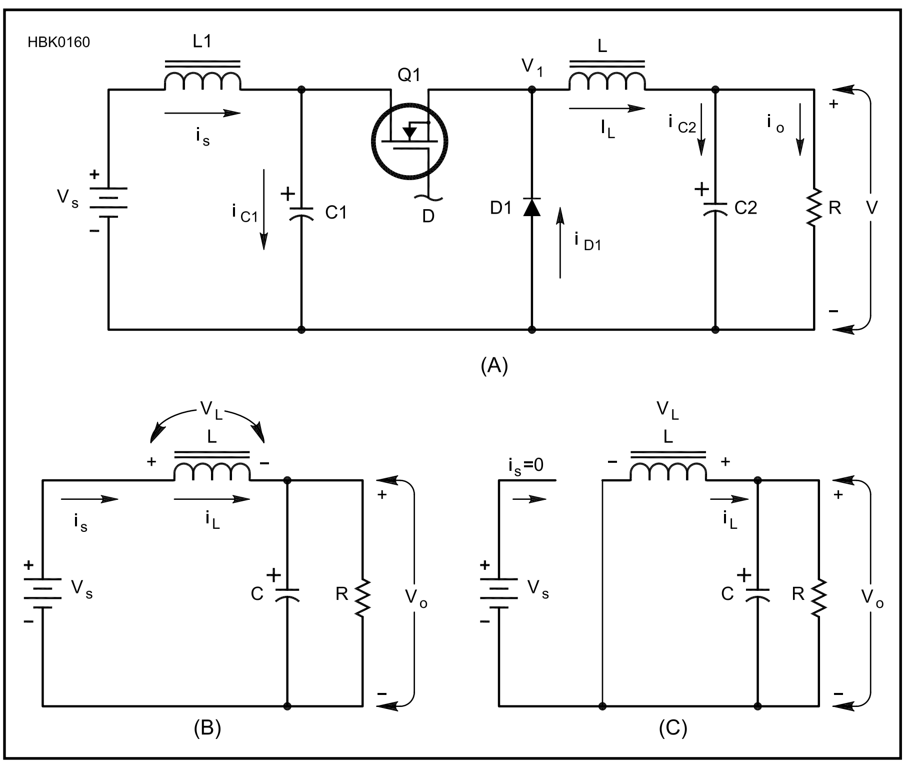


Fig 7.30 — Typical buck converter.

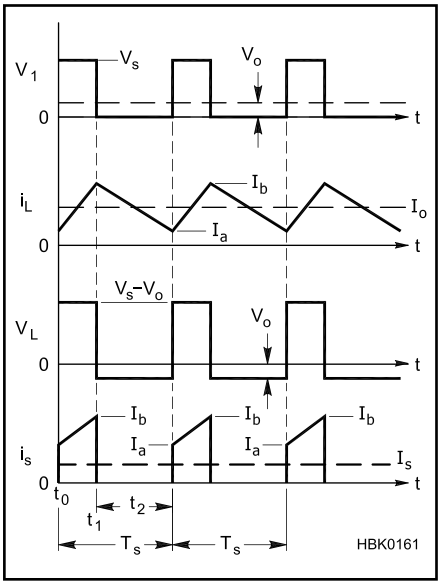


Fig 7.31 — Waveforms in a buck converter.

components are assumed to be part of V_S). V_S is connected to one end of the output inductor (L) and, because $V_O < V_S$, current flows from the source to the output delivering energy from the source to the output and also storing energy in L. At some point Q1 is turned OFF and the current flowing in L commutates (switches) to D1, as shown in Fig 7.30C. (The current in an inductor can not change instantaneously, so when Q1 turns OFF, the collapsing magnetic field in L pulls current through shunt diode D1, called a *free-wheeling diode*.) The energy in L is now being discharged into the output. This cycle is repeated at the switching frequency (f_s). The ratio of the switch ON-time to the total switching period ($T_s = t_{on} + t_{off} = 1/f_s$) is called the *duty cycle (D)*.

Typical voltage and current waveforms for a buck converter are shown in **Fig 7.31** where V_1 is the voltage at the junction of Q1, D1 and L (see Fig 7.30A).

The interval 0-t_1 corresponds to the ON time of Q1 and $T_s - t_1$ corresponds to the OFF time ($T_s = t_1 + t_2$). From the waveforms it can be seen that the current in the inductor rises while Q1 is ON and falls while Q1 is OFF. The energy in the inductor is proportional to $LI^2/2$.

The input current (I_s) is pulsating at the switching frequency. This pulsation in the input current is the reason for the input filter (L1 and C1). We need to keep this high frequency noise (the ac component of the pulsation) out of the source. All switchmode converters require some form of filter on the input to keep switching noise out of the source. Some form of filter is also required on the output to keep the switching noise out of the load. In Fig 7.30, L and C2 form the output filter. For applications requiring low ripple it would not be unusual to see an additional stage of L-C filtering added to the output.

The voltage waveform (V_1) at the input to the output filter (L and C2) is pulse-width modulated (PWM) and V_O (the dotted line in the waveform for V_1 in Fig 7.31A) is the time average of V_1. V_O is controlled or regulated by adjusting the duty cycle of Q1 in response to input voltage or output load changes. If we increase D we increase V_O up to the point where D = 1 (the switch is ON all the time) and $V_O \approx V_S$. If we never turn Q1 on (D = 0) then $V_O = 0$. Normal operation will be somewhere between these extremes.

The inductor current waveform in Fig 7.31 does not go to zero while Q1 is OFF. In other words, not all of the energy stored in L is discharged into the output by the end of each switching cycle. This is a common mode of operation for heavier loads. It is referred to as the *continuous conduction mode* or CCM, where the conduction referred to is the current in L. There is another possibility: during the time Q1 is off all of the energy in L may be discharged and for some period of time the inductor current is zero until Q1 is turned ON again. This is referred to as the *discontinuous conduction mode* or DCM. A typical converter will operate in CCM for heavy loads but as the load is reduced, at some point the circuit operation will change to DCM. CCM is often referred to as the "heavy load" condition and DCM as the "light load" condition.

This distinction turns out to be very important because the behavior of the circuit, for both small signal and large signal, is radically different between the two modes. **Fig 7.32** shows the relationship between the duty cycle (D) of Q1 and the ratio of the output voltage to the input voltage ($M = V_O/V_S$) as a function of τ_L in Fig 7.32:

$$\tau_L = \left(\frac{L}{R}\right) f_S \qquad (12)$$

τ_L is just a convenient way to make Fig 7.32 more general by tying together the variables L, R and f_s. Smaller values for τ_L correspond to lighter loads (larger values for R). As can been seen in Fig 7.32, for very light loads the higher values for D have little effect on the output/input voltage ratio. This is basically charging of the output capacitor (C2) to the peak value of $V_1 \approx V_S$.

For CCM operation, M is a linear function of D. As we vary the load, V_O will remain relatively constant. But when we go into DCM, the relation between D and M is no longer linear and in addition is heavily dependent on the load: V_O will vary as the load is changed unless the duty cycle is varied to compensate. This kind of behavior is typical of all switching regulators, even those not directly related to the buck converter. In fact, we have seen this behavior already in the section on choke input filters in Fig 7.17 where the output voltage is close to the peak input voltage for light loads and decreases as the load increases until a point is reached (I_C) where V_O stabilizes.

In a buck converter the value for the critical inductance (L_C) can be found from:

$$L_c = R\left(\frac{1-M}{2 f_S}\right) \qquad (13)$$

This looks just like equation 6 (in the earlier section on choke-input filters) with $A = 2f_s / (1 - M)$! The two phenomena are the same: peak charging versus averaging of V_1.

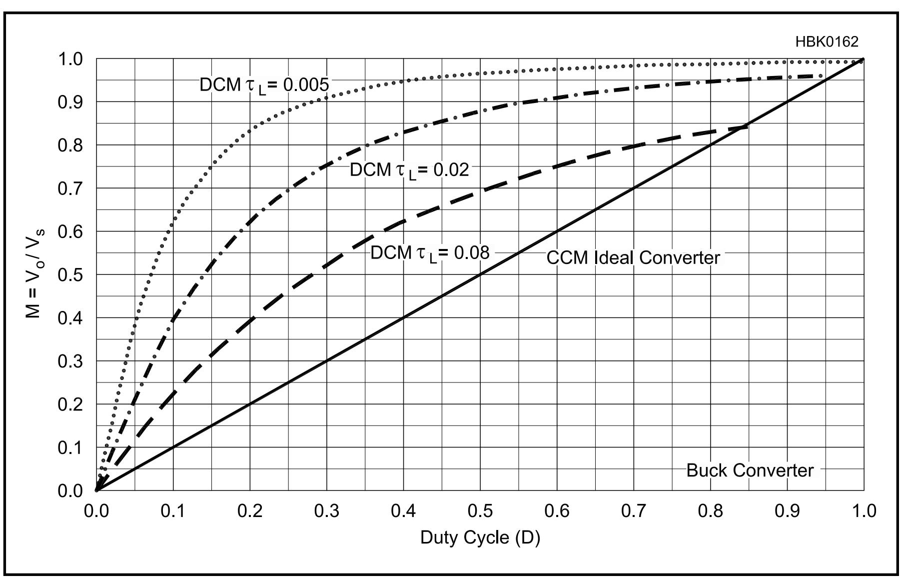


Fig 7.32 — Change in output voltage to input voltage ratio (M=VO/VS) as a function of switch duty cycle (D).

7.11.2 The Boost Converter

A *boost converter* circuit is shown in **Fig 7.33A**. This circuit is called a "boost" converter because the output voltage is always greater than or equal to the input voltage.

When Q1 is ON (Fig 7.33B), L is connected in parallel with the source (V_S) and energy is stored in it. During this time interval load energy is supplied from C. When Q1 turns OFF (Fig 7.33C), L is connected via D1 to the output and the energy in L, plus additional energy from the source, is discharged into the output. The value for L, V_S and length of time it is charged determines the energy stored in L. Again we have the possibilities that either some (CCM) or all (DCM) the energy in L is discharged during the OFF-time of Q1. The variation in the ratio of the output-to-input voltage (M) with duty cycle is shown in **Fig 7.34**.

As we saw in the buck converter, the conduction mode of L is important. In an ideal converter operating in CCM, the output voltage is substantially independent of load and $M \approx 1 / (1 - D)$. In realistic boost converters there is an important limit on the CCM value for M. In an ideal boost converter you could make the boost ratio (M) as large as you wish but in real converters the parasitic resistances associated with the components will limit the maximum value of M as indicated by the dashed line for CCM operation. The exact shape of this part of the control function will depend on the ratio of the parasitic resistance within the converter to the load resistance (see Reference 2).

There is also another very important practical effect of this limitation on the peak value for M. When the duty cycle is increased beyond the point of maximum M (this occurs at D = 0.9 in Fig 7.34), the sign of the slope of the control function changes so that the control loop will change from negative feedback to positive feedback. This can cause the converter to latch up under some load conditions if the control circuit allows D to exceed the value at maximum M. In boost converters the design of the control circuit must limit the maximum value for D so that latch-up is not possible although this may be difficult for overload conditions.

As in the case of the buck converter, in DCM the control function for M is very different from what it is in CCM and is strongly dependent on the load R. One advantage of the DCM operation is that the limitation on the maximum value for M because of parasitic circuit resistances is not nearly so pronounced. By operating in DCM it is possible to have M > 5.

An important limitation of the boost converter is that with Q1 turned OFF, you have no control over V_O. V_O will simply equal V_S. In addition, when V_S is turned on there will be an inrush current through D1, into C which cannot be controlled by Q1. In the case of the buck converter, if Q1 is kept OFF when V_S is turned on, there will be no inrush current charging the output filter capacitor (C2). In the buck converter C2 can be charged slowly by ramping up the duty cycle of Q1 during turn-on but this is not possible in the boost converter.

7.11.3 Buck-Boost and Flyback Converters

Fig 7.35 shows an example of a *buck-boost converter*. The name "buck-boost" comes from the fact that the magnitude of the output voltage can be either greater or less than the input voltage.

When S1 is ON, V_S is applied across L and energy is stored in it. During the ON time of Q1, D1 is reverse-biased and non-conducting. The output voltage (V_O) across the load (R) is supported solely from the energy stored in C. When S1 is OFF, the energy in L is discharged into C through D1. Note that the polarity of V_O is inverted from V_S: *positive V_S means negative V_O*. The relationship between V_O and V_S as a function of duty cycle is shown in **Fig 7.36**.

This graph closely resembles that for the boost converter (Fig 7.34) with one important exception: |M| begins at zero. This allows the

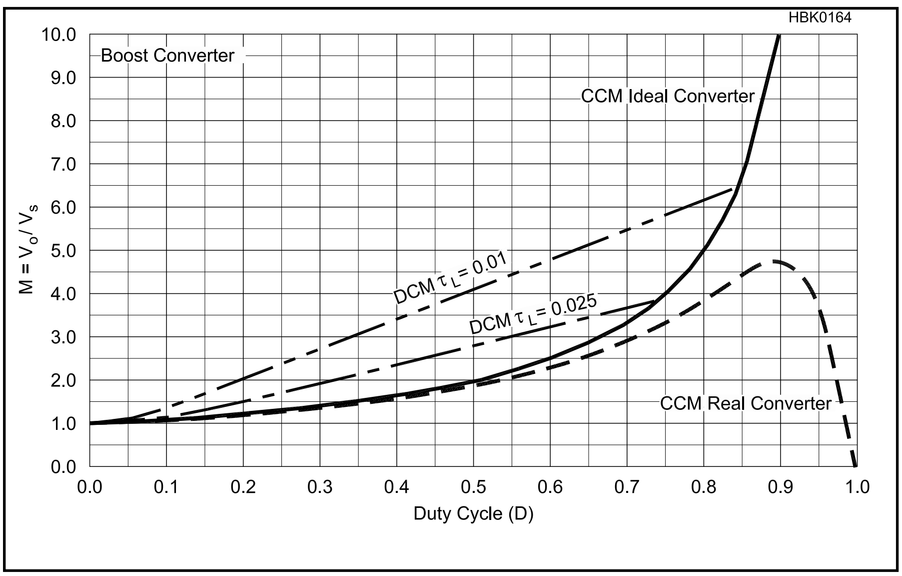


Fig 7.34 — DC control characteristic for a boost converter. $M = V_O/V_S$, D = duty cycle of Q1 and $\tau_L = f_s (L/R)$.

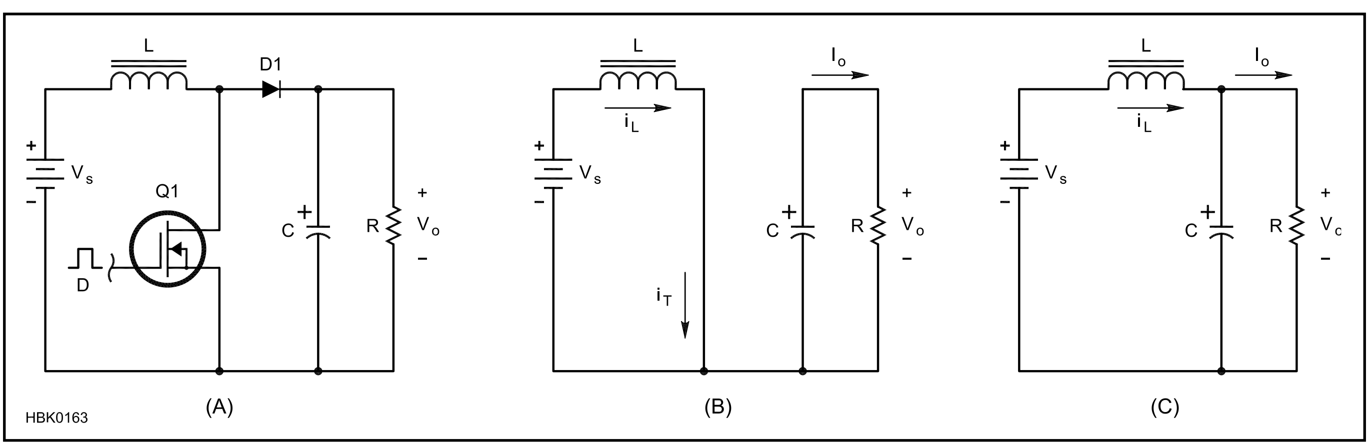


Fig 7.33 — Typical boost converter.

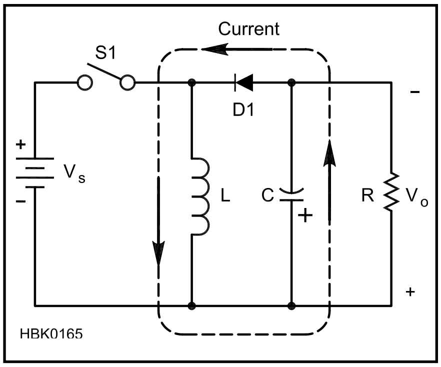


Fig 7.35 — An example of a buck-boost converter.

magnitude of the output voltage to be either below or above the source voltage, hence the name "buck-boost."

FLYBACK CONVERTERS

Simple buck-boost converters are occasionally used, but much more often it is the transformer-coupled version of this converter, referred to as a *flyback converter*, that is employed. The relationship between the flyback and buck-boost converters is illustrated in **Fig 7.37**. In Fig 7.37A we have a standard buck-boost converter, the only change from Fig 7.35 being two parallel and equal windings on the inductor. In Fig 7.37B we remove the links at the top and bottom of the two parallel windings, converting the inductor into a transformer with primary and secondary windings. The only change is that when S1 is ON, current flows in the primary of the transformer and when S1 is OFF, the current flows in the secondary winding delivering the stored energy to the output. At this point the circuit operation is the same except that we have introduced primary-to-secondary galvanic isolation.

We are now free to change the turns ratio from 1:1 to anything we wish. We can also change the polarity of the output voltages and/or add more windings with other voltages and additional loads as shown in Fig 7.37C, a typical example of a flyback converter. These are most often used in the power range of a few watts to perhaps 200 W. For higher power levels other circuits are generally more useful.

The advantages of the flyback converter lie in its simplicity. It requires only one power switch and one diode on each of the output windings. The inductor is also the isolation transformer so you have only one magnetic

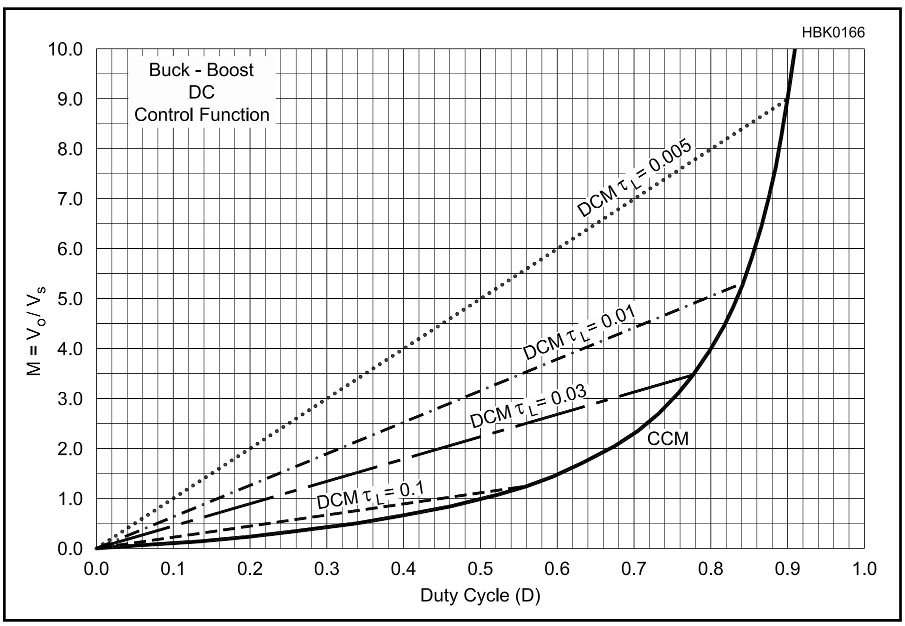


Fig 7.36 — DC control function for an ideal buck-boost converter $M = V_O/V_S$, D = duty cycle of Q1 and $\tau_L = f_s (L/R)$.

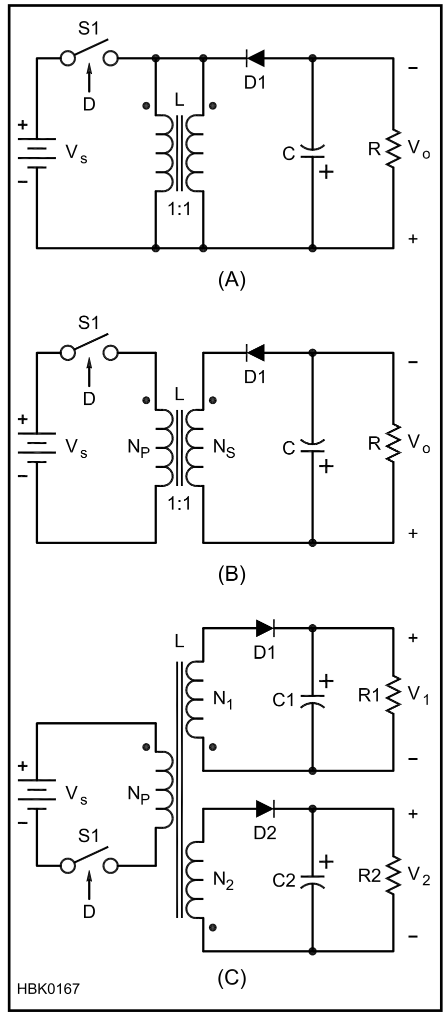


Fig 7.37 — The relationship between buck-boost and flyback converters.

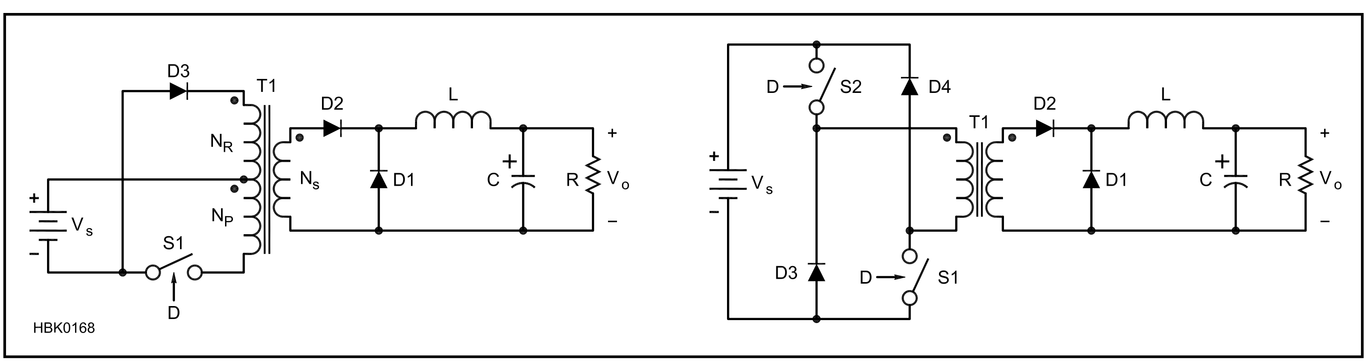


Fig 7.38 — Example of a forward converter.

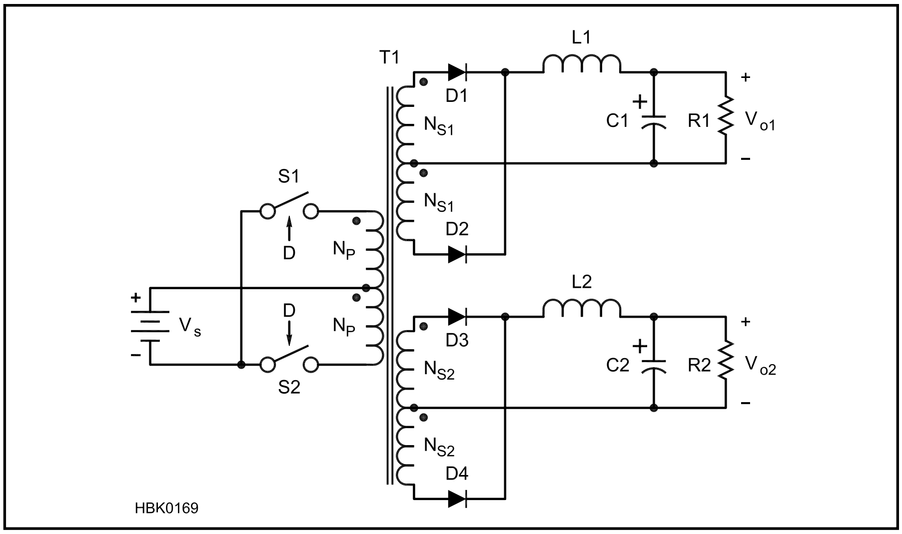


Fig 7.39 — Example of a parallel quasi-square wave dc-dc converter.

component. The disadvantage of this circuit is that both in the input and output current waveforms are pulsating. The result is that more filtering is required and the filter capacitors are exposed to high RMS currents relative to the power level.

7.11.4 The Forward Converter

The buck converter has many useful properties but it lacks input-to-output galvanic isolation, the ability to produce output voltages higher than the input voltage, and/or multiple isolated output voltages for multiple loads. We can overcome these drawbacks by inserting a transformer between the switch (S1) and the shunt diode (D1). To make this simple idea work however, we also have to add a diode in series with the output of the transformer (D2), and a third winding (N_R) with another diode (D3) to the transformer. This is done to provide a means for resetting the core (returning the magnetic flux to zero) by the end of each switching cycle. The result is the *forward converter* shown in **Fig 7.38**.

The circuit in A is the one just described. The variation in B uses two switches (S1 and S2, which switch ON and OFF *simultaneously*) instead of one but eliminates the need for a reset winding on the transformer. For the circuit in A and a given input voltage, the voltage across the switch (in the OFF state) will be equal to V_S plus the reset voltage during the OFF-time. Typically the peak switch voltage will be about $2V_S$. In the circuit shown in B the switch voltage is limited to V_S which is very helpful at higher line voltages. These circuits behave just like a buck converter except you can now have multiple isolated outputs with arbitrary voltages and polarities. These circuits are typically used in converters with power levels of 100 to 500 W.

7.11.5 Parallel, Half and Full-Bridge Converters

As the power level increases it becomes advantageous to use more switches in a somewhat more complex circuit. For applications where the input voltage is low (<100V) and the current high, the push-pull *quasi-square wave* circuit shown in **Fig 7.39** is often used.

S1 and S2 are switched on alternately with duty cycles <0.5. The output voltages are controlled by the duty cycle, D. The peak switch currents are equal to the input current but the peak switch voltages will be $2V_S$.

This circuit is still just a buck converter, but with an isolation transformer added that allows multiple outputs with different voltages above or below the input voltage. It would be very common to have a 5 V, high-current output and ±12 V, lower-current outputs, for example. This converter is typically used for operation from vehicular power with loads up to several hundred watts.

Operating directly from rectified ac util-

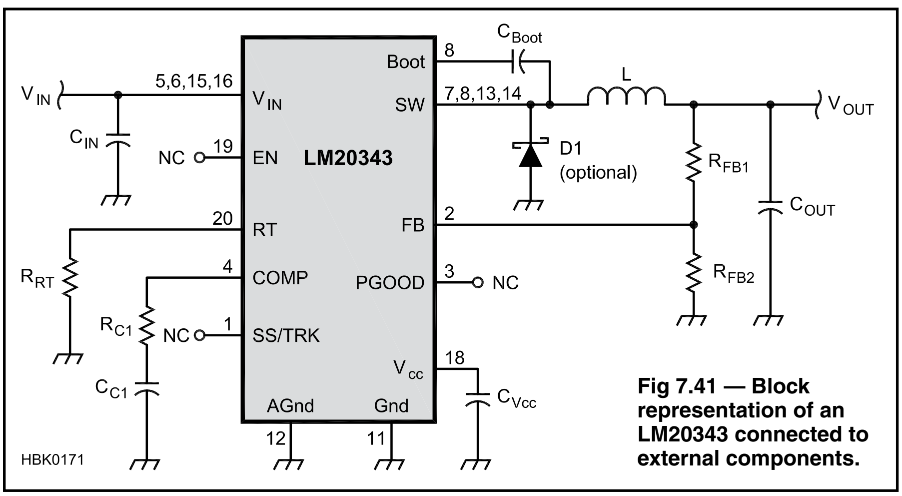


Fig 7.41 — Block representation of an LM20343 connected to external components.

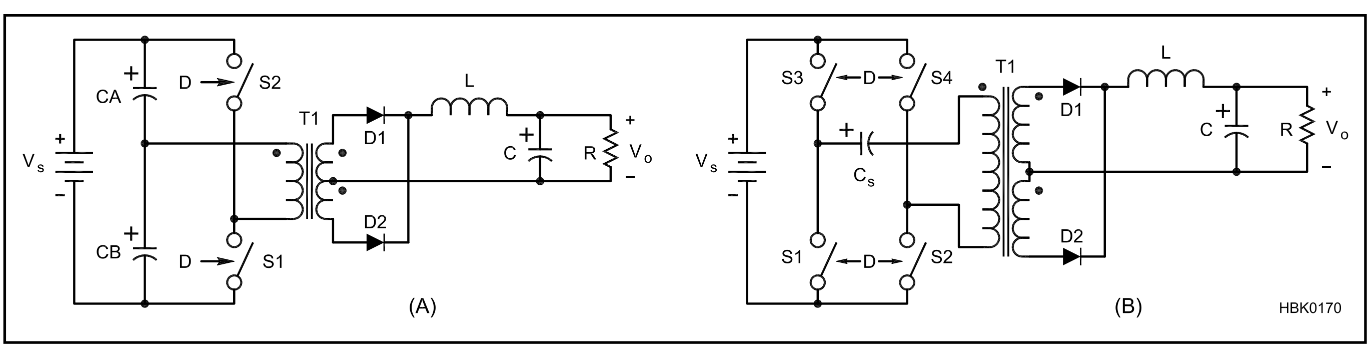


Fig 7.40 — Examples of half and full-bridge quasi-square wave dc-dc converters.

ity power usually means that V_S will be 200 V or more. For these applications, **Fig 7.40** shows how the switches are configured in either a half- (Fig 7.40A) or full-bridge (Fig 7.40B) circuit.

In A, S1 and S2 switch alternately and are pulse-width modulated (PWM) to control the output. CA and CB are large capacitors that form a voltage divider with $V_S/2$ across each capacitor. The peak switch voltages will be equal to V_S but the switch currents will be $2I_S$. The peak voltage across the primary winding will be $V_S/2$. This circuit would typically be used for off-line applications with output powers of 500 W or so.

In B, S1 and S4 switch simultaneously alternating with S2 and S3 which also switch simultaneously. The output is controlled by PWM. The peak switch voltage will be equal to V_S and the peak switch current close to I_S (the peak value is a little higher due to ripple on the inductor current which is reflected back into the primary winding). The full-bridge circuit is typically used for power levels of 500 W to several kW. C_S is present in the full-bridge circuit to prevent core saturation due to any asymmetry in the primary PWM voltage waveforms.

7.11.6 Building Switching Power Supplies

Selecting a switching circuit or topology is just the first step in building a practical switching power supply. In the actual circuit you will have to sample the output voltage, provide a voltage reference against which to compare the output, include a modulator that will convert the error voltages to a variable-duty cycle signal and finally provide correct drive to the power switches. Fortunately, all of these functions can be provided with readily available integrated circuits and a few external components. These ICs typically come with extensive applications information.

Particularly for low power applications (<100 W) there are ICs that provide all the control functions *and* the power switch. In some cases a power diode is also included. **Fig 7.41** gives an example of one of these ICs. Similar IC's are made by many different manufacturers.

The IC provides most of the components for a buck converter but some external components are still needed: filter capacitors (C_{IN}, C_{OUT}, C_{VCC}), the output inductor (L), output voltage sensing (R_{FB1} and R_{FB2}) and components to set the switching frequency (R_{C1} and C_{C1}). The shunt diode (D1) is marked as

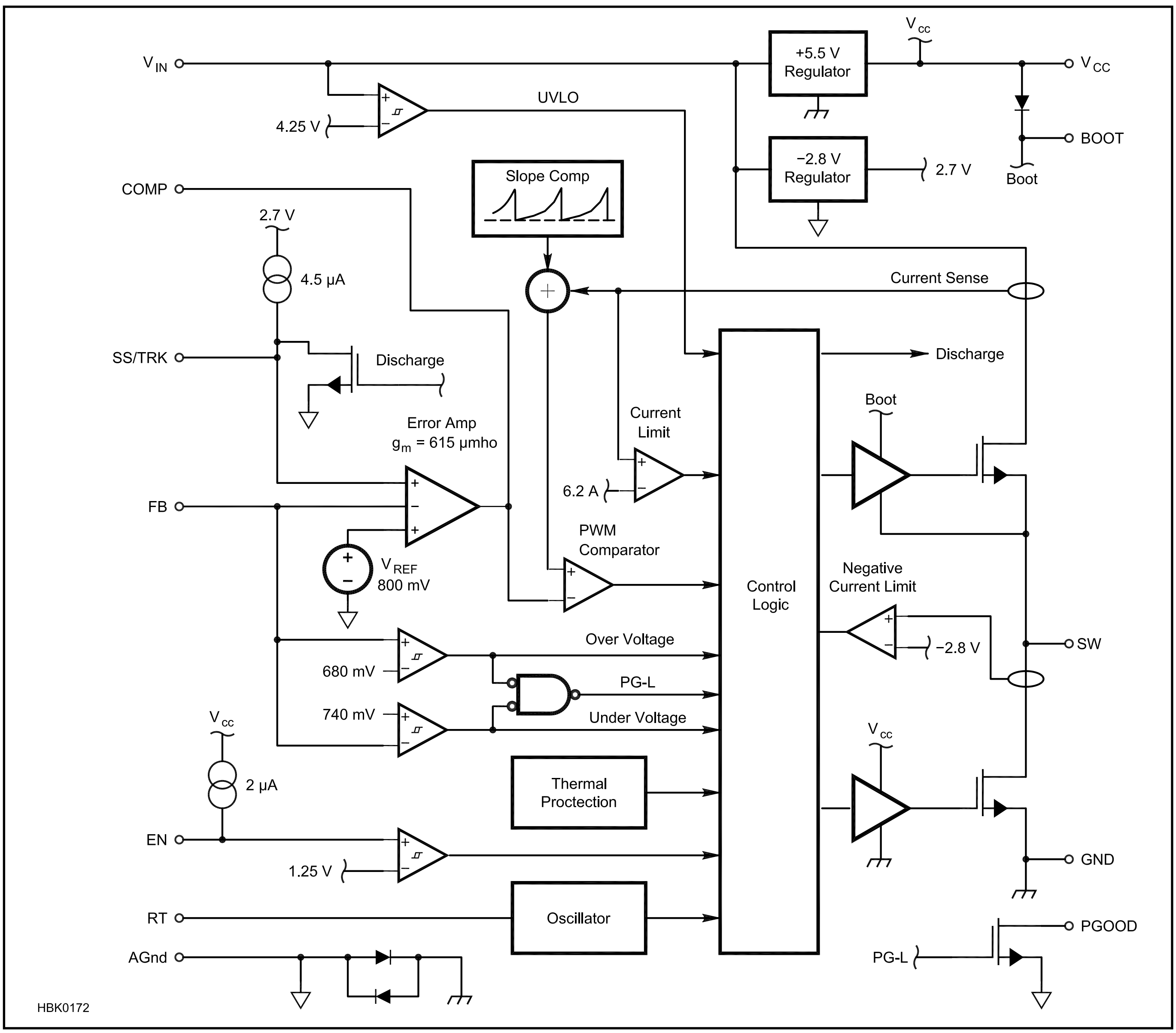


Fig 7.42 — Internal block diagram for the LM20343.

optional because there is an internal switch that can perform this function for lower powers. We can get a more detailed look at the circuitry within the IC from **Fig 7.42**.

In addition to the control functions there are also some protective functions such as input under- and overvoltage protection, over-temperature (thermal) protection, over-current limiting and slow-starting (slowly ramping up D) to limit inrush current at turn-on. ICs similar to this one are available for boost, buck-boost and flyback converters, as well as many other topologies.

For higher power applications, the power components are external to the control IC but additional features may be added. **Figs 7.43** and **7.44** are examples of typical general-purpose IC controllers. There are also ICs available that implement more complex control schemes such as current mode control and feed-forward compensation.

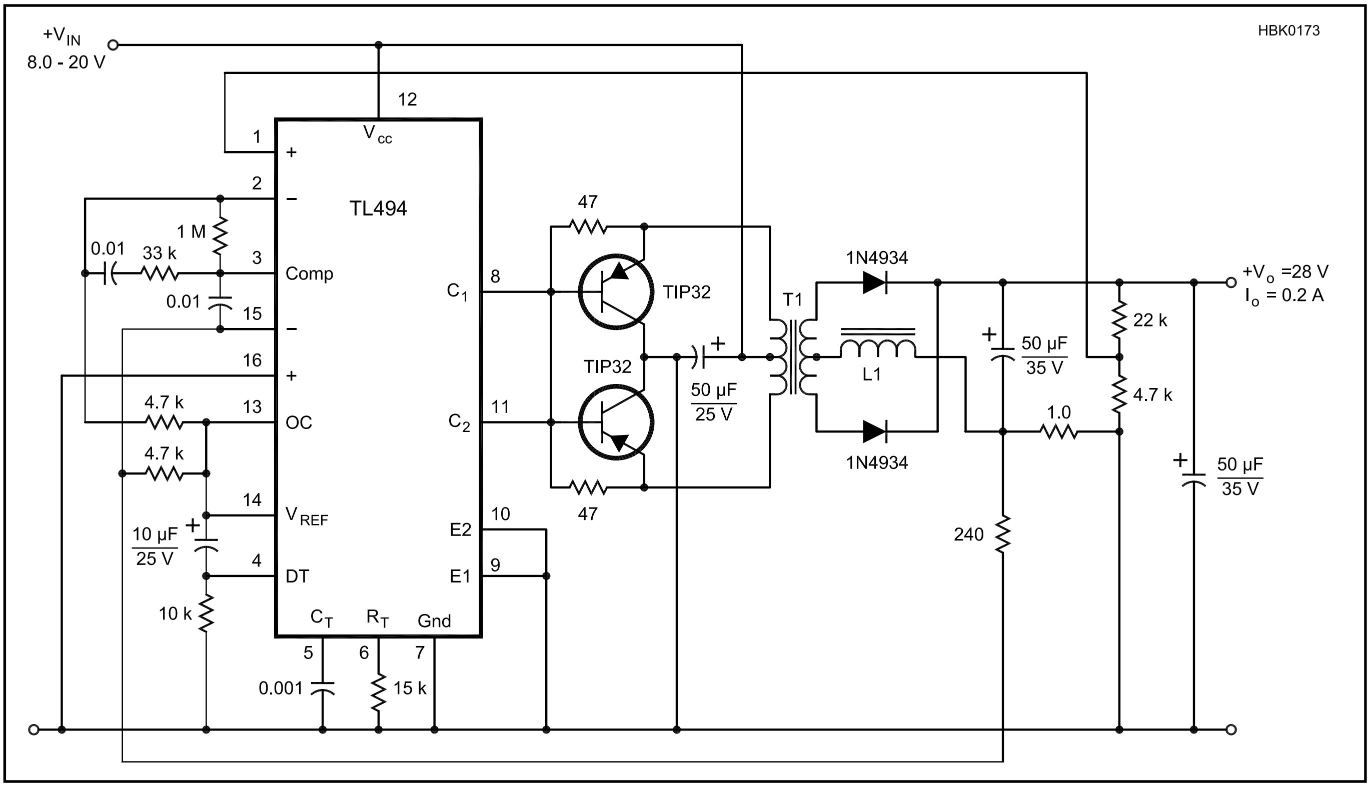


Fig 7.43 — An example of the TL494 applied to a push-pull dc-dc converter.

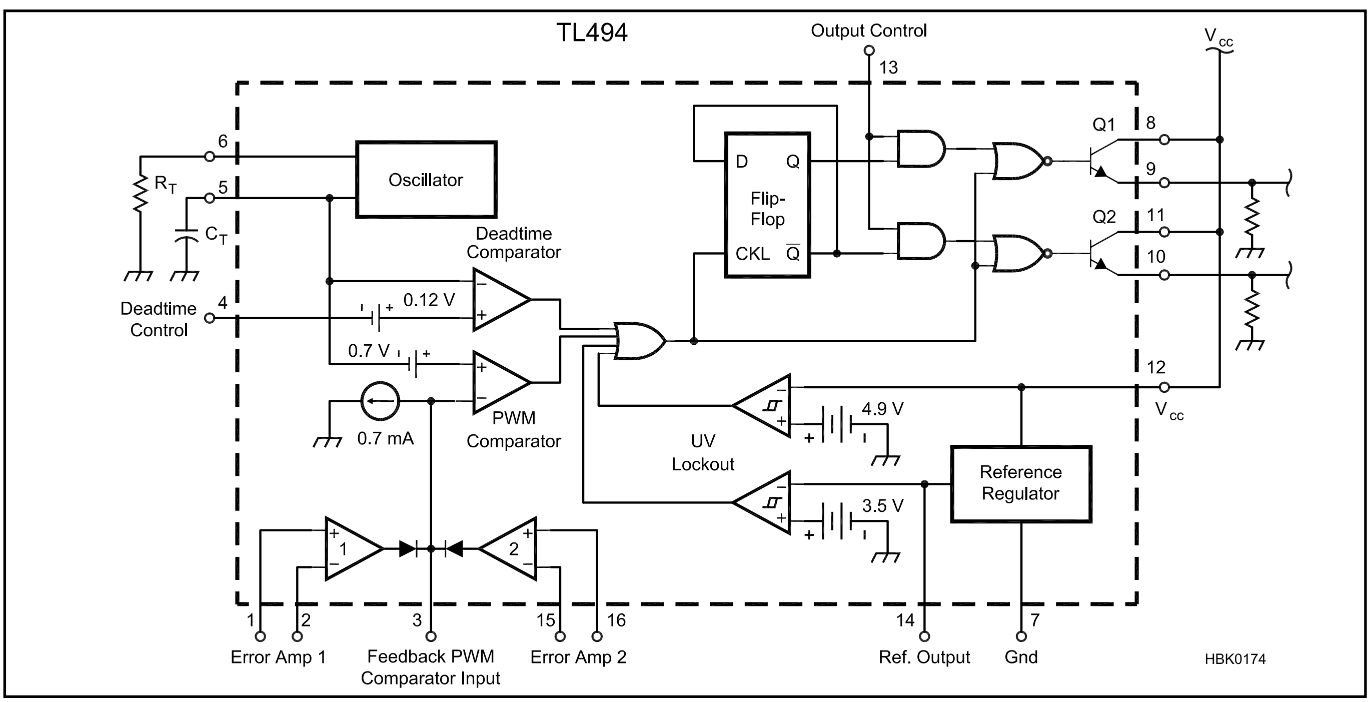


Fig 7.44 — Internal block diagram representation for the TL494.

Switchmode Converter Design Aids

By Chuck Mullett, KR6R

Today, there are hundreds of ICs on the market, from dozens of manufacturers, aimed at controlling the power conversion process. These devices have grown from their simplest forms when introduced over 20 years ago to include a long list of ancillary functions aimed at reducing the supporting circuitry and enhancing the quality and efficiency of the power conversion process. As a result, the task of designing these ICs into the final power conversion circuit has actually become more complex. To the circuit designer who is below the expert level, the task can seem daunting, indeed.

In answer to this problem, many IC manufacturers provide computer-aided design tools that greatly simplify the task of using their products. These tools take several forms, from the most rudimentary cookbook-style guides, to full-fledged circuit simulation tools like *SPICE*. The purpose of these tools is to help the designer pick appropriate resistors, capacitors and magnetic components for the design, and also to estimate the stresses on the power-handling components.

These design tools fall generally into two types: equation-based design tools and iterative circuit simulators. The differences are that the equation-based tools simply automate the basic design equations of the circuit, providing recommended choices for the components and then solving equations to compute the voltage, current and power stresses on the components. In some cases, thermal analysis is also included.

The simulators, on the other hand, use detailed mathematical models for the components and provide both dc and ac simulation of the circuits. This allows the user to see the dynamic behavior of the circuit. The simulation includes details during the switching intervals and shows rise times, parasitic effects caused by component capacitance, internal resistance and other characteristics.

These simulators use iterative mathematical processes, so they can take tens of seconds or even minutes to do the analysis, even on a modern, high-performance computer. Usually, the IC manufacturers provide fairly rudimentary versions of these simulators, capable of analyzing a basic circuit with a dozen or so components. The free *SPICE*-based software from Linear Technology Corporation (*LTSpice*, **www.linear.com**) is capable of modeling larger circuits. More capable simulators are available from software vendors, but the cost can run from hundreds of dollars to tens of thousands.

In modern power converter design, the magnetic components — transformers, power inductors and filter inductors — can be a challenge to the designer not versed in magnetic design. As a result, the vendors of these parts usually furnish design tools created specifically to the design of these components and/or the proper choice of off-the-shelf versions.

The reader is encouraged to explore design aids available from the Web sites of manufacturers (**Table 7.A1**). The following is far from complete, but should give the reader valuable insight into the vast array of available tools at his disposal. Remember also that manufacturers of passive components such as capacitors, resistors, heat sinks and thermal hardware may also have helpful and informative design aids on their Web sites.

Table 7.A1
Switchmode Converter Design Aids

Vendor	*Main Web Site*	*Design Tools*
Fairchild Semiconductor	**www.fairchildsemi.com**	From the home page, select "Design Center"
Infineon	**www.infineon.com**	From the home page Search window, select "Website" and "Search Technical Documents", then select "Power Management IC's" from the pull-down menu. From the subsequent list, select "Application Notes"
International Rectifier	**www.irf.com/indexsw.html**	**www.irf.com/design-center/mypower/**
Intersil	**www.intersil.com**	**www.intersil.com/design/**
National Semiconductor	**www.national.com/analog**	**www.national.com/analog/power/simple_switcher**
ON Semiconductor	**www.onsemi.com**	**onsemi.com/PowerSolutions/supportDoc.do?type=tools**
Texas Instruments	**www.power.ti.com**	From the Power Management page, click "Tools and Software"
Linear Technology Corp	**www.linear.com**	Search for "LTSpice" in the home page window to access the simulation tools page.

7.11.7 Switchmode Control Loop Issues

We do not have the space in this book to explain all the control issues associated with switchmode converters, but it is important to recognize that designing stable, high performance control loops for switchmode converters is a much more complex task than it would be for a series pass-regulator or an audio amplifier. All we can do here is to alert the reader to some of the issues and suggest consulting Reference 2 with detailed explanations and examples. In addition, there are numerous books and applications notes on switchmode converter design that go into the necessary detail.

The complexities arise from the inherent behavior of switchmode circuits. This behavior is often non-intuitive and sometimes even bizarre. As we've seen already, the dc behavior differs dramatically between CCM and DCM modes of operation. This difference is even more pronounced in the small-signal control-to-output characteristics. It is very common to see poles and zeros in the control system's response (see the **Analog Basics** chapter) that move with input voltage, duty cycle and/or load. Fixed double-poles can change to moving single-poles.

Moving right-half-plane zeros (this is a zero with decreasing phase-shift instead of increasing phase-shift with frequency) are inherent in CCM operation of boost and buck-boost circuits and their derivatives. There can also be large signal instabilities such as oscillations at sub-harmonics of the switching frequency and occasionally chaotic response to large load current or line voltage changes.

This is not intended to discourage readers from working with switching converters, but to simply alert them in advance of some of the difficulties. Once a basic understanding is achieved, the design of switching converters can be very interesting and rewarding.

7.12 High-Voltage Techniques

The construction of high-voltage supplies requires special considerations in addition to the normal design and construction practices used for lower-voltage supplies. In general, the builder needs to remember that physical spacing between leads, connections, parts and the chassis must be sufficient to prevent arcing. Also, the series connection of components such as capacitor and resistor strings needs to be done with consideration for the distribution of voltage stresses across the components. High-voltages can constitute a safety hazard and great care must be taken to limit physical access to components and wiring while high potentials are present.

7.12.1 High-Voltage Capacitors

For reasons of economy and availability, electrolytic capacitors are frequently used for output filter capacitors. Because these capacitors have relatively low voltage ratings (<600 V), in HV applications it will usually be necessary to connect them in series strings to form an equivalent capacitor with the capability to withstand the higher applied voltage. Electrolytic capacitors have relatively high leakage currents (low leakage resistance) especially at higher temperatures.

To keep the voltages across the capacitors in the series string relatively constant, equal-value bypassing resistors are connected across each capacitor. These *equalizing resistors* should have a value low enough to equalize differences in capacitor leakage resistance between the capacitors, while high enough not to dissipate excessive power. Also, capacitor bodies need to be insulated from the chassis and from each other by mounting them on insulating panels, thereby preventing arcing to the chassis or other capacitors in the string. The insulated mounting for the capacitors is often plastic or other insulating material in board form. A typical example from a commercial amplifier that implements some the guidelines given in this section is shown in **Fig 7.45**.

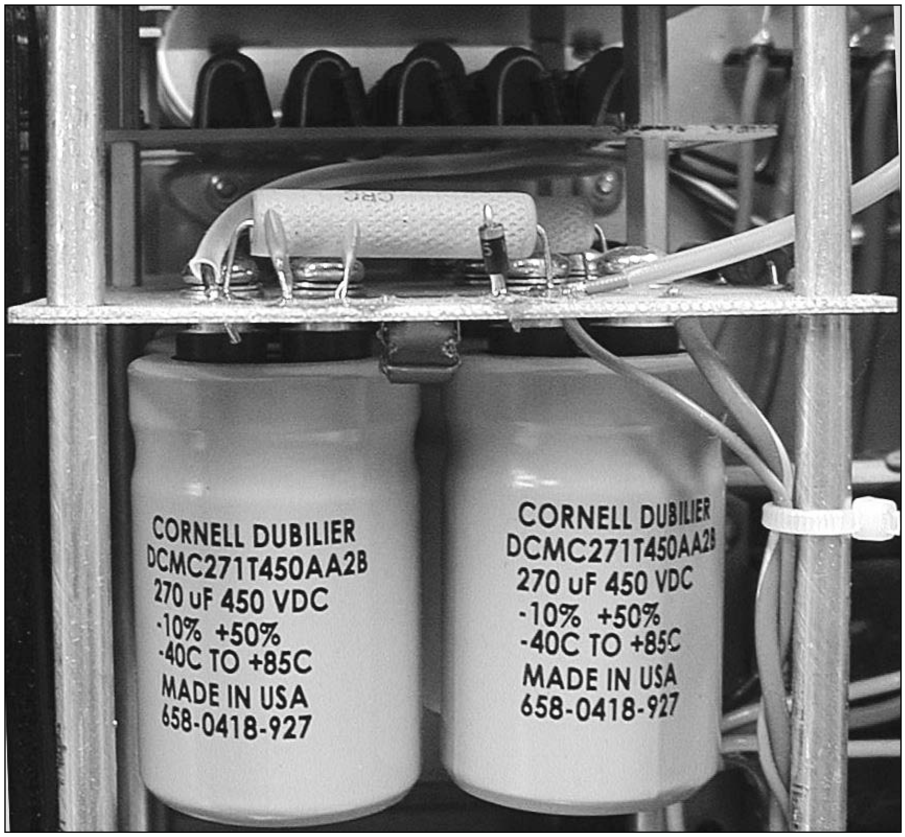


Fig 7.45 — Example of the filter capacitor/bleeder resistor installation in a high voltage power supply.

Equalizing resistors are needed because of differences in dc leakage current between different capacitors in the series string. The data sheet for an electrolytic capacitor will usually give the dc leakage current at 20 °C in the form of an equation. For example:

$$I_{leakage} = 3\sqrt{CV} \quad (14)$$

where

C = the value in µF

V = the working voltage

$I_{leakage}$ = leakage current in µA.

Keep in mind that the leakage current will increase as the capacitor temperature rises above 20 °C. A value of 3 to 4 times the 20 °C value would not be unusual because of normal heating.

We can use an approximation which includes allowance for heating to determine the maximum value of the divider resistors:

$$R \le \frac{V_r - V_m}{I_{leakage}} \quad (15)$$

where

V_r = the voltage rating of the capacitor

V_m = the voltage across the capacitor during normal operation.

A typical 3000-V power supply might use eight 330-µF, 450-V capacitors in series. In that case, V_r = 450 V, V_m = 375 V and $I_{leakage}$ = 1.2 mA. This would make R = 63 kΩ or a total of 504 kΩ for the resistor string. With 3000-V across the resistor string, the total power dissipation would be 18 W or about 2.25 W per resistor. To be conservative you should use resistors rated for at least 4 W each.

The equalizing resistors may dissipate significant power and become quite warm. It is important that these resistors do not heat the capacitors they are associated with, as this will increase the capacitor leakage current. You also have to be careful that the heat from the resistors is not trapped under the plastic support panels. The best practice is to place the resistors above the capacitors and their mounting structure allowing the heat to rise unobstructed as shown in Fig 7.45.

OIL-FILLED CAPACITORS

For high voltages, oil-filled paper-dielectric capacitors are superior to electrolytics because they have lower internal impedance at high frequencies, higher leakage resistance and are available with much higher working voltages. These capacitors are available with values of several microfarads and have working voltage ratings of thousands of volts. On the other hand, they can be expensive, heavy and bulky.

Oil-filled capacitors are frequently offered for sale at flea markets at attractive prices. One caution: It is best to avoid older oil-filled capacitors because they may contain polychlorinated biphenyls (PCBs), a known cancer-causing agent. Newer capacitors have eliminated PCBs and have a notice on the case to that effect. Older oil-filled capacitors should be examined carefully for any signs of leakage. Contact with leaking oil should be avoided, with careful washing of the hands after handling. Do not dispose of any oil-filled capacitors with household trash, particularly older units. Contact your local recycling agency for information about how to dispose of them properly.

7.12.2 High-Voltage Bleeder Resistors

Bleeder resistors across the output are used to discharge the stored energy in the filter capacitors when the power supply is turned off and should be given careful consideration. These resistors provide protection against shock when the power supply is turned off and dangerous wiring is exposed. A general rule is that the bleeder should be designed to reduce the output voltage to 30 V or less within 30 seconds of turning off the power supply.

Take care to ensure that the maximum voltage rating of the resistor is not exceeded. In a typical divider string, the resistor values are high enough that the voltage across the resistor is not dissipation limited. The voltage limit is typically related to the insulation intrinsic to the resistor. Resistor maximum voltage ratings are usually given in the manufacturer's data sheet and can be found online. Two major resistor manufacturers are Ohmite Electronics (**www.ohmite.com**) and Stackpole Electronics (**www.seielect.com**).

A 2-W carbon composition resistor will have a maximum voltage rating of 500 V. As a rough estimate, larger wire-wound power resistors are typically rated at 500 V_{RMS} per inch of length — but check with the manufacturer to be sure.

The bleeder will consist of several resistors in series. Typically wire-wound power resistors are used for this application. One additional recommendation is that two separate (redundant) bleeder strings be used, to provide safety in the event one of the strings fails. When electrolytic capacitors are used, the equalizing resistors can also serve as the bleeder resistor but they should be redundant. Again, give careful attention to keeping the heat from the resistors away from the capacitors as shown in Fig 7.45.

In the example given above for calculating the equalizing resistor value, eight 330-µF capacitors in series created the equivalent of a 41-µF capacitor with a 3000-V rating. The total resistance across the capacitors was 8 × 63 kΩ = 504 kΩ. This gives us a time constant (R × C) of about R × C ≈ 21 seconds. To discharge the capacitors to 30 V from 3000 V would take about four time constants (about 84 seconds) which is well over a minute. To get the discharge time down to 30 seconds would require reducing the equalizing resistors to 25 kΩ each. The bleeder power dissipation would then be:

$$P = \frac{V^2}{R} = \frac{3000^2}{200,000} = 45 \text{ W} \qquad (16)$$

For a 2 or 3 kV power supply, this is a reasonable value but it is still a significant amount of power and you need to make sure the resulting heat is properly managed.

7.12.3 High-Voltage Metering Techniques

Special considerations should be observed for metering of high-voltage supplies, such as the plate supplies for linear amplifiers. This is to provide safety to both personnel and to the meters themselves.

To monitor the current, it is customary to place the ammeter in the supply return (ground) line. This ensures that both meter terminals are close to ground potential. Placing the meter in the positive output line creates a hazard because the voltage on each meter terminal would be near the full high-voltage potential. Also, there is the strong possibility that an arc could occur between the wiring and coils inside the meter and the chassis of the amplifier or power supply itself. This hazardous potential cannot exist with the meter in the negative leg.

Another good safety practice is to place a low-voltage Zener diode across the terminals of the ammeter. This will bypass the meter in the event of an internal open circuit in the meter. A 1-W Zener diode will suffice since the current in the metering circuit is low.

The chapter on **RF Power Amplifiers** contains examples of how to perform current and voltage metering in high voltage supplies and amplifiers. In the past, amplifier articles have shown the meters mounted on plastic boards with stand-off insulators behind a plastic window in the amplifier or power supply front panel. While this can work it is not considered good practice today. It is usually possible to arrange the metering so that the meters are close to ground potential and may be safely mounted on the front panel of either the amplifier or the power supply.

For metering of high voltage, the builder should remember that resistors to be used in multiplier strings will often have voltage-breakdown ratings well below the total voltage being sampled. Usually, several identical series resistors will be used to reduce voltage stress across individual resistors. A basic rule of thumb is that these resistors should be limited to a maximum of 200 V, unless rated otherwise. For example, in a 2000-V power supply, the voltmeter multiplier should have a string of 10 resistors connected in series to distribute the voltage equally. The comments on resistor voltage rating in the sections on capacitor equalization and bleeder resistors apply to this case, as well.

7.12.4 High-Voltage Transformers and Inductors

Usable transformers and filter inductors can often be found at amateur flea markets but frequently these are very old units, often dating from WWII. The hermetically-sealed military components are likely to still be good, especially if they have not been used. Open-frame units, with insulation that has been directly exposed to the atmosphere and moisture for long periods of time, should be considered suspect. These units can be checked by running what is called a "hipot test" (high potential test). This involves a low current, variable output voltage power supply with a high-value resistor in series with the output. Unfortunately this equipment is seldom available to amateurs. Some motor repair shops will have an insulation tester called a "Megger" and may be willing to perform an insulation test on a transformer or inductor for you. This is not a completely definitive test but will certainly detect any gross problems with the insulation.

An alternative would be to perform the transformer tests discussed earlier with a variable autotransformer on the ac input. Slowly increase the input voltage until full line voltage is reached and let the transformer run for an hour or two while watching to see if there are any signs of failure. Doing this test in a dark room makes it easier to see any visible corona discharge, another sign of insulation problems. Because the transformer terminals will be at full voltage great care should be taken to avoid contact with the HV terminals. Some form of transparent insulating shield for the test setup is necessary.

7.12.5 Construction Techniques for High-Voltage Supplies

Layout and component arrangement in HV supplies requires some additional care beyond those for lower voltage projects. The photographs in the **RF Power Amplifiers** chapter are a good start, but there are some points which may not be obvious from the pictures.

SHARP EDGES

Sharp points, board edges and/or hardware with ragged edges can lead to localized intensification of the electric field's strength, resulting in a possible breakdown. One common offender is the component leads on the soldered side of a printed-circuit board. These are usually soldered and then cut off leaving a small but often very sharp point. The best way to handle this is to cut the lead as close to the board as practical and form a small solder ball or mound around the cut-off lead end. An example of these suggestions is given in **Fig 7.46**. The protruding component wire near the top would have a high field gradient. Below that we see a wire cut short with a rounded mound of solder covering the end of the wire.

The ends of bolts with sharp threads often protrude beyond the associated nut. One way to eliminate this is to use the dome-style nuts (cap nuts) that capture the end of the bolt or screw, forming a nicely rounded surface. An example of a cap nut is shown in Fig 7.46.

Sheet metal screws, with their needle-sharp ends, should be avoided if possible. Sheet metal screws used to close metal housings at high potential should have the screw tips inside the enclosure, which would be normal in most cases. You must also be careful of the tips of sheet metal screws that protrude on the inside of an outer enclosure. If these are in proximity to circuits at high potential, they can also lead to arcing. Keeping sheet metal screw tips well away from high voltage circuitry is the best defense.

Small pieces of sheet metal that may be part of a structure at high potential should have their edges rounded off with a file. Copper traces near the edges of circuit boards do not benefit from rounding, however. In fact, filing may make the edge sharper, ragged and more prone to breakdown. In critical areas, a small solid round wire can be soldered to the edges of a copper trace to form a rounded edge as illustrated at the right side of Fig 7.46.

INSULATORS

Some portions of the circuit may be mounted on insulators or plastic sheets. A new, clean insulator should easily withstand 10 kV per inch without creepage or breakdown across the insulating surface, but over time that surface will accumulate dirt and dust. Reducing the high-voltage stress across insulators to 5 kV per inch would be more conservative.

In theory the spacing between two smooth surfaces across an air gap can be much smaller, perhaps 20 to 30 kV per inch or even higher. But given the uncertainties of layout and voltage gradients around hardware, wiring, components and board edges, sticking with 5 kV per inch is good idea even for air gaps. This separation will seldom cause construction layout problems unless you are trying to build a very compact unit.

FUSES

Sometimes a fuse will be placed in series with a high voltage output to provide protection in case of load arcing. These fuses pose special problems. When a fuse blows, the fuse element will at least melt and perhaps even vaporize. There may be an interval when most, if not all, the output voltage appears across the fuse but the fuse has not stopped conducting. That is, there can be a sustained arc in the fuse.

Fuses have voltage ratings that are the maximum voltages across the fuse for which the arc can be expected to quench quickly. The standard 0.25 × 1.25 inch fuses used in the input ac line are typically rated for 250 V. While no doubt these ratings are conservative, this type of fuse cannot be expected to reliably clear an arc with 2 to 3 kV across the fuse and should not be used in series with a high-voltage output.

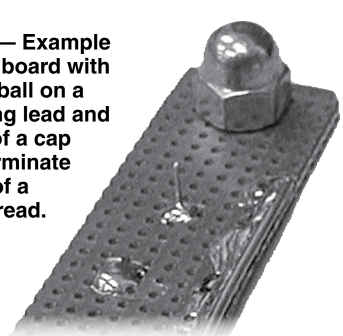


Fig 7.46 — Example of an HV board with a solder ball on a protruding lead and the use of a cap nut to terminate the end of a screw thread.

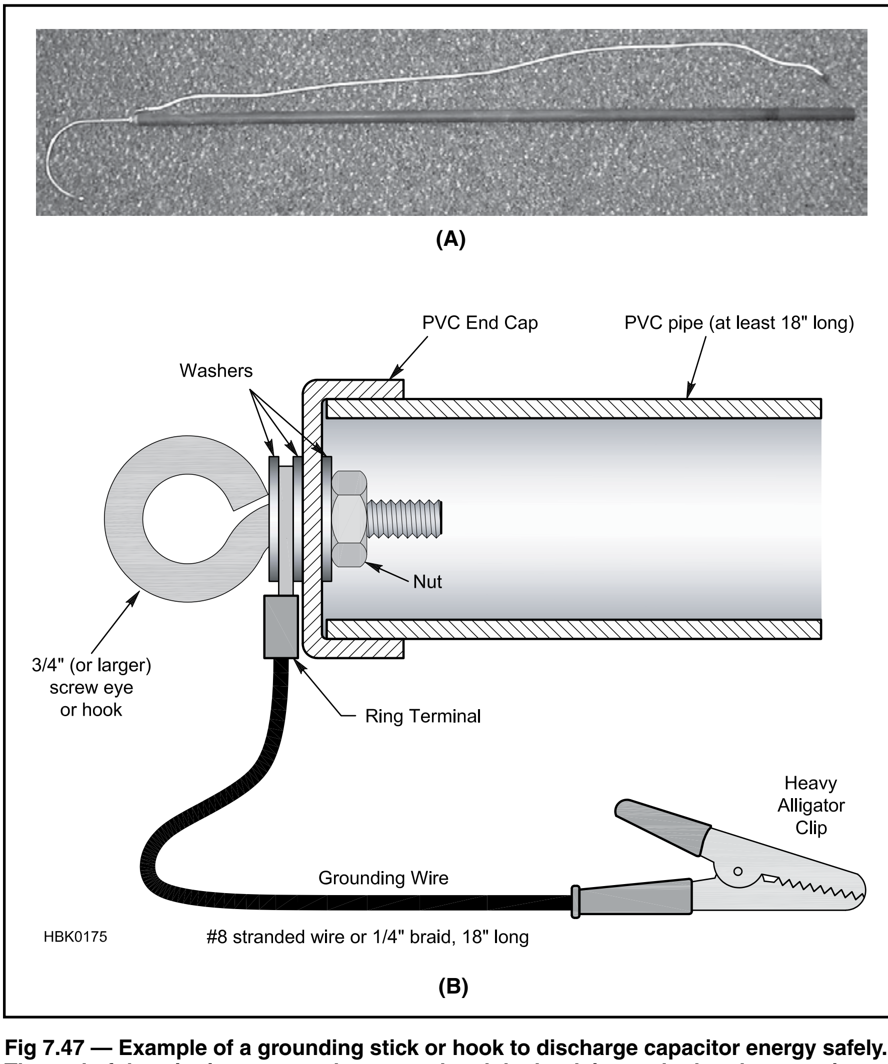


Fig 7.47 — Example of a grounding stick or hook to discharge capacitor energy safely. The end of the wire is connected to ground and the hook is touched to the capacitor terminals to discharge them. A diagram showing how to construct a grounding stick is shown at B.

Fuses with very high voltage ratings do exist however. Unfortunately, for the most part these HV fuses are for utility applications. They are physically large and not available in low current ratings. Fuses specifically rated for low currents (<2A) and several kV exist but may be very hard to locate. Even if you have a suitable fuse, care must be taken that the fuse holder can withstand the voltage. Normally high voltage fuses are mounted with end clips on an insulated board.

7.12.6 High-Voltage Safety Considerations

The voltages present in HV power supplies are potentially lethal. Every effort must be made to restrict physical access to any high potentials. A number of steps can be taken:

1) Build the power supply within a closed box, preferably a metal one.

2) Install interlocks on removable panels used for access to the interior. An interlock is usually a normally-open microswitch in series with the input power line. The microswitch is positioned so that it can be closed only when the overlying panel has been secured in place. When the panel is removed the switch moves to the open position, removing power from the supply.

3) As noted previously, the use of a bleeder resistor to rapidly discharge the capacitors is mandatory.

4) To further protect the operator when accessing the inside of a high-voltage power supply, a grounding hook like that shown in **Fig 7.47A** should be used to positively discharge each capacitor by touching the hook to each capacitor terminal. The hook is connected to the chassis via the wire shown and held by the insulated handle. It is normal practice to leave the hook in place across the capacitors while the supply is being worked on, just in case primary power is inadvertently applied. When the work is finished, remove the hook removed and replace the covers on the power supply.

Fig 7.47B shows how to construct a grounding stick of your own. A large screw eye or hook is substituted for the hook. It is important that the grounding wire be left *outside* the handle so that any hazardous voltages or currents will not be present near your hand or body.

7.13 Batteries

The availability of solid-state equipment makes it practical to use battery power under portable or emergency conditions. Handheld transceivers and instruments are obvious applications, but it's common for amateurs to power 100-W class transceivers from batteries for Field Day or emergency operation.

A battery is a group of chemical cells, usually series-connected to give some desired multiple of the *cell voltage*. The cell voltage is usually in the range of 1-4 V. Each chemical system used in the cell gives a particular nominal voltage that will vary with temperature and state of charge. This must be taken into account to make up a particular battery voltage. For example, four 1.5-V carbon-zinc cells make a 6-V battery and six 2-V lead-acid cells make a 12-V battery. **Table 7.1** shows the voltage and energy capacity of common types and sizes of batteries.

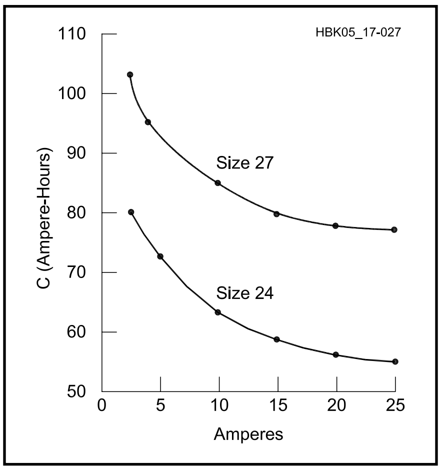


Fig 7.48 — Output capacity as a function of discharge rate for two sizes of lead-acid batteries.

7.13.1 Battery Capacity

The common rating of battery capacity is in ampere-hours (Ah) or milliampere-hours (mAh), the product of current and time. The symbol "C" is commonly used (where C is the nominal capacity in Ah). For example, C/10 would be the current continuously available for 10 hours. The value of C changes with the discharge rate and might be 110 at 2 A but only 80 at a heavier load of 20 A. **Fig 7.48** gives capacity-to-discharge rates for two standard-size vehicular lead-acid batteries. Battery capacity varies from 35 mAh for some of the small hearing-aid batteries, to more than 450 Ah for a large deep-cycle battery.

Because the current is produced by an electro-chemical reaction, cold batteries have less charge available, and some attempt to keep a battery warm before use is worthwhile. A battery may lose 70% or more of its capacity at cold extremes, but it will recover when warmer. All batteries have some tendency to freeze, but those with full charges are less susceptible. A fully charged lead-acid battery is safe to –30 °F (–34 °C) or colder. Storage batteries may be warmed somewhat by charging. Blowtorches or other open flame should never be used to heat any type of battery.

The output voltage of all batteries drops as they discharge. A practical discharge limit occurs when the load will no longer operate satisfactorily on the lower output voltage near the "discharged" point. Equipment intended for "mobile" use may be designed for an average of 13.6 V and a peak of perhaps 15 V, but may not operate well below 12 V. To obtain full use of battery charge, equipment should operate properly (if not at full power) on as little as 10.5 V with a nominal 12 to 13.6-V rating. The amount by which the battery's voltage drops depends on the type of battery and the rate of discharge. The battery manufacturer can provide information on the expected amount of voltage drop versus remaining charge at different discharge rates.

When no current is drawn from a battery, the chemical reaction essentially stops until current is required. A small amount of chemical action does continue, so stored batteries eventually will *self-discharge*. The time taken for degradation without battery use is called *shelf life*. The suggested end-of-life date is frequently marked on the side of primary batteries. Heavy-duty and industrial batteries usually have a longer shelf life.

The shelf-life of modern batteries has improved significantly, but when stored for long periods of time they will still self-discharge. One way to extend the shelf life is to keep the batteries in a refrigerator. (Do not freeze batteries as any water inside them may expand and destroy the battery.) This was standard practice with the older carbon-zinc batteries and while not as necessary with modern batteries, storage in a cool and dry place can still help extend the storage life.

7.13.2 Chemical Hazards of Batteries

In addition to the precautions given for each type of battery, the following precautions apply to all battery types. Always follow the manufacturer's advice! Extensive application information can found on manufacturer's Web sites.

Hydrogen gas escaping from storage batteries can be explosive. Keep flames or lighted tobacco products away. Use and charge batteries in well-ventilated areas to prevent hydrogen gas from building up.

No battery should be subjected to unnecessary heat, vibration or physical shock. The battery should be kept clean. Frequent inspection for leaks is a good idea. Electrolyte that has leaked or sprayed from the battery should be cleaned from all surfaces. The electrolyte is chemically active and electrically conductive, and may ruin electrical equipment. Acid may be neutralized with sodium bicarbonate (baking soda), and alkalis (found in NiCd batteries) may be neutralized with a weak acid such as vinegar. Both neutralizers will dissolve in water and should be quickly washed off. Do not let any of the neutralizer enter the battery.

Keep a record of the battery use, and include the last output voltage and, for lead-acid storage batteries, the hydrometer reading. This allows prediction of useful charge remaining, and the recharging or procuring of extra batteries, thus minimizing failure of battery power during an excursion or emergency.

Batteries can contain a number of hazardous materials such as lead, cadmium, mercury or acid, and some thought is needed for their disposal at the end of useful life. Municipal and county waste disposal sites and recycling centers will usually accept lead acid batteries because they can readily be recycled. Other types of batteries are typically not recycled and should be treated as hazardous waste. Most disposal sites and recycling centers will have occasional special programs for accepting household hazardous waste. Hardware and electronic stores may have battery recycling programs, as well. Take advantage of them.

7.13.3 Battery Types

PRIMARY BATTERIES

A *primary* battery is intended for one-time use and is then discarded. Sealed primary cells usually benefit from intermittent (rather than continuous) use. The resting period allows completion of chemical reactions needed to dispose of byproducts of the discharge.

One of the most common type of primary battery is the *alkaline* cell, in which chemical oxidation occurs during discharge. The alkaline battery has a nominal voltage of 1.5 V. Larger cells are capable of producing more mAh and less voltage drop than smaller cells. Alkaline cells have replaced the older carbon-zinc batteries that have the same nominal voltage, but substantially lower capacity.

Lithium primary batteries have a nominal voltage of about 3 V per cell and by far the best capacity, discharge, shelf-life and temperature characteristics. Their disadvantages are high cost and the fact that they cannot be readily replaced by other types in an emergency. The lithium-thionyl-chloride battery is a primary cell, and should not be recharged under any circumstances. The charging process vents hydrogen, and a catastrophic hydrogen explosion can result. Even accidental charging caused by wiring errors or a short circuit should be avoided.

Zinc-air (1.5 V), *silver oxide* (1.5 V), and *mercury* (1.4 V) batteries are very good where nearly constant voltage is desired at low currents for long time periods. Their main use is in hearing aids (zinc-air, primarily), though they may be found in other mass-produced devices such as household smoke alarms. Mercury batteries are often found as voltage references in older voltmeters.

Primary batteries should not be recharged for two reasons: It may be dangerous because of heat generated within sealed cells, and even in cases where there may be some success, both the resulting charge and life are limited. One type of alkaline battery is rechargeable, and is so marked.

SECONDARY (RECHARGEABLE) BATTERIES

A *storage* or *secondary* battery may be recharged many times. Rechargeable batteries are very popular, both as individual cells and in *battery packs* used with low-power radios, portable instruments, laptop computers and power tools. A battery pack consists of several battery cells connected in series and packaged as a single battery. In the following discussion, "battery" refers to both individual cells and battery packs.

Table 7.1
Battery Types and Characteristics

Battery Style	*Chemistry Type*	*Fully-Charged Voltage*	*Energy Rating (average)*
AAA	Alkaline — Disposable	1.5 V	1100 mAh
AA	Alkaline — Disposable	1.5 V	2600-3200 mAh
AA	Carbon-Zinc — Disposable	1.5 V	600 mAh
AA	Nickel-Cadmium (NiCd) — Rechargeable	1.2 V	700 mAh
AA	Nickel-Metal Hydride (NiMH) — Rechargeable	1.2 V	1500-2200 mAh
C	Alkaline — Disposable	1.5 V	7500 mAh
D	Alkaline — Disposable	1.5 V	14000 mAh
9V	Alkaline — Disposable	9 V	580 mAh
9V	Nickel-Cadmium (NiCd) — Rechargeable	9 V	110 mAh
9V	Nickel-Metal Hydride — Rechargeable	9 V	150 mAh
Coin Cells	Lithium — Disposable	3 - 3.3 V	25-1000 mAh
Packs	Lithium Ion – Rechargeable	3.3 - 3.6 V per cell	Varies

Nickel-Cadmium

A common type of secondary battery is the nickel-cadmium (NiCd or "nye-cad"), with a nominal voltage of 1.2 V per cell. Carefully used, these are capable of 500 or more charge and discharge cycles. For best life, the NiCd battery must not be fully discharged. Where there is more than one cell in the battery, the most-discharged cell may suffer polarity reversal, resulting in a short circuit, or seal rupture. All storage batteries have discharge limits, and NiCd types should not be discharged to less than 1.0 V per cell. There is a popular belief that it is necessary to completely discharge NiCd cells in order to recharge them to full capacity. Called the "memory effect," while this was possibly true of earlier cells, this effect is not seen in modern batteries.

Caution is required in the replacement of carbon-zinc or alkaline cells by NiCd storage cells and vice versa. Eight alkaline cells will give 12 V, while 10 of the same size NiCd cells are required for the same voltage. If a 10-cell battery holder is used, the equipment should be designed for 15 V in case the alkaline units are plugged in.

Nickel-cadmium cells are not limited to "D" cells and smaller sizes. They also are available in larger varieties ranging to mammoth 1000-Ah units having carrying handles on the sides and caps on the top for adding water, similar to lead-acid types. These large cells are sold to the aircraft industry for jet-engine starting, and to the railroads for starting locomotive diesel engines. They also are used extensively for uninterruptible power supplies. Although expensive, they have very long life. Surplus cells are often available through surplus electronics dealers, and these cells often have close to their full rated capacity.

Advantages for the ham in these vented-cell batteries lie in the availability of high discharge current to the point of full discharge. Also, cell reversal is not the problem that it is in the sealed cell, since water lost through gas evolution can easily be replaced. Simply remove the cap and add distilled water. Tap water or bottled drinking water should never be added to either nickel-cadmium or lead-acid cells, since dissolved minerals in the water can hasten self-discharge and interfere with the electrochemical process.

Lead-Acid

The most widely used high-capacity rechargeable battery is the lead-acid type. In addition to high-capacity units which are universally available, small sealed "gel-cell" batteries with capacities of a few hundred mAh and up are also becoming widely available and can be very useful. In automotive service, the battery is usually expected to discharge partially at a very high rate (engine starting), and then to be recharged promptly while the alternator is also carrying the electrical load. If the conventional auto battery is allowed to discharge fully from its nominal 2 V per cell to 1.75 V per cell, fewer than 50 charge and discharge cycles may be expected, with reduced storage capacity.

The most attractive battery for extended high-power electronic applications is the so-called "deep-cycle" battery, which is intended for such uses as powering electric fishing motors and the accessories in recreational vehicles. Size 24 and 27 batteries furnish a nominal 12 V and are about the size of small and medium automotive batteries. These batteries may furnish between 1000 and 1200 Wh per charge at room temperature. When properly cared for, they may be expected to last more than 200 cycles. They often have lifting handles and screw terminals, as well as the conventional truncated-cone automotive terminals. They may also be fitted with accessories, such as plastic carrying cases, with or without built-in chargers.

"Discharged" condition for a 12-V lead-acid battery should not be less than 10.5 V. It is also good to keep a running record of hydrometer readings, but the conventional readings of 1.265 as charged and 1.100 as discharged apply only to a long, low-rate discharge. Heavy loads may discharge the battery with little reduction in the hydrometer reading.

Lead-acid batteries with liquid electrolyte usually fall into one of three classes — conventional, with filling holes and vents to permit the addition of distilled water lost from evaporation or during high-rate charge or discharge; maintenance-free, from which gas may escape but water cannot be added; and sealed. Generally, deep-cycle batteries have filling holes and vents. As in the case of the NiCd batteries, use only distilled water to top off the electrolyte.

New lead-acid batteries are usually sold "dry charged" with a separate container for the electrolyte. In a dry charged battery the chemical composition of the cell plates is the same as for a fully charged battery but the electrolyte is omitted. This allows for much longer storage time before the battery is sold or used, keeping the battery from deteriorating. New batteries should be filled with electrolyte and allowed to stand for at least half an hour before use. They should then be charged at about 15 A for 15 minutes or so. The capacity of the battery will build up slightly for the first few cycles of charge and discharge, and then have fairly constant capacity for many cycles. Slow capacity decrease may then be noticed.

Lead-acid batteries are also available with gelled electrolyte. Commonly called *gel cells*, these may be mounted in any position if sealed, but some vented types are position sensitive. Also popular are AGM (absorbed glass mat) deep-cycle batteries that are rugged, sealed, maintenance free units that do not outgas dangerous levels of hydrogen when charging.

Nickel-Metal Hydride

The nickel-metal hydride (NiMH) battery is quite similar to the NiCd, but the cadmium electrode is replaced by one made from a porous metal alloy that traps hydrogen (therefore the name metal hydride). NiMH cells do not contain any dangerous substances, while both NiCd and lead-acid cells contain quantities of toxic heavy metals. NiCd cells have been largely replaced by NiMH and other newer, less toxic, technologies in new equipment.

Many of the basic characteristics of NiMH cells are similar to NiCds. For example, the voltage is very nearly the same, they can be slow-charged from a constant-current source, and they can safely be deep-cycled.

There are also some important differences: The most attractive feature is a much higher capacity for the same cell size — often nearly twice as much as NiCd batteries! The typical AA NiMH cell has a capacity of 2000 mAh or greater, compared to the 600 to 830 mAh for the same size NiCd. The fast-charge process is different for NiMH batteries. A fast charger designed for NiCd will not correctly charge NiMH batteries, but many commercial fast chargers are designed for both types of batteries. Check the charger specifications and setting before attempting to recharge any batteries.

The internal resistance (discussed below) of NiMH cells is somewhat higher than that of NiCd cells, resulting in reduced performance at very high discharge currents. This can cause slightly reduced power output from a handheld transceiver powered by a NiMH pack, but the effect is barely noticeable and the higher capacity and resulting longer run time far outweigh this drawback. NiMH batteries outperform NiCd batteries whenever high capacity is desired, while NiCd batteries still have advantages when delivering very high peak currents, such as for power tools. At least one manufacturer warns that the self-discharge of NiMH cells is higher than for NiCd, but again, in practice this can hardly be noticed.

At the time of this writing, many cell phones and portable consumer electronics equipment use NiMH batteries, and manufacturers offer NiMH packs for Amateur Radio applications. Standard-sized NiMH cells are widely available from the major electronic parts suppliers.

Lithium-Ion

The lithium-ion battery (Li-ion) is another alternative to NiCd batteries. For the same energy storage, a Li-ion battery is about one-third the weight and one-half the volume of a NiCd. It also has a lower self-discharge rate. Typically, at room temperature, a NiCd cell will lose from 0.5 to 2% of its charge per day. The Lithium-ion cell will lose less than 0.5% per day and even this loss rate decreases after about 10% of the charge has been lost. At higher temperatures the difference is even greater in favor of Li-ion. The result is that lithium-ion cells are a much better choice for standby operation where frequent recharge is not available.

One major difference between NiCd and Li-ion cells is the cell voltage. The nominal voltage for a NiCd cell is about 1.2 V. For the Li-ion cell it is 3.6 V with a maximum cell charging voltage of 4 V. You cannot substitute Li-ion cells directly for NiCd cells. You will need one Li-ion cell for three NiCd cells. Chargers intended for NiCd batteries must not be used with Li-ion batteries, and vice-versa.

Table 7.2
Battery Energy Storage Capability

Battery Type	*Wh/kg*	*Joules/kg*	*Wh/liter*
Lead-acid	41	146,000	100
Alkaline long-life	110	400,000	320
Carbon-zinc	36	130,000	92
NiMH	95	340,000	300
NiCd	39	140,000	140
Lithium-ion	128	460,000	230

7.13.4 Battery Discharge Planning

As shown in the previous section, discharge profiles will vary with the type of battery and the characteristics of the load. Both the type of battery and the characteristics of the load need to be considered to get the maximum from a given battery. Another important characteristic especially for portable operation is the energy density of a given battery. This varies widely between types. **Table 7.2** makes a comparison between several different types of batteries. The energy densities are given in terms of both weight and volume. For a fixed volume or physical size of the battery, discharge planning must also take into account the substantial differences between different types of batteries.

Transceivers usually drain a battery at two or three rates: one for receiving, one for transmit standby and one for key-down or average voice transmit. Considering just the first and last of these (assuming the transmit standby is equal to receive), average two-way communication would require the low rate 75% of the time and the high rate 25% of the time. The ratio may vary somewhat with voice. The user may calculate the percentage of battery charge used in an hour by the combination (sum) of rates. If, for example, 20% of the battery capacity is used in an hour, the battery will provide five hours of communications per charge. In most actual on-air situations, the time spent listening is much greater than that spent transmitting.

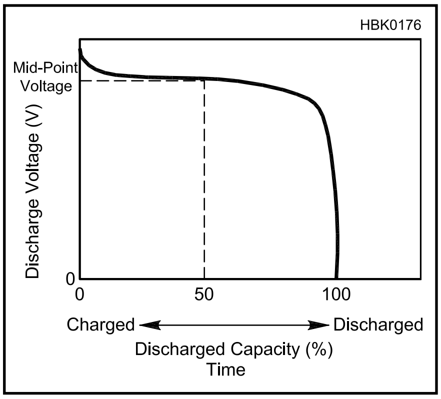


Fig 7.49 — Typical NiCd battery discharge characteristic at a constant load current.

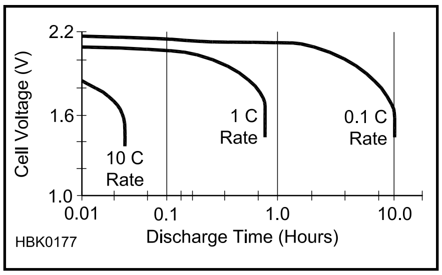


Fig 7.50 — Typical lead acid battery discharge characteristic.

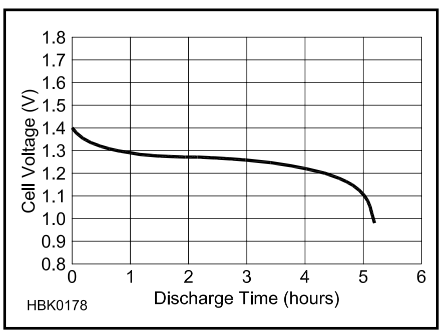


Fig 7.51 — Typical NiMH battery discharge characteristic.

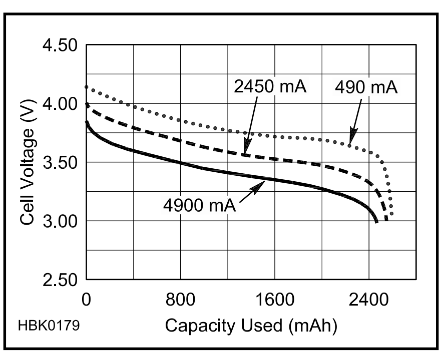


Fig 7.52 — Typical lithium-ion battery discharge characteristic at three discharge rates.

7.13.5 Battery Internal Resistance

Battery internal resistance is very important to handheld transceiver users. This is because the internal resistance is in series with the battery's output and therefore reduces the available battery voltage at the high discharge currents demanded by the transmitter. The result is reduced transmitter output power and power wasted in the cell itself by internal heating. Because of cell-construction techniques and battery chemistry, certain types of cells typically have lower internal resistance than others.

The NiCd and NiMH cells are the best battery type for high discharge current capability. The NiCd is slightly better, but NiMH battery construction is rapidly becoming competitive with NiCd. Both battery types maintain their low internal resistance throughout the *discharge curve* (a graph of output voltage versus charge remaining). **Figs 7.49** through **7.52** show typical discharge characteristics for four different types of batteries.

Alkaline primary cells have higher internal resistance than NiCd and NiMH. When these cells are used with handheld transceivers, it is not uncommon to have lower output power because of the lower battery voltage under load. Even with fresh cells, the low battery indicator may come on as the cell voltage drops more than for comparable NiCd or NiMH batteries..

The lead-acid cell, which is used in larger belt-hung battery packs and for temporary or emergency power, is pretty close to the alkaline cell for internal resistance, but only at full charge. Unlike the NiCd, the electrolyte in the lead-acid cell enters into the chemical reaction. During discharge, the specific gravity of the electrolyte gradually drops as it approaches water, and the conductivity decreases. Therefore, as the lead-acid cell approaches a discharged state, the internal resistance increases. Larger battery capacities are usually used (approximately 2 Ah) and so the effects of the higher internal resistance are consequently reduced.

The highest internal resistance is found in the common carbon-zinc flashlight cell. With the transmit current demand levels of handheld radios, these cells have limited utility although they may be useful for low-current, receive-only applications.

7.13.6 Battery Charging

Each different type of rechargeable battery must be charged according to the manufacturer's recommended charge rate and duration. Batteries often require a slight overcharge, based on the battery capacity, C. If, for instance, the specified charge rate is 0.1 C (the 10-hour rate), 12 or more hours may be needed for the charge. Once full-

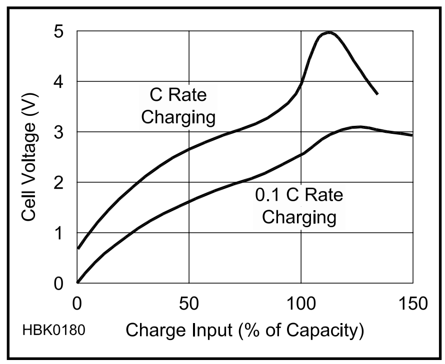


Fig 7.53 — Typical NiCd battery charge characteristic at two different rates.

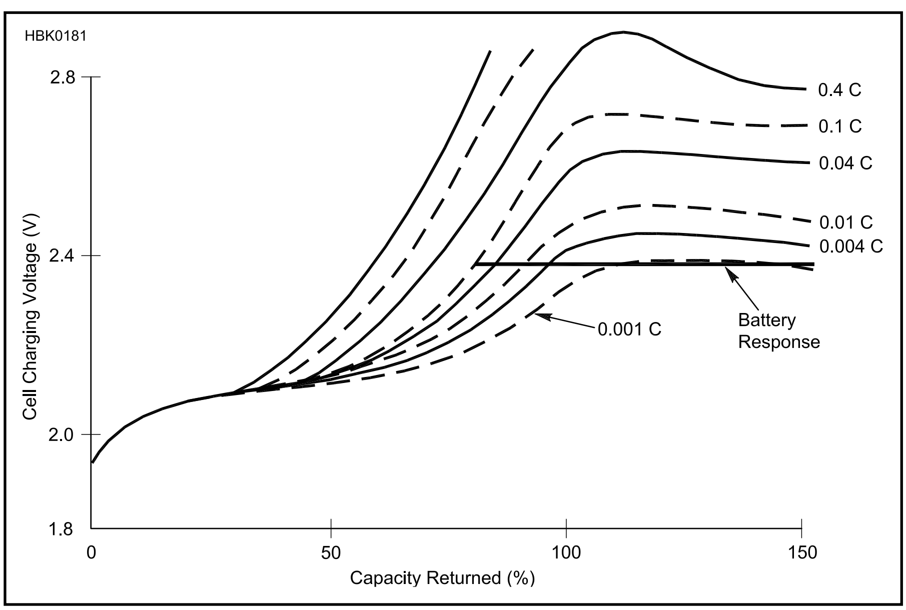


Fig 7.54 — Typical lead-acid battery charge characteristic at various rates.

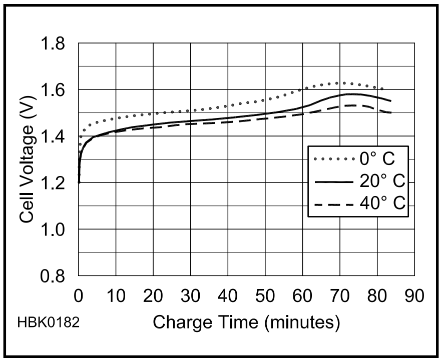


Fig 7.55 — Typical NiMH battery charge characteristic at various temperatures.

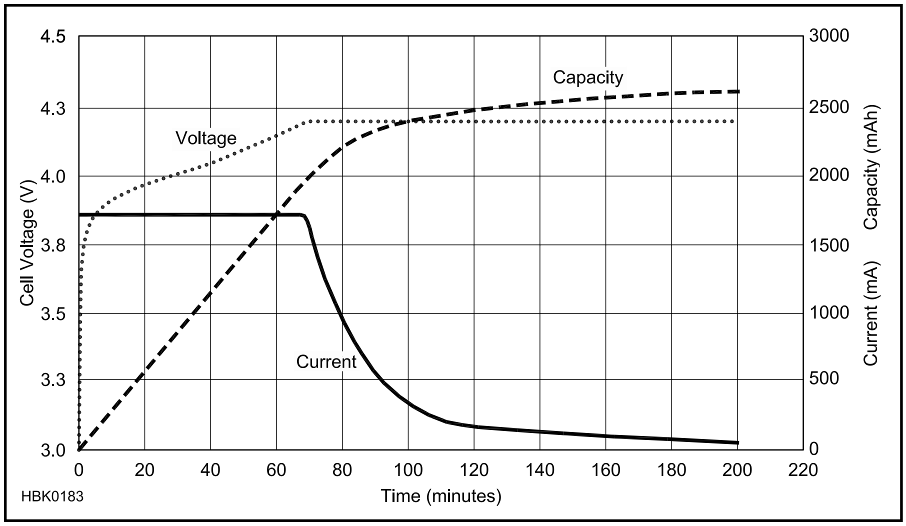


Fig 7.56 — Typical lithium-ion battery charge characteristic.

charge is reached, the battery charger should switch to a maintenance mode that varies with battery type. Do not attempt to charge batteries in a charger not specified for that type of battery. The result may be damage to the battery, the charger, or both. If you are attempting to charge batteries while still in a radio, using the wrong type of batteries can damage the radio, as well. Typical charge profiles for several types of batteries are shown in **Figs 7.53** through **7.56**.

CHARGING LEAD-ACID BATTERIES

Deep-cycle lead-acid cells are best charged at a slow rate, while automotive types may safely be given higher rate charges. This depends on the amount of heat generated within each cell, and cell venting to prevent pressure build-up. Some batteries have built-in temperature sensing, used to stop or reduce charging before the heat rise becomes a danger. Fast charges do not usually allow gas recombination, so some of the battery water will escape in the form of gas. If the water level falls below a certain point, acid hydrometer readings are no longer reliable. If the water level falls to the level of the battery's internal plates, permanent battery damage may result.

Gelled-electrolyte lead-acid batteries provide 2.4 V/cell when fully charged. Damage results, however, if they are overcharged. (Avoid *constant-current* or *trickle charging* unless battery voltage is monitored and charging is terminated when a full charge is reached.) Voltage-limited charging is best for these batteries. A proper charger maintains a safe charge-current level until 2.3 V/cell is reached (13.8 V for a 12-V battery). Then, the charge current is tapered off until 2.4 V/cell is reached. Once charged, the battery may be safely maintained at the *float* level, 2.3 V/cell. Thus, a 12-V gel-cell battery can be "floated" across a regulated 13.8-V system as a battery backup in the event of power failure. These batteries are readily available since they are used in household and commercial burglar/fire alarm systems as well as in uninterruptible power supplies (UPSs).

CHARGING NiCd AND NiMH BATTERIES

NiCd batteries have a flat voltage-versus-charge characteristic until full charge is reached; at this point the battery voltage rises abruptly. With further charging of the NiCd battery, the electrolyte begins to break down and oxygen gas is generated at the positive (nickel) electrode and hydrogen at the negative (cadmium) electrode.

Since the cell should be made capable of accepting an overcharge, battery manufacturers typically prevent the generation of hydrogen by increasing the capacity of the cadmium electrode. This allows the oxygen formed at the positive electrode to reach the metallic cadmium of the negative electrode and reoxidize it. During overcharge, therefore, the cell is in equilibrium. The positive electrode is fully charged and the negative electrode less than fully charged, so oxygen evolution and recombination "wastes" the charging power being supplied.

To ensure that all cells in a NiCd battery reach a fully-charged condition, NiCd batteries should be charged by a constant current at about a 0.1-C current level. This level is about 50 mA for the AA-size cells used in most handheld radios. This is the optimum rate for most NiCd cells since 0.1 C is high enough to provide a full charge, yet it is low enough to prevent over-charge damage and provide good charge efficiency.

Although *fast-charge-rate* (three to five hours, typically) chargers are available for hand-held transceivers, they should be used with care. The current delivered by these units is capable of causing the generation of large quantities of oxygen in a fully charged cell. If the generation rate is greater than the oxygen recombination rate, pressure will build in the cell, forcing the vent to open and the oxygen to escape. This can eventually cause drying of the electrolyte, and then cell failure. The cell temperature can also rise, which can shorten cell life. To prevent overcharge from occurring, fast-rate chargers should have automatic charge-limiting circuitry that will switch or taper the charging current to a safe rate as the battery reaches a fully charged state.

Overcharging NiCds in moderation causes little loss of battery life. Continuous overcharge, however, may generate a voltage depression when the cells are later discharged. For best results, charging of NiCd cells should be terminated after 15 hours at the slow rate. Better yet, circuitry may be included in the charger to stop charging, or reduce the current to about 0.02 C when the 1.43-V-per-cell terminal voltage is reached.

Although many of the same concerns and suggestions apply to NiMH batteries, there are some important differences. After full-charge is reached, the NiMH cell electrodes discharge slightly. Many chargers use this slight drop in battery voltage as a signal that full charge has been reached. Another sign that the battery is fully charged is a rise in internal temperature. If the full charging rate is maintained, internal temperature can become dangerously high, leading to battery damage or explosion. Trickle- or float-charging is not recommended for NiMH batteries.

SOLAR CHARGING SYSTEMS

Price and availability make solar panels an attractive way to maintain the charge on your batteries. Relatively small, low-power solar arrays provide a convenient way to charge a NiCd or sealed-lead-acid battery for emergency and portable operation. This is especially popular for QRP operating and for stations on boats or used while camping. You should always connect some type of charge controller between the battery and the solar array. This will prevent overcharging the battery, and the possible resulting battery damage.

7.13.7 DC-AC Inverters

For battery-powered operation of ac-powered equipment, dc-ac inverters are used. An inverter is a dc-to-ac converter that provides 120-V ac. Inverters come with varying degrees of sophistication. The simplest type of inverter switches directly at 60 Hz to produce a square-wave output. This is no problem for lighting and other loads that don't care about the input waveform. However, some equipment will work poorly or not at all when supplied with square wave power because of the high harmonic content of the waveform.

The harmonic content of the inverter output waveform can be reduced by the simple expedient of reducing the waveform duty cycle from 50% (for the square wave) to about 40%. For many loads, such as computers and other electronic devices, this may still not be adequate, and so many inverters use waveform shaping to approximate a sine-wave output. The simplest of these methods is a resonant inductor-capacitor filter. This adds significant weight and size to the inverter. Most modern inverters use high-frequency Pulse-width modulation (PWM) techniques to synthesize the 60 Hz sinusoidal output waveform, much like a switching power supply. See **Fig 7.57**.

Inverters are usually rated in terms of their VA or "volt-ampere product" capability although sometimes they will be rated in watts. Care is required in interpreting inverter ratings. A purely resistive load operating from a sinusoidal voltage source will have a sinusoidal current flowing in phase with the voltage. In this case, VA, the product of the voltage (V) and the current (A), will equal the actual power in watts delivered to the load so that the VA and the watt ratings are the same. Some loads, such as motors and many rectifier power supplies, will shift the phase of load current away from the source voltage or the load current will flow in short pulses as shown earlier for capacitor input filters. In these cases (which are very common), the VA product for that type of load can be much larger than the delivered power in watts. In the absence of detailed knowledge of the load characteristics it is prudent to select an inverter with a VA or wattage capability of 25% or more above the expected load.

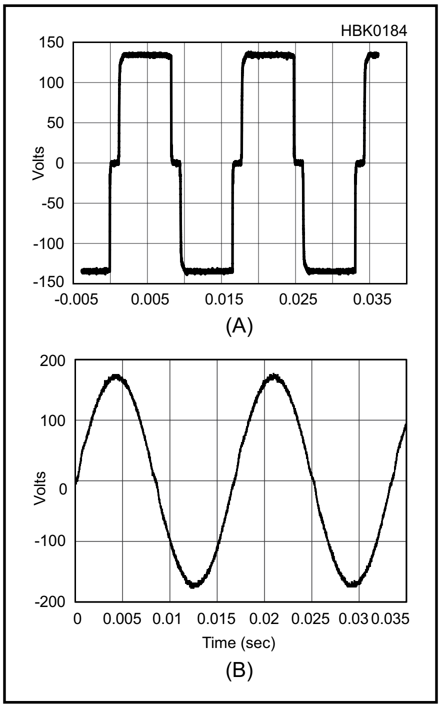


Fig 7.57 — Output waveforms of typical dc-ac inverters. At A, the output of a modified sine wave inverter. Note the stepped square waves. At B, the output of a "pure sine wave" inverter. Note the close approximation of a commercial ac sine wave.

7.13.8 Selecting a Battery for Mobile Operation

There are two basic modes of mobile operation; *in-motion*, and *stationary*. Each mode has unique power requirements and thus different battery requirements. To satisfy these needs, there are three basic battery types. It is important to understand the differences between the types and to apply them properly.

Standard vehicle batteries are referred to as *SLI* (starting, lights, ignition) or just *starter* batteries. Their primary function is to start the engine and then act as a power filter for the alternator which is the actual long-term power source. The most important rating of an SLI battery is the *cold cranking amps* (CCA) rating — the number of amps that the battery can produce at 32 °F (0 °C) for 30 seconds.

Batteries designed for repeated cycles of charging and discharging use are often called *deep-cycle* although the term is widely over-used. A true deep-cycle battery is designed to be repeatedly discharged to 20% remaining capacity. The term "deep cycle" is a misnomer, as all lead-acid batteries are considered discharged when their output voltage drops below 10.5 V at some specified current draw as outlined by the Battery Council Institute (BCI — **www.batterycouncil.org**). At 10.5 V, a six-cell lead-acid battery is considered discharged.

Marine batteries are designed to be stored without charging for up to two years, yet still maintain enough power to start a marine engine. Contrary to common practice, they're not really designed for extended low-current power delivery. Marine batteries often have hybrid characteristics between SLI and deep-cycle batteries.

To differentiate true deep-cycle batteries from SLI and marine batteries, examine a battery's *reserve capacity* (RC). A battery's RC rating is the number of minutes that the battery can deliver 25 A while maintaining an output voltage above 10.5 V. RC batteries typically have RC ratings 20% or more higher than SLI batteries and perhaps 50% higher under ICAS (intermittent, commercial, and amateur service) conditions. A deep-cycle battery has a lower CCA rating than an SLI battery due to its internal construction that favors long-term power delivery over high-current starting loads.

With these facts in mind, we can now select the correct battery for our style of mobile operating. For in-motion operation with power outputs up to 200 W, a second trunk-mounted battery is seldom needed if the power cabling wire size is chosen correctly. (See the **Assembling a Station** chapter, section 29.2, for information on wire sizes in mobile applications.)

When an amplifier is added for higher output power levels, it is often less expensive to add a second trunk-mounted battery than to install larger cables to the main battery. In these cases, the second battery can be of almost any type, as long as it is lead-acid. The second battery should be connected in parallel to the vehicle's main SLI battery. The battery's ampere-hour rating should be close to that of the vehicle's main SLI battery.

All secondary wiring should be properly fused, as outlined in the **Assembling a Station** chapter's on mobile installations. The use of relays and circuit breakers should be avoided. Remember, should a short circuit occur, good-quality lead-acid batteries can deliver upwards of 3000 A which exceeds the break circuit ratings of most relays and circuit breakers. A better solution is a FET switch such as those made by Perfect Switch (**www.perfectswitch.com**).

Assuming the second battery is mounted inside the vehicle's passenger compartment or in the trunk, it should be an AGM type. AGM (Absorption Glass Mat) batteries do not outgas explosive hydrogen gas under normal operating conditions. Flooded (liquid electrolyte) batteries should *never* be used in an enclosed environment.

For stationary operation, select a battery with a large RC rating because it will not be continuously charged. There are two main considerations; the ampere-hour rating (Ah) and the reserve capacity rating, typically listed as C/8, C/10, or C/20, with units of hours. (C is the battery's capacity in Ah.) Dividing the Ah rating by the load amperage (8, 10 or 20 A) will give you the reserve capacity in hours, but the actual ampere-hours any given battery can deliver before the voltage reaches 10.5 V (nominal discharge level) will vary with the load, both average and peak. Heavier loads will reduce the actual ampere-hours available.

Automotive batteries are arranged in BCI group sizes (**www.rtpnet.org/teaa/bcigroup.html**), from 21 through 98. Generally speaking, the larger the group size the larger the battery and the higher the Ah rating. For example, size 24 (small car) has an average rating of 40 Ah, and size 34 (large car) has an average rating of 55 Ah. Exact ratings, including their reserve capacity, are available from the manufacturers' Web sites listed below. A good rule of thumb is to select a battery as physically large as you have room for, consistent with the highest RC rating for any given Ah rating.

Batteries are heavy, and need to be properly secured inside a battery box or by using factory-supplied brackets. For example, a BIC group 34 (average SLI size) battery weighs about 55 pounds. Some battery models (such as the Optima) come supplied with mounting brackets and terminal protection covers. Even though battery boxes aren't always needed, they should be used as a safety precaution to prevent accidental contact with the terminals and can protect the battery from external items. Battery restraints should be adequate to provide 6Gs of lateral and 4Gs of vertical retention, ruling out sheet metal screws and most webbing material. Use the proper brackets!

There are three other considerations: isolating the battery electrically, recharging the battery, and output voltage regulation. Diode-based battery isolators are not all equal. Models with FET bypass switches are the preferred type because of the low voltage-drop across the FET.

If you have wired the battery in parallel with the vehicle's main SLI battery, recharging is taken care of whenever the vehicle is running. If you plan on operating in stationary mode, you'll need a separate recharging system. Most vehicle factory-installed trailer wiring systems also include a circuit for charging RV or boat "house" batteries. Check with your dealer's service personnel about these options.

Voltage regulators, commonly called "battery boosters," are almost a necessity for stationary operation. A model with a low-voltage cutoff should be used to avoid discharging the battery below 10.5 V, as discharging a lead-acid battery beyond this point drastically reduces its charge-cycle life — the number of full-charge/full-discharge cycles. (See the November 2008 *QST* Product Review column.)

For additional information on battery ratings, sizes, and configuration, visit these Web sites:

optimabatteries.com, www.exide.com
www.interstatebatteries.com
www.lifelinebatteries.com

7.14 Glossary of Power Supply Terms

Bleeder — A resistive load across the output or filter of a power supply, intended to quickly discharge stored energy once the supply is turned off.

Boost converter — A switchmode converter in which the output voltage is always greater than or equal to the input voltage.

Buck converter — A switchmode converter in which the output voltage is always less than or equal to the input voltage.

Buck-boost converter — A switchmode converter in which the magnitude of the output voltage can be either greater or less than the input voltage.

C-rate — The charging rate for a battery, expressed as a ratio of the battery's ampere-hour rating.

CCA (cold cranking amps) — A measure of a battery's ability to deliver high current to a starter motor.

Circular mils — A convenient way of expressing the cross-sectional area of a round conductor. The area of the conductor in circular mils is found by squaring its diameter in mils (thousandths of an inch), rather than squaring its radius and multiplying by pi. For example, the diameter of 10-gauge wire is 101.9 mils (0.1019 inch). Its cross-sectional area is 10380 CM, or 0.008155 square inches.

Core saturation (magnetic) — That condition whereby the magnetic flux in a transformer or inductor core is more than the core can handle. If the flux is forced beyond this point, the permeability of the core will decrease, and it will approach the permeability of air.

Crowbar — A last-ditch protection circuit included in many power supplies to protect the load equipment against failure of the regulator in the supply. The crowbar senses an overvoltage condition on the supply's output and fires a shorting device (usually an SCR) to directly short-circuit the supply's output and protect the load. This causes very high currents in the power supply, which blow the supply's input-line fuse.

Darlington transistor — A package of two transistors in one case, with the collectors tied together, and the emitter of one transistor connected to the base of the other. The effective current gain of the pair is approximately the product of the individual gains of the two devices.

DC-DC converter — A circuit for changing the voltage of a dc source to ac, transforming it to another level, and then rectifying the output to produce direct current.

Deep-cycle — A battery designed for repeated charge-discharge cycles to 20% of remaining capacity.

Equalizing resistors — Equal-value bypassing resistors placed across capacitors connected in series for use in a high-voltage power supply to keep the voltages across the capacitors in the string relatively constant.

Fast recovery rectifier — A specially doped rectifier diode designed to minimize the time necessary to halt conduction when the diode is switched from a forward-biased state to a reverse-biased state.

Flyback converter — A transformer-coupled version of the **buck-boost converter**.

Forward converter **—** A **buck converter** with multiple isolated outputs at different voltage levels and polarities.

Foldback current limiting — A special type of current limiting used in linear power supplies, which reduces the current through the supply's regulator to a low value under short circuited load conditions in order to protect the series pass transistor from excessive power dissipation and possible destruction.

Ground fault (circuit) interrupter (GFI or GFCI) — A safety device installed between the household power mains and equipment where there is a danger of personnel touching an earth ground while operating the equipment. The GFI senses any current flowing directly to ground and immediately switches off all power to the equipment to minimize electrical shock. GFCIs are now standard equipment in bathroom and outdoor receptacles.

Input-output differential — The voltage drop appearing across the series pass transistor in a linear voltage regulator. This term is usually stated as a minimum value, which is that voltage necessary to allow the regulator to function and conduct current. A typical figure for this drop in most three-terminal regulator ICs is about 2.5 V. In other words, a regulator that is to provide 12.5 V dc will need a source voltage of at least 15.0 V at all times to maintain regulation.

Inverter — A circuit for producing ac power from a dc source.

Li-ion — Lithium-ion, a type of rechargeable battery that is about ⅓ the weight and 1/2 the volume of a **NiCd** battery of the same capacity.

Low dropout regulator — A three-terminal regulator designed to work with a low minimum input-output differential value.

Marine — A battery designed to retain significant energy over long periods of time without being continuously charged.

NiCd — Nickel cadmium, a type of rechargeable battery.

NiMH — Nickel metal hydride, a type of rechargeable battery that does not contain toxic substances.

Peak inverse voltage (PIV) — The maximum reverse-biased voltage that a semiconductor is rated to handle safely. Exceeding the peak inverse rating can result in junction breakdown and device destruction.

Power converter — Another term for a power supply.

Power processor — Another term for a power supply.

Primary battery — A battery intended for one-time use and then discarded.

RC (reserve capacity) — A measure of a battery's ability to deliver current over long periods.

Regulator — A device (such as a Zener diode) or circuitry in a power supply for maintaining a constant output voltage over a range of load currents and input voltages.

Resonant converter — A form of dc-dc converter characterized by the series pass switch turning on into an effective series-resonant load. This allows a zero current condition at turn-on and turn-off. The resonant converter normally operates at frequencies between 100 kHz and 500 kHz and is very compact in size for its power handling ability.

Ripple — The residual ac left after rectification, filtration and regulation of the input power.

RMS — *R*oot *M*ean *S*quare. Refers to the effective value of an alternating voltage or current, corresponding to the dc voltage or current that would cause the same heating effect.

Secondary battery — A battery that may be recharged many times. Also called a *storage battery*.

Secondary breakdown — A runaway failure condition in a transistor, occurring at higher collector-emitter voltages, where hot spots occur due to (and promoting) localization of the collector current at that region of the chip.

Series pass transistor, or pass transistor — The transistor(s) that control(s) the passage of power between the unregulated dc source and the load in a regulator. In a linear regulator, the series pass transistor acts as a controlled resistor to drop the voltage to that needed by the load. In a switch-mode regulator, the series pass transistor switches between its ON and OFF states.

SLI (starter, lights, ignition) — An automotive battery designed to start the vehicle and provide power to the lighting and ignition systems.

SOA (Safe Operating Area) — The range of permissible collector current and collector-emitter voltage combinations where a transistor may be safely operated without danger of device failure.

Spike — See transient

Surge — A moderate-duration perturbation on a power line, usually lasting for hundreds of milliseconds to several seconds.

Switching regulator — Another name for a switchmode converter.

Switchmode converter — A high-efficiency switching circuit used for dc-dc power conversion. Switching circuits are usually much smaller and lighter than conventional 60 Hz, transformer-rectifier circuits because they operate at much higher frequencies — from 25 to 400 kHz or even higher.

Three-terminal regulator — A device used for voltage regulation that has three leads (terminals) and includes a voltage reference, a high-gain error amplifier, temperature-compensated voltage sensing resistors and a pass element.

Transient — A short perturbation on a power line, usually lasting for microseconds to tens of milliseconds.

Varistor — A surge suppression device used to absorb transients and spikes occurring on the power lines, thereby protecting electronic equipment plugged into that line. Frequently, the term MOV (*M*etal *O*xide *V*aristor) is used instead.

Volt-Amperes (VA) — The product obtained by multiplying the current times the voltage in an ac circuit without regard for the phase angle between the two. This is also known as the apparent power delivered to the load as opposed to the actual or real power absorbed by the load, expressed in watts.

Voltage multiplier — A type of rectifier circuit that is arranged so as to charge a capacitor or capacitors on one half-cycle of the ac input voltage waveform, and then to connect these capacitors in series with the rectified line or other charged capacitors on the alternate half-cycle. The voltage doubler and tripler are commonly used forms of the voltage multiplier.

Voltage regulation — The change in power supply output voltage with load, expressed as a percentage.

7.15 References and Bibliography

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1) Landee, Davis and Albrecht, *Electronic Designer Handbook*, 2nd edition, (McGraw-Hill, 1977), page 12-9. This book can frequently be found in technical libraries and used book stores. The power supply section is well worth reading.

2) Severns and Bloom, *Modern DC-to-DC Switchmode Power Converter Circuits,* (Van Nostrand Reinhold, 1984, ISBN: 0-442-21396-4). A reprint of this book is currently available at the Power Sources Manufacturers Association Web site, **www.psma.com**.

3) Fair-Rite Web site, **www.fair-rite.com**

4) **www.mag-inc.com**

OTHER RESOURCES

M. Brown, *Power Supply Cookbook,* (Butterworth-Heinemann, 2001).

J. Fielding, ZS5JF, *Power Supply Handbook,* (RSGB, 2006).

K. Jeffrey, *Independent Energy Guide,* (Orwell Cove Press, 1995). Includes information on batteries and inverters.

7.16 Power Supply Projects

Construction of a power supply can be one of the most rewarding projects undertaken by a radio amateur. Whether it's a charger for batteries, a low-voltage, high-current monster for a new 100-W solid-state transceiver, or a high-voltage supply for a new linear amplifier, a power supply is basic to all of the radio equipment we operate and enjoy. Final testing and adjustment of most power-supply projects requires only a voltmeter, and perhaps an oscilloscope — tools commonly available to most amateurs.

General construction techniques that may be helpful in building the projects in this chapter are outlined in the **Construction Techniques** chapter. Other chapters in the *Handbook* contain basic information about the components that make up power supplies.

Safety must always be carefully considered during design and construction of any power supply. Power supplies contain potentially lethal voltages, and care must be taken to guard against accidental exposure. For example, electrical tape, insulated tubing ("spaghetti") or heat-shrink tubing is recommended for covering exposed wires, components leads, component solder terminals and tie-down points. Whenever possible, connectors used to mate the power supply to the outside world should be of an insulated type designed to prevent accidental contact.

Connectors and wire should be checked for voltage and current ratings. Always use wire with an insulation rating higher than the working voltages in the power supply. For supply voltages above 300 V, use wire with insulated rated accordingly. The **Component Data and References** chapter contains a table showing the current-carrying capability of various wire sizes. Scrimping on wire and connectors to save money could result in flashover, meltdown or fire.

All fuses and switches should be placed in the hot circuit(s) only. The neutral circuit should not be interrupted. Use of a three-wire (grounded) power connection will greatly reduce the chance of accidental shock. The proper wiring color code for 120-V circuits is: black — hot; white — neutral; and green — ground. For 240-V circuits, the second hot circuit generally uses a red wire.

POWER SUPPLY PRIMARY-CIRCUIT CONNECTOR STANDARD

The International Commission on Rules for the Approval of Electrical Equipment (CEE) standard for power-supply primary-circuit connectors for use with detachable cable assemblies is the CEE-22. The CEE-22 has been recognized by the ARRL and standards agencies of many countries. Rated for up to 250 V, 6 A at 65 °C, the CEE-22 is the most commonly used three-wire (grounded), chassis-mount primary circuit connector for electronic equipment in North America and Europe. It is often used in Japan and Australia as well.

When building a power supply requiring 6 A or less for the primary supply, a builder would do well to consider using a CEE-22 connector and an appropriate cable assembly, rather than a permanently installed line cord. Use of a detachable line cord makes replacement easy in case of damage. CEE-22 compatible cable assemblies are available with a wide variety of power plugs including most types used overseas.

Some manufacturers even supply the CEE-22 connector with a built-in line filter. These connector/filter combinations are especially useful in supplies that are operated in RF fields. They are also useful in digital equipment to minimize conducted interference to the power lines.

CEE-22 connectors are available in many styles for chassis or PC-board mounting. Some have screw terminals; others have solder terminals. Some styles even contain built-in fuse holders.

7.16.1 Four-Output Switching Bench Supply

This project by Larry Cicchinelli, K3PTO, describes the four-output bench power supply shown in **Fig 7.58** with three positive outputs and one negative output. The three positive outputs use identical switching regulator circuits that can be set independently to any voltage between 3.3 V and 20 V at up to 1 A. The fourth output is a negative regulator capable of about 250 mA. As built, the supply has two fixed outputs and two variable outputs, but any module can be built with variable output. (Construction diagrams and instructions, a complete parts list, and additional design details are included on this *Handbook's* CD-ROM.)

The only dependency among the outputs is that they are all driven by a single transformer. The transformer used is rated at 25 V and 2 A — good for 50 W. Assuming that the regulator IC being used has a 75% efficiency, a total of about 37 W is available from the power supply outputs.

One of the features of a switching regulator is that you can draw more current from the outputs than what the transformer is supplying — at a lower voltage, of course — as long as you stay within the 37 W limit and maximum current for the regulator. Most of the discussion in this article will be about the positive regulators as the negative regulator was an add-on after the original system was built.

POSITIVE REGULATOR

Fig 7.59 is the circuit for the positive regulator modules — a buck-type regulator. There are several variations of the circuit, any of which you can implement.

- L2 and C4 are optional. These two components implement a low-pass filter that will decrease high frequency noise that might otherwise appear at the output.
- The pads for R1 will accommodate a small, multi-turn potentiometer. You can insert one here or you can use the pads to connect a panel-mounted potentiometer.
- If you want a fixed output you can simply short out R1 and use R2 by itself.
- You can also insert a fixed resistor in the R1 position in the case where the calculated value is non-standard and you want to use two fixed resistors.

The formula for setting output voltage us-

Fig 7.58 — The front panel of the four-output switching supply.

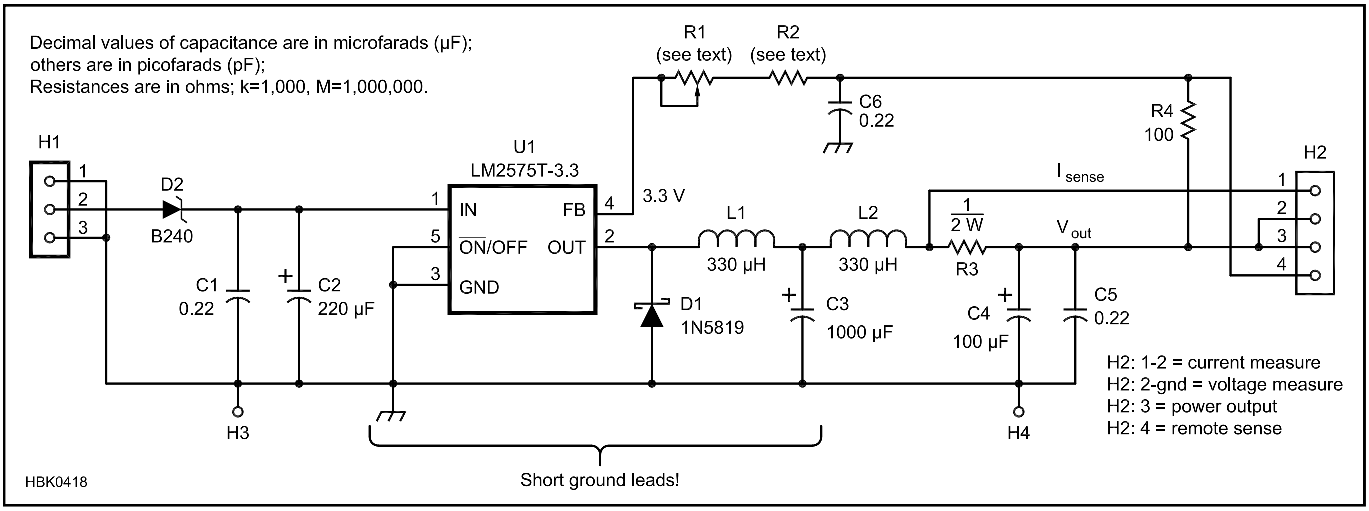


Fig 7.59 — The positive buck-type switchmode regulator uses the LM2575-3.3, a fixed-voltage regulator, with an external voltage-set resistor (R1 + R2). See text for details of the calculations needed to determine the value of R1 and R2. As noted in the text, these values are the total resistance for both parts, and can be made from one fixed resistor, one variable resistor or a combination. Some common values (R1 + R2 total) are: For a 12 V fixed supply, 7.1 kΩ; for 5 V, 1.4 kΩ; for a 3.3 to 20 V variable supply, 0-13.7 kΩ (use a 15 kΩ pot); for 5 to 15 V, 1.4-9.6 kΩ (use 1 kΩ fixed-value resistor and a 1 kΩ pot). A full parts list is included on the *Handbook* CD-ROM.

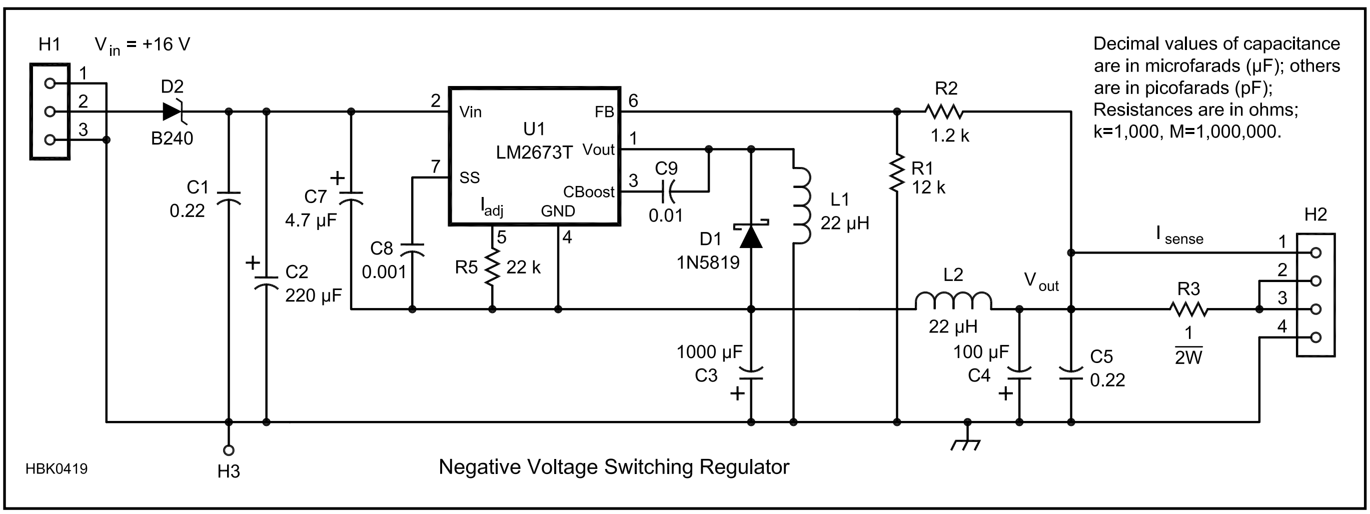


Fig 7.60 — The negative regulator uses the LM2673T in a buck-boost circuit. This circuit inverts the output voltage from the input voltage. A full parts list is included on the *Handbook* CD-ROM.

ing the 3.3 V version of the regulator is based on knowing the current (in mA) through the regulator's internal voltage divider = 3.3 V / 2.7 kΩ = 1.22 mA. The sum of R1 and R2 must cause the voltage at the regulator FB pin to equal 3.3 V. Thus, R1 + R2 in kΩ = $(V_{out} - 3.3) / 1.22$ and $V_{out} = 1.22\,(R1 + R2) + 3.3$. If R1 = R2 = 0, a direct connection from the output voltage to the FB pin, the calculation results in an output of 3.3 V. The leakage current of the Error Amplifier in the regulator is somewhat less then 25 nA so it can be ignored. The values for R1 and R2 are shown above the schematic for Fig 7.59.

The only critical parts are R1 and R2 which form the voltage dividers for the regulator module. Even their values can be changed, within reason, as long as the ratios are maintained. If you want to have an accurate, fixed output voltage, select a value for R2 that is lower than the calculated value and use a potentiometer for R1 to set the voltage exactly. The value of C3 is not especially critical; however, it should be a low-ESR (equivalent series resistance) type that is intended for use in switchmode circuits.

NEGATIVE REGULATOR

The negative regulator is a buck-boost configuration — it converts a positive voltage into a negative one — see **Fig 7.60**. This design uses many of the same component values as the positive regulators except the regulator IC is an LM2673 to improve circuit stability. The author was unable to implement the current measuring circuit within a feedback loop. Several configurations introduced a significant low frequency noise component to the output voltage. There was also some 50 kHz noise present on the output, but an additional low-pass filter on the output reduced it considerably.

REMOTE SENSING

Many power supplies use remote sensing to electronically compensate for the voltage drop in the wires carrying current to the load. Even with relatively short wires, there can be significant voltage drop between the regulator and its load. There is provision for remote sensing in this circuit described in the sup-

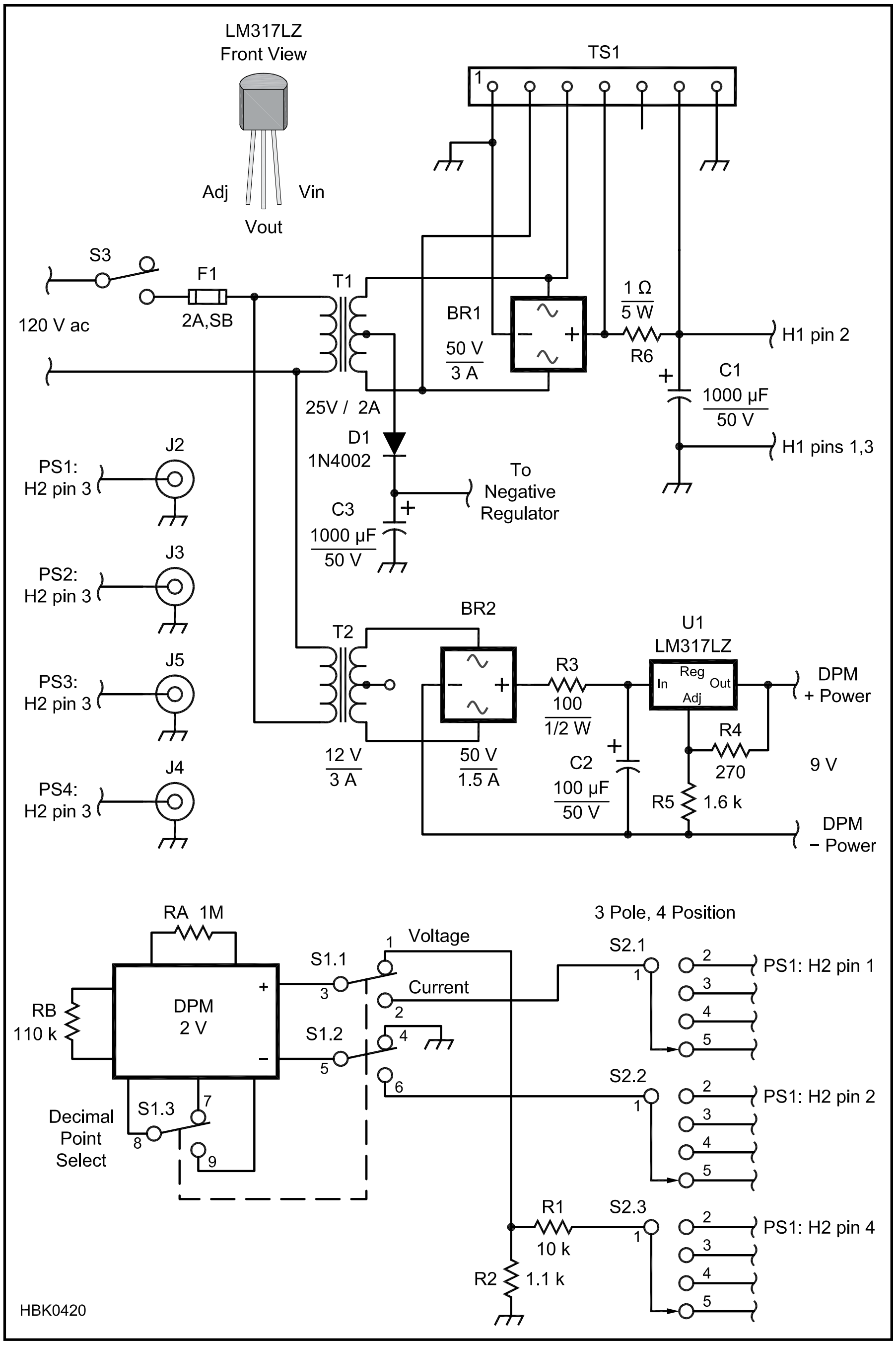


Fig 7.61 — Rectifier and metering schematic. The panel meter is switched between the four modules with a rotary switch (S2) and between voltage and current with a 3PDT toggle (S1). A separate rectifier provides power for the negative supply and a separate three-terminal regulator circuit provides power to the DPM. A full parts list is included on the *Handbook* CD-ROM.

port information for this project on the book's CD-ROM.

If you are not going to use remote sensing then you should insert a jumper in place of R4 in Fig 7.59. R4 (100 Ω) is there for protection just in case the remote sense connection is missing. If you do not want to use remote sensing you can simplify the digital panel meter (DPM) switch wiring to use a two-pole switch instead of the three-pole model listed. In this case, do not use S2.2 and connect S1.2 to the common of S2.3 instead of S2.2.

RECTIFIER CIRCUIT

Fig 7.61 shows the connections among the parts of the system: regulator boards, the digital panel meter (DPM), and rectifier circuit. The components used for the main rectifier circuit are mounted on a terminal strip. You can see the terminal strip and R6 at the top-left of **Fig 7.62**.

Barely visible at the top left, C1 is mounted underneath the terminal strip. The leads of the bridge rectifier are soldered into the holes where the terminals are riveted to the insulating strip. One of the four leads, the negative output, is soldered to a grounded terminal.

THE DIGITAL PANEL METER

Another feature of the unit is the DPM which can be switched to measure the output voltage (H2 pin 1 to ground) as well as the current draw (voltage between H2 pins 1 and 2) for each of the positive supplies. Fig 7.61 shows the 3-pole, 4-position rotary switch (S2) that selects which power supply to monitor and a 3PDT toggle switch (S1) that selects between measuring voltage and current.

In order to measure the voltage drop across the 1-Ω current sense resistors, the DPM needs either an isolated power supply or some more circuitry. This system uses an isolated power supply. A series regulator is used simply because they are somewhat easier to implement and the DPM has a very low current requirement. All components except the transformer are mounted on a piece of perforated board that you can see just below the transformer in Fig 7.62.

Since the 1.2 mA current for the feedback circuit flows through the current sense resistor it will be included in the value displayed by the DPM when current is selected.

The DPM also has a set of jumpers that allow you to set the decimal point location. As can be seen in Fig 7.61, one pole of the toggle switch selects its location.

CONSTRUCTION DETAILS

Both the DPM and the regulators use pin headers for all of the connections that come off the boards (see the parts list for details). This allows assembly of the subsystems without having to consider any attached wires. Wire lengths can be determined later, then install the mating connectors on the wires and simply push them onto the pins.

The printed circuit board (PCB) for the positive regulators contains four identical circuits. The boards could be separated, but a single board made mounting the board and heat sinking the regulator ICs easier. Regulator ICs are mounted on the bottom side instead of on the top and folded over with the flat side parallel to the PCB and farthest from the board (see **Fig 7.63**). If you fold the ICs identically, you can use their mounting holes to fasten them to the side of the enclosure. This not only is a convenient method of mounting the board it also gives the ICs a good heat sink and ground! Fig 7.62 shows the completed assembly bolted to the side of the enclosure.

The negative regulator was an addition to the positive-output system, so it is installed separately from the positive regulators on the bottom of the enclosure. If you already have a positive voltage power supply you can build the negative regulator circuit and simply connect its input to the output of your existing supply.

A caution regarding the circuit boards is in order. PC boards are available from FAR Circuits (**www.farcircuits.net**), a company that provides a lot of boards for ham-related projects. Their boards do not have plated through holes so you will have to be sure that you solder the through-hole components on *both* sides of the board.

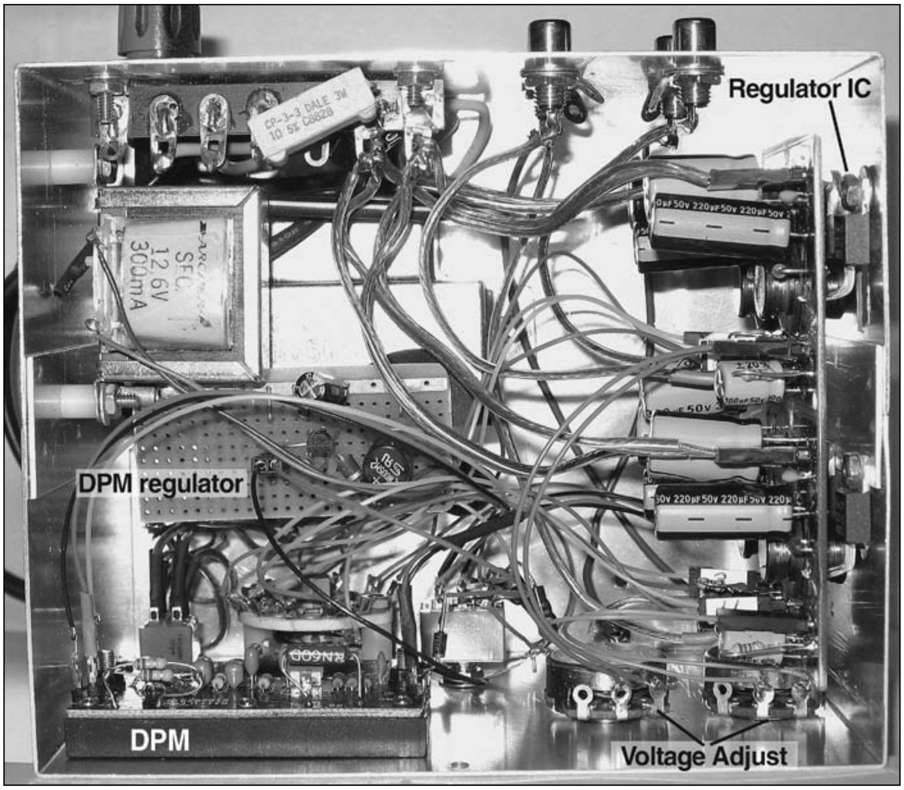


Fig 7.62 — Internal supply wiring. The positive regulator boards are attached to the enclosure wall by the mounting tabs of the regulator ICs. This provides good heat sinking for the regulators.

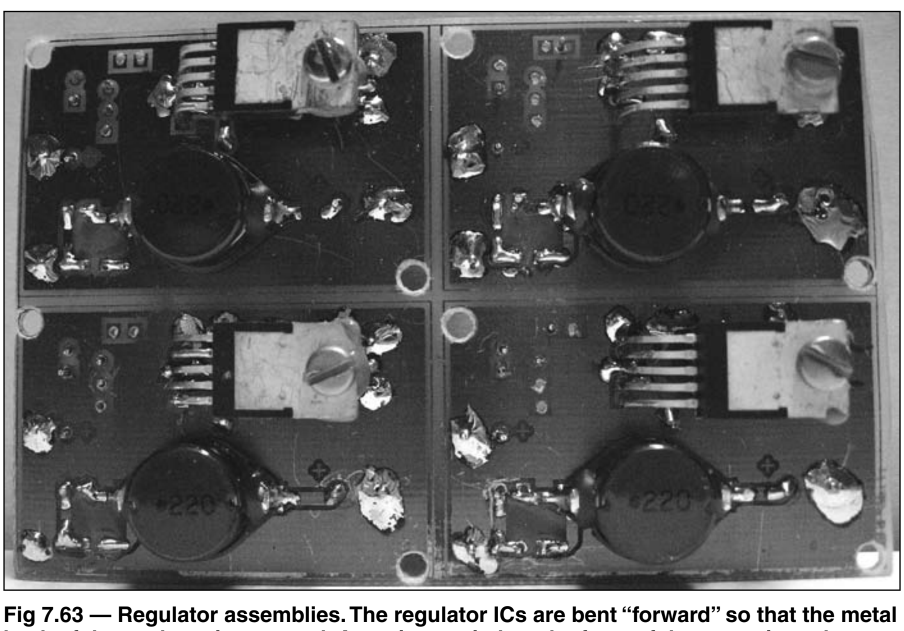


Fig 7.63 — Regulator assemblies. The regulator ICs are bent "forward" so that the metal back of the package is exposed. A nut is epoxied on the front of the mounting tab so that a screw can be inserted through the enclosure wall to mount the regulator.

Artwork for the PC board layout, Gerber files, and a drill file are available on the CD-ROM included with this book. The schematic capture software *DipTrace* was used in the development of this project. Source files for the schematic and PCB files are also available on the CD-ROM.

RFI

The power supply, in its aluminum enclosure, was put into a completely shielded box with an 18 inch antenna within a few inches of it. An 8-Ω, 20-W resistor was installed on the #2 output with the output voltage adjusted to +8 V. Using a calibrated EMI measurement system, the only emission that would have failed FCC Part 15 testing occurs around 88 MHz as shown in **Fig 7.64**.

7.16.2 12-V, 15-A Linear Power Supply

The supply shown in **Fig 7.65** is a linear 12-V, 15-A design by Ed Oscarson, WA1TWX, with adjustable output voltage and current limiting. Supply regulation is excellent, typically exhibiting a change of less than 20 mV from no load to 15 A. This basic design, with heftier components and additional pass transistors, can deliver over 30 A. (All numbered notes, additional circuit design information, a discussion of how to change the supply voltage and/or current ratings, construction notes, and a complete parts list are available on the CD-ROM included with this *Handbook.*)

CIRCUIT DESCRIPTION

Fig 7.66 is the supply's schematic. The ac line input is fused by F1, switched on and off by S1 and filtered by FL1. F1 and S1 are rated at about ¼ of the output current requirement (for 15-A output, use a 4- or 5-A slow-blow fuse or a similarly rated circuit breaker). FL1 prevents any RF from the secondary or load from coupling into the power line and prevents RF on the power line from disturbing supply operation. If your ac power line is clean, and you experience no RF problems, you can eliminate FL1, but it's inexpensive insurance.

When discharged, filter capacitor C1 looks like a short circuit across the output of rectifier U2 when ac power is applied. That usually subjects the rectifier and capacitor to a large inrush current, which can damage them. Fortunately a simple and inexpensive means of inrush-current limiting is available. Keystone Carbon Company (and others) produce a line of inrush-current limiters (thermistors) for this purpose. The device (RT1) is placed in series with one of the transformer primary leads. RT1 has a current rating of 6 A,[1] and

[1]See the *Handbook* CD ROM for all numbered notes and additional construction information.

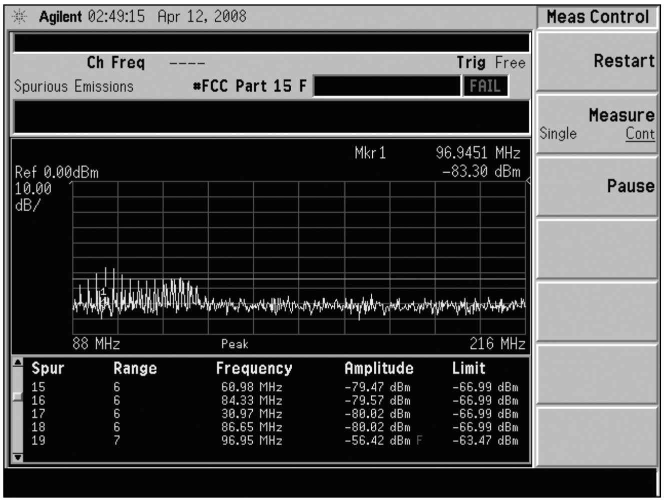


Fig 7.64 — Spectrum of the supply's RF emissions. The only frequency at which emissions exceed FCC Part 15 limits is near 88 MHz.

a cold resistance of 5 Ω. When it's hot, RT1's resistance drops to 0.11 Ω. Such a low resistance has a negligible effect on supply operation. Thermistors run *hot* so they must be mounted in free air, and away from anything that can be damaged by heat.[2,3]

The largest and most important part in the power supply is the transformer (T1). If purchased new, it can also be the most costly. Fortunately, a number of surplus dealers offer power transformers that can be used in this supply.

T1 produces 17 V ac RMS at 20 A; the center tap is not used. Bridge rectifier U2 provides full-wave rectification. Full-wave rectification reduces the ripple component of current that flows in the filter capacitor, resulting in less power dissipation in the capacitor's internal resistance. U2's voltage rating should be at least 50 V, and its current rating about 25% higher than the normal load requirement; a 2-A bridge rectifier will do. U2 is secured to the chassis (or a heat sink) because it dissipates heat.

C1 is a computer-grade electrolytic. Any capacitor value from 15,000 to 30,000 µF will suffice. This version uses a 19,000 µF, 40-V capacitor. The capacitor's voltage rating should be at least 50% higher than the expected no-load rectified de voltage. In this supply, that voltage is 25 V, and a 40-V capacitor provides enough margin,

R5, a 75-Ω, 20-W bleeder resistor, is connected across C1's terminals to discharge the supply when no load is attached or one is removed, Any resistance value from 50 Ω to 200 Ω is fine; adjust the resistor's wattage rating appropriately.

At the terminals of C1, we have a dc voltage, but it varies widely with the load applied. When keying a CW transmitter or switching a rig from receive to full output, 5-V swings can result. The dc voltage also has an ac ripple component of up to 1.5 V under full load. Adding a solid-state regulator (U1) provides a stable output voltage even with a varying input and load.

VOLTAGE REGULATOR IC AND PASS TRANSISTORS

The LM723 used at U1 has a built-in voltage reference and sense amplifier, and a 150-mA drive output for a pass-transistor array. U1's voltage reference provides a stable point of comparison for the internal regulator circuitry. In this supply, it's connected to the non-inverting input of the voltage-sense op amp. The reference is set internally to 7.15 V, but the absolute value is not critical because an output-voltage adjustment (R12) is provided. What is important is that the voltage is stable, with a specified variation of 0.05% per 1000 hours of operation. This is more than adequate for the supply.

For the regulator to work properly, its ground reference must be at the same point as the output ground terminal. The best way to ensure this is to use the output GROUND terminal (J4) as a single-point ground for all of the supply grounds. Run wires to J4 from each component requiring a ground connection. Fig 7.66 attempts to show this graphically through the use of parallel connections to a single circuit node.

The output pass transistor array consists of a TIP112 Darlington-pair transistor (Q5) driving three 2N3055 power transistors (Q1-Q3). This two-stage design is less efficient than connecting the power transistors directly to the LM723, but Q5 can provide considerably more base current to the 2N3055s than the 150-mA maximum rating of the LM723. You can place additional 2N3055s in parallel to increase the output current capacity of the supply.

This design is not fussy about the pass transistors or the Darlington transistor used. Just ensure all of these devices have voltage ratings of at least 40 V. Q5 must have a 5-A (or greater) collector-current rating and a beta of over 100. The pass transistors should be rated for collector currents of 10 A or more, and have a beta of at least 10.[5]

Resistors R17, R18 and R19 prevent leakage current through the collector-base junction from turning on the transistor by diverting it around the base-emitter junction. When the pass transistors are hot, at the V_{CE} encountered in this design, the leakage current can be as high as 3 mA. The resulting drop across the 33-Ω resistors is 0.1 V — safely below the turn-on value for V_{BE}.

When unmatched transistors are simply connected in parallel they usually don't equally share the current.[6] By placing a low-value resistor in each transistor's emitter lead (emitter-ballasting resistors, R1-R3), equal current sharing is ensured. When a transistor with a lower voltage drop tries to pass more current, the emitter resistor's voltage drop increases,

Fig 7.65 — The completed 12-V, 15-A supply. If individual meters are not available, a DMM can be substituted.

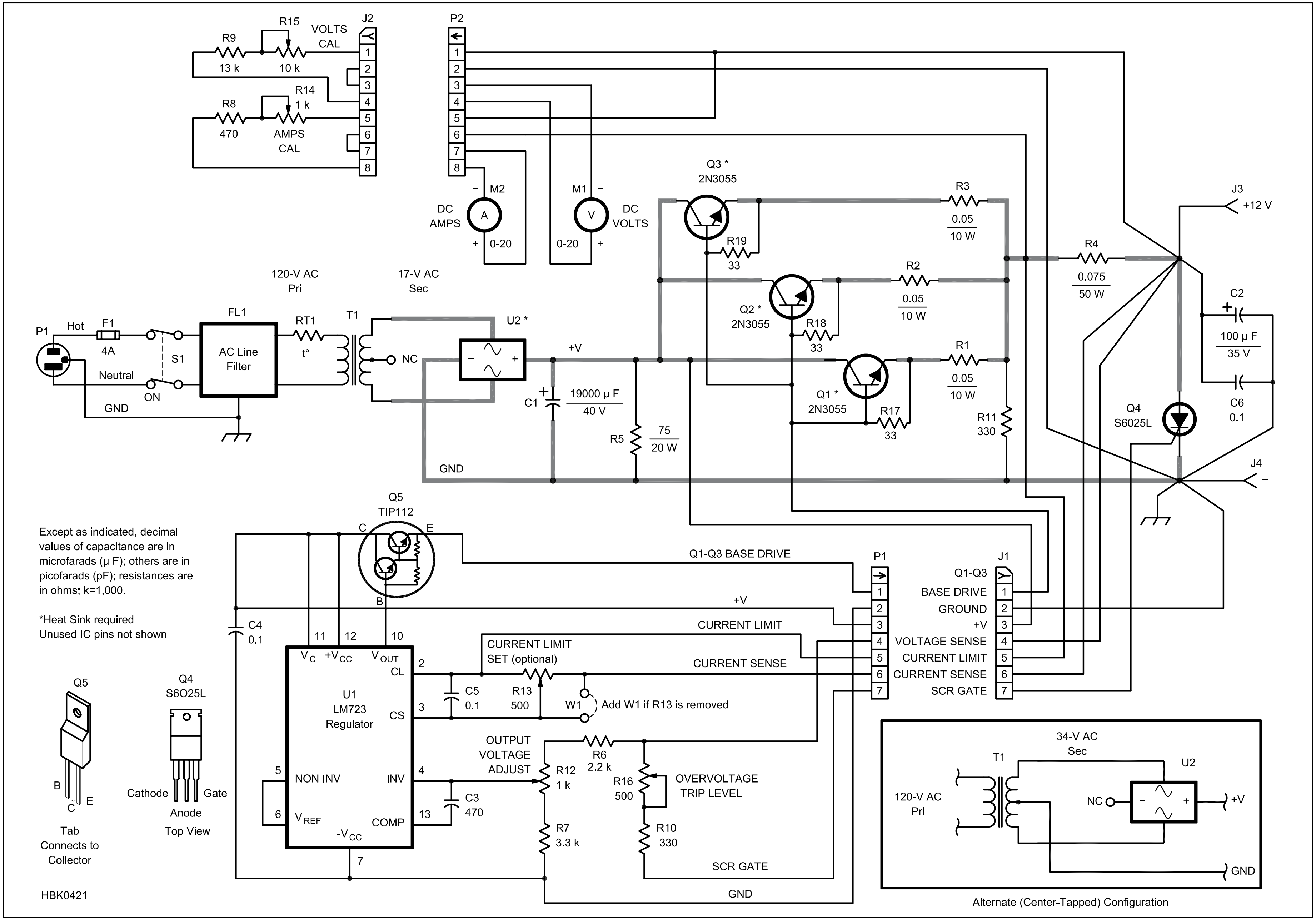


Fig 7.66 — Schematic of the power supply. A parts list may be found on the *Handbook* CD-ROM. Equivalent parts can be substituted. The bold lines indicate high-current paths that should use heavy-gauge (#10 or #12 AWG) wire. This schematic graphically shows wiring to a single-point ground; see text. The majority of the parts used in this supply are available as surplus components.

allowing the other transistors to provide more current. Because the voltage-sense point is on the load side of the resistors, the transistors are forced to dynamically share the load current.

With a 5-A emitter current, 0.25 V develops across each 0.05-Ω resistor, producing 1.25 W of heat. Ideally, a resistor's power rating should be at least twice the power it's called upon to dissipate. To help the resistors dissipate the heat, mount them on a heat sink, or secure them to a metal chassis (as shown in **Fig 7.67**). You can use any resistor with a value between 0.065 and 0.1 Ω, but remember that the power dissipated is higher with higher-value resistors (10 W resistors are used here).

At the high output currents provided by this supply, the pass transistors dissipate considerable power. With a current of 5 A through each transistor-and assuming a 9-V drop across the transistor — each device dissipates 45 W. Because the 2N3055's rating is 115 W when used with a properly sized heat sink, this dissipation level shouldn't present a problem. If the supply is to be used for continuous-duty operation, increase the size of the heat sink and mount it with the fins oriented vertically to assist in air circulation.

The output-voltage sense is connected through a resistive divider to the negative input of U1. U1 uses the difference between its negative and positive inputs to control the pass transistors that in turn provide the output current. C3, a compensation capacitor, is connected between this input and a dedicated compensation pin to prevent oscillation. The output voltage is adjusted by potentiometer R12 and two fixed-value resistors, R6 and R7.[7] The voltage-sense input is connected to the supply's positive output terminal, J3.

Current sensing is done through R4, a 0.075-Ω, 50-W resistor connected between the emitter-ballasting resistors and J3. R4's power dissipation is much higher than that of R1, R2 or R3 because it sees the total output current. At 15 A, R4 dissipates 17 W. At 20 A, the dissipated power increases to 30 W.

U1 provides current limiting via two sense inputs connected across R4. Limiting takes place when the voltage across the sense inputs is greater than 0.65 V.[8] For a 15-A maximum output-current limit, this requires a 0.043-Ω resistor. By using a larger-value sense resistor and a potentiometer, you can vary the current limit. Connecting potentiometer R13 across R4 provides a current-limiting range from full limit voltage (8.7 A limit) to no limit voltage. This allows the current limit to be fine-tuned, if needed, and also permits readily available resistor values (such as the author's 0.075-Ω resistors) to be used. A current limit of 20 A is at the top end of the ammeter scale.

R20 maintains a small load of 35-40 mA depending on power supply output voltage. This reduces the effect of leakage current in the pass transistors and keeps the regulator's feedback action active, even if no external load is connected.

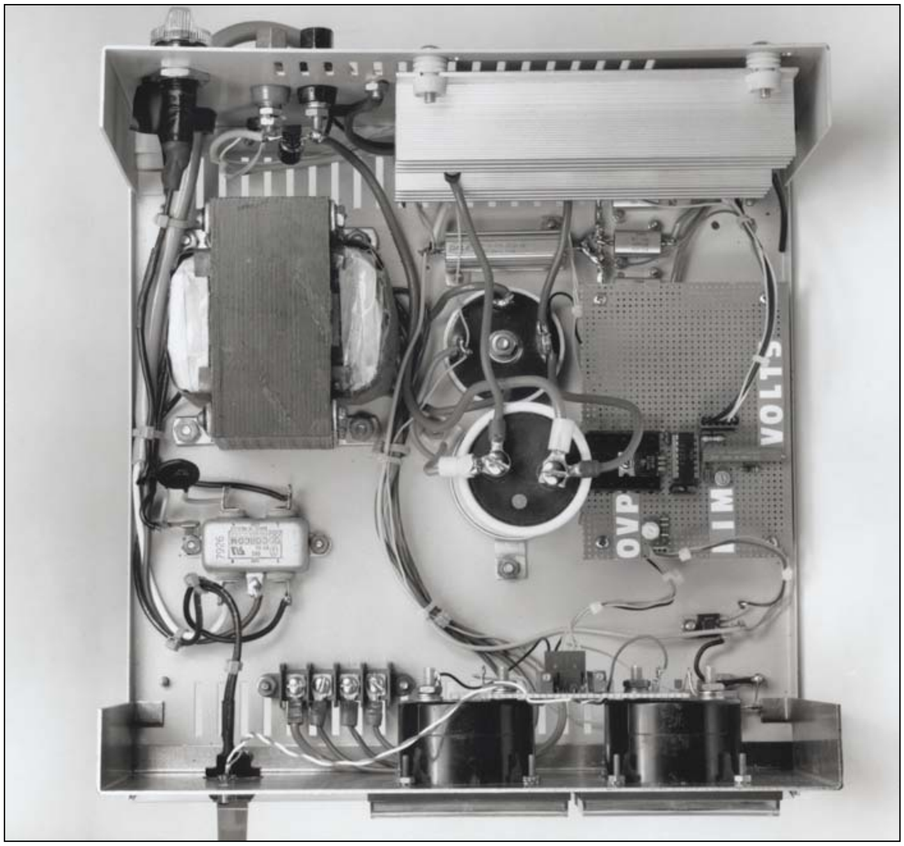


Fig 7.67 — An inside view of the neatly constructed 12-V, 15-A power supply. There's plenty of room in this cabinet to accept the components comfortably and allow hands and tools to get at them. Vents in the bottom and rear of the cabinet provide some measure of convection cooling.

METERING

Voltmeter M1 is a surplus meter. R8 and potentiometer R15 provide for voltmeter calibration. If the correct fixed-value resistor is available, R15 can be omitted. The combined value of the resistor and potentiometer is determined by the full-scale current requirement of the meter used.[9]

Ammeter M2 is actually a voltmeter (also surplus) that measures the potential across R4. The positive side of M2 connects to the high side of R4. R8 and potentiometer R14 connect between the positive output terminal (J3) and the negative side of M2 to provide calibration adjustment. The values of R8 and R14 are determined by the coil-current requirements of the meter used. (Digital panel meters or a dedicated DMM can be used instead of separate analog meters.)

OUTPUT WIRING AND CROWBAR CIRCUIT

The supply output is connected to the outside world by two heavy-duty banana jacks, J3 and J4. C2, a 100-µF capacitor, is soldered directly across the terminals to prevent low-frequency oscillation. C6, a 0.1-µF capacitor, is included to shunt RF energy to ground. Heavy-gauge wire must be used for the connections between the pass transistors and J3 and between chassis ground and J4. The voltage-sense wire must connect directly to J3 and U2's ground pin must connect directly to J4 (see Fig 7.66). This provides the best output voltage regulation.

An over-voltage crowbar circuit prevents the output voltage from exceeding a preset limit. If that limit is exceeded, the output is shunted to ground until power is removed. If the current-limiting circuitry in the supply is working properly, the supply current-limits to the preset value. If the current limiting is not functioning, the crowbar causes the ac-line fuse to blow. Therefore, it's important to use the correct fuse size: 4 to 5 A for a 15-A supply.

The crowbar circuit is a simple design based on an SCR's ability to latch and conduct until the voltage source is removed. The SCR (Q4) is connected across output terminals J3 and J4. (The SCR can also be connected directly across the filter capacitor, C1, for additional protection.) R10 and potentiometer R16 in series with the Q4's gate provide a means of adjusting the trip voltage. The prototype crowbar is set to conduct at 15 V. The S6025L SCR is rated at 25 A and should be mounted on a metal chassis or heat sink. (Note: Some SCRs are isolated from their mounting tabs, others are not. The S6025L and the 65-ampere S4065J are isolated types. If the SCR you use is not isolated, use a mica washer or thermal pad to insulate it from the chassis or heat sink.)

The bold lines in Fig 7.66 indicate high-current paths that should use heavy-gauge (#10 or #12) wire. Traces that are connected to the output terminals in the schematic by individual lines should be connected directly to the terminals by individual wires. This establishes a 4-wire measurement, where the heavy wires carry the current (and have voltage drops) and the sense wires carry almost no current and therefore voltage errors are not caused by voltage drops in the wiring. If desired, the sense wire can be carried out to the load, but that may introduce noise into the sense feedback circuit, so use caution if that is done.

CONSTRUCTION

Fig 7.67 shows the inside of the prototype supply. The larger components are chassis mounted; two perf-boards contain the majority of the low-power parts of the supply. This includes the potentiometers for the regulator, meters and over-voltage adjustments. A template to make a PC board that contains all of the parts mounted on the two perf-boards is available in the **Templates** folder of this book's CD-ROM.

Q1, Q2, and Q3 must be mounted on a heat sink. The one used measures 1¼ × 6 × 3⅝ inches (HWD), the minimum size recommended. If you can find a heat sink with vertically oriented fins, so much the better. Use a small amount of heat-sink compound between the transistors and the heat sink. Because the transistor collectors are at a potential of + 25 V, they must be insulated from the heat sink, or, as in this supply, the entire heat sink can be isolated from the chassis. (It is better to use a grounded heat sink design because of the potential for short-circuits if the external heat sink is electrically "hot." — *Ed.*) If there is adequate ventilation, you can mount the heat sink inside the chassis. The three emitter-ballasting resistors and the sense resistor can be secured to the chassis rear or bottom, but they should be located near the transistors to which they are connected. Orient the components so that they can be easily soldered to the common output connections.

Also mounted on the back panel are a line-cord strain relief and fuse holder. Use a strain relief to prevent the cord from being pulled out of the chassis. (It is preferable to use a standard CEE-22 plug and socket as discussed at the beginning of this section — *Ed.*) Mount the fuse holder directly above the line cord. The output terminals, J3 and J4, are placed in the same area; use heavy-duty banana jacks or terminal blocks. If FL1 is used, mount it on the chassis. Install an insulated terminal strip near the filter to hold the inrush-current limiter,

Once all of the major components are installed, some of the wiring can be done. Wire the line cord to the fuse with the black (hot) lead at the center, and the outer ring connected to the power switch. It's important to connect the green (ground) line-cord wire to the chassis for safety. Connect the transformer secondary directly to the bridge rectifier. Use #12 AWG wire to connect the rectifier output to the filter capacitor. Use crimp-on or solder-on terminal lugs as needed, as at the filter capacitor connections. Connect C1's negative lead directly to the output GROUND terminal, J4. Connect a length of #12 AWG wire from J4 (or C1) to the chassis. J4 is the single-point ground for the rest of the system. The positive connection will be made later. Attach bleeder resistor R5 to C1.

At this point, you should test the basic dc supply. When ac power is applied, about 20 to 28 V dc should be present across C1's terminals. This potential is dependent on the transformer used, but should not exceed 30 V dc. Turn off the supply.

Next, wire the output pass transistors. If the transistors are insulated from the heat sink, use #10 AWG wire to connect the collectors together. Leave an 8- to 10-inch pigtail for later connection to C1's positive terminal. If the transistors are mounted directly to the heat sink, the pigtail can be connected to the heat sink.

Use #20 AWG wire to connect the transistor base leads together and provide a pigtail for attachment to U1. Using #12 AWG wire, connect the emitters of Q1-Q3 to their respective emitter-balancing resistors. Solder together the remaining emitter-resistor leads and use #10 AWG wire to connect them to R4. Solder the other side of R4 to J3, the positive output terminal. Connect R17, R18, and R19 directly across the transistor base and emitter terminals. Connect R11 directly from the junction of R4 and the emitter resistors to ground.

Next, attach the 100-µF (C2) and 0.1-µF capacitors (C6) across the output terminals. Keep the leads as short as possible, especially those of C6.

Once the regulator board is wired, attach its mating connector to the appropriate points on the chassis. The voltage-sense wire and ground wires must connect directly to the appropriate output terminals. Use #20 or #22 AWG wire for the voltage-sense wire and #18 AWG for the ground.

Attach the current-limit and current-sense wires directly to R4. This is essential for proper regulation and current limiting. There is little current in the wires, so use #20 or #22 AWG wire here and for the power, SCR gate and base-drive connections.

Q4 connects across J3 and J4. Q4's gate is attached to the PC board SCR gate connection at J1, pin 7. Set potentiometer R16 to its maximum resistance or disconnect the SCRs gate prior to testing the supply.

TESTING

Initial testing is done without a load. Use a 2-A fuse at F1 to protect the components in case of problems. If any of the steps do not produce the expected results, check the circuit wiring.

Connect a voltmeter to the output terminals. Turn on the supply. The voltmeter should read between 8 and 15 V. Adjust R12 for a 12-V output (13.8 V if supplying a radio). Adjust R15 for a 13.8-V reading on the front-panel meter. Turn off the supply.

Connect a 12-Ω, 20-W resistor to J3 and attach the other end through an ammeter to J4. (The ammeter must be capable of reading a current now of more than 1 A.) Turn on the supply and measure the output current which should be 1 A. Adjust R14 until ammeter M3 displays 1 A. Turn off the supply.

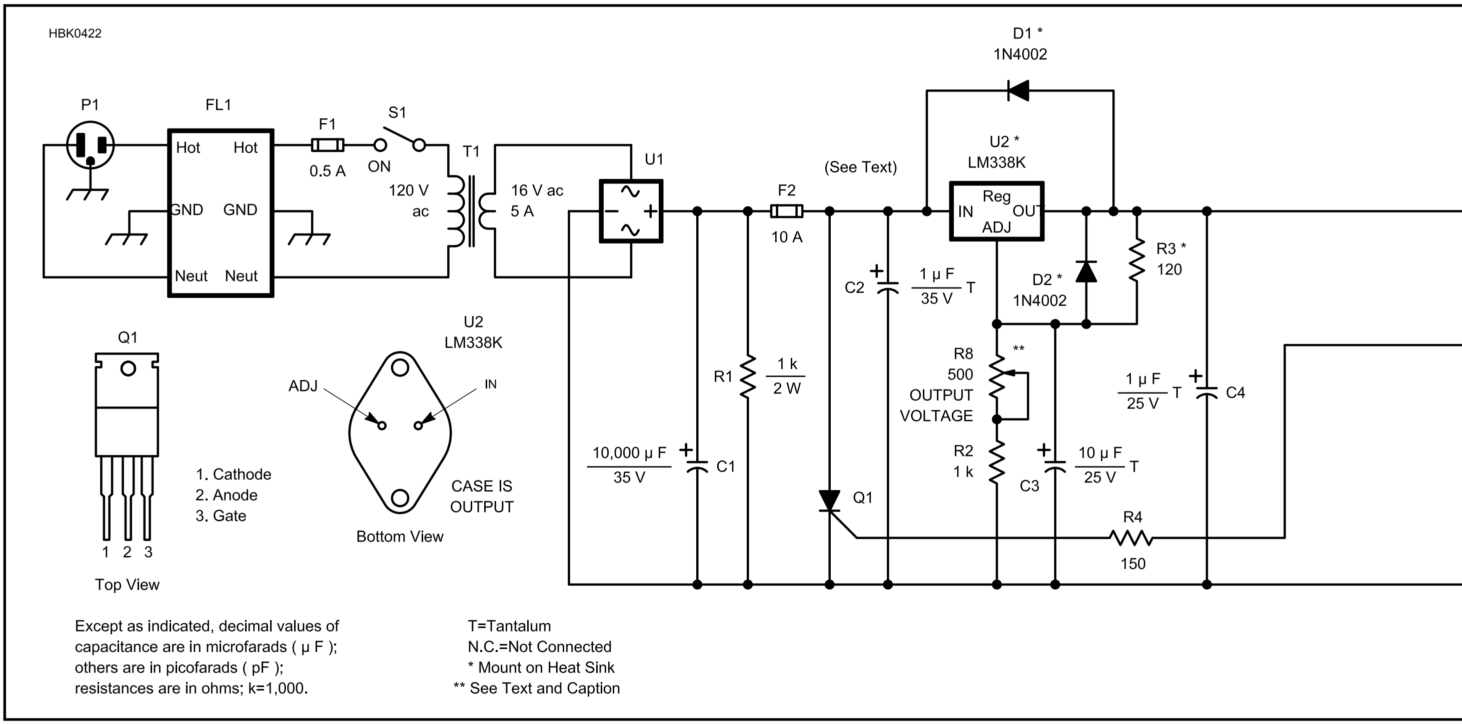
The next test requires a 0.5-Ω load resistor. Use a high-power-dissipation resistor. To provide additional cooling, immerse the resistor in a plastic container of clean water. Connect the resistor to the supply in place of the 12-Ω load resistor previously used. If the ammeter is left in series with the load, it must be capable of reading a current flow of at least 10 A. The front-panel ammeter may also be used to measure the current. Adjust the CURRENT LIMIT SET potentiometer (R13) to the position where the wiper is at the same end of the potentiometer as the terminal that is connected to the output side of R4. This sets the current limit to 8.7 A (if R4 is a 0.075-Ω resistor).[12] Turn on the supply. The ammeter should indicate about 9 A. If it doesn't, immediately turn off the supply. Check the wiring of the current-limiting circuit, including R13.

If the ammeter reading was okay, remove the series-connected ammeter and connect the 0.50-Ω load resistor across the output terminals. Turn on the supply and adjust CURRENT LIMIT SET pot R13 for a 20-A current indication (or the desired limit point). Turn off the supply.

At this point, the output voltage and the current limit are set. You can recalibrate M2 with a 5- or 10-A load to get better meter resolution when adjusting R14.

The following sequence assumes that the desired output voltage is 12 V, and the over-voltage trip point is 15 V. Using R12, set the output voltage to 15. Decrease the resistance of R16 until the SCR trips. When this happens, turn off the power. With the power off, adjust R12 to decrease the voltage. Turn on the supply and readjust R12 for 12 V.

Now, regulation needs to be checked. Connect a voltmeter across the output terminals with no load connected to the supply. Turn



on the supply and record the output voltage. Connect a 10- or 15-A load to the supply and record the voltage. Turn off the supply. The difference between the no-load and 10- or 15-V load voltages should be less than 50 mV. (It is typically less than 20 mV on the prototype.) Higher voltage differences could be caused by the current limit being activated (if near the limit point), or by problems in the sense or single-point ground wiring.

7.16.3 13.8-V, 5-A Linear Power Supply

The power supply shown in **Fig 7.68**, designed by Ben Spencer, G4YNM, provides 13.8 V dc at 5 A, suitable for many low-power transceivers and accessories. It features time-dependent current limiting and short-circuit protection, thermal overload protection within the safe operating area of the regulator IC, and overvoltage protection for the equipment it powers.

Construction, testing and calibration are straightforward, requiring no special skills or equipment. Many of the components can be found in junk boxes, or purchased at hamfests or from mail-order suppliers.

CIRCUIT DESCRIPTION

Fig 7.69 is the power-supply schematic. Incoming ac-line current is filtered by a chassis-mounted line filter (FL1) and, after passing through the fuse (F1), is routed S1 to T1.

U1 rectifies, and C1 filters, the ac output of T1. U2 is an LM338K voltage regulator. This IC features a continuous output of 5 A, with a guaranteed peak output of 7 A, on-chip thermal and safe-operating-area protection for itself, and current limiting. U2's output voltage is set by two resistors (R2 and R3) and a trimmer potentiometer (R8), which allows for adjustment over a small range. U2's input and output are bypassed by C2, C3, and C4. D1 and D2 protect U2 against these capacitors discharging through it.

Overvoltage protection is provided by an SCR, Q1, across the regulator input. Normally, Q1 presents an open circuit, but under fault conditions, it's triggered and short-circuits the unregulated dc input to ground. This discharges C1 and blows F2, reducing the possibility of damage to any connected equipment.

U3, an overvoltage-protection IC, continuously monitors the output voltage. When the output voltage rises above a predetermined level, U3 starts charging C5. If the overvoltage duration is sufficiently long, U3 triggers Q1. This built-in delay (about 1 ms) allows short transient noise spikes on the output voltage to be safely ignored while still triggering the SCR if a true fault occurs. The monitored voltage is set by R5 and R6 and trimmer potentiometer R9, which allows for adjustment over a limited range.

D3 protects the supply from reverse-polarity discharge from connected equipment. The presence of output voltage is indicated by an LED, DS1. R7 is a current limiting resistor for DS1.

CONSTRUCTION

How you construct your supply depends on the size of the components and enclosure you use. General physical layout is not important, although there are a couple of areas that require some attention. In the unit shown in **Fig 7.70**, FL1, the fuse holders, S1, the heat sink. DS1 and the binding posts are mounted on the front and rear enclosure panels. T1 and the PC board are secured to the enclosure's bottom plate. C1's mounting clamp is attached to the rear panel. Bleeder resistor R1 is connected directly across C1's terminals. D3 is soldered directly across the output binding posts.

U2, D1, D2 and R3 are all mounted on the

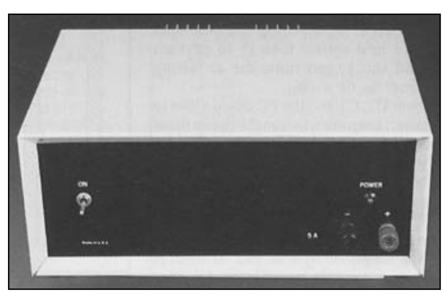


Fig 7.68 — Completed 13.8-V, 5-A power supply

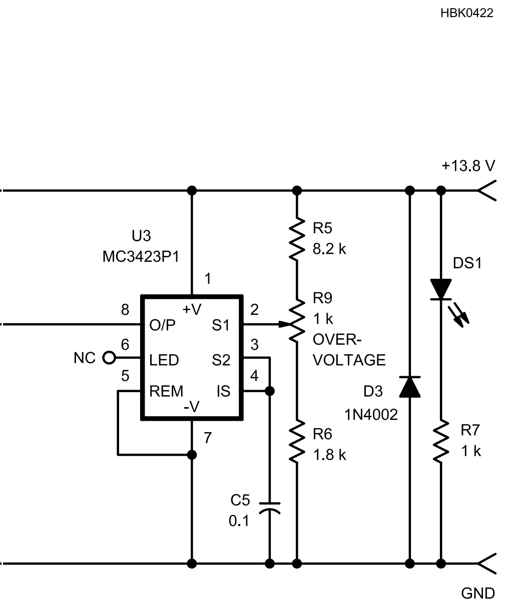


Fig 7.69 — Schematic for the 13.8-V, 5-A power supply. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. A PC board and U3 are available from FAR Circuits (www.farcircuits.net). The PC board has mounting holes and pads to allow for handling different trimmer-potentiometer footprints. A PC board template is available on the CD-ROM for this book.

C1 — 10,000 µF, 35-V electrolytic.
C2, C4 — 1 µF, 35-V tantalum.
C3 — 10 µF, 35-V tantalum.
C5 — 0.1 µF, 25-V ceramic disc.
D1-D3 — 1N4002.
DS1 — Red LED.
F2 — Fast-acting 10-A fuses; three required (see text).
F1 — Fast-acting 0.5-A fuse.
FL1— Ac-line filter.
Q1 — BT152 400-V, 25-A SCR in TO-220A package (NTE5554)
R8 — 500 Ω, single-turn trimmer potentiometer.
R9 — 500 Ω or 1 kΩ, single-turn trimmer potentiometer.
S1 — SPST panel-mount switch.
T1 — 120-V primary, 16- to 20-V, 5-A secondary.
U1 — 100-PIV, 6-A bridge rectifier.
U2 — LM338K 5-A adjustable power regulator in a TO-3 package.
U3 — MC3423P1 overvoltage protection IC.

Misc: two panel-mount fuse holders; line cord; heat sinks for TO-3 case transistors; TO-3 mounting kit and heat-sink grease; black and red binding posts; chassis or cabinet; PC board; hardware, rubber hoods, heat-shrink tubing or electrical tape for F1 and FL1, hook-up wire.

heat sink as shown in **Fig 7.71**. It's important to keep R3 attached as closely as possible to U2's terminals to prevent instability. Use a TO-3 mounting kit and heat-conductive grease or thermal pad to electrically isolate U2 from the heat sink.

Cover all ac-input wiring (use insulated wire and heat-shrink tubing) to prevent electrical shock and route the ac wiring away from the dc wiring. The rear panel assembly is shown in **Fig 7.72**.

Mount U2, C1, and the PC board close to each other and keep the wire runs between these components as short as possible. Excessively long wire runs may lead to unpredictable behavior.

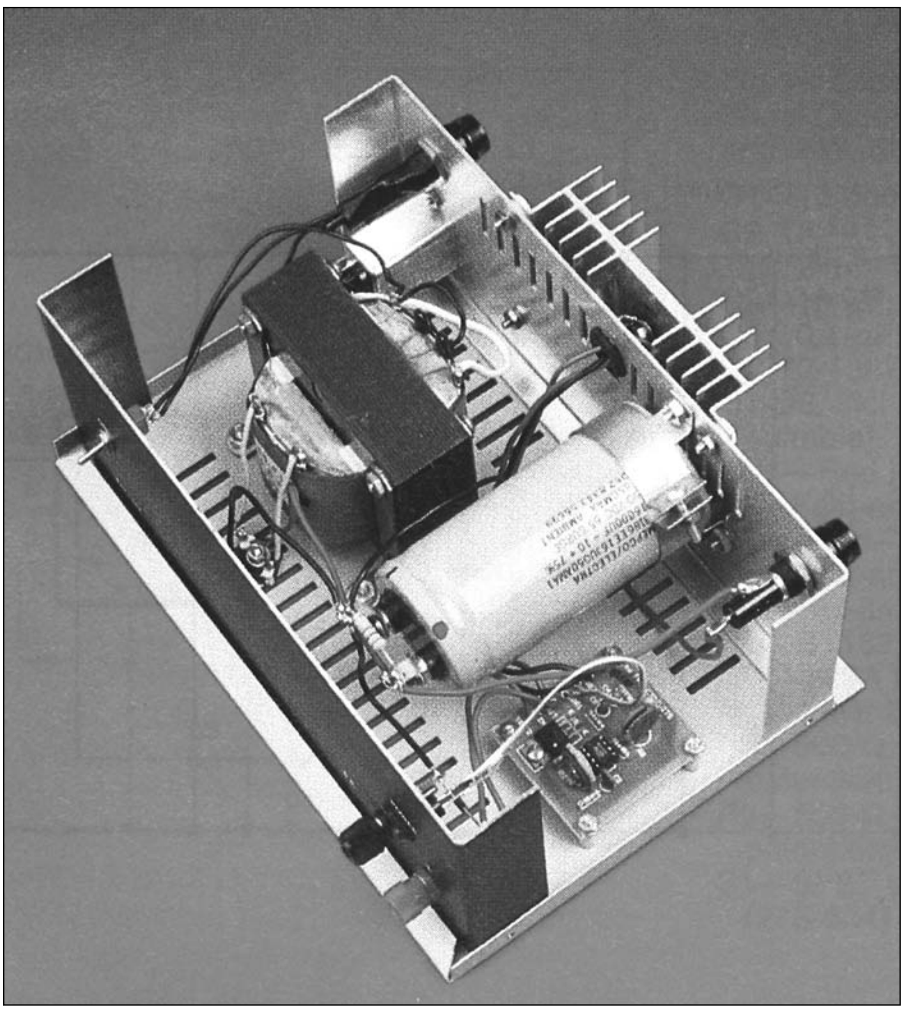


Fig 7.70 — Physical layout of the 13.8-V, 5-A power supply. On the rear panel, left, are the ac-line filter and F1. The regulator's heat sink is at the middle of the panel and F2 is to the right. At the bottom of the enclosure, in front of T1, is the diode bridge rectifier. Because C1 is too tall to mount vertically within the Hammond #14260 cabinet, its mounting clamp is secured to the inside rear panel. Immediately to the right of C1 is the PC board. On the front panel are the on/off switch, LED power-on indicator and output-voltage binding posts

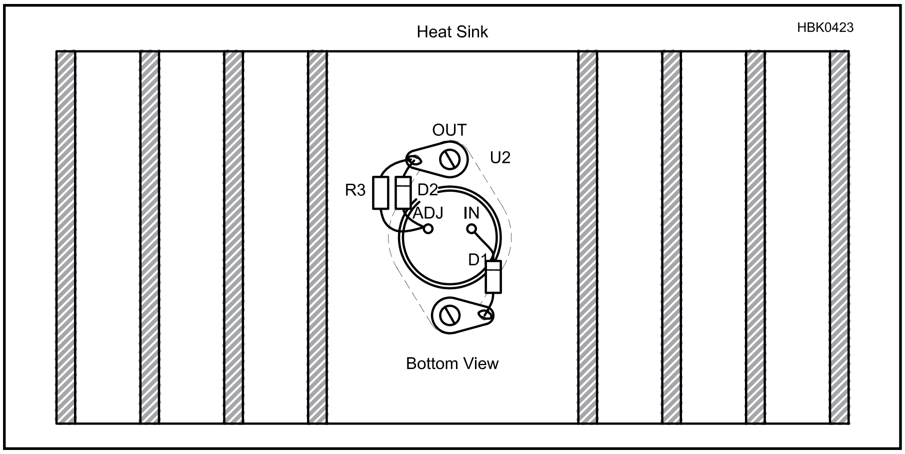


Fig 7.71 — Here's the regulator's heat-sinking arrangement, showing U2, D1, D2 and R3. U2 is electrically isolated from the heat sink by a TO-3 mounting kit.

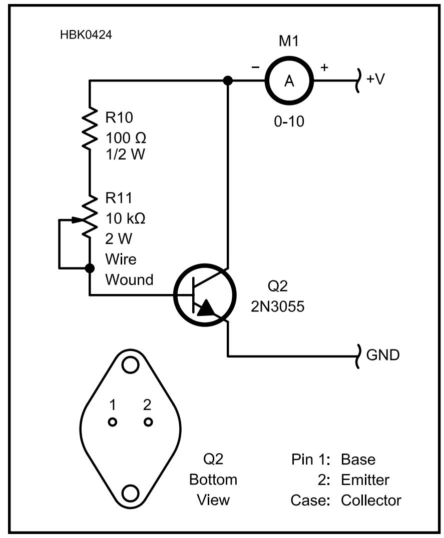


Fig 7.73 — An active resistive load to use in testing the supply. Adjustment of R11 is sensitive.

M1 — Multimeter or ammeter capable of measuring 10 A.
Q2 — 2N3055; mount on heat sink.

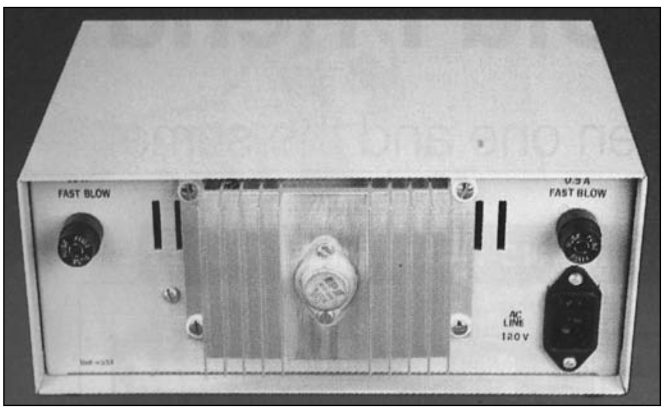


Fig 7.72 — A rear view of the power supply enclosure. Louvers in the enclosure bottom and on the rear panel provide convective cooling.

TEST AND CALIBRATION

An accurate multimeter covering ranges of 30 V dc and 10 A dc is required. A variable resistive load with a power rating of 100 W is also needed; this can he made using a heat-sink-mounted 2N3055 power transistor and a couple of components as shown in **Fig 7.73**.

First, set R8 (OUTPUT VOLTAGE) fully clockwise and R9 (OVERVOLTAGE) fully counterclockwise. Insert a fuse in the dc line at F2. Connect the ac line, turn on S1 and check that DS1 lights. Measure the output voltage: it should be about 12 V. Adjust R8 counterclockwise until you obtain 14.2 V output; this sets the trip voltage.

While monitoring the output voltage, gradually adjust R9 clockwise until the voltage suddenly falls to zero. This indicates that the SCR has triggered and blown F2. Disconnect the ac line cord from the wall socket. *Don't make any adjustment to R9!* Instead, adjust R8 fully clockwise.

With the ac line cord removed, check that F2 is open. Replace F2 with a new fuse (now you know why two of the three fuses are called for). Reconnect the ac line cord, and while continually monitoring the output voltage, gradually adjust R8 until Q1 again triggers at 14.2 V, blowing F2. If you find the adjustment of R9 to be too sensitive, use a 500-Ω potentiometer in its place and reduce the value of R5 (if necessary) to provide the required adjustment range.

Again disconnect the line cord from the wall socket, set R8 fully clockwise, replace F2 (there's the third fuse!) and reset R8 for 13.8 V. This completes the voltage calibration and overvoltage protection tests. The power supply is now set to 13.8 V output, with the overvoltage protection set for 14.2 V.

Adjust the variable resistive load to maximum resistance and connect it to the power supply output in series with the ammeter. Turn on the power supply and gradually adjust R10 until a current of 5 A flows. Decrease the resistance further and check that the current limits between 5.5 A and 0.5 A.

Finally, be thoroughly unpleasant and apply a short circuit via the ammeter. Check that the current-limiting feature operates correctly. The prototype limited at approximately 3.5 A. Disconnect the ac line cord and test equipment, dose up the case and your power supply is ready for service.

SUMMARY

The prototype supply powers a 25-W transmitter that continually draws 4.5 A. Under these conditions, the entire power supply gets hot, so ventilation holes were eventually drilled into the case. (On this prototype, the heat sinks are mounted inside the cabinet). The heat sink gets hot enough to scorch the skin of the unwary, so be careful!

7.16.4 High-Voltage Power Supply

This two-level, high-voltage power supply was designed and built by Dana G. Reed, W1LC. It was designed primarily for use with an RF power amplifier using a triode in class AB2 grounded-grid operation. The supply is rated at a continuous output current of 1.5 A, and will easily handle intermittent peak currents of 2 A. The 12-V control circuitry, and the low-tap setting of the plate transformer secondary, make it straightforward to adapt the design to homemade tube amplifiers.

The step-start circuit is straightforward and ensures that the rectifier diodes are current-limited when the power supply is first turned on. A 6-kV meter is used to monitor high-voltage output.

Fig 7.74 is a schematic diagram of the bi-level supply. An ideal power supply for a high-power linear amplifier should operate from a 240-V circuit, for best line regulation. A special, hydraulic/magnetic circuit breaker also serves as the disconnect for the plate transformer primary. Don't substitute a standard circuit breaker, switch or fuses for this breaker; fuses won't operate quickly enough to protect the amplifier or power supply in case of an operating abnormality. The 100-kΩ, 3-W bleeder resistors are of stable metal-oxide film design. These resistors are wired across each of the 14 capacitors to equalize voltage drops in the series-connected bank. This choice of bleeder resistor value provides a lighter load (less than 25 W total under high-tap output) and benefits mainly the capacitor-bank filter by yielding much less heat as a result. A reasonable, but longer bleed-down time to fully discharge the capacitors results — about nine minutes after power is removed. A small fan is included to remove any excess heat from the power supply cabinet during operation.

POWER SUPPLY CONSTRUCTION

The power supply can be built in a 23½ × 10¾ × 16-inch cabinet. The plate transformer is quite heavy at 67 lbs, so use ⅛-inch aluminum for the cabinet bottom and reinforce it with aluminum angle for extra strength and stability. The capacitor bank will be sized for the specific capacitors used. This project employed ⅜-inch thick polycarbonate for reasonable mechanical stability and excellent high-voltage isolation. The full-wave bridge consists of four commercial diode block assemblies.

POWER SUPPLY OPERATION

When the front-panel breaker is turned on, a single 50-Ω, 100-W power resistor limits primary inrush current to a conservative value as the capacitor bank charges. After approximately two seconds, step-start relay K1 actuates, shorting the 50-Ω resistor and allowing full line voltage to be applied to the plate transformer. No-load output voltages under low- and high-tap settings as configured and shown in Fig 7.74 are 3050 V and 5400 V, respectively. Full-load levels are somewhat lower, approximately 2800 V and 4900 V. If a tap-select switch is used as described in the schematic parts list, it should only be switched when the supply is off.

7.16.5 Reverse-Polarity Protection Circuits

*The following material was collected from various public-domain sources by Terry Fletcher, WAØITP(**www.wa0itp.com/revpro.html**) and published in the* QRP Quarterly*, Spring 2012 issue. (**www.qrparci.org**)*

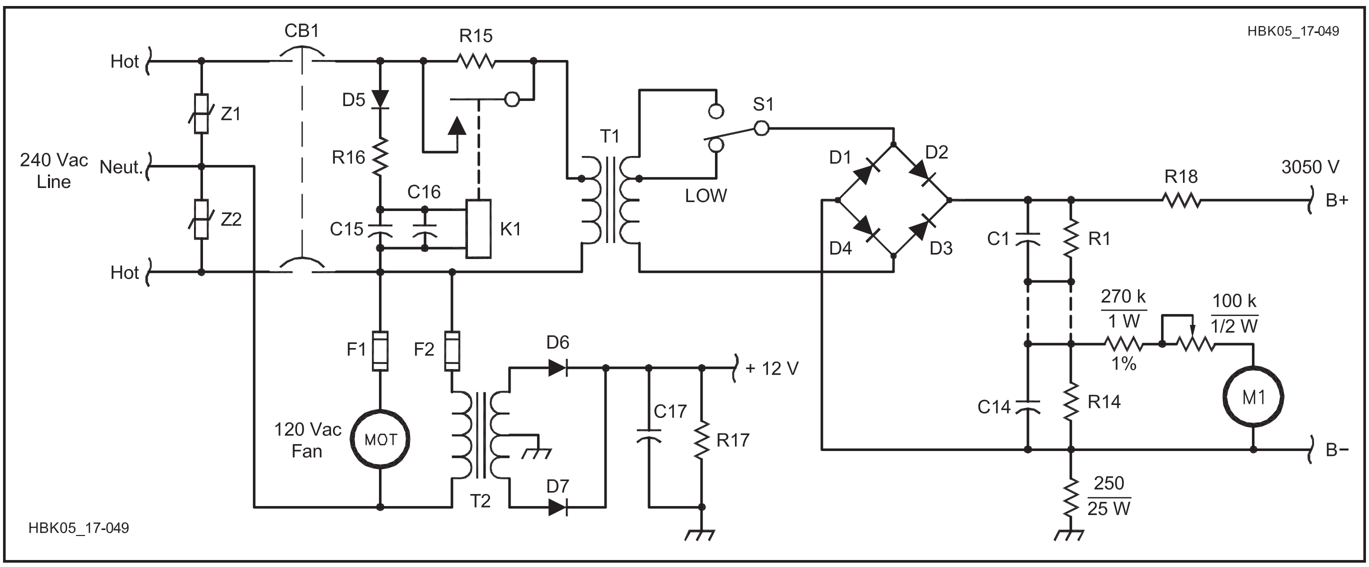


Fig 7.74 — Schematic diagram of the 3050-V/5400-V high-voltage power supply.

C1-C14 — 800 µF, 450 V electrolytic.
C15, C16 — 4700 µF, 50 V electrolytic.
C17 — 1000 µF, 50 V electrolytic.
CB1 — 20-A hydraulic/magnetic circuit breaker (Potter & Brumfield W68-X2Q12-20 or equiv). 40-A version required for commercial applications/ service (Potter & Brumfield W92-X112-40).
D1-D4 — Commercial diode block assembly: K2AW HV14-1, from K2AW's Silicon Alley
D5 — 1000-PIV, 3-A, 1N5408 or equiv.
D6, D7 — 200-PIV, 3-A, 1N5402 or equiv.
F1, F2 — 0.5 A, 250 V (Littelfuse 313 Series, 3AG glass body or equiv).
K1 — DPDT power relay, 24-V dc coil; both poles of 240-V ac/25-A contacts in parallel (Potter & Brumfield PRD-11DYO-24 or equiv).
M1 — High-voltage meter, 6-kV dc full scale. (Important: Use a 1-mA or smaller meter movement to minimize parallel-resistive loading at R14. Also, select series meter-resistor and adjustment-potentiometer values to calibrate your specific meter. Values shown are for a 1-mA meter movement.)
MOT1 — Cooling fan, 119mm, 110120-V ac, 30-60 CFM, (EBM 4800Z or equiv).
R1-R14 — Bleeder resistor, 100 kΩ, 3 W, metal oxide film.
R15 — 50 Ω, 100 W.
R16 — 3.9 kΩ, 25 W.
R17 — 30 Ω, 25 W.
R18 — 20 Ω, 50 W.
S1 — Ceramic rotary, 2-pos. tap-select switch (optional). Voltage rating between tap positions should be at least 2.5 kV. Mount switch on insulated or ungrounded material such as a metal plate on standoff insulators, or an insulating plate, and use only a *nonconductive* or otherwise *electrically-isolated* shaft through the front panel for safety
T1 — High-voltage plate transformer, 220/230-V primary, 2000/3500-V, 1.5-A CCS JK secondary (Peter W. Dahl Company, Hipersil C-Core; www.harbachelectronics.com). Primary 220-V tap fed with nominal 240-V ac line voltage to obtain modest increase in specified secondary voltage levels
T2 — 120-V primary, 18-V CT, 2-A secondary (Mouser 41FJ020).
Z1-Z2 — 130-V MOV.

DC power is the standard for most amateur radios and accessories, usually a nominal 12 V (10.5 to 13.8 V). There are many different types of connectors used for dc power — from screw terminals to custom-molded multi-pin designs. This makes it easy to accidentally apply power with reversed polarity and damage equipment. Even a few milliseconds of reversed power can be sufficient to destroy a semiconductor or burn out a narrow PCB trace. This can be a particular problem when using 9 V batteries as it is easy to reverse the snap-on connector when changing or installing a battery.

The following collection of circuits illustrates ways to protect equipment from reverse-polarity dc power. The suitability of the circuits depends on the equipment and power source. All dc power sources, particularly batteries, should be fused or current-limited to mitigate fire hazards and other damage from overheating wires and other conductors.

Fig 7.75 shows several passive circuits that dissipate some power due to the series forward voltage drop of the diodes. At currents above 1 A, the power dissipated can easily exceed 1 W and the maximum junction temperature of the diode can be exceeded without some sort of thermal protection or heat sinking. (The shunt diode circuit in Fig 7.75D does not dissipate power.) **Fig 7.76** shows two methods of using an electromechanical relay that avoid the forward voltage drop of diode-based protection circuits. Which circuit you choose depends on the type of equipment and constraints on power dissipation and voltage drop.

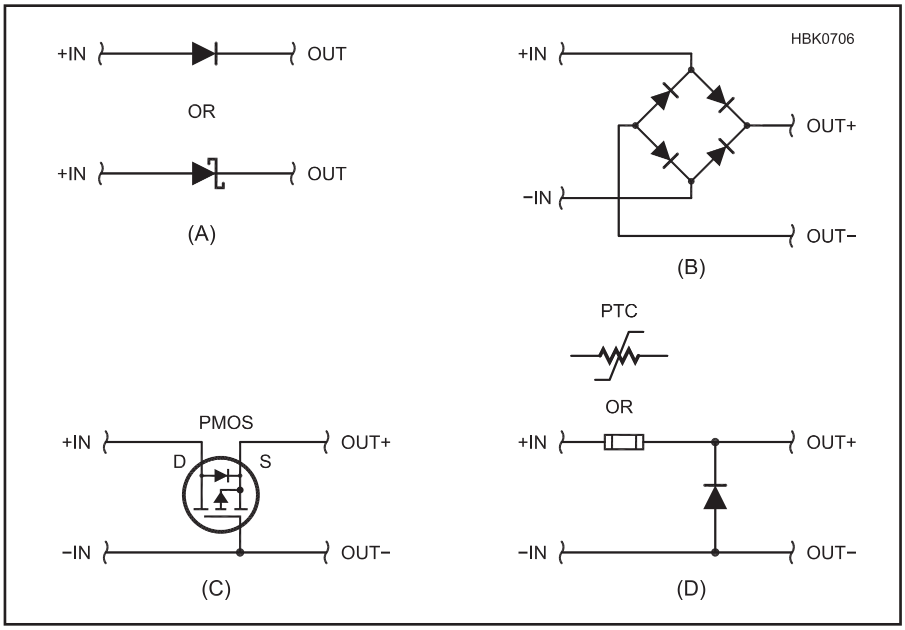


Fig 7.75 — Passive circuits for reverse-polarity protection. Series diode (A). Full-wave rectifier (B). PMOS MOSFET with integral body diode shown as separate component (C). Shunt diode (D). Schottky barrier diodes may be used in all circuits as a substitute for silicon junction rectifiers. See text for circuit comparison.

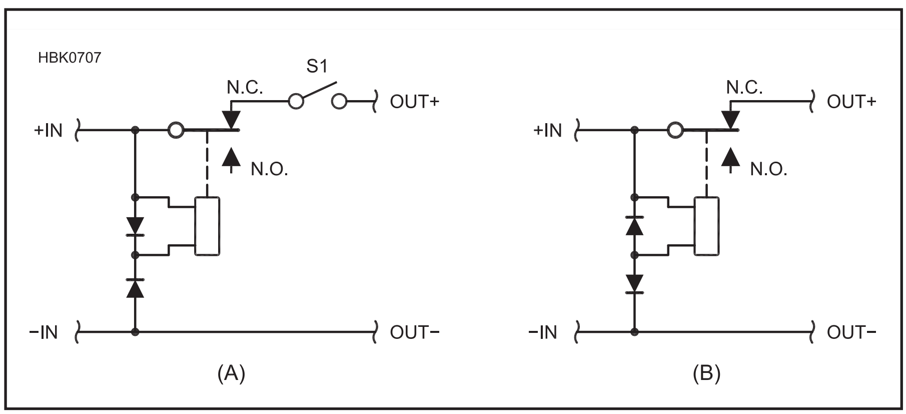


Fig 7.76 — Relay-based circuits for reverse-polarity protection. Normally-closed contacts (A) and normally-open contacts (B). See text for circuit comparison.

PASSIVE CIRCUITS

Blocking diode (Fig 7.75A) — A series diode is very simple and inexpensive. Its PIV rating should be at least twice the expected applied voltage — 50 V PIV is a good minimum value for automotive and 12 V dc use. Its maximum average forward current rating should be several times the expected maximum steady-state current draw.

Remember that a silicon junction diode's forward voltage drop, V_f, is at least 0.6 V and can approach 1.0 V at high forward current. This can result in significant power dissipation ($P = V_f \times I_f$) and cause the diode's maximum rated junction temperature to be exceeded unless some means of cooling the diode is provided.

The forward voltage drop of the diode will also reduce the voltage available to the equipment being powered. This will raise the minimum allowable power supply voltage for the equipment to operate properly. For example, if a piece of equipment is rated to operate properly at or above 11 V, a series diode with $V_f = 0.6$ V raises the minimum allowable power supply voltage to 11 + 0.6 = 11.6 V. This may be significant in battery-powered installations.

The Schottky barrier diode shown as an alternate may be a better choice due to its forward voltage drop being lower by several tenths of a volt, reducing power dissipation. A Schottky diode may be used in place of any of the diodes in Figs 7.75 and 7.76. Be sure the reverse current leakage of the Schottky diode is acceptable.

Full-wave bridge rectifier (Fig 7.75B) — The full-wave circuit has the advantage of always supplying voltage with the proper polarity to the equipment being powered. Full-wave rectifiers are also available as integrated packages, making them easy to install. Remember that the forward voltage drop and power dissipation of this circuit will be twice that of the single series diode because two diodes are always in series with supply current.

PMOS P-channel MOSFET (Fig 7.75C) — This circuit uses a P-channel enhancement-mode (PMOS) MOSFET that conducts current with the gate connected as shown. PMOS devices have low on-resistance ($R_{ds(on)}$) and high maximum current ratings. Devices with on-resistance of 0.050 Ω and lower are commonly available. For more information about using PMOS and NMOS devices for polarity protection see Maxim Electronics Application Note 636, "Reverse-Current Circuitry Protection" (**www.maxim-ic.com/app-notes/index.mvp/id/636**). N-channel devices may also be used in the current return or ground lead, but opening the return connection can create other problems inside the equipment and with other devices on the same power circuits.

Shunt diode with fuse (Fig 7.75D) — These configurations use a single diode that acts to blow a fuse (either a fusible-link or positive temperature coefficient PTC resettable device) if reverse polarity voltage is applied. The advantage of this circuit is that no power is dissipated by the diode during normal operation. The diode must be sufficiently rated to handle the high surge current from shorting the power source and have current ratings significantly higher than the fuse current rating.

If the diode fails shorted or in a low-resistance state, it will continue to blow the fuse until replaced. If the diode fails open or high-resistance, it will no longer protect the circuit. If shunt diode protection is used and the fuse opens, check the diode to be sure it has not failed as well.

RELAY-BASED CIRCUITS

Relay-based circuits have the advantage of little to no voltage drop, even at high currents as long as the contact ratings are sufficient. The circuits are more complex than the diode-based circuits in the preceding section but can generally handle more current and are not damaged by reverse polarity voltages. The circuits reset themselves automatically.

Relay with normally-closed contacts (Fig 7.76A) — There is no current drain through the relay coil until reverse-polarity is applied. However, there will be a few milliseconds during which reverse polarity voltage is applied if no power switch (S1) is used or the power switch is closed. This is generally enough time for damage to occur so this circuit is only recommended if a power switch is used to turn the equipment ON and OFF.

Relay with normally-open contacts (Fig 7.76B) — The relay contacts close and supply power to the equipment only when applied voltage has the proper polarity. The relay coil draws current continuously during normal operation. This may be unacceptable for low-power and battery-powered equipment.

7.16.6 Automatic Sealed Lead-Acid Battery Charger

After experiencing premature failure of the battery in his Elecraft K2 transceiver, Bob Lewis, AA4PB, began searching for an automatic battery charger. Although this charger was designed specifically for use with the Power-Sonic PS-1229A sealed lead-acid (SLA) battery used in the Elecraft K2 transceiver, its design concepts have wide ranging applications for battery operated QRP rigs of all types. Comments pertaining to the SLA batteries and chargers apply across the board and the charger described here can be used with any similar battery.

SLAs are commonly called gel-cells because of their gelled electrolyte. As with all things, to obtain maximum service life from an SLA battery, it needs to be treated with a certain degree of care. SLA batteries must be recharged on a regular basis; they should not be undercharged or overcharged. If an SLA battery is left unused, it will gradually self-discharge.

USING A THREE-MODE CHARGER

The author first attempted to use a commercial automatic three-mode charger. However, most three-mode chargers work by sensing current and are never intended to charge a battery under load.

Three-mode chargers begin the battery charging process by applying a voltage to the battery through a 500-mA current limiter. This stage is known as *bulk-mode* charging. As the battery charges, its voltage begins to climb. When the battery voltage reaches 14.6 V the charger maintains the voltage at that level and monitors the battery charging current. This is known as the *absorption mode*, sometimes called the *overcharge mode*. By this time, the battery has achieved 85% to 95% of its full charge. As the battery continues to charge — with the voltage held constant at 14.6 V — the charging current begins to drop.

When the charging current falls to 30 mA, the three-mode charger switches to *float mode* and lowers the applied voltage to 13.8 V. At 13.8 V, the battery becomes self-limiting, drawing only enough current to offset its normal self-discharge rate. This works great — until you attach a light load to the battery, such as a receiver. The K2 receiver normally draws about 220 mA. When the charger detects a load current above 30 mA, it's fooled into thinking that the battery needs charging, so it reverts to the absorption mode, applying 14.6 V to the battery. If left in this condition, the battery is overcharged, shortening its service life.

UC3906 IC CHARGERS

Chargers using the UC3906 SLA charge-controller IC work just like the three-mode charger described earlier except that their return from float mode to absorption mode is based on voltage rather than current. Typically, once the charger is in float mode it won't return to absorption mode until the battery voltage drops to 90% of the float-mode voltage (or about 12.4 V). Although this is an improvement over the three-mode charger, it still has the potential for overcharging a battery to which a light load is attached.

First, let's look at the situation where a UC3906-controlled charger is in absorption mode and you turn on the K2 receiver, applying a load. The battery is fully charged, but because the load is drawing 220 mA, the charging current never drops to 30 mA and the charger remains in absorption mode, thinking that it is the battery that is asking for the current. As with the three-mode charger, the battery is subject to being overcharged.

If we remove the load by turning off the K2, the current demand drops below 30 mA and the charger switches to float mode (13.8 V). When the K2 is turned on again, because the charger is able to supply the 220 mA for the receiver, the battery voltage doesn't drop, so the charger stays in float mode and all is well. However, if the transmitter is keyed (increasing the current demand), the charger can't supply the required current, so it's taken from the battery and the battery voltage begins to drop. If we unkey the transmitter before the battery voltage reaches 12.4 V, the charger stays in float mode. Now it takes much longer for the charger to supply the battery with the power used during transmit than it would have if the charger had switched to absorption mode.

Let's key the transmitter again, but this time keep it keyed until the battery voltage drops below 12.4 V. At this point, the charger switches to the absorption mode. When we un-key the transmitter, we're back to the situation where the charger is locked in absorption mode until we turn off the receiver.

WHY WORRY?

So why this concern about overcharging an SLA battery? At 13.8 V, the battery self-limits, drawing only enough current to offset its self-discharge rate (typically about 0.001 times the battery capacity, or 2.9 mA for a 2.9 Ah battery). An SLA battery can be left in this float-charge condition indefinitely without overcharging it. At 14.6 V, the battery takes more current than it needs to offset the self-discharge. Under this condition, oxygen and hydrogen are generated faster than they can be recombined, so pressure inside the battery increases. Plastic-cased SLA batteries such as the PS-1229A have a one-way vent that opens at a couple of pounds per square inch pressure (PSI) and release the gases into the atmosphere. This results in drying the gelled electrolyte and shortening the battery's service life. Both undercharging and overcharging need to be avoided to get maximum service life from the battery.

Continuing to apply 14.6 V to a 12-V SLA battery represents a relatively minor amount of overcharge and results in a gradual deterioration of the battery. Applying a potential of 16 V or excessive bulk-charging current to a small SLA battery from an uncontrolled solar panel can result in serious overcharging. Under these conditions, the overcharging can cause the battery to overheat, which causes it to draw more current and result in *thermal runaway*, a condition that can warp electrodes and render a battery useless in a few hours. To prevent thermal runaway, the maximum current and the maximum voltage need to be limited to the

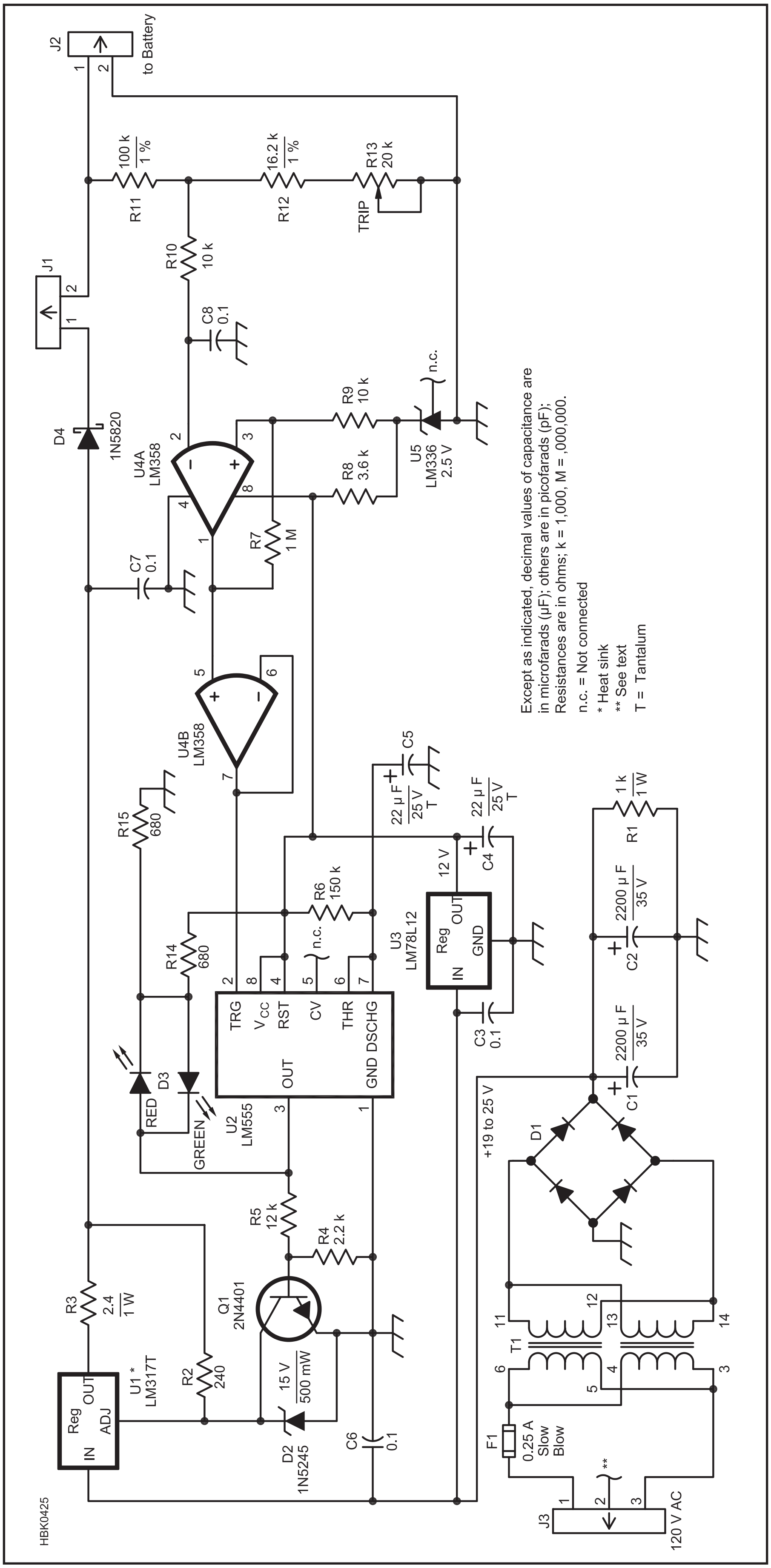


Fig 7.77 — Schematic of the SLA charger. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. Equivalent parts can be substituted; n.c . indicates no connection. A PC board is available from FAR Circuits (www.farcircuits.net).

C1, C2 — 2200 µF, 35 V electrolytic.
C3, C6, C7, C8 — 0.1 µF, 50 V metalized-film.
C4, C5 — 22 µF, 25 V tantalum.
D1 — 400 V, 4 A bridge rectifier.
D2 — 1N5245 Zener diode, 15 V, 500 mW.
D3 — Bicolor LED, red-green.
D4 — 1N5820 5chottky diode.
F1 — 0.25 A slow-blow fuse.
J1 — 2-pin header, PC mount.
J2 — 2-pin connector, PC mount.
J3 — 3-pin connector, PC mount.
Q1 — 2N4401 NPN transistor.
R13 — 20 kΩ multi-turn pot.
T1 — 15 V ac, 666 mA.
U1 — LM317T voltage regulator, TO-220 case.
U2 — LM555 timer.
U3 — LM78L12 voltage regulator, TO-92 case.
U4 — LM358 dual op amp.
U5 — LM336, 2.5 V voltage reference, TO-92 case.
Misc: PC board;TO-220 heat sink; five ¼-inch, #4-40 stand-offs; two fuse-holder clips, PC mount; two-pin shunt; two-pin connector housing; three-pin connector housing; four housing pins; enclosure

battery manufacturer's specifications.

DESIGN DECISION

To avoid the potential of overcharging a battery with an automatic charger locked up by the load, the author decided to design a charger that senses battery voltage rather than current in order to select the proper charging rate. A 500-mA current limiter sets the maximum bulk rate charge to protect the battery and the charger's internal power supply.

Like the three-mode chargers, when a battery with a low terminal voltage is first connected to the charger, a constant current of 500 mA flows to the battery. As the battery charges, its voltage begins to climb. When the battery voltage reaches 14.5 V, the charger switches off. With no charge current flowing to the battery, its voltage now begins to drop. When the current has been off for four seconds, the charger reads the battery voltage. If the potential is 13.8 V or less, the charger switches back on. If the voltage is still above 13.8 V, the charger waits until it drops to 13.8 V before turning on. The result is a series of 500-mA current pulses varying in width and duty cycle to provide an average current just high enough to maintain the battery in a fully charged condition. Because the repetition rate is very low (a maximum of one current pulse every four seconds) no RFI is generated that could be picked up by the K2

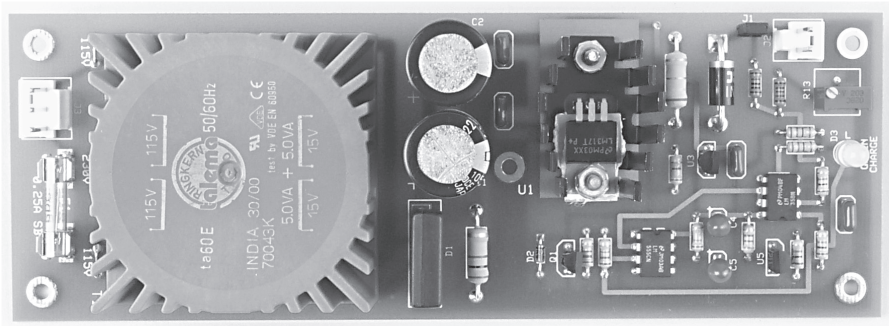


Fig 7.78 — Complete SLA charger PC board.

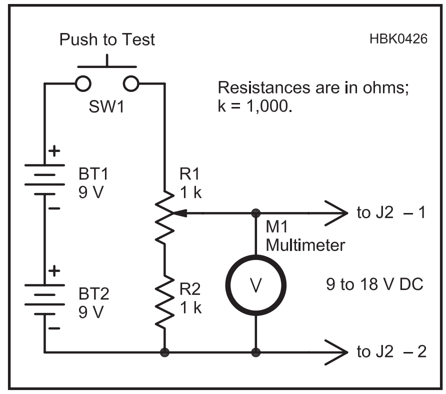


Fig 7.79 — Test voltage source for the battery charger.

receiver. Because the K2's critical circuits are all well-regulated, slowly cycling the battery voltage between 13.8 V and 14.5 V has no ill effects on the transmitted or received signals.

As the battery continues to charge, the pulses get narrower and the time between pulses increases (a lower duty cycle). Now when the K2 receiver is turned on and begins drawing 220 mA from the battery, the battery voltage drops more quickly so the pulses widen (the duty cycle increases) to supply a higher average current to the battery and make up for that taken by the receiver. When the K2 transmitter is keyed, it draws about 2 to 3 A from the battery. Because the charger is current limited to 500 mA, it is not able to keep up with the transmitter demands. The battery voltage drops and the charger supplies a constant 500 mA. The battery voltage continues to drop as it supplies the required transmit current. When the transmitter is un-keyed, the battery voltage again begins to rise as the charger replenishes the energy used during transmit. After a short time, (depending on how long the transmitter was keyed) the battery voltage reaches 14.5 V and the pulsing begins again. The charger is now fully automatic, maintaining the battery in a charged condition and adjusting to varying load conditions.

The great thing about this charging system is that during transmit the majority of the required 2 to 3 A is taken from the battery. When you switch back to receive, the charger is able to supply the 220 mA needed to run the receiver and deliver up to 280 mA to the battery to replenish what was used during transmit. This means that the power source need only supply the average energy used over time, rather than being required to supply the peak energy needed by the transmitter. (You don't need to carry a heavy 3-A regulated power supply with your K2.) As long as you don't transmit more than about 9% of the time, this system should be able to power a K2 indefinitely.

Have you ever noticed that sometimes when your handheld radio has a low battery and you drop it into its charger you hear hum on the received signals? This charger's power supply is well filtered to ensure that there is no ripple or ac hum to get into the K2 under low battery voltage conditions.

CIRCUIT DESCRIPTION

The charger schematic is shown in **Fig 7.77**; the unit is dubbed the PCR12-500A, short for Pulsed-Charge Regulator for 12-V SLA batteries with maximum bulk charge rates of 500 mA. U1, an LM317 three-terminal voltage regulator, is used as a current limiter, voltage regulator and charge-control switch. A 15-V Zener diode (D2) sets U1 to deliver a no-load output of 16.2 V. R3 sets U1 to limit the charging current to 500mA. When Q1 is turned on by the LM555 timer (U2), the ADJ pin of U1 is pulled to ground, lowering its output voltage to 1.2 V. D4 effectively disconnects the battery by preventing battery current from flowing back into U1. A Schottky diode is used at D4 because of its low voltage drop (0.4 V).

An LM358 (U4A) operates as a voltage comparator. U5, an LM336, provides a 2.5-V reference to the positive input (pin 3) of U4. R11, R12 and R13 function as a voltage divider to supply a portion of the battery voltage to pin 2 of U4A. R13 is adjusted so that when the battery terminal voltage reaches 14.5 V, the negative input of U4A rises slightly above the 2.5-V reference and its output switches from +12 V to 0 V. When this happens, the 1-MΩ resistor (R7) causes the reference voltage to drop a little and provide some hysteresis. The battery voltage must now drop to approximately 13.8 V before U4A turns back on.

U4B is a voltage follower. It pulls the trigger input (pin 2) of U2 to 0 V, causing its output to go to 12 V. U4B's output remains at 12 V until C5 has charged through R6 (approximately four seconds) and the trigger has been released by U4A sensing the battery dropping to 13.8 V or less. While the output of U2 is at 12 V, emitter/base current for Q1 flows via R5 and Q1's collector pulls U1's ADJ pin to ground, turning off the charging current.

The output of U2 also provides either +12 V or 0 V to the bicolor LED, D3. R14 and R15

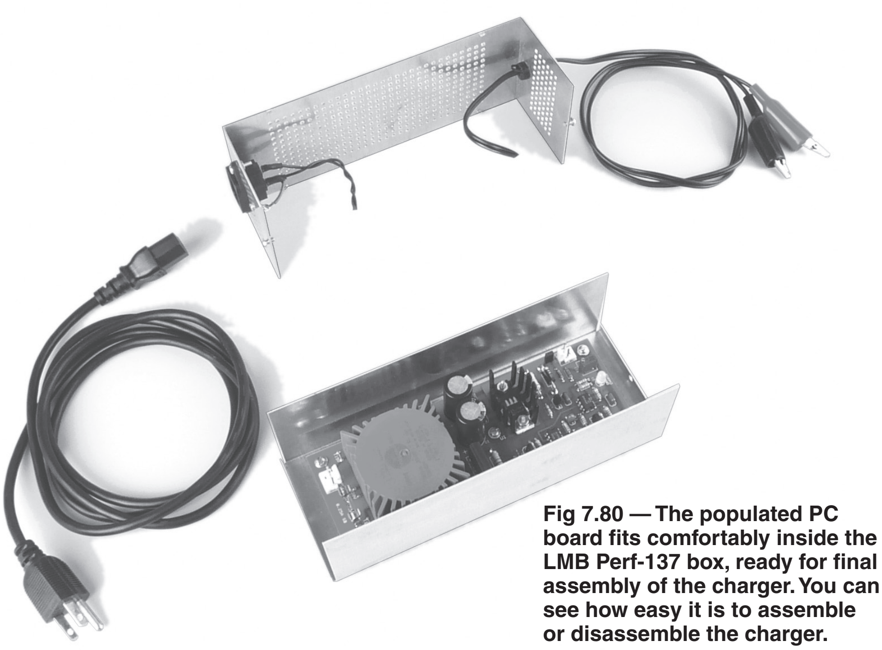


Fig 7.80 — The populated PC board fits comfortably inside the LMB Perf-137 box, ready for final assembly of the charger. You can see how easy it is to assemble or disassemble the charger.

form a voltage divider to provide a reference voltage to D3 such that D3 glows red when U2's output is +12 V and green when U2's output is at 0 V. When ac power is applied but U1 is switched off and not supplying current to the battery, D3 glows red. When U1 is on and supplying current to the battery, D3 is green. As the battery reaches full charge, D3 blinks green at about a four-second rate. As the battery charge increases, the on time of the green LED decreases and the off time increases. A fully charged battery may show green pulses as short as a half-second and the time between pulses may be 60 seconds or more.

T1, D1, C1, and C2 form a standard full-wave-bridge power supply providing an unregulated 20 V dc at 500 mA. U3, an LM78L12 three-terminal regulator, provides a regulated 12-V source for the control circuits.

Note that the mounting tab on U1 is not at ground potential. U1 should be mounted to a heat sink with suitable electrically insulated but thermally conductive mounting hardware to avoid short circuits. Suitable mounting hardware is included with the PC board.

OTHER BULK-CHARGE RATES

The maximum bulk-charge rate is set by the value of R3 in the series regulator circuit. The formula used to determine the value of this resistor is R (Ω) = 1200 / 1 (mA). T1 must be capable of supplying the bulk charge current and U1 must be rated to handle this current. The LM317T used here is rated for a maximum current of 1.5 A, provided it has a heat sink sufficiently large enough to dissipate the generated heat. If you increase the bulk-charge rate, you'll definitely need to increase the size of the on-board heat sink. Mounting U1 directly to the housing (be sure to use an insulator) may be a good option.

TRANSFORMER SUBSTITUTION

The transformer used for T1 is small in size and offers PC-board mounting. You can substitute any transformer rated at 15 or 16 V ac (RMS) at 500 mA or more. You may find common frame transformers to be more readily available. You can mount such a transformer to an enclosure wall and route the transformer leads to the appropriate PC-board holes.

CONSTRUCTION

There is nothing critical about building this charger. You can assemble it on a prototyping board, but a PC board and heat sink are available (**www.farcircuits.net**; see **Fig 7.78)** The specially ordered heat sink supplied with the PC board is ¼-inch higher than the one identified in the parts list and results in slightly cooler operation of U1. The remaining parts are available from Digi-Key.

Be sure to space R1 and R3 away from the board by ¼ inch or so to provide proper cooling. R13 can be a single-turn or a multi-turn pot. You'll probably find a multi-turn pot makes it easier to set the cutoff voltage to exactly 14.5 V.

R13 Adjustment

To check for proper operation and to set the trip point to 14.5 V dc, we need a test-voltage source variable from 12 to 15 V dc. A convenient means of obtaining this test voltage is to connect two 9-V transistor-radio batteries in series to supply 18 V as shown in **Fig 7.79**. Connect a 1-kΩ resistor (R2) in series with a 1-kΩ potentiometer (R1) and connect this series load across the series batteries with the fixed-value resistor to the negative lead. The voltage at the pot arm should now be adjustable from 9 to 18 V. During the following procedure, be sure to adjust the voltage with the test supply connected to the charger at 12 V because the charger loads the test-voltage supply and causes the voltage to drop a little when it's connected.

Remove the jumper at J1 and apply ac line voltage to the unit at J3. Turn R13 fully counterclockwise. D3 should glow green. Connect the test voltage to J2 and adjust R1 of Fig 7.79 for an output of 14.5 V. Slowly adjust R13 clockwise until D3 glows red. To test the circuit, wait at least four seconds, then gradually reduce the test voltage until D3 turns green. At that point, the test voltage should be approximately 13.8 V. Slowly increase the test voltage again until D3 turns red. The test voltage should now read 14.5 V. If it is not exactly 14.5 V, make a minor adjustment to R13 and try again. The aim of this adjustment is to have D3 glow red just as the test voltage reaches 14.5 V.

To test the timer functioning, remove the test voltage from J2 and set it for about 15 V. Momentarily apply the test voltage to J2. D3 should turn red for approximately four seconds, then turn green. The regulator is now calibrated and ready for operation. Remove the test voltage and ac power and install the jumper at J1.

The prototype used an 8 × 3 × 2.75-inch LMB Perf137 box (Digi-Key L171-ND) to house the charger as shown in **Fig 7.80**. An alternative enclosure is the Bud CU482A Convertabox, which measures 8×4×2 inches (available from Mouser). If you use the Convertabox, be sure to add some ventilation holes directly above the board-mounted heat sink. The LMB Perf box comes with a ventilated cover. If you are inclined to do some metal work, you could build your own enclosure using aluminum angle stock and sheet and probably reduce the size to perhaps 8 × 3 × 2 inches. If you use a PC-board-mounted power transformer, watch out for potential shorts between the transformer pins (especially the 120-V ac-line pins) and the case. If you use a metal enclosure, connect the safety ground (green) wire of the ac-line cord directly to the case.

OPERATION

It is very important that this charger be connected directly to the SLA battery with no diodes, resistors or other electronics in between the two. The charger works by reading the battery voltage, so any voltage drop across an external series component results in an incorrect reading and improper charging. For example, the Elecraft K2 has internal diodes in the power-input circuit, so it's necessary to add a charging jack to the transceiver that provides a direct connection to the battery. Now the K2 can be left connected to the charger at all times and be assured that its internal battery is fully charged and ready to go at a moment's notice.

7.16.7 Overvoltage Protection for AC Generators

When using portable generators, there is always a possibility of damage to expensive equipment as a result of generator failure, especially from overvoltage. If the generator supplying power to this equipment puts out too much voltage, you run the risk of burning up power supplies or other electronic components. This project, by Jerry Paquette, WB8IOW, addresses the problem of increased voltage (not lower voltage) or surges and spikes lasting for a few microseconds.

Using a portable generator overvoltage protection device 120-V circuit and ground-fault circuit interrupter (GFCI) shown in **Fig 7.81** is good insurance. This overvoltage protection device must be used in conjunction with a GFCI at each station! (More information on GFCIs may be found in the **Safety** chapter.)

CIRCUIT DESCRIPTION

Refer to **Fig 7.82** for this description. R1 places an intentional fault on the load side of the GFCI. With the value resistor used, the fault is limited to 10 mA. (The normal tripping threshold of a GFCI is 5 mA. This current forces the GFCI to trip in just a few milliseconds. This circuit will not function at all without the use of a GFCI. A GFCI must be used at each station. If a single GFCI were used at the generator, rather than one at each location, premature tripping could occur. Several hundred feet of extension cords could have enough leakage to trip the GFCI.

You can see that the GFCI has separate lines (inputs) and loads (outputs). GFCI input terminals must be connected to the generator output. The GFCI ground must be tied to the ground of the generator. The load (computers, radios, etc) will plug into the GFCI or are wired to the load side of the GFCI. The primary of T1 is wired to the load side of

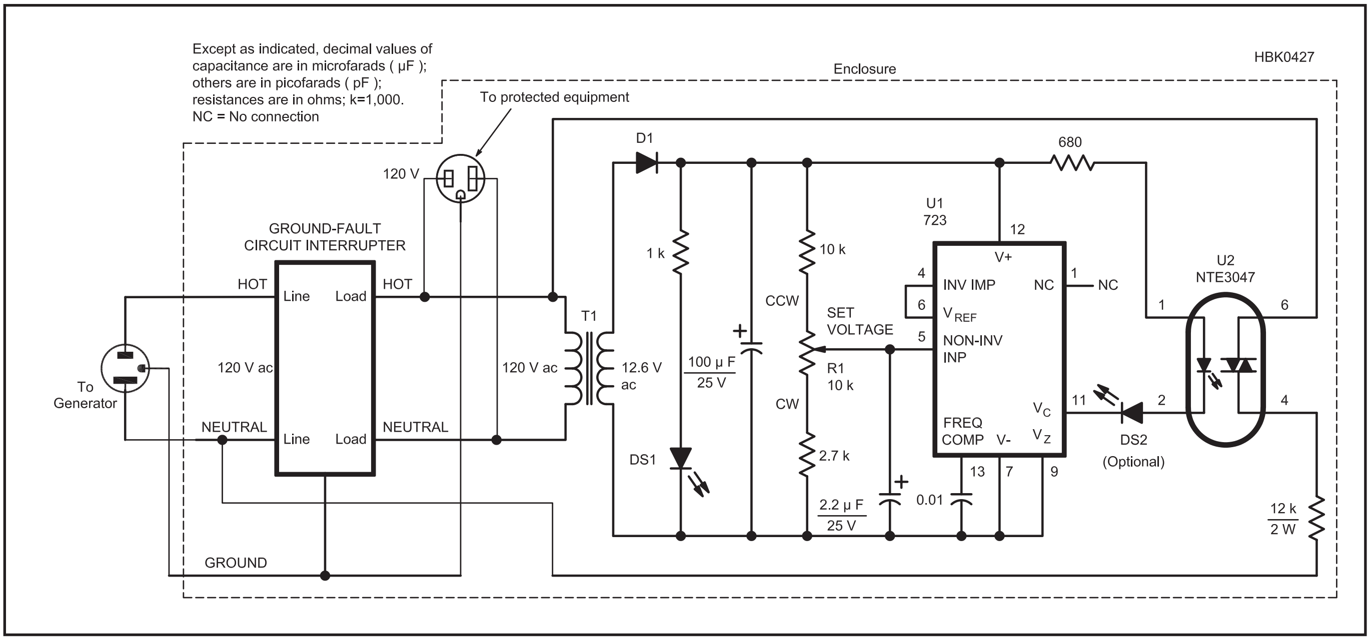


Fig 7.81 — Schematic of the Field Day equipment overvoltage-protection circuit. This circuit must be used in conjunction with a ground fault circuit interrupter (GFCI). A separate GFCI must be installed at each station. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. A PC board is available from FAR Circuits (www.farcircuits.net).

D1 — 200 PIV, 1 A diode; 1N4003 or equiv.
DS1, DS2 — Small LEDs.
R1 — 10 kΩ board-mounted, multi-turn potentiometer
T1 — 12.6 V ac transformer (see text).
U1 — 723 adjustable voltage regulator IC.
U2 — Optoisolator with TRIAC output; NTE3047 or equiv.

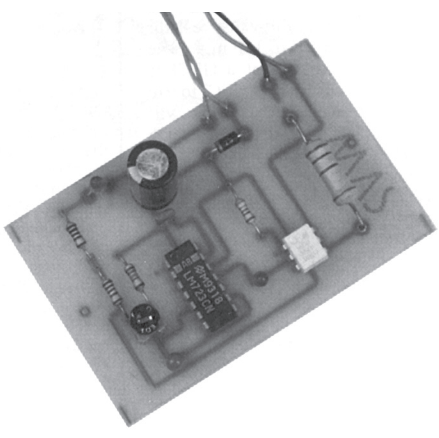


Fig 7.82 — The completed over-voltage protection PC board.

the GFCI. The 12-kΩ/2-W resistor, however, must be wired to neutral on the *line* side of the GFCI in order for it to trip when used with generators that have windings isolated from ground. For safety, construct the entire unit in a single enclosure including the GFCI and its wiring. The generator connection can be made through a wired plug or using a male receptacle mounted on the enclosure.

T1 can be any 120- to 12.6-V transformer capable of delivering 100 mA or more. Mounting of this transformer varies depending on the type used. All remaining components mount on a circuit board. D1 rectifies the ac from T1 and the 100-µF capacitor filters the dc. This voltage provides the power to the 723 voltage regulator.

Two fixed resistors and a potentiometer form the voltage-divider network supplying voltage to the LM723 input, pin 5. R1, the board-mounted potentiometer, has only three leads, but there are four pads on the circuit board, to accommodate different styles of pots. The 2.2-µF capacitor provides a slight delay, to prevent false tripping when the circuit is powered up. The 0.01-µF capacitor from pin 13 of the 723 to the negative supply bus should always be used. When the voltage at pin 5 goes higher than the reference voltage at pins 4 and 6, pin 11 goes low, turning on the trip indicator LED DS2 and the optical coupler LED. LED current is limited by the l-kΩ resistor. The optical coupler turns on the TRIAC, which creates a 10-mA fault current between the hot wire and ground of the GFCI. DS2 will remain lit as an indicator until the 100 µF capacitor is discharged.

ADJUSTMENT

Adjustment is simple. You'll need a variable ac transformer (Powerstat or Variac). Turn R1 fully clockwise and use the variable transformer to adjust input to 130 V ac. Turn the pot counterclockwise until the GFCI trips.

7.16.8 Overvoltage Crowbar Circuit

If the regulator circuitry of a power supply should fail — for example, if a pass transistor should short — the unregulated supply voltage could appear at the output terminals. This could cause a failure in the equipment connected to the supply. The overvoltage "crowbar" circuit shown here has been shown to be effective as a last line of defense against power supply overvoltage failures. When an overvoltage condition is detected, the heavy-duty SCR is fired, becoming a short circuit — as if a crowbar (representing any over-sized conductor capable of handling the supply's short-circuit output current) had been connected across the supply, thus the name.

This circuit shown in **Fig 7.83** was originally designed for the 28-V power supply project available on the CD-ROM for this book. The use of the MC3423 overvoltage protection IC provides quicker triggering and more reliable gate drive to the SCR than comparable Zener-based circuits. Power supply builders can incorporate the overvoltage protection circuit into any dc power supply with the required component value adjustments.

The crowbar circuit is usually connected with the power input and the sense input connected together at the supply positive output and COMMON to the supply negative output. Another option is to connect power input and

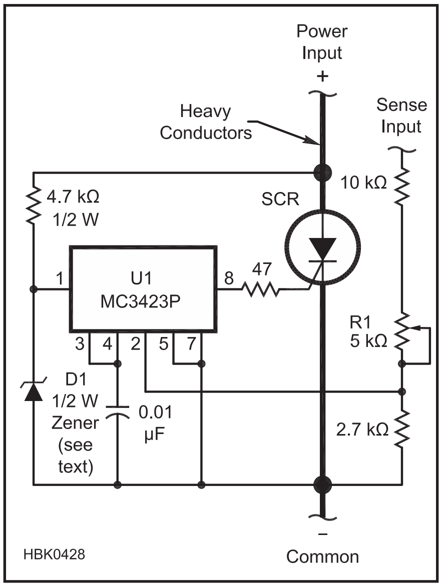


Fig 7.83 — Schematic of the over-voltage crowbar circuit. Unless otherwise specified, resistors are ¼-W, 5%-tolerance carbon-composition or film units. A PC board is available from FAR Circuits (www.farcircuits.net).

SCR — C38M stud-mount (TO-65 package).

U1 — MC3423P or NTE7172 voltage protection circuit, 8-pin DIP.

D1 —Zener diode, ½-W (see text).

R1 — 5 kΩ, PC board mount trimmer potentiometer.

common across the rectifier filter capacitor and the sense input to the power supply output. In either case, the power input and common wires should be adequately sized to handle the full short-circuit current and are shown as heavy lines on the schematic.

The crowbar circuit functions as follows: the 4.7 kΩ resistor and Zener diode D1 create a supply voltage for the MC3423. U1 will function properly with a supply voltage of 4.5-40 V. Use a Zener diode with a voltage rating a few volts below that of the crowbar circuit positive-to-negative power input voltage. For example, if the crowbar is connected to a 12-V supply output, D1 voltage should be 6 to 9 V. The exact value is not critical.

U1 contains a 2.5-V reference and two comparators. When the voltage at pin 2 (sense terminal) reaches 2.5 V, the output voltage (pin 8) changes from the negative input voltage to the positive input voltage. This drives the gate of the SCR through the 47-Ω resistor. The trip voltage is set by the resistive divider across the + and – inputs:

$$V_{trip} = 2.5\left(1 + \frac{10\text{ k}\Omega + R}{2.7\text{ k}\Omega}\right)$$

The application notes for the MC3423 recommend that the resistance from the sense input to the negative input be less than 10 kΩ for minimum drift, suggesting the value of 2.7 kΩ. The value of 10 kΩ for the fixed portion of the adjustable resistance is selected for V_{trip} = 15 V at the midpoint of the 5 kΩ potentiometer (R1) travel. For other trip voltages, the fixed resistor value should be the closest standard value to

$$R_{fixed} = 2.7\text{ k}\Omega\left(\frac{V_{trip}}{2.5} - 1\right) - 2.5\text{k}\Omega$$

assuming the potentiometer value remains at 5 kΩ.

When the SCR turns on, it short-circuits the inputs, causing any protective fuses or circuit breakers to open. The SCR will stay on until the current through drops below the "keep alive" threshold, at which point the SCR turns off. The SCR will stay on even if the input voltage to U1 drops below 4.5 V.

If the crowbar circuit fires due to RFI, an additional 0.01-µF capacitor should be connected from pin 2 of U1 to common and another across the D1.

More information may be found in the MC3423 datasheet and "Semiconductor Consideration for DC Power Supply Voltage Protector Circuits," ON Semi AN004E/D, both available from ON Semiconductor.

Additional Projects and Information

Additional power supply projects and supporting files are included on the *Handbook* CD-ROM.