A 12-V, 15-A Power Supply

This husky and attractive power supply is easy to build and is an ideal power platform for your solid-state transceiver or workbench.

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Most amateurs have seen low-voltage, high-current regulated supplies, from the switching supplies (*switchers*) used in computers to three-terminal linear regulators (*linears*) used in many pieces of ham gear. Although switching technology is in vogue today, switchers generally produce lots of RF noise and exhibit limited dynamic load regulation. On the other hand, linear regulators offer good dynamic load regulation and generate little RF noise. Linears are usually heavier and less efficient than switchers, however. For amateur use, weight—in most cases—is not an issue, and the loss of efficiency, which translates to higher dissipated power, can be tolerated. Therefore, the linear regulator is still the most common design in amateur use.

This supply is a linear 12-V, 15-A design 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.

Circuit Description

Fig 1 is the supply's schematic. The ac-line input is fused by F1, switched on and off by S1 and filtered by FL1. For safety, F1 and S1 are mandatory. F1 and S1 are rated at about one-fourth 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) market 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 a cold resistance of 5 ohms. When it's hot, RT1's resistance drops to 0.11 ohm. Such a low resistance has a negligible effect on supply operation. Thermistors run *HOT!*[2] They must be mounted in free air, and away from anything that can be damaged by heat.

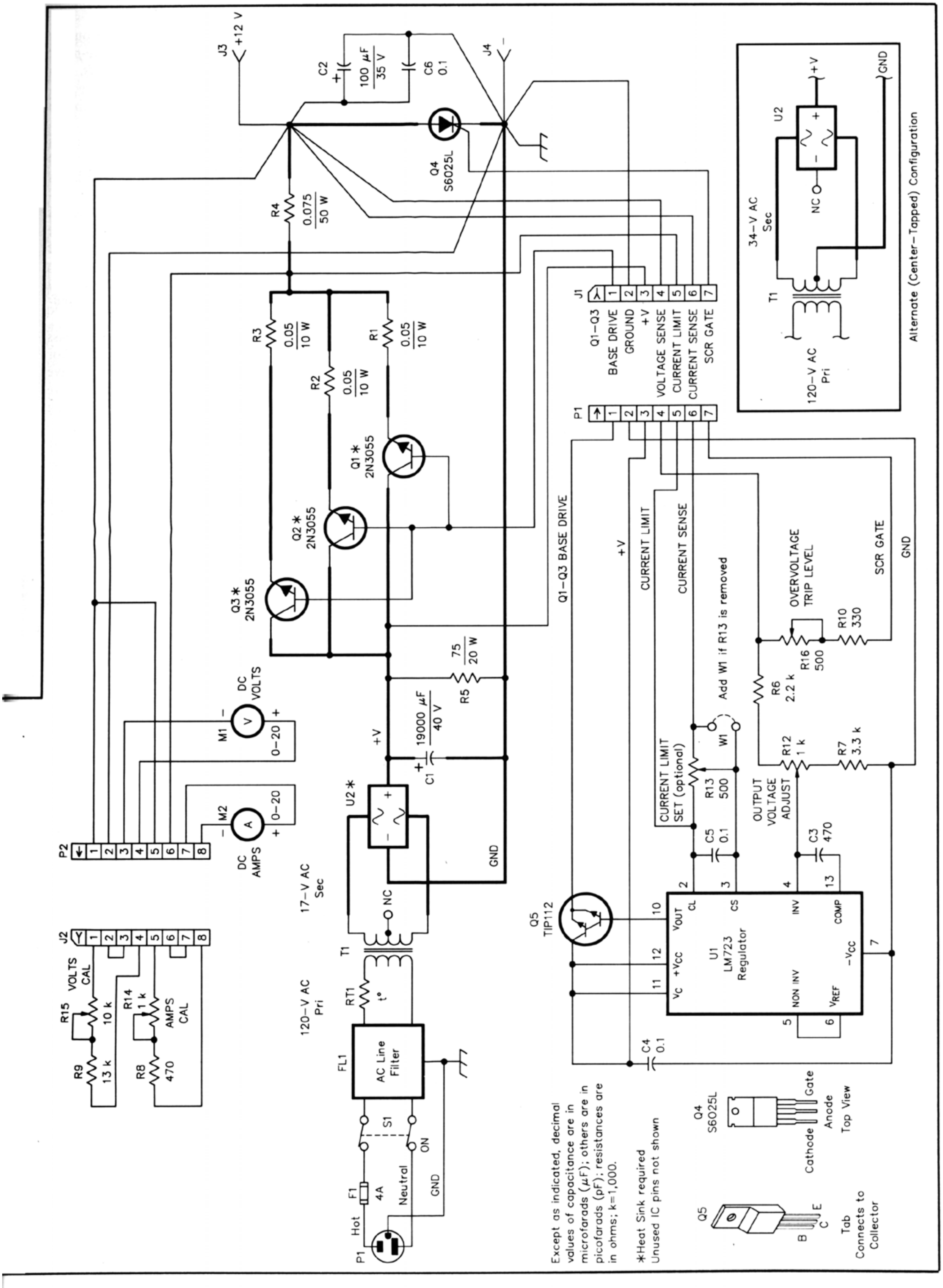
[1]Notes appear on page 41.

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 (see Table 1) offer power transformers that can be used in this supply.

Two parameters important to the power-

Fig 1—Schematic of the power-supply. Equivalent parts can be substituted. Unless otherwise specified, resistors are ¼-W, 5%- tolerance carbon-composition or film units. The bold lines indicate high-current paths that should use heavy-gauge (#10 or #12) wire. This schematic graphically shows wiring to a single-point ground; see text. The majority of the parts used in this supply are surplus components.

C1—19,000 µF, 40-V computer-grade electrolytic capacitor.
C2—100 µF, 35-V capacitor.
C3—470 pF, 50 V.
C4, C5—0.1 µF, 50-V.
F1—120-V, 4-A, Littlefuse SLO-BLO fuse.
FL1—6-A CORCOM ac line filter (surplus model #6H1 used; new model is #6EH1).
J1, J2—8-position SIP female jack.
J3, J4—Heavy-duty binding posts (one red, one black).
M1—0-to 20-V dc voltmeter (1-mA movement, 1-kΩ coil).
M2—0- to 20-A ammeter (1-mA movement, 1-kΩ coil).
P1, P2—8-position male SIP plug.
P3—3-wire ac plug and line cord.
Q1, Q2, Q3—2N3055 NPN power transistor.
Q4—TIP112 NPN Darlington power transistor.
Q5—S6025L 25-A SCR.
R1, R2, R3—0.05-Ω, 5%-tolerance, 10-W.
R4—0.075 Ω, 5%-tolerance, 50-W.
R5—75-Ω, 5%-tolerance, 20-W.
R6—2.2 kΩ.
R7—3.3 kΩ.
R8—470 Ω.
R9—13 kΩ.
R10—1 kΩ.
R11—330 Ω, ½ W.
R12—1-kΩ multiturn trimmer potentiometer.
R13—500-Ω multiturn trimmer potentiometer.
R14—500-Ω multiturn trimmer potentiometer.
R15—10-kΩ multiturn trimmer potentiometer.
R16—500-Ω multiturn trimmer potentiometer.
RT1—Thermistor, 5-Ω no-load resistance, 0.11-Ω I_{max} resistance, 6-A I_{max} (Digi-Key KC004L-ND; KC003L-ND can be substituted).
S1—DPDT toggle.
U1—LM723C voltage regulator (14-pin DIP).
U2—50-V, 25-A bridge rectifier.
Misc: Enclosure (Hammond Manufacturing #1426Q used here), fuse holder, 14-pin DIP socket, PC board (see Note 10).



Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k=1,000.
*Heat Sink required
Unused IC pins not shown
Alternate (Center-Tapped) Configuration
U1 LM723 Regulator
Q5 TIP112
Q4 S6025L
Q1 * 2N3055
Q2 * 2N3055
Q3 * 2N3055
AC Line Filter
120-V AC Pri
17-V AC Sec
34-V AC Sec
DC VOLTS
DC AMPS
VOLTS CAL
AMPS CAL
CURRENT LIMIT SET (optional)
OUTPUT VOLTAGE ADJUST
OVERVOLTAGE TRIP LEVEL
Add W1 if R13 is removed
Q1-Q3 BASE DRIVE
GROUND
+V
VOLTAGE SENSE
CURRENT LIMIT
CURRENT SENSE
SCR GATE
GND
Hot
Neutral
Cathode
Anode
Gate
Top View
Tab Connects to Collector

Table 1
Parts Sources

A&A Engineering 2521 W La Palma Ave, Unit K Anaheim, CA 92801 Tel: 714-952-2114 Fax: 714-952-3280	Transformers, power transistors, bridge rectifiers, heat sinks, capacitors, cabinets.
All Electronics Corp PO Box 567 Van Nuys, CA 91408-0567 Tel: 800-826-5432, 818-997-1806 Fax: 818-781-2653	Transformers, power transistors, bridge rectifiers, heat sinks.
Digi-Key 701 Brooks Ave South PO Box 677 Thief River Falls, MN 56701-0677 Tel: 800-344-4539, 218-681-6674 Fax: 218-681-3380	Thermistors, SCRs, transistors, and most other small parts.
R&D Electronics 1224 Prospect Ave Cleveland, OH 44115 Tel: 800-642-1123, 216-621-1121 Fax: 216-621-8628	Transformers, power transistors, bridge rectifiers, heat sinks.

supply design are the transformer secondary's voltage and current capability. For a 12-V supply, a secondary voltage of about 15 to 17 V ac under load is adequate. If a center-tapped transformer is used (you have to alter the rectifier connection and ground the center tap; see the inset of Fig 1), the secondary has to deliver twice that (30 to 34 V CT). Transformers with higher secondary voltages can be used, but the power dissipated (wasted) in the pass transistors (Q1-Q3) increases proportionally. If you're fortunate, you may find a transformer that has an additional secondary winding—or a tapped primary winding—that offers the ability to fine-tune the secondary voltage. For intermittent duty (such as SSB or CW), the secondary winding's current rating should be at least equal to the required output current. If continuous duty is required of the supply (such as in FM or RTTY service), increase the secondary current rating requirement by about 25%.

T1 produces 17 V ac at 20 A; the center tap is not used. Bridge rectifier U2 provides full-wave rectification. Full-wave rectification produces a low ripple component on the filtered dc, which results in dissipating little power in the filter capacitor. U2's voltage rating should be at least 50 V, and its current rating about 25% higher than the normal load requirement; a 25-A bridge rectifier will do. U2 is secured to the chassis (or a heat sink) because it dissipates heat.[3]

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

As mentioned earlier, C1 dissipates power proportional to the ripple voltage. With a 15-A load and a measured ripple voltage of 1.5, that amounts to 32 W.[4] Therefore, the *physical* size of the capacitor is important, too. A physically larger capacitor is better able to dissipate the power.

R5, a 75-ohm, 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 ohms to 200 ohms 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. The LM723 used at U1 is an older chip that provides voltage regulation and current limiting with few external components. Additional components can be added to provide output metering. 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 noninverting 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 1 attempts to show this graphically.

The output pass-transistor array consists of a TIP112 Darlington 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. 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]

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, its emitter resistor's voltage drop increases, 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 2). I used 10-W resistors because that's what I had available. 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.

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.

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.[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 my 0.075-Ω resistors) to be used. I normally set the current limit at 20 A because that's the top end of the ammeter scale.

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.

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 1). 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.

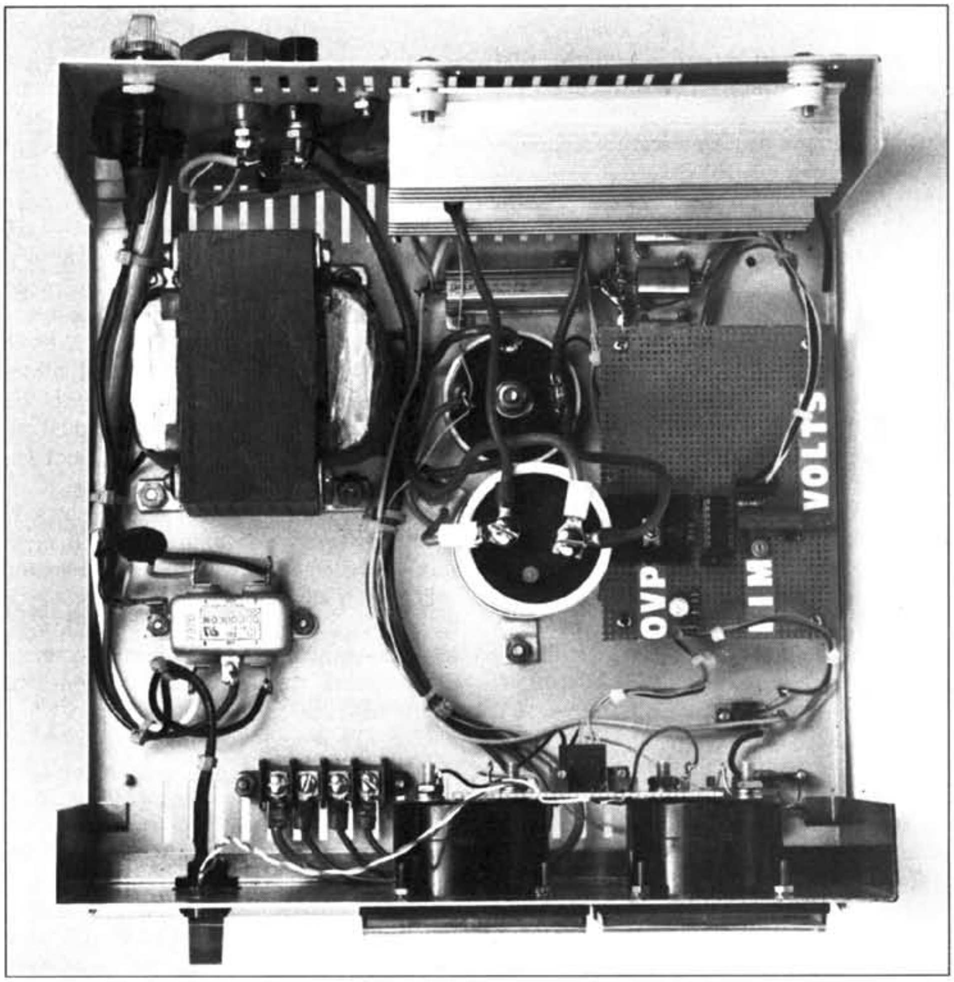


Fig 2—An inside view of the neatly constructed 12-V, 15-A power supply. There's plenty of room in this cabinet to accept the components comfortably and allow hands and tools to get at them. Vents in the bottom and rear of the cabinet provide some measure of convection cooling. For safety, all exposed ac-line leads and connections are insulated with shrink tubing. Wire bunches are secured with plastic cable ties. The terminal block at the lower left serves only to provide a resting place for unused transformer secondary-winding leads. Along the left of the cabinet are the ac-line filter, thermistor (it looks like a large, black ceramic capacitor) and power transformer. The fuse holder and line cord are behind the transformer, as are the dc output terminals J3 and J4. The black objects across the terminals are C2 and C6. Pass transistors Q1-Q3 are hidden by the heat sink fins. Note the insulated standoffs supporting the heat sink. The metal-cased power resistors are secured to the enclosure bottom beneath the heat sink. Bridge rectifier U2 is between the heat sink and C1. The parts mounted on the perfboard secured to the chassis near U2 and C1, and the perfboard mounted behind the meters, are now contained on one PC board (see Note 10). The small three-terminal device between the perf board and front panel is the SCR, Q4. Bleeder resistor R5 is secured to the chassis beneath M2.

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. R10 and potentiometer R16 in series with the Q4's gate provide a means of adjusting the trip voltage. I set the crowbar in my supply to conduct at 15 volts. 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 to insulate it from the chassis or heat sink.)

The bold lines in Fig 1 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 do not have errors caused by voltage drops in the wiring. If desired, the sense wires 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 2 shows the inside of the prototype supply. The larger components are chassis mounted; two perfboards contain the majority of the low-power parts of the supply. This includes the potentiometers for the regulator, meters and over-voltage adjustments. A PC board is available that contains all of the parts mounted on the two perf boards.[10, 11]

Start construction by selecting a cabinet or chassis adequate to contain the components. Not only must the chassis have sufficient room inside, it must also be sturdy enough to support the weight of the components without deforming. I used an attractive Hammond Manufacturing #1426Q cabinet that measures 5.5 × 11 × 11.7 inches (HWD).

Once the chassis is selected, lay out the front panel. Drill and punch the necessary holes, apply the appropriate labels and coat the panel with clear acrylic paint.

I placed the transformer on the left side of the chassis for best access to the power switch. Wires from an unused secondary winding and the center tap are routed to an out-of-the-way location on the chassis and connected to a terminal strip, isolating them from each other and the surrounding components. On the chassis bottom, U2 is positioned near T1, as are C1 and R5. Identify the position of the regulator PC board. Don't locate components too near the rear of the chassis because the heat sink and/or circuit wiring need clearance.

Q1, Q2 and Q3 must be mounted on a heat sink. The one I used measures 1-1/4 × 6 × 3-5/8 inches (HWD). It's the minimum size I'd recommend using. 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 my supply, the entire heat sink can be isolated from the chassis. 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. 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 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 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 wire to connect together the collectors. 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 wire to connect together the transistor base leads, and provide a pigtail for attachment to U1. Using #12 wire, connect the emitters of Q1-Q3 to their respective emitter-balancing resistors. Solder together the remaining emitter-resistor leads and use #10 wire to connect them to R4. Solder the other side of R4 to J3, the positive output terminal.

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 wire for the voltage-sense wire and #18 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 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 SCR's 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 to bring the output voltage to 12. Adjust R15 for a 12-V reading on the 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 flow 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 (I use a discarded margarine container) of 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, and the over-voltage trip point is 15. 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 volts.

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-A load voltages should be less than 50 mV. (It is typically less than 20 mV on my 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.

Summary

This supply was originally designed to power a 100-W, solid-state amplifier for 10-meter FM operation. It operated fine in that application for a number of years. At that time, I used only two of the three pass transistors in the supply and was able to get 100 W output from the amplifier without overtaxing the supply. I added the third pass transistor to increase the output current to meet the requirements of some of the newer HF rigs.

The supply now powers VHF equipment in my shack and doubles as a lab bench supply. For applications that need less current, you can build the supply using a single pass transistor. This should prove capable of powering even the newer VHF rigs, many of which now provide 40 to 50 W output. I've had no problems powering a 25-W, 2-meter rig from the supply; with only two pass transistors installed, it doesn't even get warm. I've not experienced any significant RF problems with the supply. The 100-μF and 0.1-μF capacitors across the output terminals shunt any RF on the power lines to ground before it gets inside the supply. If RF problems do arise, better grounding of the equipment in the shack—or better antenna matching—is probably called for.

As presented here, the supply is as modular as possible. This allows you to add (or delete) some parts as cost or needs warrant. The transformer size, number of pass transistors, current limiting, over-voltage protection and metering circuits can all be modified to support your requirements. With a little ingenuity, the core of this supply could find its way into many useful applications in the shack. You might want to modify the supply to provide a 5-V output for a logic supply, or change the voltage-adjust circuit to provide a variable output for use as a bench supply. Whatever the application, this supply can provide you with a reliable power source for years to come.

Notes

[1]Limiters with higher current ratings are available.

[2]ARRL Lab measurements show the surface temperature of the thermistor to be about 100 °C.

[3]U2's heat dissipation is calculated by:

$$P_{(watts)} = 0.7 \times I \times 2 = 21 \qquad \text{(Eq 1)}$$

where I is the maximum delivered current (15 A), 0.7 is a typical diode voltage drop, and 2 is for the two diodes in the bridge that are simultaneously conducting.

[4]In order to determine the power, we also need either the ripple current, or the impedance of the capacitor. The impedance of a 19,000-μF (0.019-F) capacitor at the 120-Hz ripple frequency (120 Hz because of full-wave rectification) is:

$$Z = \frac{1}{(2\pi \times 120 \text{ Hz} \times 0.019 \text{ F})} = 0.07\ \Omega \qquad \text{(Eq 2)}$$

Plugging the 0.07-Ω impedance into the power equation with the measured 1.5 ripple voltage and a 15-A load yields

$$P_{watts} = \frac{V^2}{Z} = \frac{(1.5)^2}{0.07\ \Omega} = 32 \qquad \text{(Eq 3)}$$

Thirty-two watts is a lot of power, so you can see the physical size of the capacitor is important. The larger the capacitor, the better it can dissipate that power.

[5]Simply dividing the maximum required output current (15 A) by the transistor beta, you can determine the drive requirement of the LM723. If we assume betas of 10 for the 2N3055s and 100 for the Darlington, the drive current is required from the LM723 is:

$$I = \frac{\left(\frac{15 \text{ A}}{\beta_{2N3055}}\right)}{\beta_{TIP112}} = \frac{\left(\frac{15}{10}\right)}{100} = 0.015 \text{ A} \qquad \text{(Eq 4)}$$

or 150 mA.

[6]This is caused by manufacturing-process variations in the transistor die that result in different voltage drops across the part. These differences are very small, but can result in large variations in current flow among devices.

[7]The voltage range is determined by the equations:

$$V_{out} \text{ Upper} = V_{ref} \left(\frac{(3.3 + 1 + 2.2)}{3.3}\right) \qquad \text{(Eq 5)}$$

$$V_{out} \text{ Lower} = V_{ref} \left(\frac{(3.3 + 1 + 2.2)}{(3.3 + 1)}\right) \qquad \text{(Eq 6)}$$

where 3.3 is the 3.3-kΩ resistance of R7, 2.2 is the 2.2-kΩ resistance of R6 and 1 is the 1-kΩ resistance of R12. U1's reference voltage (V_{ref}) is nominally 7.15 This results in an output-voltage range of 10.8 to 14.1 V dc. The range can be increased by increasing the value of R12. (For example, a 2.5-kΩ potentiometer yields 9.8 to 17.3 V.)

[8]Therefore, the value of the current-limiting resistor for a fixed output is

$$R_{limit} = \frac{0.65 \text{ V}}{I_{out}} \qquad \text{(Eq 7)}$$

[9]For a 1-mA meter movement and 20-V full-scale reading, the resistance should be:

$$R = \frac{20 \text{ V}}{1 \text{ mA}} - R_{meter} \qquad \text{(Eq 8)}$$

where R_{meter} is the dc resistance of the meter movement. See the Test Equipment and Measurements chapter of The *Handbook* for information on how to determine the meter's internal resistance. In recent editions, this is Chapter 25.

[10]Bare PC boards ($15), assembled PC boards ($30) and kits of parts containing the PC board and board-mounted parts ($25) are available from Single Chip Solutions, PO Box 680, New Hartford, CT 06057. Please add $3.50 for shipping and handling charges; Connecticut residents add sales tax.

[11]A PC-board template package is available free of charge from the ARRL. Please address your request for the OSCARSON 12-V, 15-A POWER SUPPLY PC-BOARD TEMPLATE to the Technical Department Secretary, ARRL, 225 Main St, Newington, CT 06111. Please include a business-size SASE.

[12]If the sense resistor you use has a value other than 0.075 Ω, the minimum-current limit is calculated by

$$I_{limit} = \frac{0.65 \text{ V}}{R_{sense}} \qquad \text{(Eq 9)}$$

where R_{sense} is the value of the sense resistor in ohms. QST

ARRL Handbook CD
Template File

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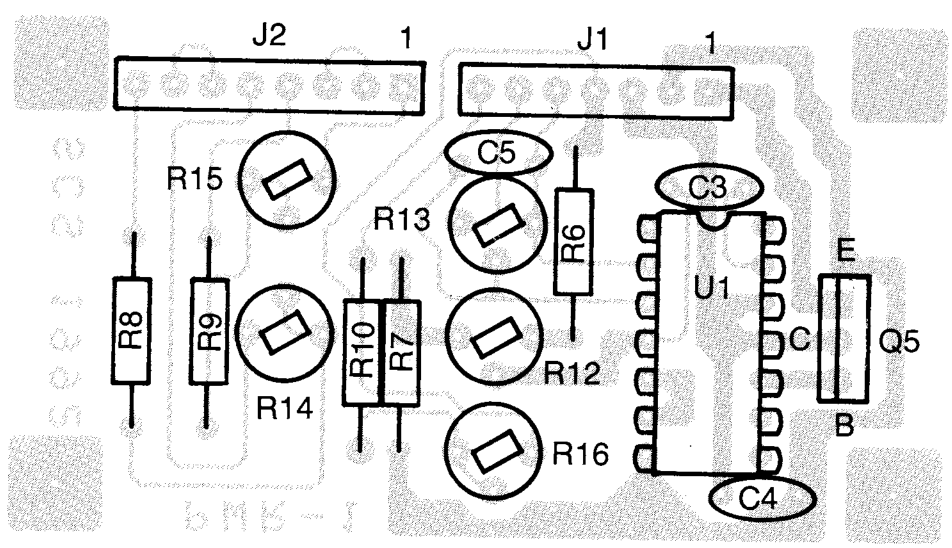
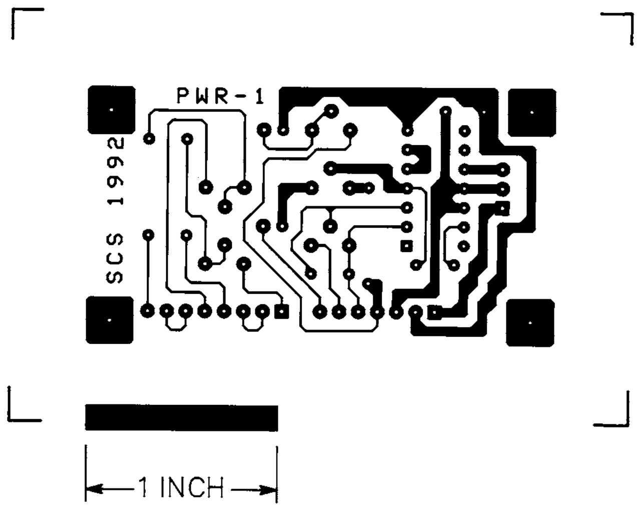
Chapter: 7
Topic: 13.8 V, 15 A power supply project by WA1TWX

Template contains:

PC board etching pattern
Parts placement diagram

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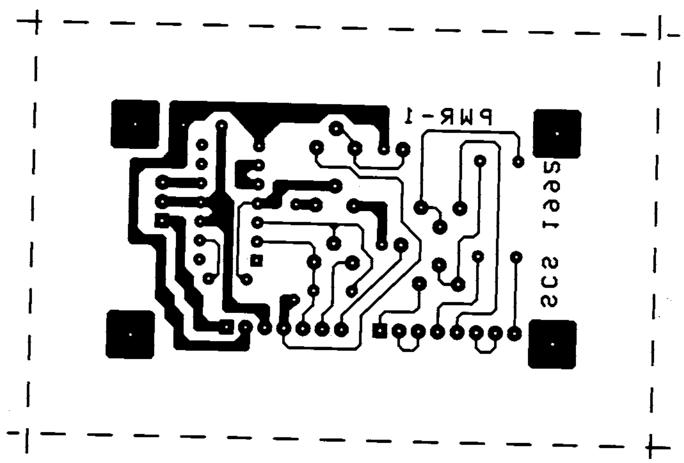
SCS 1992
1 INCH



ARRL Handbook CD
Template File

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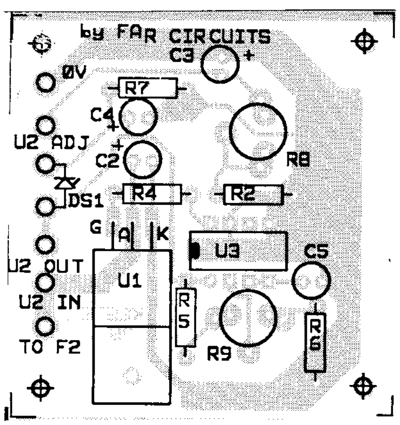
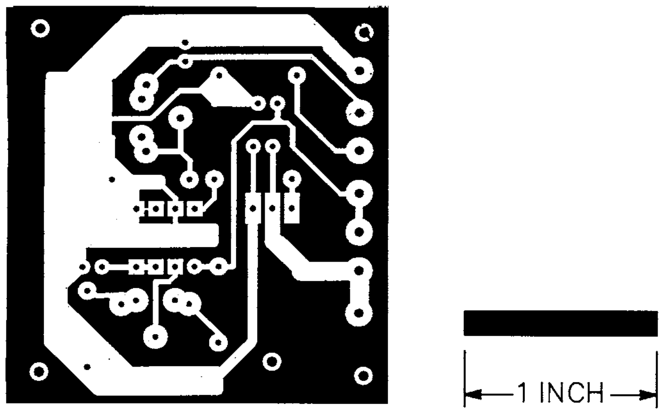
Chapter: 7
Topic: 13.8 V, 5 A power supply project by G4YNM

Template contains:

PC board etching pattern
Parts placement diagram

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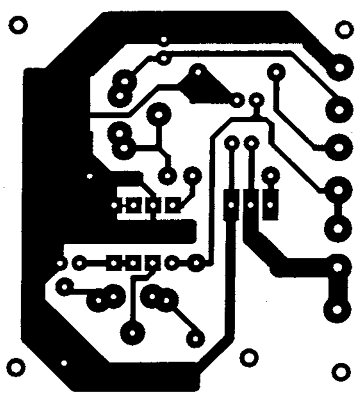

28-V, HIGH-CURRENT POWER SUPPLY

Many modern high-power transistors used in RF power amplifiers require 28-V dc collector supplies, rather than the traditional 12-V supply. By going to 28 V (or even 50 V), designers significantly reduce the current required for an amplifier in the 100-W or higher output class. The power supply shown in **Fig 17.44** through **Fig 17.48** is conservatively rated for 28 V at 10 A (enough for a 150-W output amplifier) — continuous duty! It was designed with simplicity and readily-available components in mind. Mark Wilson, K1RO, built this project in the ARRL lab.

CIRCUIT DETAILS

The schematic diagram of the 28-V supply is shown in **Fig 17.45**. T1 was designed by Avatar Magnetics specifically for this project. The primary requires 120-V ac, but a dual-primary (120/240 V) version is available. The secondary is rated for 32 V at 15 A, continuous duty. The primary is bypassed by two 0.01-µF capacitors and protected from line transients by an MOV.

U1 is a 25-A bridge module available from a number of suppliers. It requires a heat sink in this application. Filter capacitor C1 is a computer-grade 22,000-µF electrolytic. Bleeder resistor R1 is included for safety because of the high value of C1; bleeder current is about 12 mA.

There is a tradeoff between the transformer secondary voltage and the filter-capacitor value. To maintain regulation, the minimum supply voltage to the regulator circuitry must remain above approximately 31 V. Ripple voltage must be taken into account. If the voltage on the bus drops below 31 V in ripple valleys, regulation may be lost.

Fig 17.44 — The front panel of the 28-V power supply sports only a power switch, pilot lamp and binding posts for the voltage output. There is room for a voltmeter, should another builder desire one.

In this supply, the transformer secondary voltage was chosen to allow use of a commonly available filter value. The builder found that 50-V electrolytic capacitors of up to about 25,000 µF were common and the prices reasonable; few dealers stocked capacitors above that value, and the prices increased dramatically. If you have a larger filter capacitor, you can use a transformer with a lower secondary voltage; similarly, if you have a transformer in the 28- to 35-V range, you can calculate the size of the filter capacitor required. Equation 3, earlier in this chapter in the Filtration section, shows how to calculate ripple for different filter-capacitor and load-current values.

The regulator circuitry takes advantage

of commonly available parts. The heart of the circuit is U3, a 723 voltage regulator IC. The values of R8, R9 and R10 were chosen to allow the output voltage to be varied from 20 to 30 V. The 723 has a maximum input voltage rating of 40 V, somewhat lower than the filtered bus voltage. U2 is an adjustable 3-terminal regulator; it is set to provide approximately 35 V to power U3. U3 drives the base of Q1, which in turn drives pass transistors Q2-Q5. This arrangement was selected to take advantage of common components. At first glance, the number of pass transistors seems high for a 10-A supply. Input voltage is high enough that the pass transistors must dissipate about 120 W (worst case), so thermal considerations dictate the use of four transistors. See the **Real-World Component Characteristics** chapter for a complete discussion of thermal design. If you use a transformer with a significantly different secondary potential, refer to the thermal-design tutorial to verify the size heat sink required for safe operation.

R9 is used to adjust supply output voltage. Since this supply was designed primarily for 28-V applications, R9 is a "set and forget" control mounted internally. A 25-turn potentiometer is used here to allow precise voltage adjustment. Another builder may wish to mount this control, and perhaps a voltmeter, on the front panel to easily vary the output voltage.

The 723 features current foldback if the load draws excessive current. Foldback current, set by R7, is approximately 14 A, so F2 should blow if a problem occurs. The output terminals, however, may be shorted indefinitely without damage to any power-supply components.

If the regulator circuitry should fail, or if a pass transistor should short, the unregulated supply voltage will appear at the output terminals. Most 28-V RF transistors would fail with 40-plus volts on the collector, so a prospective builder might wish to incorporate the overvoltage protection circuit shown in **Fig 17.46** in the power supply. This circuit is optional. It connects across the output terminals and

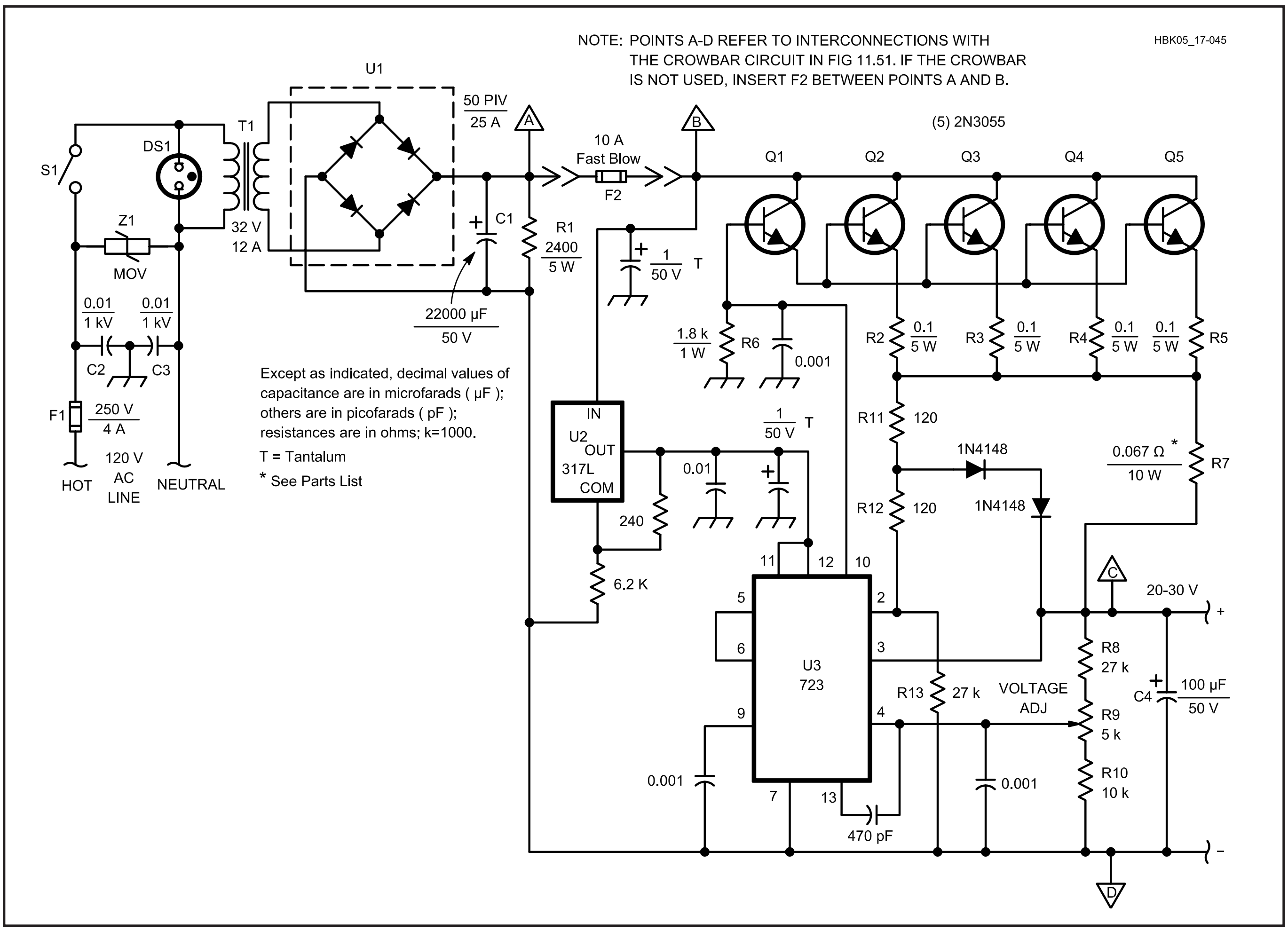

Fig 17.45 — Schematic diagram of the 28-V, high-current power supply. Resistors are ¼-W, 5% types unless otherwise noted. Capacitors are disc ceramic unless noted; capacitors marked with polarity are electrolytic.

C1 — Electrolytic capacitor, 22000 μF, 50 V (Mallory CG223U050X4C or equiv., available from Mouser Electronics)
C2, C3 — ac-rated bypass capacitors.
C4 — Electrolytic capacitor, 100 μF, 50 V
DS1 — Pilot lamp, 120-V ac
Q1-Q5 — NPN power transistor, 2N3055 or equiv
R2-R5 — Power resistor, 0.1 Ω, 5 W (or greater), 5% tolerance
R7 — Power resistor, 0.067 Ω, 10 W (or greater), made from three 0.2-Ω, 5-W resistors in parallel
T1 — Power transformer. Primary, 120-V ac; secondary, 32 V, 15 A. (Avatar Magnetics AV-430 or equiv. Dual primary version is part #AV-431. Available from Heritage Transformers Co.)
U1 — Bridge rectifier, 50 PIV, 25 A
U2 — Three-terminal adjustable voltage regulator, 100 mA (LM-317L or equiv.) See text.
U3 — 723-type adjustable voltage regulator IC, 14-pin DIP package (LM-723, MC1723, etc)
Z1 — 130-V MOV

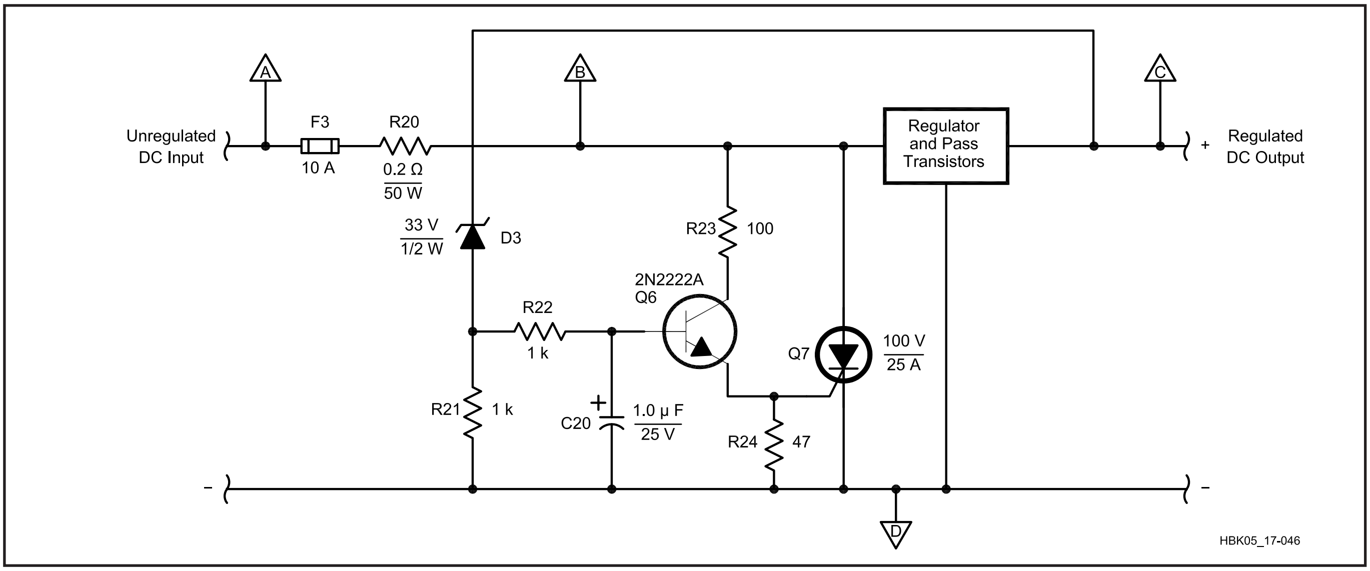

Fig 17.46 — Schematic diagram of the overvoltage protection circuit. Resistors are ¼-W, 5% carbon types unless noted.
D3 — 33 V, ½-W Zener (NTE 5036A or equiv.)
Q6 — NPN Transistor (2N2222A or equiv.)
Q7 — 100 V, 25A SCR (NTE 5522 or equiv.)

may be added or deleted with no effect on the rest of the supply. If you choose to use the "crowbar," make the interconnections as shown. Note that R20 and F3 of Fig 17.46 are added between points A and B of Fig 17.45. If the crowbar is not used, connect F2 between points A and B of Fig 17.45.

The crowbar circuit functions as follows: The Zener-hold off diode (D3) blocks the positive regulated voltage from appearing at the base of Q6 until its avalanche voltage is exceeded. In the case of the device selected, this voltage level is 33 V, which provides for small overshoots that might occur with sudden removal of the output load (switching off a load, for instance).

In the event the output voltage exceeds 33 V, D3 will conduct, and forward bias Q6 through R22 and C20, which eliminates short duration transients and noise. When Q6 is biased on, trigger current flows through R23 and Q6 into the gate of SCR Q7, turning it on and shorting the raw dc source, forcing F3 to blow. Since some SCRs have a tendency to turn themselves on at high temperature, resistor R24 shunts any internal leakage current to ground.

CONSTRUCTION

Fig 17.47 shows the interior of the 28-V supply. It is built in a Hammond 1401K enclosure. All parts mount inside the box. The regulator components are mounted on a small PC board attached to the rear of the front panel. See **Fig 17.48**. Most of the parts were purchased at local electronics stores or from major national suppliers. Many parts, such as the heat sink, pass transistors, 0.1-Ω power resis-

Fig 17.47 — Interior of the 28-V, high-current power supply. The cooling fan is necessary only if the pass transistors and heat sink are mounted inside the cabinet. See text.

tors and filter capacitor can be obtained from scrap computer power supplies found at flea markets.

Q2-Q5 are mounted on a Wakefield model 441K heat sink. The transistors are mounted to the heat sink with insulating washers and thermal heat-sink compound to aid heat transfer. TO-3 sockets make electrical connections easier. The heat-sink surface under the transistors must be absolutely smooth. Carefully deburr all holes after drilling and lightly sand the edges with fine emery cloth.

A five-inch fan circulates air past the heat sink inside the cabinet. Forced-air cooling is necessary only because the heat sink is mounted inside the cabinet. If the heat sink was mounted on the rear panel with the fins vertical, natural convection would provide adequate cooling and no fan would be required.

U1 is mounted to the inside of the rear panel with heat-sink compound. Its heat sink is bolted to the outside of the rear panel to take advantage of convection cooling.

U2 may prove difficult to find. The 317L is a 100-mA version of the popular 317-series 1.5-A adjustable regulator. The 317L is packaged in a TO-92 case, while the normal 317 is usually packaged in a larger TO-220 case. Many electronics suppliers sell them, and direct replacements are available from many local electronics shops. If you can't find a 317L, you can use a regular 317.

R7 is made from two 0.1-Ω, 5-W resistors connected in parallel. These resistors get warm under sustained operation, so they are mounted approximately 1⁄16 inch above the circuit board to allow air to circulate and to prevent the PC board from becoming discolored. Similarly, R6 gets warm to the touch, so it is mounted away from the board toallow air to circulate. Q1 becomes slightly warm during sustained operation, so it is mounted to a small TO-3 PC board heat sink.

Not obvious from the photograph is the use of a single-point ground to avoid ground-loop problems. The PC-board ground connection and the minus lead of the supply are tied directly to the minus terminal of C1, rather than to a chassis ground.

The crowbar circuit is mounted on a small heat sink near the output terminals. Q7 is a stud-mount SCR and is insulated from the heat sink. The other components are mounted on a small circuit board attached to the heat sink with angle brackets.

Although the output current is not extremely high, #14 or #12 wire should be used for all high-current runs, including the wiring between C1 and the collectors of Q2-Q5; between R2-R5 and R7; between F2 and the positive output terminal; and between C1 and the negative output terminal. Similar wire should be used between the output terminals and the load.

TESTING

First, connect T1, U1 and C1 and verify that the no-load voltage is approximately 44 V dc. Then, connect unregulated voltage to the PC board and pass transistors. Leave the gate lead of Q6 disconnected from pin 8 of U4 at this time. You should be able to adjust the output voltage between approximately 20 and 30 V. Set the output to 28 V.

Next, short the output terminals to verify that the current foldback is working. Voltage should return to 28 when the shorting wire is disconnected. This completes testing and setup.

The supply shown in the photographs dropped approximately 0.1 V between no load and a 12-A resistive load. During testing in the ARRL Lab, this supply was run for four hours continuously with a 12-A resistive load on several occasions, without any difficulty.

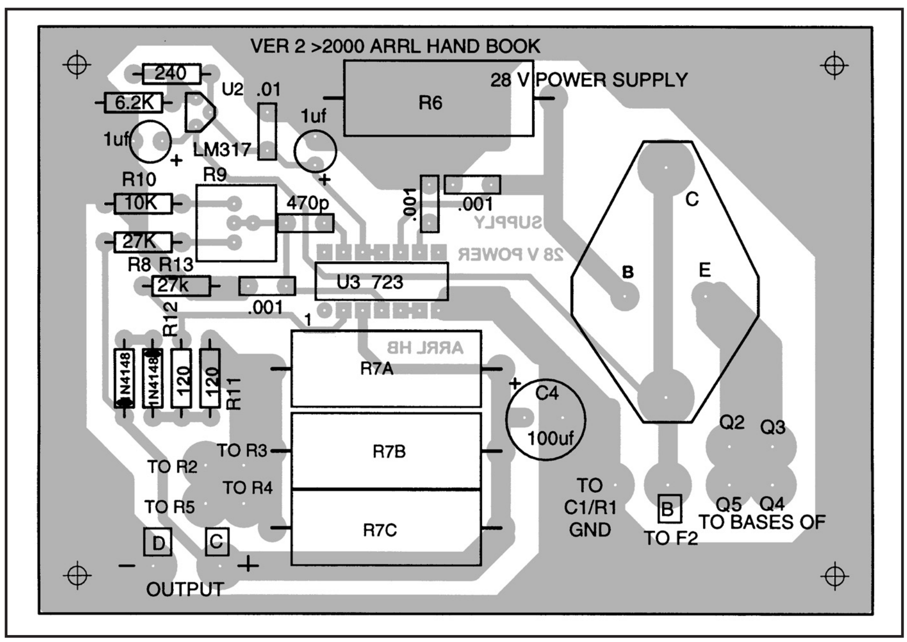

Fig 17.48 — Parts placement diagram for the 28-V power supply. A full-size etching pattern is in the Templates section of the *Handbook* CD-ROM.

ARRL Handbook CD

Template File

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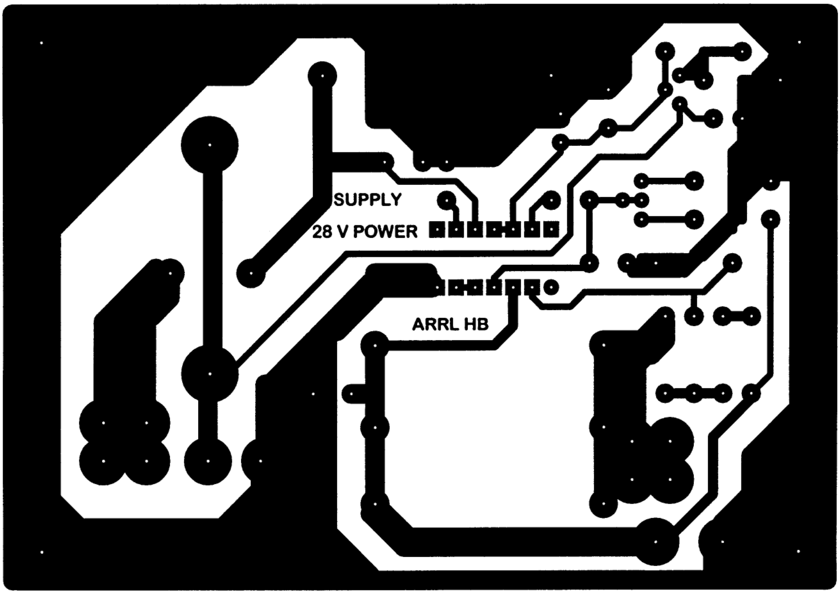
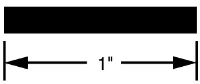
Chapter:	7
Topic:	28-V, High-Current Power Supply

Template contains:

PC board etching pattern.

Parts placement diagram.


SUPPLY
28 V POWER
ARRL HB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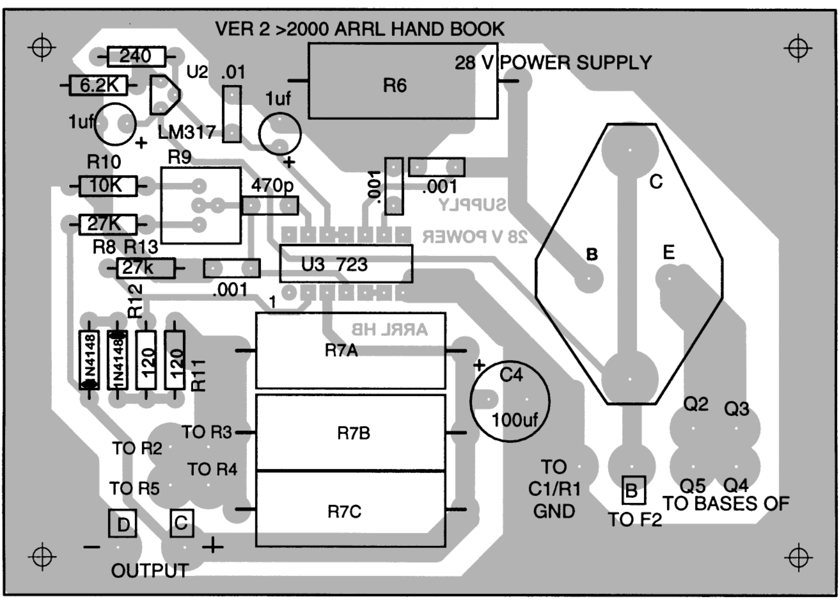

VER 2 >2000 ARRL HAND BOOK
28 V POWER SUPPLY
240
6.2K
U2
.01
LM317
1uf
1uf
R6
R10
R9
10K
470p
.001
.001
27K
R8 R13
27k
.001
U3 723
1
R12
1N4148
1N4148
120
120
R11
R7A
R7B
R7C
C4
100uf
C
B
E
Q2
Q3
Q5
Q4
TO BASES OF
TO C1/R1 GND
TO F2
TO R3
TO R2
TO R4
TO R5
D
C
OUTPUT

James C. Garland, W8ZR

102 Spur Ranch Rd, Santa Fe, NM 87540-7574; **w8zr@arrl.net**

A Deluxe High Voltage Supply

If you have several RF power amplifiers, here is one power supply to run them all.

A half-century ago Larry Kleber, K9LKA, published one of the most popular construction articles in the history of Amateur Radio. Appearing first in November 1961 *QST* and subsequently in eight editions of *The ARRL Handbook*, plus other ARRL publications, the article featured single-band kilowatt amplifiers that shared a common high voltage power supply. Despite the obvious financial benefits of sharing a power supply — typically the most costly part of a high power linear amplifier — this practice has seldom been imitated, even though amplifier building continues to be extremely popular among "homebrew" devotees.

Sharing a high voltage power supply is harder than it might seem. In K9LKA's original design, 2000 V was simultaneously applied to all connected amplifiers, with the filament and screen voltage (the amplifiers used 813 tetrodes with "instant-on" filaments) switched only to the selected amplifier. Today this practice would not only be considered unsafe, but technical advances in amplifier design now necessitate a more sophisticated approach requiring flashover protection, avoidance of metering interaction problems, and so forth.

Despite the obvious convenience of relegating a heavy high voltage power supply to an inconspicuous spot behind an operating desk, the practice of using a separate power supply has lost favor among commercial linear amplifier manufacturers. Once a staple of commercial designs, as in the venerable RL Drake linear amplifiers of the 1970s and '80s, external HV power supplies significantly increase manufacturing costs, in large part because of safety and liability concerns. Safely routing several thousand volts through the rat's nest of cables behind the typical operating desk is a serious and expensive enterprise, and manufacturers are no longer willing to assume the risk of using a single unshielded wire to do the job. (Although used in many commercial designs several decades ago, no manufacturer would today consider using the classic J.W. Millen HV connector, which has no strain relief and uses an unshielded conductor held in place by a single blob of solder.)

Figure 1 — The two high voltage power supplies are identical, except for different voltage ranges. Each rests on the floor behind an operating table and independently powers up to three legal-limit RF power amplifiers.

The high voltage power supplies described here (two power supplies were built, identical except for different output voltages) are intended to overcome these concerns, in effect bringing the benefits, convenience and economy of the 50-year-old design pioneered by K9LKA into the twenty first century. The results, shown in Figure 1, are contest-grade power supplies rated for legal limit continuous duty service in any mode, with substantial "headroom." They sit on the floor behind an operating table, each allowing independent control of one, two, or three remotely located RF decks. For example, one RF deck could be dedicated to 160 m and another to 6 m, both popular bands that vintage commercial amplifiers seldom cover. High power monoband amplifiers are relatively easy to build and design, with none of the tradeoffs and expenses necessitated by multiband designs. (A ceramic multideck high-power band switch, purchased new, can cost more than $500!)

For these power supplies, internal logic circuits handle all the switching and control functions for each RF deck, with vacuum relays designed specifically for dc voltages, safely routing high voltage only to the selected amplifier. (Each power supply is intended for single-operator use, in which only one RF deck is on-line at a time.) Simplicity of operation was an important design goal. Thus, operation of a power supply requires only two momentary-action pushbutton switches on each RF deck, one that toggles on and off the low voltage circuits (blowers, filaments, etc.) and a second that toggles the RF deck on-line or off-line. The low voltage circuits for each RF deck may be turned on simultaneously, but an interlock circuit permits only one amplifier at a time to be on-line. A power failure resets all the control circuitry to an off state, so that the supply must be manually powered up after the power is restored.

Features

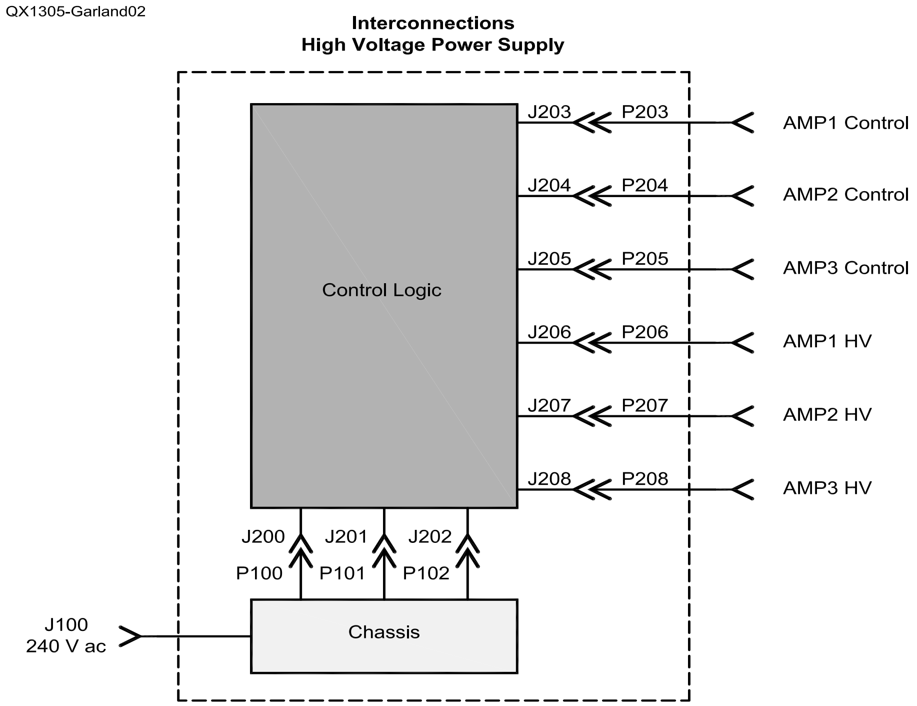
Each power supply is remotely operated by a connected RF deck via a 10 conductor shielded cable. The cable provides switched 120 V ac for powering filaments, blowers, and low-voltage circuits, as well as other connections for power and on-line switching, high voltage metering, plate current trip and reset circuits, indicator lamps, and so forth. The high voltage connection to each RF deck is made through a shielded length of RG6/U coaxial cable, using high voltage BNC connectors rated at 5000 working volts. For safety purposes, the connectors are designed with reverse polarity pins (the male pin is in the jack, rather than the plug), with recessed contacts ensuring that the grounded shield is always connected before the center conductor makes or breaks contact. Other than the 240 V ac circuit breakers and a safety "HV Enable" key-operated switch that must be closed to allow the HV circuits to operate, a power supply has no controls or switches.

Gigavac G81A vacuum relays rated for switching high voltage dc loads up to 10 kV at 5 A transfer the high voltage to the selected RF deck only when that deck is switched on-line. An internal power relay selects primary taps on the plate transformer so that different plate voltages can be automatically assigned to a selected RF deck. Thus, one power supply provides 4500 V dc to two amplifier ports (intended for RF decks using a 3CPX1500A7 triode), and 3700 V dc to the third port for an amplifier that uses 3CPX800A7 triodes. The second power supply, identical to the first except for the plate transformer and filter capacitor, is designed for lower voltage tubes running 2500 to 3000 V dc, such as the 3-500Z and the GU-74B

Circuit Description

Chassis Components

As shown in the block diagram of Figure 2,



Control (AMP1, AMP2, AMP3)
AMP 14 pin
Mouser 571-206043-1

1 - GND
2 - L1 Switched
3 - GND
4 - Ip Reset
5 - LV Control
6 - PWR Pushbutton
7 - Neutral
8 - Spare
9 - HV Meter
10 - In/Out Pushbutton
11- Standby LED
12 - GND
13 - B −
14 - GND

P101/J201 - Chassis Control
12 Pin mini MATE-N-LOC
Mouser 571-1-794066-0 (J201)
Mouser 571-7705811 (P101)

1 - L1 Switched
2 - L1 Switched
3 - Neutral
4 - AC Relay
5 - HV Meter
6 - B −
7 - Ip Sense
8 - 12 V ac
9 - 12 V ac
10 - HV/LV Select
11 - +12
12 - GND

P102/J202 - Status LEDs
Molex 0.100 with Lock 8 pin
Mouser 538-22-01-2087
Mouser 538-22-03-2081

1 - Amp1 PWR
2 - Amp1 HV
3 - Amp2 PWR
4 - Amp2 HV
5 - Amp3 PWR
6 - Amp3 HV
7 - Standby
8- +12

P203/J203, P204/J204, P205/J205, Control
12 Pin mini MATE - N - LOC
Mouser 571-1-794066-0
Mouser 571-7705811

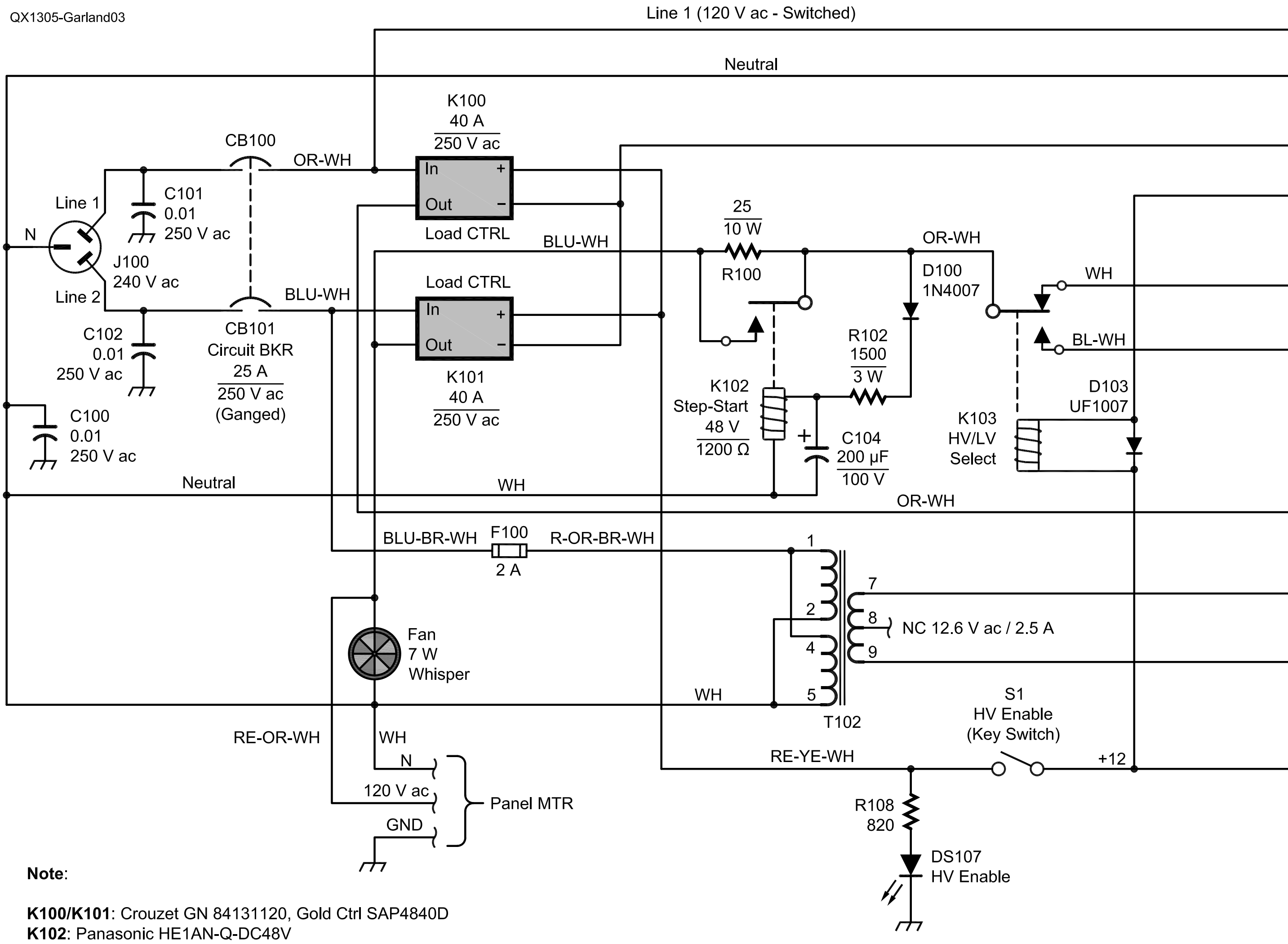
1- Neutral
2 - L1 Switched
3 - GND
4 - Ip Reset
5 - LV Control
6 - PWR Pushbutton
7 - Neutral
8 - Spare
9 - HV Meter
10 - In/Out Pushbutton
11 - Standby LED
12 - B −

HV (Amp1, Amp2, Amp3)
Kings High Voltage BNC 1 pin + GND

P100/J200 HV
HV Cable with Spade Lug (custom) 1 Pin

P206/ J206, P207/J207, P208/J208 HV
HV Cable with Spade Lug (custom) 1 Pin

Figure 2 — The heart of the power supply is a control logic circuit that arbitrates among three connected RF decks, allowing fully automatic operation.



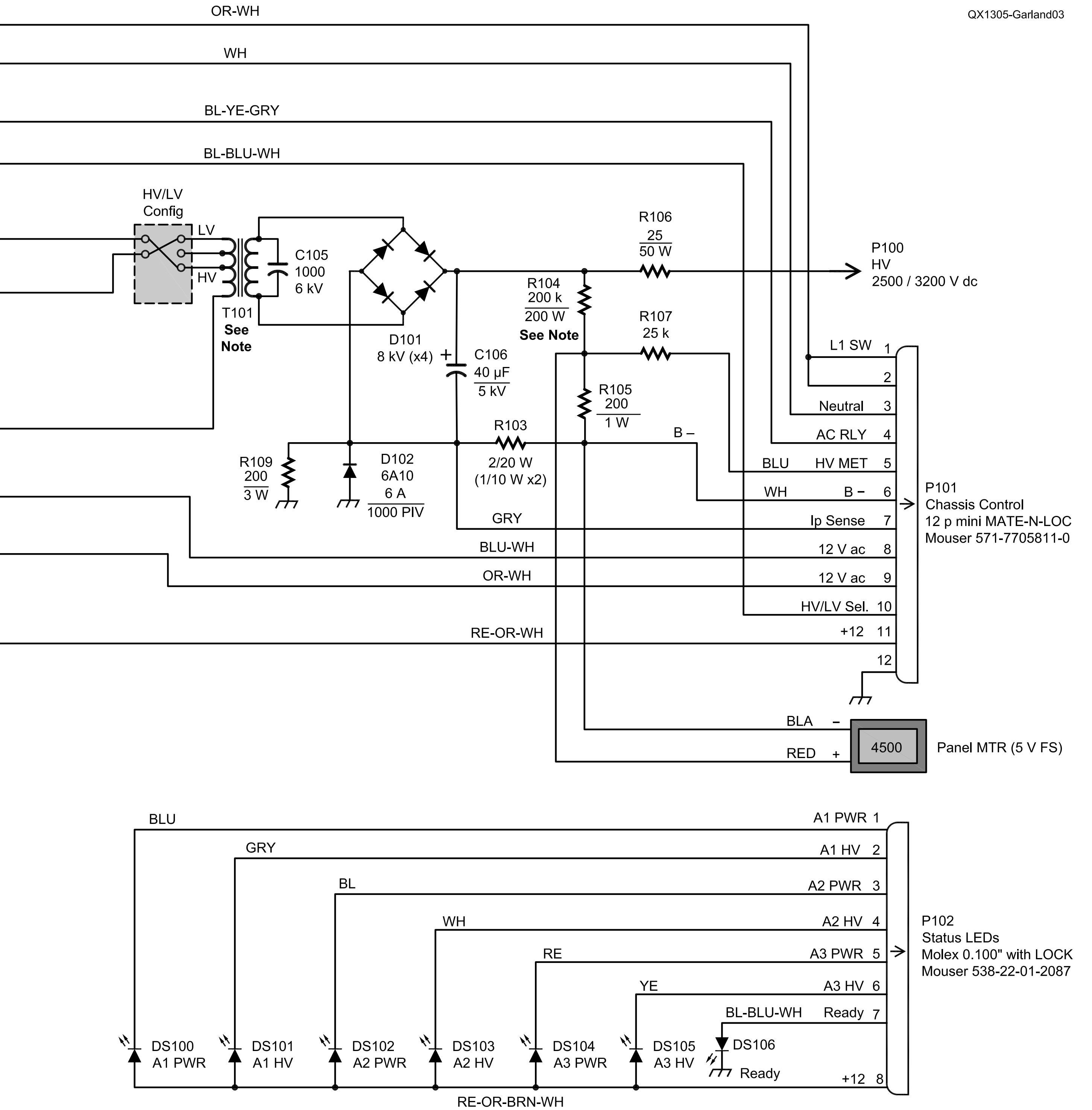
Note:

K100/K101: Crouzet GN 84131120, Gold Ctrl SAP4840D
K102: Panasonic HE1AN-Q-DC48V
K103: Tyco/P&B T92P11D22-12
R104: Two 100 k/100 W in series. 50 k/50 W okay for V < 3000 V dc
T100: Signal DP-241-6-12 (12.6 V ac at 2.5 A)
T101: Peter Dahl Custom 1.5 A CCS

v.1 - pri: P1, P2, P3 sec1: 2040 (P1)/2270 (P2)/2500 (P3)
sec2: 2695 (P1)/2995 (P2)/3300 (P3)

v.2. - pri: P1, P2, P3 sec: 1920(P1), 2050 (P2), 2250 (P3) at 245 V ac

Figure 3 — The power supply uses a capacitor input filter with a large oil-filled capacitor filtering the rectified output from the full-wave bridge rectifier. An "HV Enable" key-operated safety switch disables the plate transformer by deactivating the solid state power relays.



OR-WH
QX1305-Garland03
WH
BL-YE-GRY
BL-BLU-WH
HV/LV Config
LV
HV
C105 1000 6 kV
T101 See Note
D101 8 kV (x4)
R106 25 50 W
P100 HV 2500 / 3200 V dc
R104 200 k 200 W See Note
R107 25 k
C106 40 µF 5 kV
R105 200 1 W
R103 2/20 W (1/10 W x2)
B –
R109 200 3 W
D102 6A10 6 A 1000 PIV
L1 SW 1
2
Neutral 3
AC RLY 4
BLU HV MET 5
WH B – 6
GRY Ip Sense 7
BLU-WH 12 V ac 8
OR-WH 12 V ac 9
HV/LV Sel. 10
RE-OR-WH +12 11
12
P101 Chassis Control 12 p mini MATE-N-LOC Mouser 571-7705811-0
BLA –
RED +
4500
Panel MTR (5 V FS)
BLU A1 PWR 1
GRY A1 HV 2
BL A2 PWR 3
WH A2 HV 4
RE A3 PWR 5
YE A3 HV 6
BL-BLU-WH Ready 7
+12 8
P102 Status LEDs Molex 0.100" with LOCK Mouser 538-22-01-2087
DS100 A1 PWR
DS101 A1 HV
DS102 A2 PWR
DS103 A2 HV
DS104 A3 PWR
DS105 A3 HV
DS106 Ready
RE-OR-BRN-WH

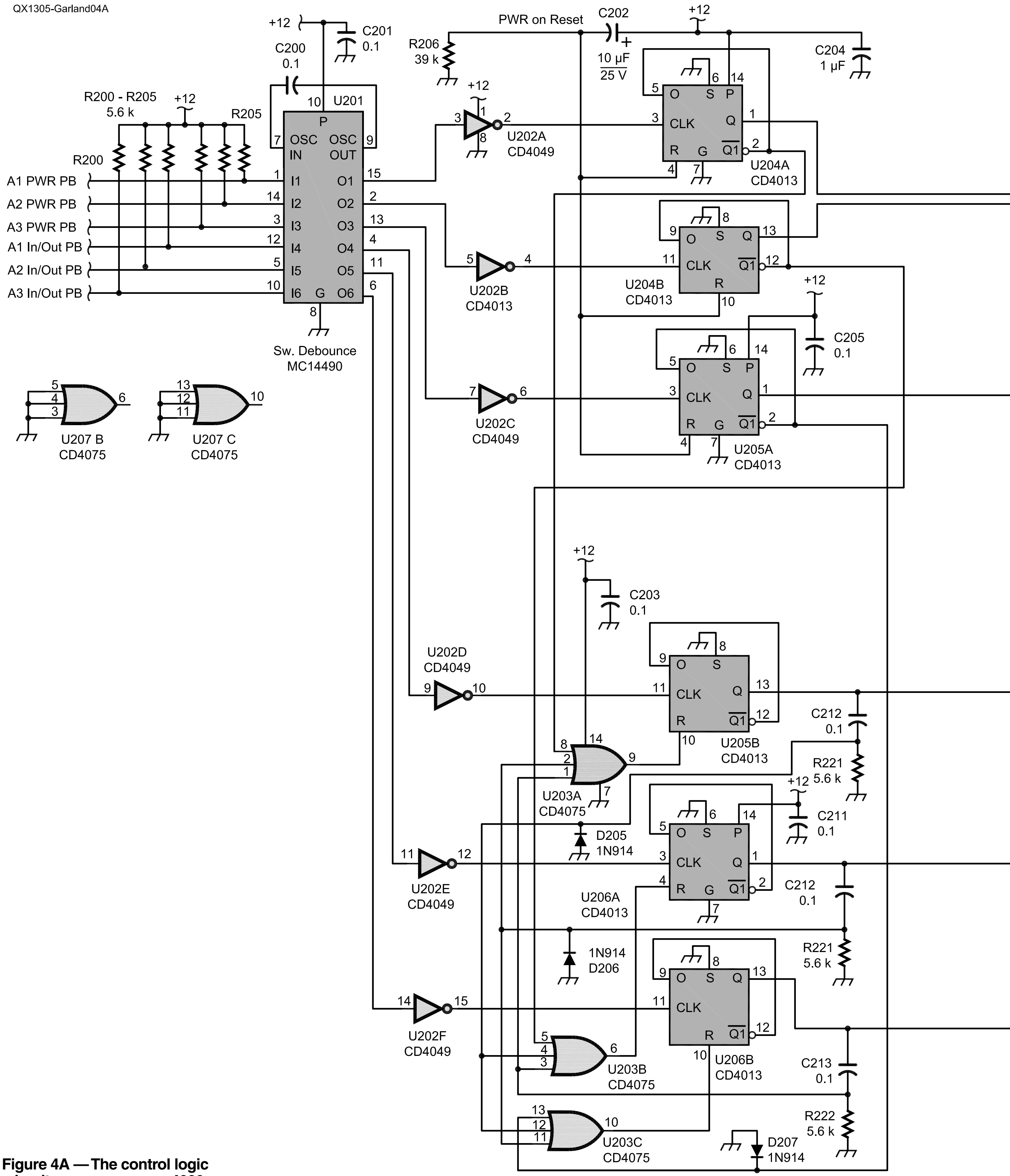
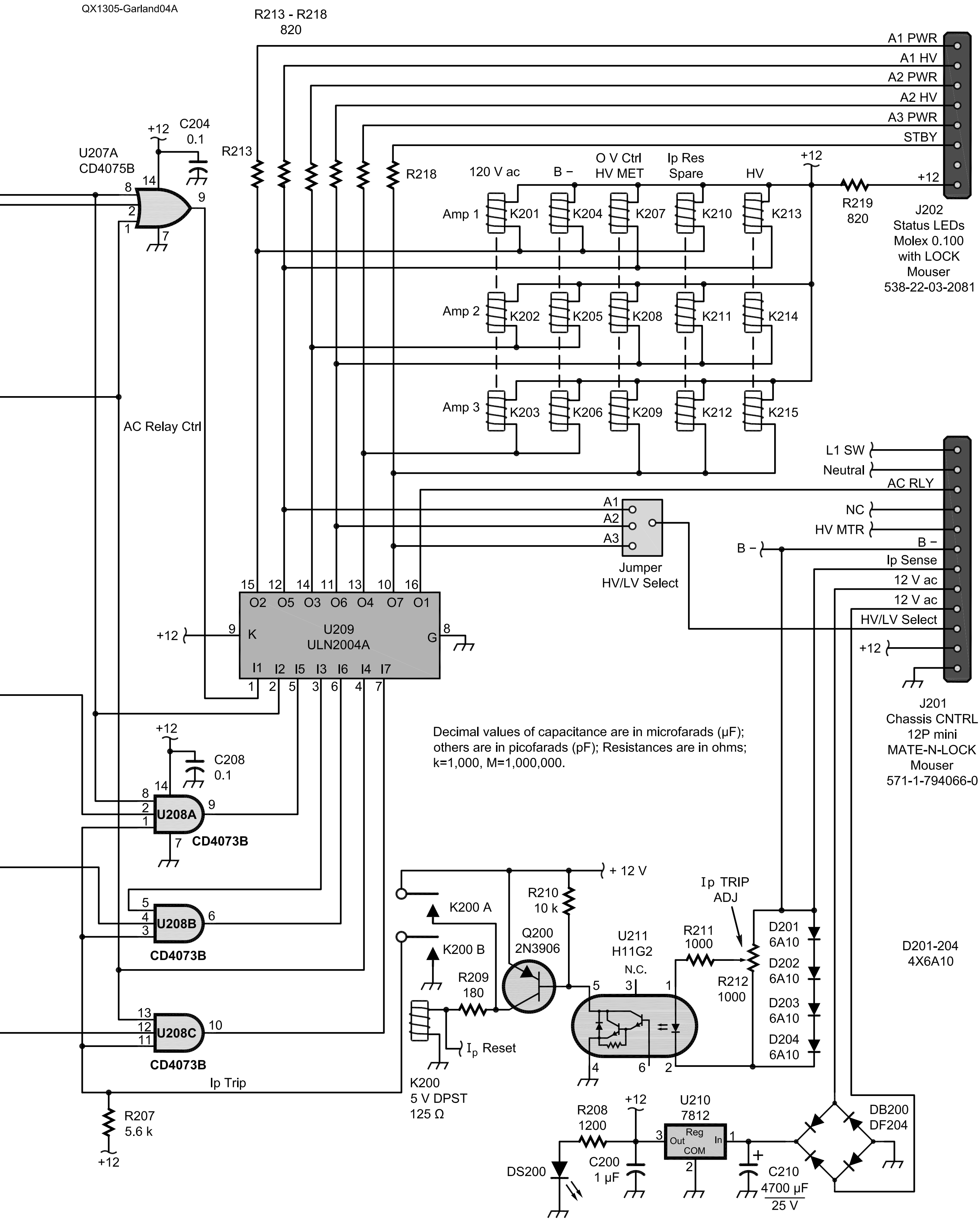


Figure 4A — The control logic circuitry uses common 4000 series CMOS integrated circuits, which operate off of 12 V dc.



Decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); Resistances are in ohms; k=1,000, M=1,000,000.

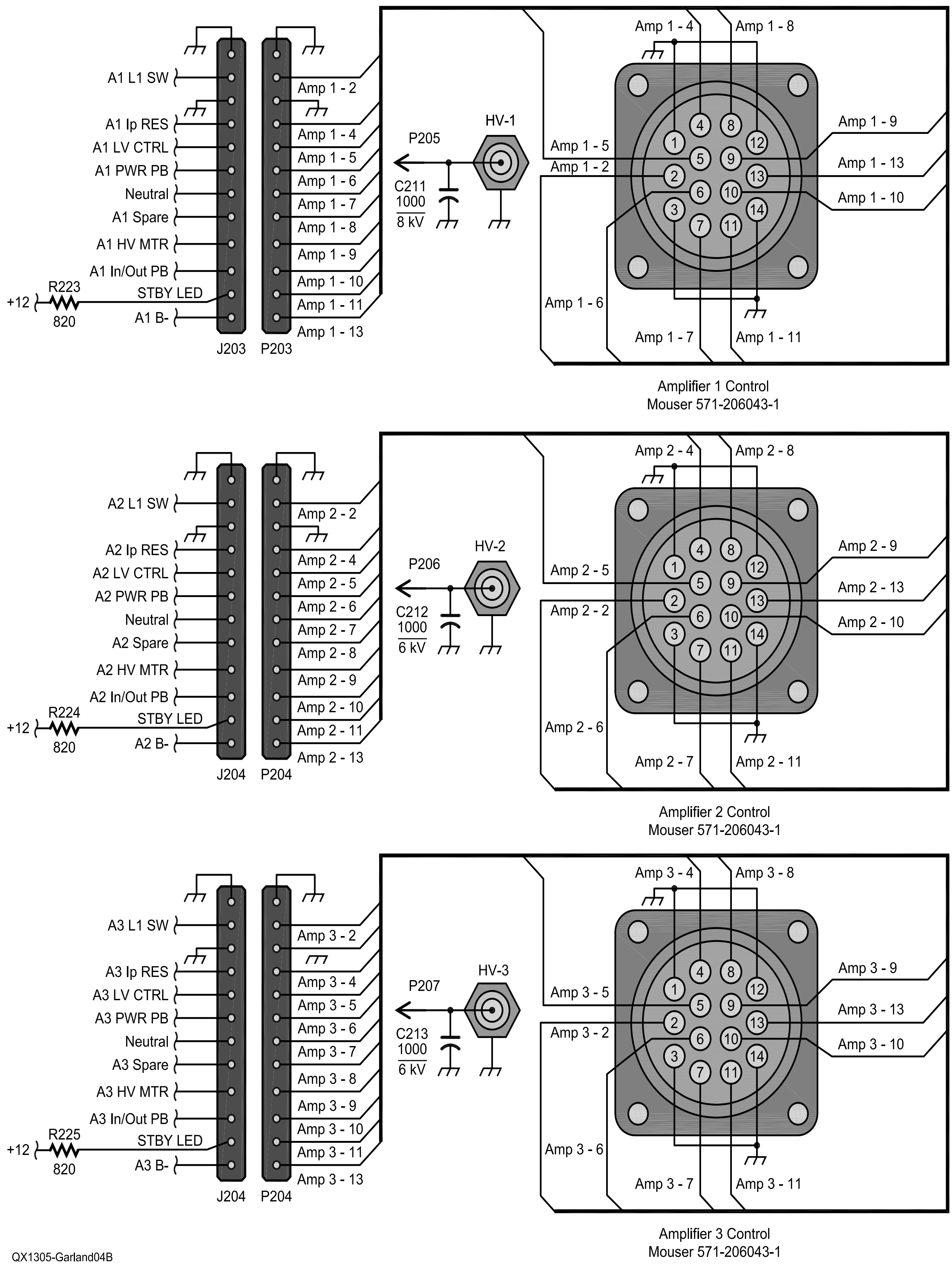
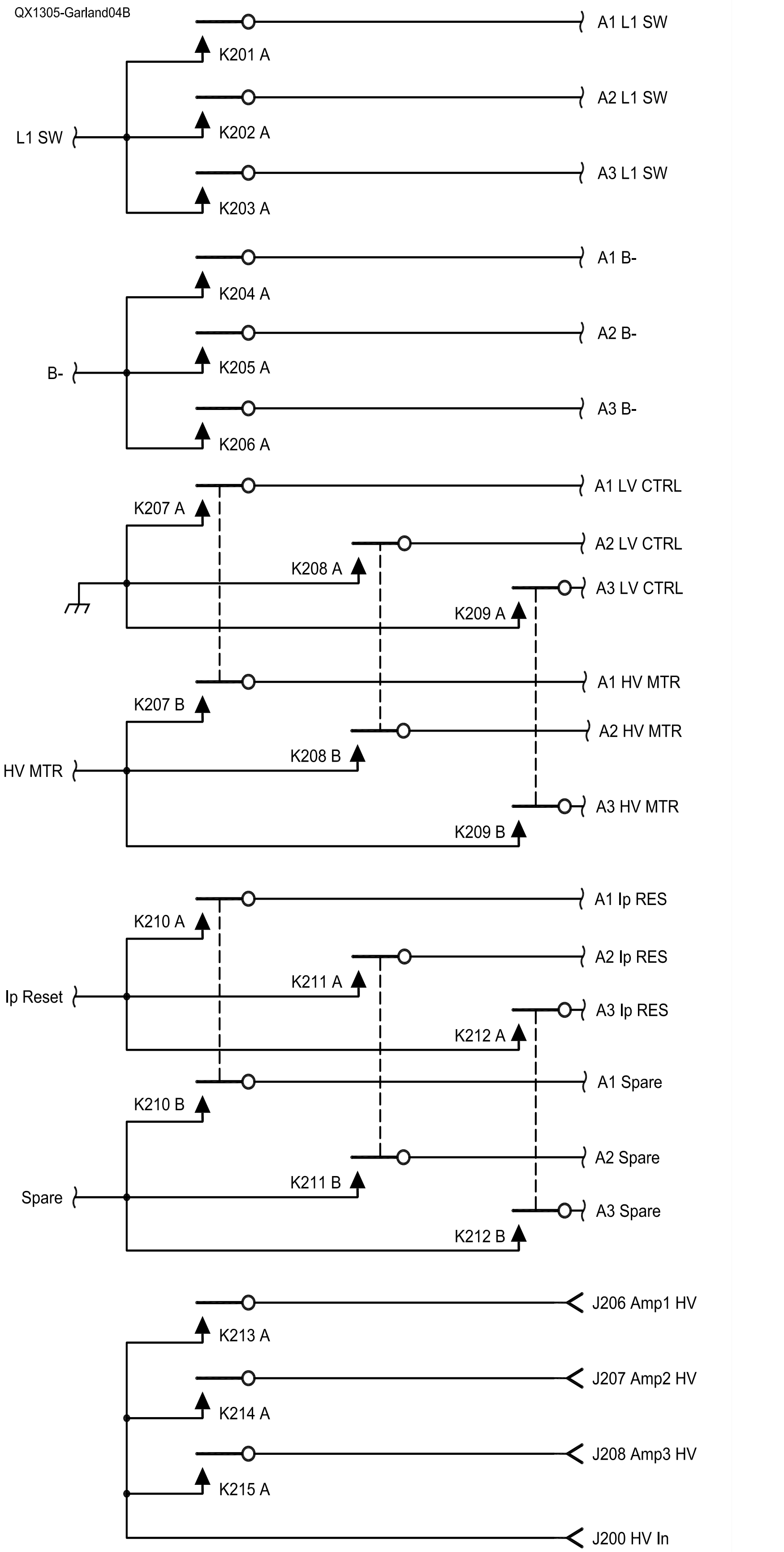


Figure 4B — Relays transfer all voltages to the selected amplifier. The high voltage switching is done using Gigavac G81A vacuum relays, rated at 10 kV.



the large chassis components of the power supply (plate transformer, rectifiers, and filter capacitor) are interfaced via three connectors to the control logic circuits contained on a single 6.0 × 7.5 inch double-sided printed circuit board. Connector pair P100/J200 carries the high voltage from the chassis-mounted components to three HV distribution relays mounted on the controller circuit board. The interconnecting cable uses 10 kV silicone-insulated test probe wire. Connector pair P101/J201 uses a 12 conductor cable and is used for all the control functions, while P102/J202 interfaces to the eight front panel LED indicators. Additional connector pairs P203/J203 through P208/J208 transfer the control lines and high voltage from the printed circuit board to the front panel control and HV connectors.

Figure 3 is a schematic diagram of the main power supply components. (Note that wire colors for the author's power supplies are shown in the diagram to facilitate servicing and circuit tracing.) As shown in the diagram, each side of the 240 V ac line is routed through ganged 25 A circuit breakers CB100/CB101 to solid state relays K100/K101. When the circuit breakers are closed, 12 V ac is applied by T102 to the control board, whose on-board regulator provides 12 V dc to operate the relays and logic circuits. The "HV Enable" key switch S1 disables the ac relays for servicing or testing purposes, but leaves alive all the other control functions. All 120 V ac components used in the power supply (muffin fan, digital panel meter) and RF decks (blowers, filaments, low voltage power supplies) use either L1 or L2 from the 240 V ac line and the N ("neutral") line. Note that it is poor design practice to ground the neutral line to the chassis, since doing so results in unpredictable and potentially dangerous paths for the power line return currents. Modern building codes often mandate a four-wire 240 V ac power cord with an integral ground wire, such as for electric dryers, but if your home has the older three-wire (L1, L2, N) configuration, then a separate station ground wire should be connected to the power supply chassis. A threaded 10-32 ground lug is provided on the front panel below the power cord for this purpose. (Note that outside the US, many countries with 240 V ac service do not use a neutral line. Builders from those countries must either use 240 V ac fans, filament transformers and so on, or else derive a "virtual" neutral from a center tap on the primary winding of the plate transformer.)

K102, R100, R102, C104 and D100 comprise a step-start circuit that limits the surge current at power-up to 10 A until filter capacitor C106 is partially charged. The intrinsic time constant of this circuit is about 0.3 s, but because D100 picks off its voltage at the downstream side of R100 the actual time delay is closer to 0.8 s. The plate transformer T101 is custom designed for the power supplies by the Peter W Dahl Company and is a versatile 67 lb (5 kVA) hypersil-wound transformer with three primary taps and two secondary taps. (Note that the secondary taps are not shown on the schematic diagram, but are selected during construction.)

[The Dahl transformer line has been available through Harbach Electronics (**www.harbachelectronics.com**) but as we prepared this article for publication Harbach announced that they would no longer handle the Dahl transformer line. Then, as we prepared this issue of *QEX* to go to press, there was an announcement that Hammond Manufacturing Company, Inc. of Cheektowaga, New York, was acquiring the line of Dahl transformers. According to the Hammond website they expect the deal to be completed and all assets transferred by March 31, 2013. Check the Hammond website for further updates (**www.hammfg.com**) — *Ed.*]

By mixing and matching taps, the higher voltage transformer can provide six RMS voltages ranging from 2000 to 3300 V ac (1920 V ac to 2250 V ac for the lower voltage transformer), each at 1.5 A CCS. Relay K103 allows each power supply to select two of these voltages. Four diode blocks, each rated at 1.5 A/15 kV comprise a bridge rectifier that rectifies the output from the transformer secondary. The rectified dc is filtered by a large 40 µF/5000 V oil-filled capacitor, C106 (50 µF/4200 V in the lower voltage power supply), which the author had on hand. Bleeder resistor R104 is made up of two 100 kΩ/100 W power resistors in series and dissipates about 100 W. D102 provides flashover protection to the metering circuits by clamping the B– return current to within 1 V of ground in the event of an arc to ground somewhere in the power supply or RF deck, while R109 anchors the B– return to ground in the unlikely event it should become disconnected from its RF deck. R103 is used to sense the power supply current and is connected to an optically isolated over-current trip circuit on the control and logic circuit board.

Control and Logic Circuits

The functions of the HV power supply control and logic circuitry are:

(1) to allow each amplifier to be independently powered on or off. When an amplifier is turned off, all power to it is removed, including all high voltage, low voltage and control circuits.

(2) to interlock each amplifier, so that only one amplifier can be placed on-line at a time. When an amplifier is brought on-line, any previously selected amplifier is taken off-line, but remains in a standby state. High voltage is applied to an amplifier only when it is on-line.

(3) to implement metering and control functions for each amplifier that are independent of one another. From the perspective of the operator, the shared power supply is essentially invisible.

(4) to control flashover surges, in order to prevent damage to the connected amplifiers.

(5) to enable simple hookup of the connected amplifiers. Each amplifier plugs into the power supply with a single control cable and a single HV cable. Any amplifier can be disconnected (unplugged) from the power supply, without affecting the operation of the remaining amplifiers. The ac power and on-line switches of each amplifier are simple momentary action SPST pushbutton switches on the front panel of each amplifier.

(6) to switch automatically primary taps on the HV power transformer, to allow the connected amplifiers to use different plate voltages.

(7) to facilitate easy construction of the HV power supply by mounting all logic, control, and switching circuitry on a single printed circuit board. Thus the construction of the power supply is only moderately more complicated than construction of an ordinary single-amplifier power supply. This means that an amplifier builder can incorporate multiple amplifier capability into a newly built power supply at reasonable effort and cost in order to allow for future needs.

Referring to the circuit diagrams of Figures 4A and 4B, the three RF amplifier decks are actuated by two momentary action pushbutton switches for each amplifier: one controls ac power (blower, filaments, LV supply) and one controls HV and enables the amplifier to be brought on-line. The buttons are debounced by U201, with C200 setting the maximum debounce time (50 ms) before the button states stabilize. R200 to R205 hold each button line high. These resistors are in parallel with 500 kΩ resistors internal to U201 and result in about 2 mA of current through each button when it is pressed. Grounding the button line activates the control circuitry.

The active-low button states are inverted by hex inverters U202, and the three ac power buttons are applied to the clock inputs of D flip-flops U204A, U204B, and U205A. Each flip-flop is configured so that it toggles its output states Q and Q' each time its button is pressed. A positive pulse is generated by R206 and C202 at power-on and is applied to the reset line of the three flip-flops, ensuring that they power up with Q = 0 and Q' = 1. The voltage pulse reaches a maximum of about 10 V about 200 ms after power is applied, ensuring a reset after the remainder of the circuitry has had time to wake up. The high state Q' = 1 of each flip-flop is passed at power-up via OR gates U203A, U203B and U203C to the reset inputs of U205B, U206A and U206B, thus ensuring that the HV logic is also powered up in a Q = 0 state.

The outputs of the six flip-flops have several functions. The Q outputs of U204A, U204B, and U205A are combined by OR-gate U207, whose output actuates the power supply's main power relays. The outputs of all six flip-flops are also applied, via 3-input AND gates U208A, U208B, and U208C, to the 8-port relay driver U209. Each port is grounded when active and can sink a maximum of 500 mA. The purpose of the AND gates is to interlock the power and HV buttons to prevent improper operation. One input of U208A, U208B and U208C is grounded when the over-current relay K200 is tripped and shuts off the HV supply of any

Table 1

Each RF amplifier connects to the power supply using a 14-pin AMP control connector. The high voltage is connected separately via a shielded length of RG6/U coaxial cable, using high voltage SHV-RP BNC connectors.

Pin Number	*Function*	*Description*
1	GND	Chassis ground
2	L1_SW	120 V ac (switched)
3	GND	Chassis ground
4	Ip_RES	Plate over-current reset (switched, ground to reset)
5	LV_CTRL	Low voltage control (switched ground for amp LV circuits)
6	PWR_PB	Amplifier momentary SPST ac power pushbutton
7	NEUTRAL	240 V ac neutral line (unswitched)
8	SPARE	Unused spare (switched)
9	HV_MTR	High voltage meter output, 0-5 V dc (switched)
10	IN/OUT_PB	Amplifier momentary SPST on-line pushbutton
11	STBY_LED	Amplifier standby LED anode, 12 mA (unswitched)
12	GND	Chassis Ground
13	B–	Amplifier B– return (switched)
14	GND	Chassis Ground

on-line amplifier. A second input ensures that the HV for any amplifier cannot be turned on unless the ac power to the amplifier has previously been turned on. The third input routes the selected amplifier to the appropriate HV relay driver port.

The HV button flip-flops U205B, U206A and U206B operate in the same manner as the ac power button flip-flops, except that each HV flip-flop is interlocked to the other two HV flip-flops via the 3-input OR gates U203A, U203B, and U203C. As mentioned previously, one input to each OR gate is used to reset the flip-flops on power-up. A second input resets each HV flip-flop whenever its corresponding ac power relay is turned off. Resetting the flip-flop in this way thus keeps the HV from inadvertently turning on if the ac power relay is subsequently energized after being turned off. In other words, the only way to turn on the HV for a particular amplifier is to actuate its HV button.

The third input of the OR gates turns off the HV of any selected amplifier whenever the HV button for another amplifier is pressed. Thus, only the most recently selected amplifier is ever on-line. This function is implemented by means of a pulse caused by the positive edge of the HV flip-flop's Q output transition, in conjunction with the RC differentiator connected to the Q output. This pulse is used to reset (turn off) any previously selected HV flip-flops. Diodes D205, D206, and D207 protect the input of the OR gates by clamping to ground the negative pulse caused by a negative-going transition of the flip-flop. If desired, the user can replace capacitors C211, C212, and C213 with wire jumpers. Doing so would then disable the automatic switch-off function and require the HV of a selected amplifier to be manually switched off before the HV of any other amplifier could be selected.

The over-current protection circuit monitors the voltage developed across a 2 Ω resistor in series with the B– return of the HV power supply (shown in Figure 2). When this voltage indicates excessive current, the optoisolator turns on Q200, latching K200 in a closed position. R212 sets the current trip threshold, and diodes D201 to D204 protect the peripheral circuitry from the momentary current surge caused by flashover in the HV supply.

At the *QEX* website, **www.arrl.org/qexfiles**, you will find the zip file **w8zr_power_supply.zip** that contains a complete schematic package and a spreadsheet with parts info for all the components on the PCB logic controller.

Mechanical and Assembly Details

Figures 5 and 6 show interior views of the power supply. As shown in Figure 6, the

Figure 5 — An interior subpanel divides the power supply into two compartments, with all the control and metering circuits housed in the smaller front compartment, and the large high voltage components in the rear compartment.

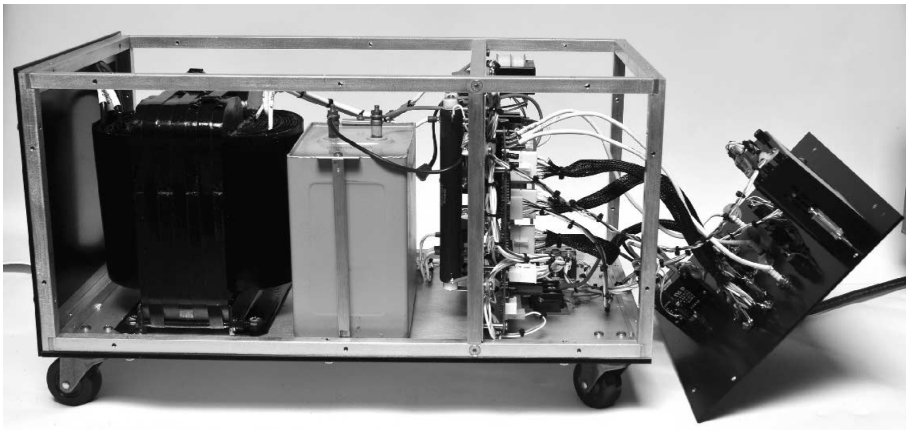


Figure 6 — The front panel detaches and tilts forward to provide access to interior components.

Figure 7 — The rectangular frame is constructed from ½ inch square aluminum stock. A single 10-32 flathead screw secures the three interlocking pieces that form each corner.

Figure 8 — The front side of the subpanel houses the controller printed circuit board, the low voltage transformer, and the step-start circuits.

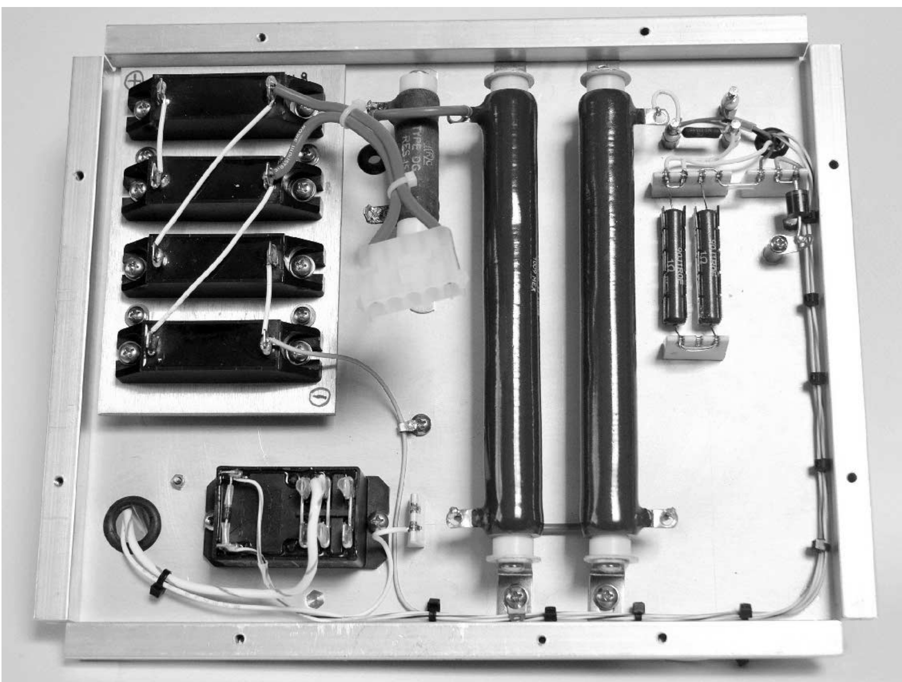


Figure 9 — The bleeder resistors, metering components and HV/LV–select relay mount on the rear of the subpanel. PTFE spacers insulate the bleeder resistors from the chassis, while the HV diode blocks are heat sinked to a $\frac{1}{8}$ inch aluminum plate.

front panel removes and tilts out of the way to enable easy access to components, should servicing ever be required. The power supply enclosure measures 12 W × 10 H × 21.5 D inches (the second power supply is only 20 inches deep) and is fabricated around a frame made from ½ inch square aluminum stock. Figure 7 shows the frame detail at the corners. The bottom plate is made from $\frac{3}{16}$ inch thick aluminum plate, while the front, rear, and top panels are fabricated from $\frac{1}{8}$ inch thick aluminum plate. The side panels are $\frac{1}{16}$ inch thick aluminum. An aluminum subpanel (Figures 8 and 9) divides the power supply into two compartments. The front compartment houses the control logic and switching circuitry, with the printed circuit board (Figure 10), step-start components, and 12.6 V ac low voltage transformer mounted on the front side of the subpanel.

The printed circuit board was designed using *Circad 98*, a commercial schematic capture and PCB layout package (**www.holophase.com**) that the author has used for many projects. "Gerber" files for the completed layout were then uploaded to Advanced Circuits (**www.4pcb.com**), which manufactures high quality printed circuit boards in small quantities at very reasonable cost. Figure 11 shows a breadboard lash up used to debug the logic circuitry before committing the design to a printed circuit board. The small plastic enclosure with the LEDs and momentary-action lever switches simulate three remote RF decks. Interested readers can view a short video demonstration of the breadboard logic circuitry at **www.youtube.com/watch?v=OycSQu55oFo**.

All point-to-point wiring in the power supply uses silicone insulated high voltage wire or color-coded PTFE (Teflon) insulated wire. PTFE is a very durable insulator and has excellent heat resistance and dielectric strength. The wire is costly, but can frequently be found at bargain-basement prices at hamfests and on-line auction and swap sites.

The HV diode blocks are heat sinked to a 4.5 × 5.75 × 0.125 inch aluminum plate, which is mounted on 0.5 inch metal standoffs on the rear side of the subpanel. The rear subpanel also holds the bleeder resistors, HV/LV-select relay, and miscellaneous other components, some of which are mounted on silver/ceramic terminal strips scavenged from old Tektronix oscilloscopes. "Pem" type threaded fasteners are used instead of nuts in order to facilitate component removal. All other hardware is stainless steel, using pan-head Phillips screws.

A 4.75 inch "whisper" muffin fan mounted on the right side of the enclosure silently exhausts warm air drawn through a ventilation cutout on the opposite side. The

High Voltage Safety Considerations

We are all so besieged these days with verbose safety warnings on mostly harmless consumer goods that it is easy to forget that some things really are dangerous. High voltage power supplies definitely fall into this category, especially since many amateurs are accustomed to solid state circuits and seldom encounter any dc voltage higher than 12 V. *This power supply produces voltages that are highly lethal.* So *please* take to heart the following ten precautions. Furthermore, don't expect to learn from your mistakes, because if you don't exercise proper precautions the first time, you're unlikely ever to have a second chance.

1. Don't let your reach exceed your grasp. This is not a project for beginners. You should not attempt to build this power supply unless you're a seasoned builder who has experience with high voltage circuitry.

2. Young amateurs should not attempt this project. Working with high voltages requires the maturity and patience that comes with age and experience.

3. Never work around high voltage when you are tired, stressed, or in a hurry.

4. Never work around high voltage after drinking alcohol. Even one beer or glass of wine can impair your judgment and make you careless.

5. Before working on a high voltage power supply, always follow these three steps: *Unplug* (the ac power cord), *discharge* (the filter capacitors) and *verify* (that the output voltage is truly zero). Time-honored practice is to use a "chicken stick" (a wooden dowel or PVC tube, with one end attached to a grounded wire) to make sure filter capacitors are completely discharged.

6. When working on a high voltage power supply, remember that a dangerous time is after the power supply has just been turned off, but before the filter capacitors have fully discharged. A 50 μF capacitor charged to 4000 V holds a potentially deadly 400 Joules of energy. Even with bleeder resistors, it can take a minute or more to discharge fully.

7. When removing a recently discharged filter capacitor from a power supply, tie the two terminals together with wire. Large high voltage capacitors can self-charge to dangerous levels if the terminals are left floating.

8. Don't stake your life on the expectation that bleeder resistors, fuses, circuit breakers, relays, and switches are always going to do their job. Even though modern components are very reliable, it is safe practice always to assume the worst.

9. Don't build this power supply if you don't understand how the circuit works. High power amplifiers and power supplies are not "plug-and-play" projects with step-by-step instructions. Builders must be knowledgeable enough to improvise, make component substitutions, and implement design changes.

10. With high voltage projects, it doesn't pay to be "penny wise and pound foolish." Use high quality components throughout and save your forty-year-old junk box parts for projects where safety and reliability are not paramount requirements.

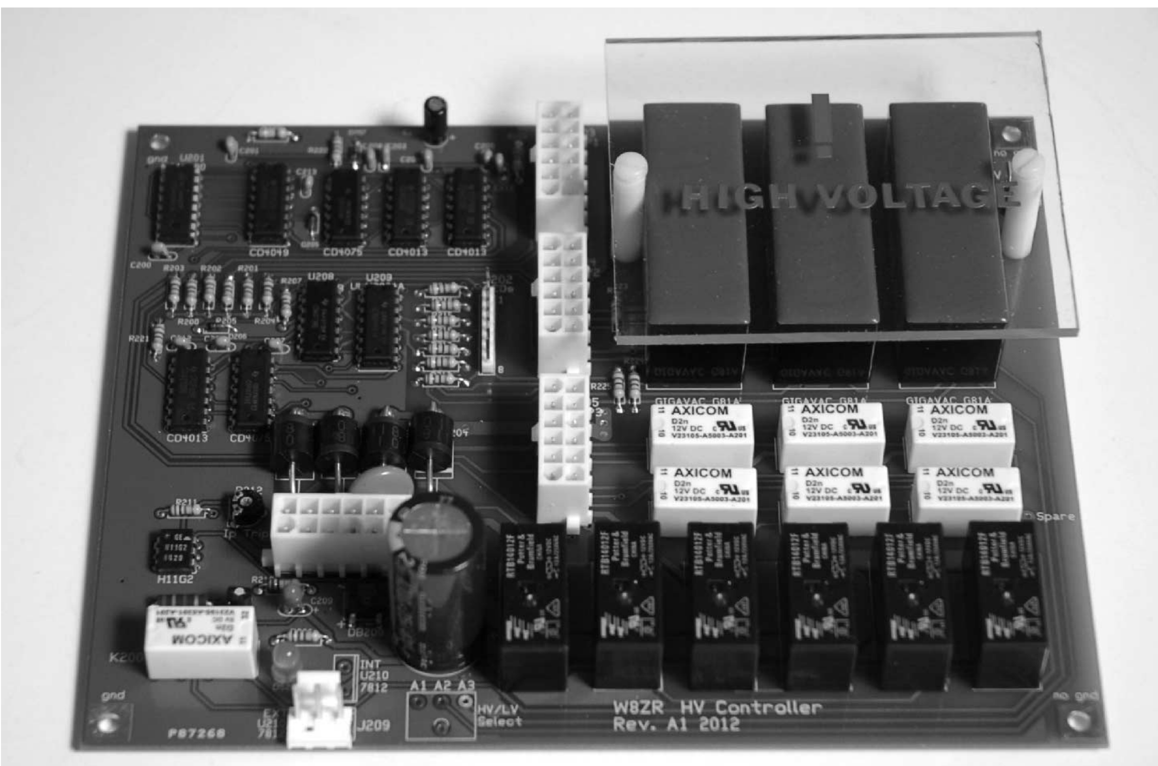


Figure 10 — A double-sided printed circuit board houses the logic and control functions, the over-current protection circuit, and the control relays. A plastic shield covers the high voltage relays to keep nearby wires and cables at a safe distance.

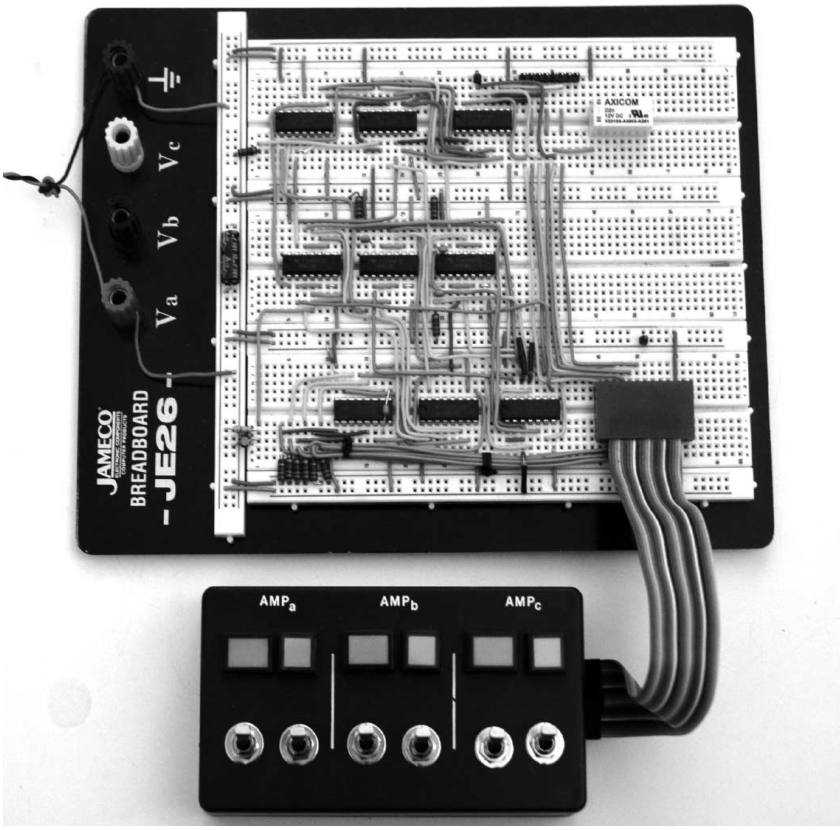


Figure 11 — This breadboard mockup of the logic circuitry preceded the design and layout of the printed circuit board.

large oil-filled capacitor sits on a rubber pad and is secured to the base plate by clamps fabricated from 3/8 inch square aluminum stock.

The enclosure panels are powder coated with a smooth black satin finish. The front panel is similarly finished, and was custom made by Front Panel Express (**www.frontpanelexpress.com**) from a CAD file supplied by the author. The panel lettering and other markings are engraved and backfilled with red and yellow paint (white paint for the lower voltage supply). Each power supply sits on two inch casters and weighs about 90 lbs.

The most tedious part of construction was fabricating the frame for the aluminum enclosure. In order for the frame to be square, tolerances for the individual pieces had to be maintained to within 0.015 inches. After the frame was completed, 60 precisely spaced holes had to be drilled and tapped into it for attaching the six panels. Obviously, other builders will likely have more sense than the author, and will spare themselves this ordeal by building the power supply into a commercial enclosure!

There are many reasons why amateurs enjoy building their own equipment. Saving money, experimenting with new circuits, learning new skills, and experiencing the satisfaction that comes from creating something innovative and useful have always motivated Amateur Radio homebrewers. For some, including the author, there is also a strong esthetic pleasure that comes from designing and building a unique piece of equipment that cuts no corners, and cannot be purchased commercially. All builders, no matter how skilled or experienced, quickly learn that there is no design that cannot be improved upon and no level of workmanship that cannot be executed more carefully. Because perfection always remains out of reach, every new project thus represents an irresistible challenge to improve one's skills and advance the state of the art. That spirit of innovation has infused Amateur Radio since its earliest days, more than a century ago, and is still alive and well today.

Jim Garland, W8ZR, holds an Amateur Extra Class license and is a former Ohio State University physics professor and president of Miami University (Ohio). He is a Life Member of the ARRL, a member of the ARRL Diamond Club and Maxim Society, and currently lives in Santa Fe, NM. His Amateur Radio website is **www.w8zr.net**.

QEX

Idea Exchange

Technical Tidbits for the QRPer

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The following item is an excerpt from the "Idea Exchange" column in the Spring 2014 issue of QRP Quarterly. *The column is edited by WA8MCQ.*

Simple Adjustable Tracking Power Supply

From Bryant Julstrom, KCØZNG—

The usual experimenter's bench power supply provides an adjustable positive voltage of up to 25V at up to 1A of current. This suffices for many circuits and projects, but others require positive and negative voltages of equal magnitude. Op-amps, for example, often require ±15V.

Fixed matched voltages are easy to provide with a pair of three-terminal regulators, one positive and one negative. A more flexible solution imitates the adjustable positive supply: a tracking power supply provides an adjustable positive voltage and an equal but negative voltage; the negative voltage tracks the positive one.

A simple way to obtain adjustable symmetric voltages extends the two-regulator fixed-voltage solution with adjustable regulators and a two-section pot, which adjusts the positive and negative voltages simultaneously. This is simple, but tracking accuracy depends on matching the resistances associated with the two regulators, including the pot's two sections. A related circuit uses an op-amp to invert the control voltage of a positive regulator to control a negative one or a pass transistor. This is elegant, but it limits the supply's voltage to slightly less than the op-amp's maximum input voltage, usually 18V. Manufacturers have produced a number of integrated circuits designed to control tracking power supplies. W1KLK described a supply that used the SG1301 back in 1973 (see Reference). A similar chip is the RC4194. These and others like them have become unobtainium in recent years.

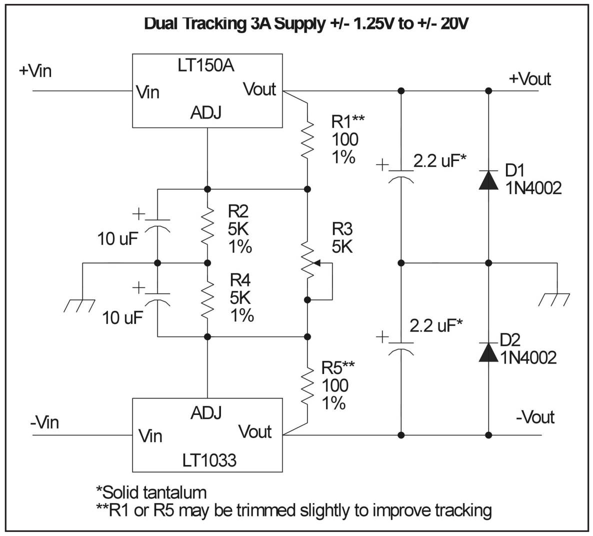

Figure 13—From the datasheet of the LT1033, a tracking regulator circuit that uses positive and negative regulators.

The Circuit

The datasheet for the LT1033, a negative three-terminal regulator from Linear Technology, presents another option: a tracking regulator circuit in which a single pot sets the output voltages of a positive and a negative regulator. Figure 13, reproduced from the datasheet, shows the circuit. It isn't necessary to use the regulators shown; I implemented the circuit using two old standbys, an LM317 positive regulator and an LM337 negative regulator, both in TO-220 packages. Note that the input voltages of the circuit are limited only by the maximum inputs of the two regulators, now ±40V.

Input to the tracking regulator is a pair of voltages and a common line between them. This requires a power transformer with a center-tapped secondary. The beefiest transformer in my junk box provided only 26 VCT, and input to the regulator of ±13V was too low. One solution uses two identical transformers with their primary windings in parallel and their secondaries in series.

I chose to use the one transformer and back-to-back half-wave voltage doublers, one on either side of the secondary winding's center tap, to provide about ±26V to the tracking regulator. This halves the current that can be drawn from the transformer, but it is rated at 2A, and 1A in each leg of the supply is sufficient for a wide range of projects. A voltmeter can be switched across either of the supply's outputs.

Figure 14 shows the circuit of the entire tracking power supply. I used 1N5402s in the voltage doublers because I had a bunch of them. As in Figure 13, the two 2.2 µF capacitors should be solid tantalum. The regulator departs from the circuit in the datasheet in two places. First, to adjust tracking, one of the two 100 ohm resistors is replaced with a 200 ohm trimpot, following a suggestion in the datasheet. Second, a 330 ohm resistor sets the regulator's minimum outputs at about ±3.6V, for reasons that will be clear in a

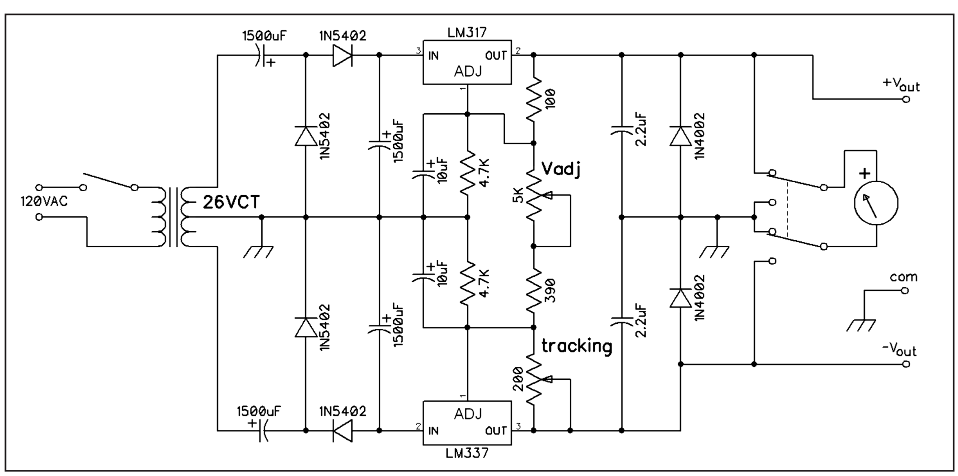

Figure 14—A tracking power supply using familiar parts, based on the regulator circuit in Figure 13.

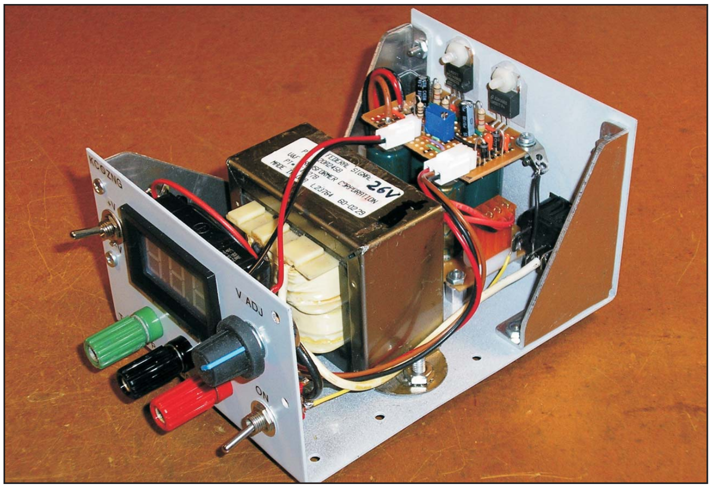

Figure 15—Interior view of the tracking power supply.

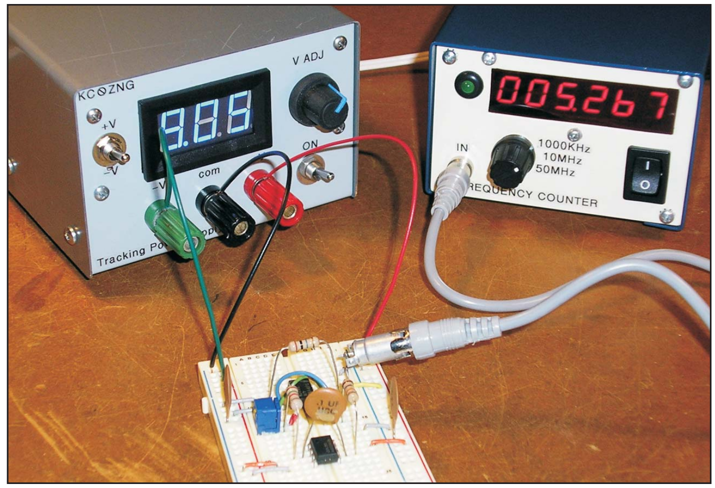

Figure 16—The tracking power supply in use.

moment. The schematic was developed using the freeware version of DipTrace™ and edited with Microsoft Paint.

Space on my bench is limited, so I wanted the supply to be no larger than necessary. Its front panel holds three binding posts, two miniature toggle switches, the voltage-adjustment pot and its knob, and a voltmeter. The meter is a three-digit LED unit from Marlin P. Jones (www.mpja.com; 30217 ME with blue digits, also available in green and red). It is powered by the voltage it measures, as long as that voltage is at least 3.6V (thus the 330 ohm resistor mentioned above).

Inside the enclosure are the transformer and two circuit boards: the two voltage doublers occupy one and the tracking regulator the other. The LM317 and LM337 are mounted at the edge of the regulator board and attached to the back panel, which then serves as a heat sink, with mica insulators, nylon bolts, and a smear of heat sink compound,. The panel also holds a snap-in IEC three-wire line connector. All the parts came from my junk box except the digital meter and the four 1500μF electrolytic capacitors in the voltage doublers.

The enclosure was bent up from two pieces of sheet aluminum, with brackets at each corner of the bottom half to stiffen the front and back panels and provide attachments for the top, which is held on with sheet metal screws. The enclosure is painted in two shades of gray, and labeled with black-on-clear tape from a Brother™ label-maker. The unit rests on four press-on rubber feet.

Figure 15 shows the unit with the top off and before the last bracket was installed. The regulator board is visible on the back panel. The doubler board is below it, mostly hidden behind the transformer. It's tight in there; cutouts on the brackets' tabs accommodate the line connector, the transformer's mounting tabs, and the switches and pot on the front panel.

Adjustment and Performance

The only adjustment in the supply is the tracking trimpot, which must be set so that the two voltages track each other accurately. This is touchy, but the two voltages can be made to match within 0.1V throughout the supply's range. The completed supply provides closely matched positive and negative voltages from ±3.6V to about ±23V, and a maximum current of 1A in each leg. Figure 16 shows the unit in use.

The tracking adjustment and the front-panel voltage adjustment are delicate. In both cases, multi-turn pots would be much easier to set. The tracking trimpot could be replaced with a series resistor and a smaller-value pot. An analog meter or a digital meter with a separate power supply would allow smaller output voltages. One could augment the supply's measurements with current metering or with simultaneous measurements in both legs, though more meters will require a larger enclosure. A higher-voltage transformer or, as mentioned, two transformers would allow output voltages of greater magnitude. However, as constructed, I've found this supply a handy addition to my bench.

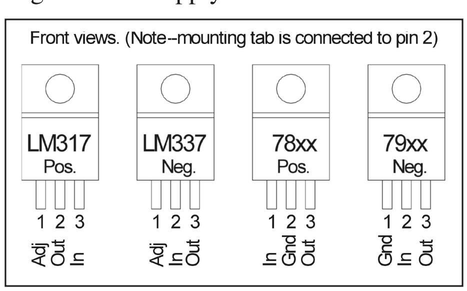

Figure 17—Pin connections of various regulators.

[WA8MCQ Note—I added Figure 17 which shows the connections of some commonly used 3 terminal regulators. I got myself in trouble a time or two over the years when I forgot that the connections of positive and negative regulators are different.]

Reference:

W1KLK: "A Dual-Polarity IC Regulator," Technical Topics, *QST*, February, 1973, p.49.

—de KCØZNG

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Keywords: DC to DC, Buck, Boost, Flyback, Inverter, PWM, Quick-PWM, Voltage Mode, Current Mode Skip, Synchronous Rectifier

Oct 19, 2000

APPLICATION NOTE 2031

DC-DC Converter Tutorial

Abstract: Switching power supplies offer higher efficiency than traditional linear power supplies. They can step-up, step-down, and invert. Some designs can isolate output voltage from the input. This article outlines the different types of switching regulators used in DC-DC conversion. It also reviews and compares the various control techniques for these converters.

Introduction

The power switch was the key to practical switching regulators. Prior to the invention of the Vertical Metal Oxide Semiconductor (VMOS) power switch, switching supplies were generally not practical.

The inductor's main function is to limit the current slew rate through the power switch. This action limits the otherwise high-peak current that would be limited by the switch resistance alone. The key advantage for using an inductor in switching regulators is that an inductor stores energy. This energy can be expressed in Joules as a function of the current by:

$$E = \frac{1}{2} * L * I^2$$

A linear regulator uses a resistive voltage drop to regulate the voltage, losing power (voltage drop times the current) in the form of heat. A switching regulator's inductor does have a voltage drop and an associated current but the current is 90 degrees out of phase with the voltage. Because of this, the energy is stored and can be recovered in the discharge phase of the switching cycle. This results in a much higher efficiency and much less heat.

What Is a Switching Regulator?

A switching regulator is a circuit that uses a power switch, an inductor, and a diode to transfer energy from input to output.

The basic components of the switching circuit can be rearranged to form a step-down (buck), step-up (boost), or an inverter (flyback). These designs are shown in **Figures 1**, **2**, **3**, and **4** respectively, where Figures 3 and 4 are the same except for the transformer and the diode polarity. Feedback and control circuitry can be carefully nested around these circuits to regulate the energy transfer and maintain a constant output within normal operating conditions.

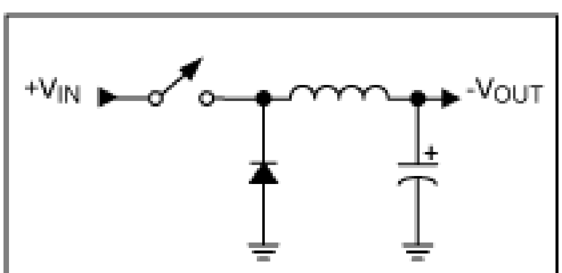

Figure 1. Buck converter topology.

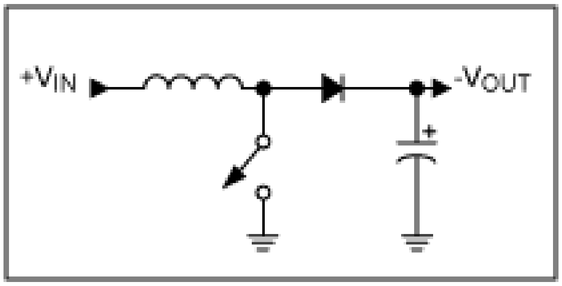

Figure 2. Simple boost converter.

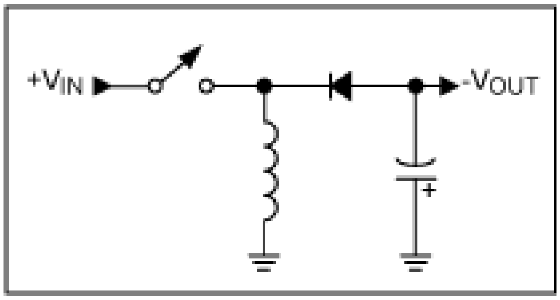

Figure 3. Inverting topology.

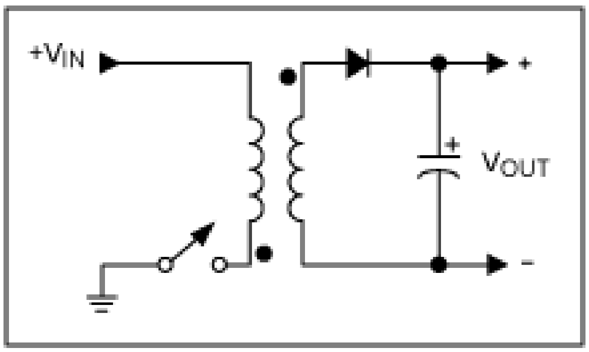

Figure 4. Transformer flyback topology.

Why Use a Switching Regulator?

Switching regulators offer three main advantages compared to a linear regulators. First, switching efficiency can be much better than linear. Second, because less energy is lost in the transfer, smaller components and less thermal management are required. Third, the energy stored by an inductor in a switching regulator can be transformed to output voltages that can be greater than the input (boost), negative (inverter), or can even be transferred through a transformer to provide electrical isolation with respect to the input (Figure 4).

Given the advantages of switching regulators, one might wonder where can linear regulators be used? Linear regulators provide lower noise and higher bandwidth; their simplicity can sometimes offer a less expensive solution.

There are, admittedly, disadvantages with switching regulators. They can be noisy and require energy management in the form of a control loop. Fortunately the solution to these control problems is found integrated in modern switching-mode controller chips.

Charge Phase

A basic boost configuration is depicted in **Figure 5**. Assuming that the switch has been open for a long time and that the voltage drop across the diode is negative, the voltage across the capacitor is equal to the input voltage. When the switch closes, the input voltage, $+V_{IN}$, is impressed across the inductor and the diode prevents the capacitor from discharging $+V_{OUT}$ to ground. Because the input voltage is DC, current through the inductor rises linearly with time at a rate proportional to the input voltage divided by the inductance.

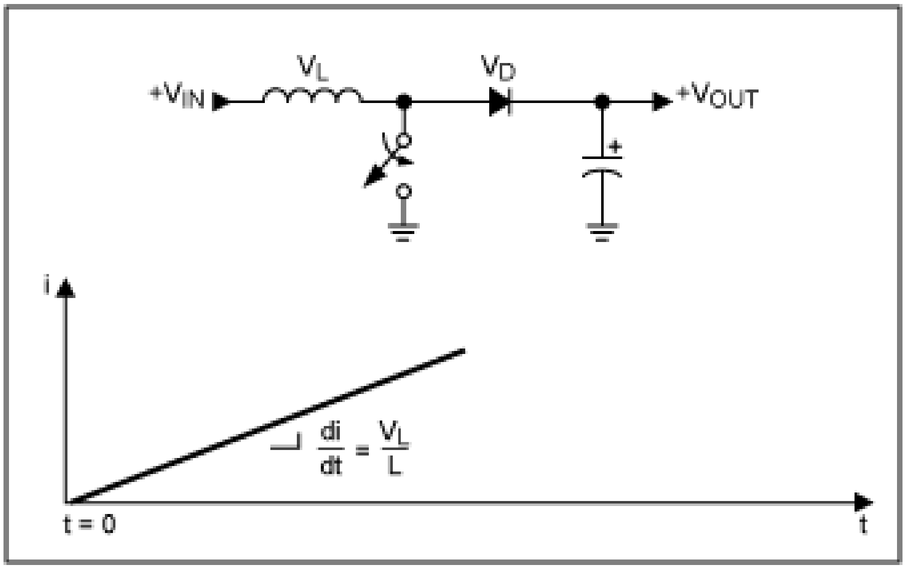

Figure 5. Charging phase: when the switch closes, current ramps up through the inductor.

Discharge Phase

Figure 6 shows the discharge phase. When the switch opens again, the inductor current continues to flow into the rectification diode to charge the output. As the output voltage rises, the slope of the current, di/dt, though the inductor reverses. The output voltage rises until equilibrium is reached or:

$V_L = L \times di/dt$

In other words, the higher the inductor voltage, the faster inductor current drops.

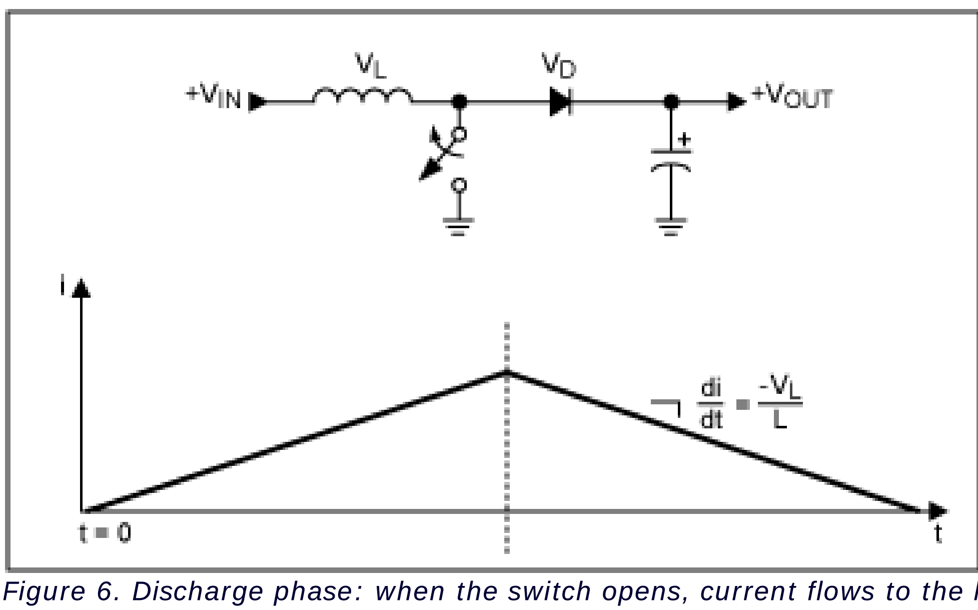

Figure 6. Discharge phase: when the switch opens, current flows to the load through the rectifying diode.

In a steady-state operating condition the average voltage across the inductor over the entire switching cycle is zero. This implies that the average current through the inductor is also in steady state. This is an important rule governing all inductor-based switching topologies. Taking this one step further, we can establish that for a given charge time, t_{ON}, and a given input voltage and with the circuit in equilibrium, there is a specific discharge time, t_{OFF}, for an output voltage. Because the average inductor voltage in steady state must equal zero, we can calculate for the boost circuit:

$V_{IN} \times t_{ON} = t_{OFF} \times V_L$

and because:

$V_{OUT} = V_{IN} + V_L$

We can then establish the relationship:

$V_{OUT} = V_{IN} \times (1 + t_{ON}/t_{OFF})$

using the relationship for duty cycle (D):

$t_{ON}/(t_{ON} + t_{OFF}) = D$

Then for the boost circuit:

$\mathbf{V_{OUT} = V_{IN} / (1-D)}$

Similar derivations can be had for the buck circuit:

$\mathbf{V_{OUT} = V_{IN} \times D}$

and for the inverter circuit (flyback):

$\mathbf{V_{OUT} = V_{IN} \times D / (1-D)}$

Control Techniques

From the derivations for the boost, buck, and inverter (flyback), it can be seen that changing the duty cycle controls the steady-state output with respect to the input voltage. This is a key concept governing all inductor-based switching circuits.

The most common control method, shown in **Figure 7**, is pulse-width modulation (PWM). This method takes a sample of the output voltage and subtracts this from a reference voltage to establish a small error signal (V_{ERROR}). This error signal is compared to an oscillator ramp signal. The comparator outputs a digital output (PWM) that operates the power switch. When the circuit output voltage changes, V_{ERROR} also changes and thus causes the comparator threshold to change. Consequently, the output pulse width (PWM) also changes. This duty cycle change then moves the output voltage to reduce to error signal to zero, thus completing the control loop.

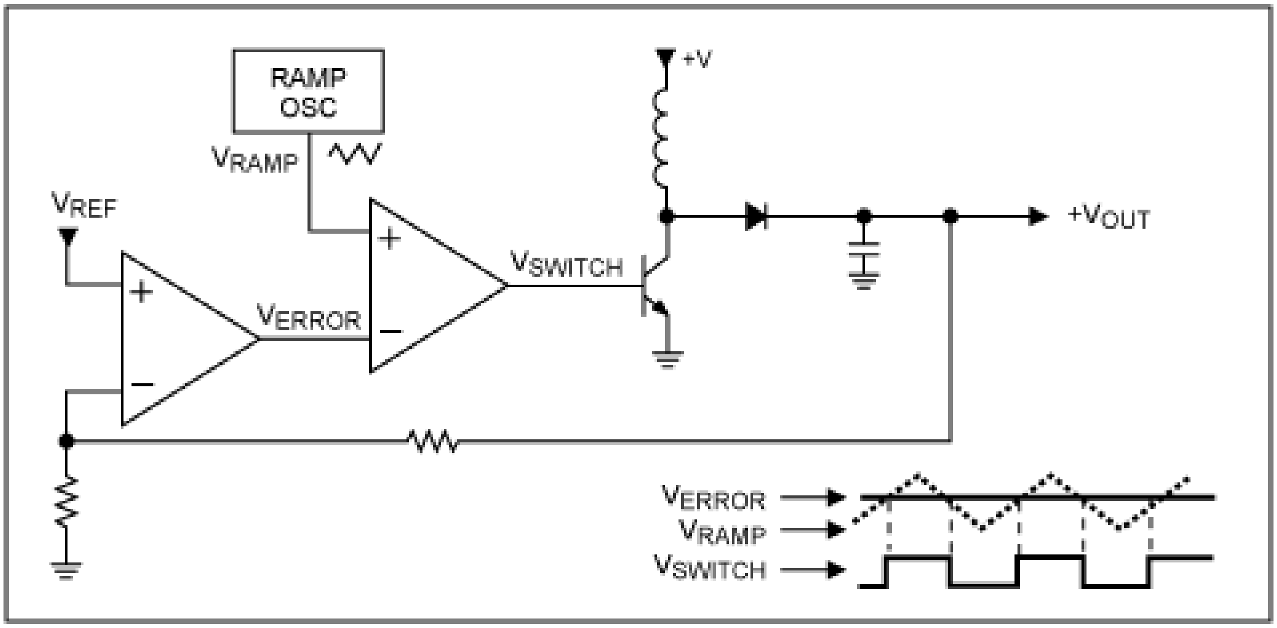

Figure 7. Varying error signal generates a pulse-width-modulated switch signal.

Figure 8 shows a practical circuit using the boost topology formed with the MAX1932. This IC is an integrated controller with an onboard programmable digital-to-analog converter (DAC). The DAC sets the output voltage digitally through a serial link. R5 and R8 form a divider that meters the output voltage. R6 is effectively out of circuit when the DAC voltage is the same as the reference voltage (1.25V). This is because there is zero volts across R6 and so zero current. When the DAC output is zero (ground), R6 is effectively in parallel with R8. These two conditions correspond to the minimum and maximum output adjustment range of 40V and 90V, respectively.

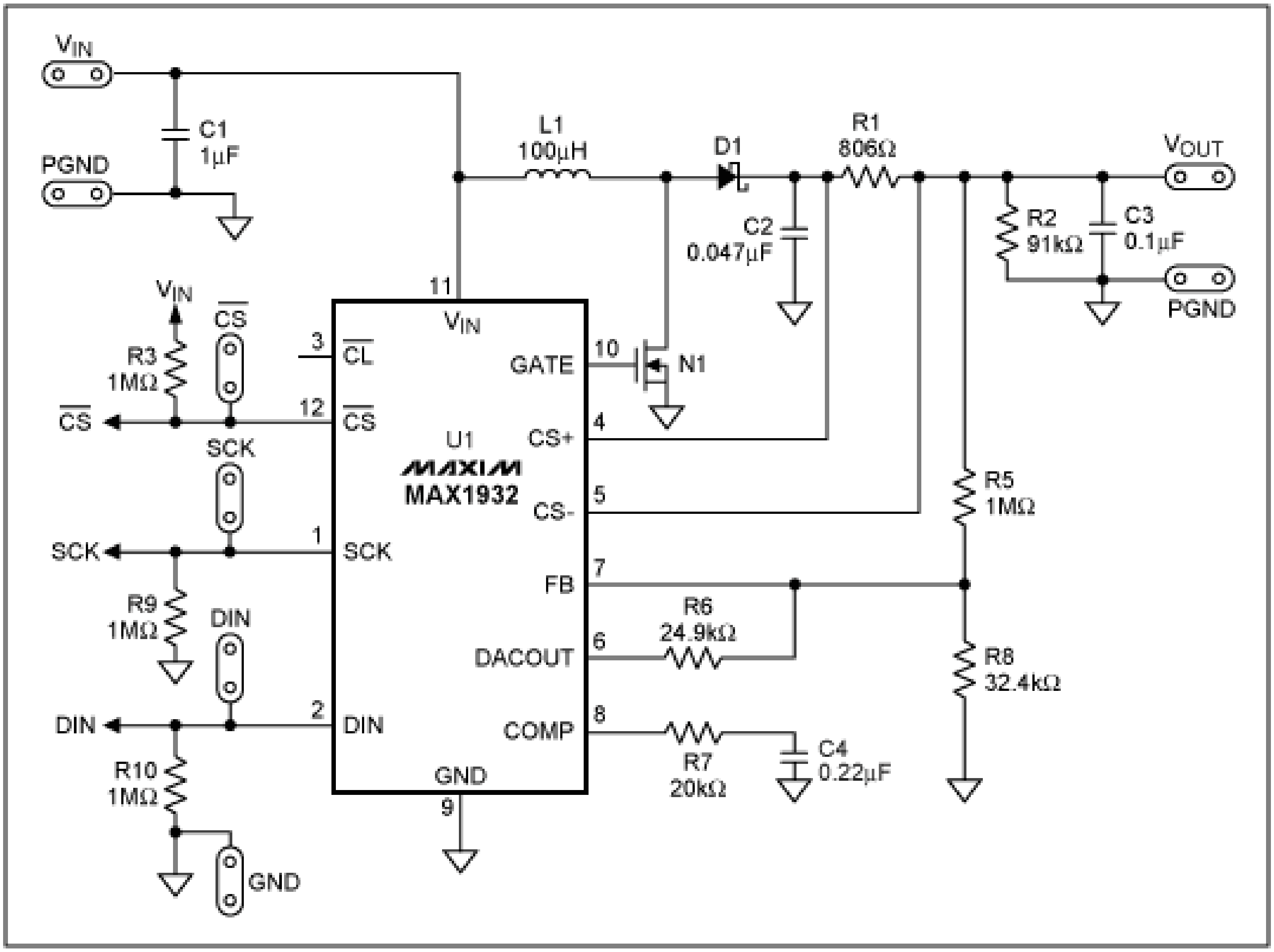

Figure 8. The MAX1932 provides an integrated boost circuit with voltage-mode control.

Next, the divider signal is subtracted from the internal 1.25V reference and then amplified. This error signal is then output on pin 8 as a current source. This, in conjunction with the differential input pair, forms a transconductance amplifier. This arrangement is used because the output at the error amp is high impedance (current source), allowing the circuit's gain to be adjusted by changing R7 and C4. This arrangement also provides the ability to trim the loop gain for acceptable stability margins. The error signal on pin 8 is then forwarded to the comparator and output to drive the power switch. R1 is a current-sense resistor that meters the output current. When the current is unacceptably high, the PWM circuit shuts down, thereby protecting the circuit.

The type of switching (topology) in Figures 7 and 8 is classified as a voltage-mode controller (VMC) because the feedback regulates the output voltage. For analysis we can assume that if the loop gain is infinite, the output impedance for an ideal voltage source is zero. Another commonly used type of control is current-mode control (CMC). This method regulates the output current and, with infinite loop gain, the output is a high-impedance source. In CMC, the current loop is nested with a slower voltage loop, as shown in **Figure 9**; a ramp is generated by the slope of the inductor current and compared with the error signal. So, when the output voltage sags, the CMC supplies more current to the load. The advantage of CMC is its ability to manage the inductor current. In VMC the inductor current is not metered. This becomes a problem because the inductor, in conjunction with the output filter capacitor, forms a resonant tank that can ring and even cause oscillations. Current mode control senses the inductor current to correct for inconsistencies. Although difficult to accomplish, carefully selected compensation components can effectively cancel out this resonance in VCM.

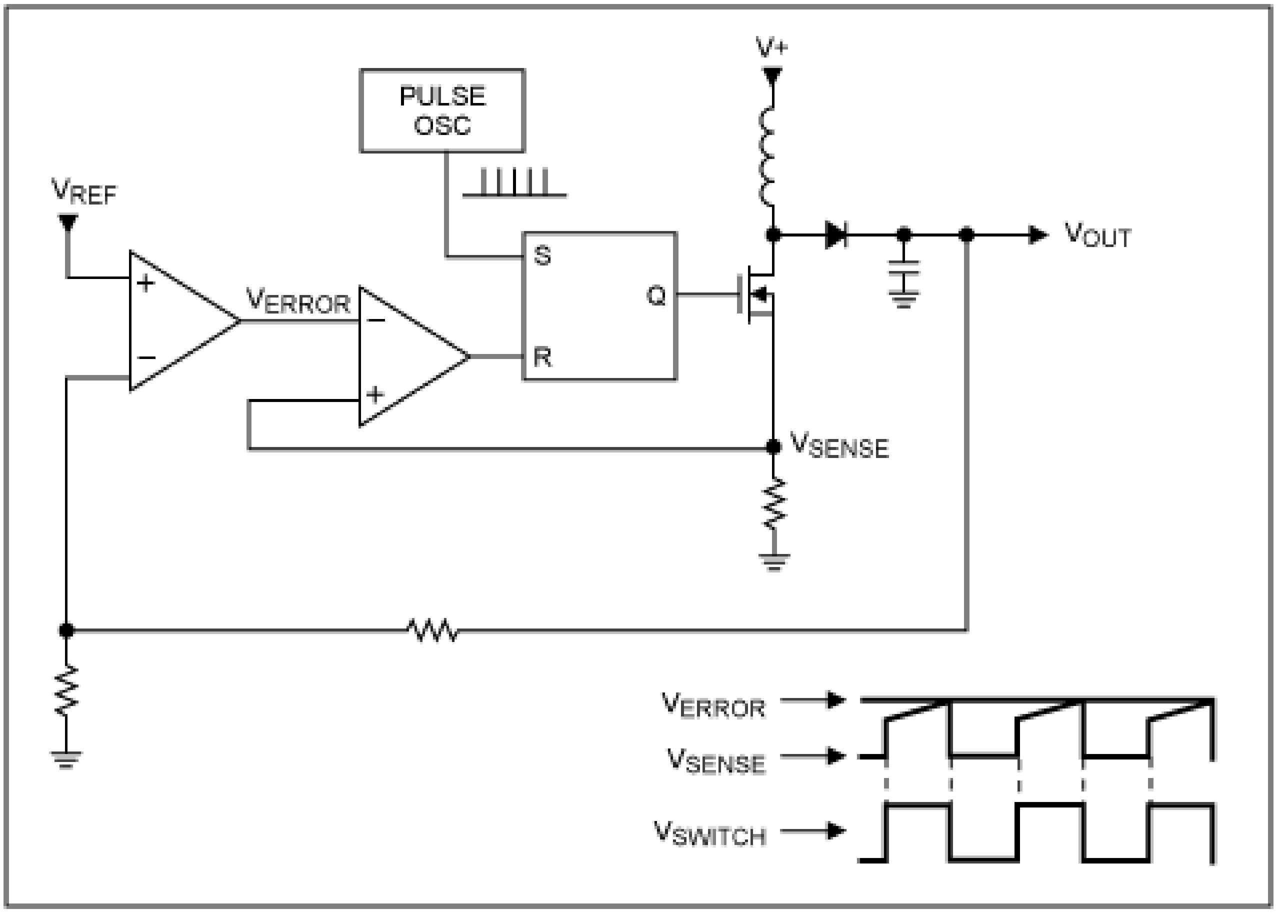

Figure 9. Current-mode pulse-width modulation.

The circuit in **Figure 10** uses CMC with the MAX668 controller. This boost circuit is similar to Figures 7 and 8 except that R1 senses the inductor current for CMC. R1 and some internal comparators provide a current limit. R5 in conjunction with C9 filters the switching noise on the sense resistor to prevent false triggering of the current limit. The MAX668's internal current-limit threshold is fixed; changing the resistor, R1, adjusts the current-limit setting. The resistor, R2, sets the operating frequency. The MAX668 is a versatile integrated circuit that can provide a wide range of DC-DC conversions.

The external components of the MAX668 can have high-voltage ratings that provide greater flexibility for high-power applications. For portable applications that require less power, the MAX1760 and MAX8627 are recommended. These latter devices use internal FETs, and sense the current by using the FETs' resistance to measure inductor current (no sense resistor required).

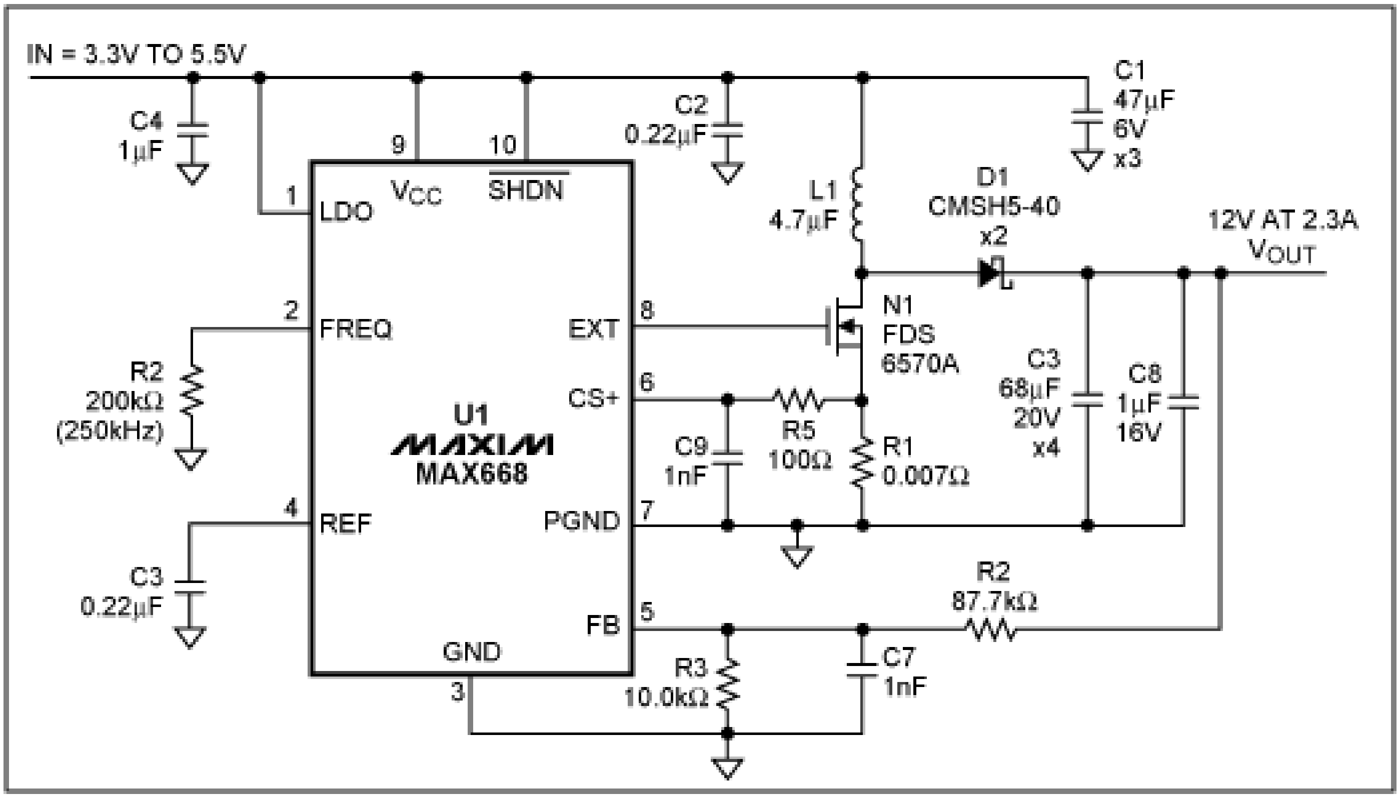

Figure 10.MAX668 for current-mode-controlled boost circuit.

Figure 11 shows a simplified version of Maxim's Quick-PWM™ architecture. To analyze this buck circuit, we start with the feedback signal below the regulating threshold defined by the reference. If there are no forward current faults, then the t_{ON} one-shot timer that calculates the on-time for DH is turned on immediately along with DH. This t_{ON} calculation is based on the output voltage divided by the input, which approximates the on-time required to maintain a fixed switching frequency defined by the constant K. Once the t_{ON} one-shot timer has expired, DH is turned off and DL is turned on. Then if the voltage is still below the regulating threshold, the DH immediately turns back on. This allows the inductor current to rapidly ramp up to meet the load requirements. Once equilibrium with the load has been met, the average inductor voltage must be zero. Therefore we calculate:

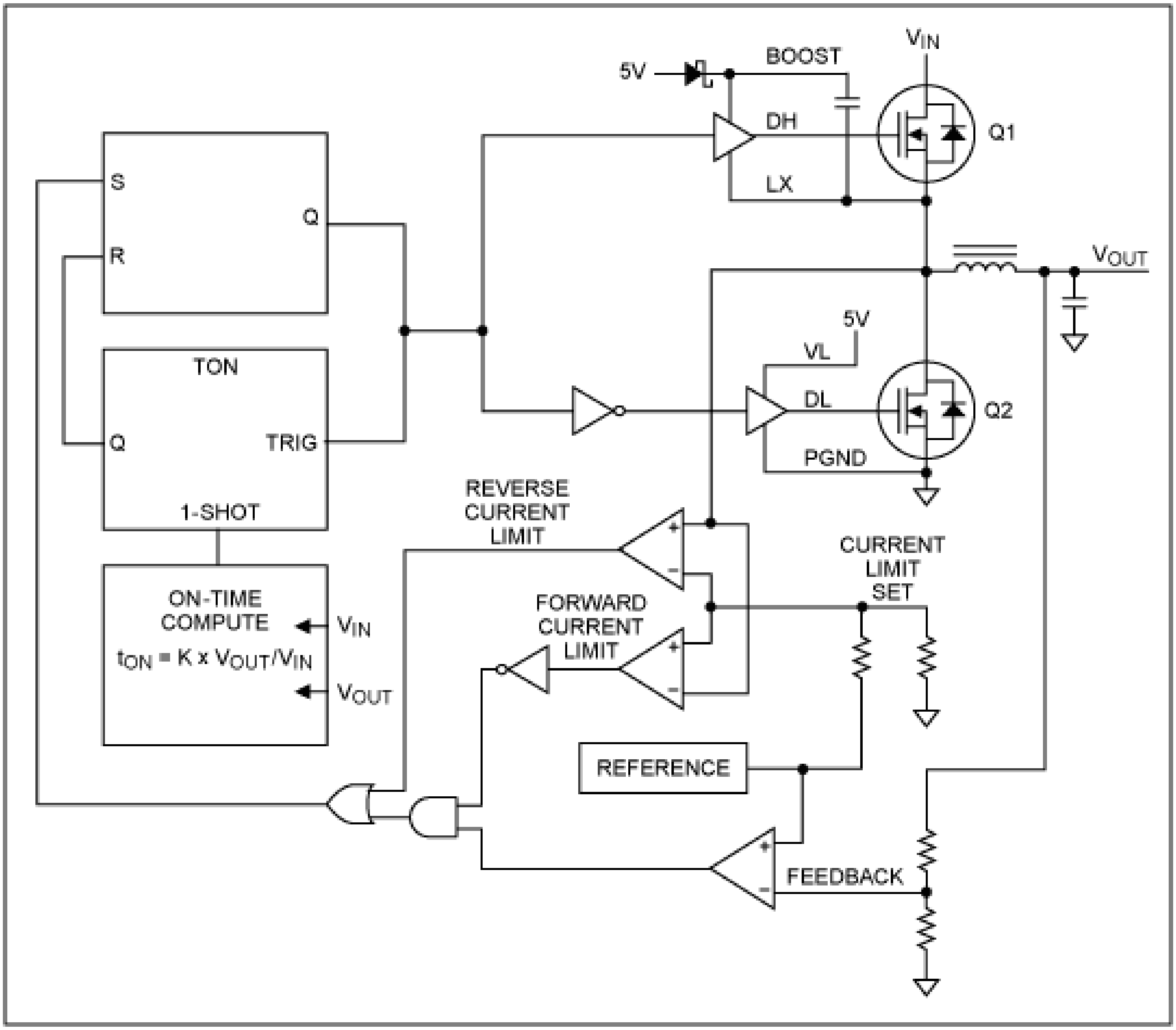

Figure 11. Simplified block diagram of Maxim's Quick-PWM control.

$t_{ON} \times (V_{IN} - V_{OUT}) = t_{OFF} \times V_{OUT}$

Rearranging:

$V_{OUT}/(V_{IN}-V_{OUT}) = t_{ON}/t_{OFF}$

Adding 1 to both side and collecting terms:

$V_{OUT}/V_{IN} = t_{ON}/(t_{ON} + t_{OFF})$

Because the duty factor is D:

$t_{ON}/(t_{ON} + t_{OFF}) = D$

For the buck circuit:

$\mathbf{D = V_{OUT} / V_{IN}}$

Maxim's proprietary Quick-PWM control method offers some advantages over PWM. Quick-PWM control generates a new cycle when the output voltage falls below the regulation threshold. Consequently, heavy transients force the output to fall, immediately firing a new on-cycle. This action results in a 100ns load-step response. It is also important to note that unlike the buck circuit in Figure 1, Figure 11 uses a MOSFET (Q2) instead of a diode for the discharge path. This design reduces the losses associated with the diode drop; the on-resistance of the MOSFET channel doubles as a current sense. Because output-voltage ripple is required to stimulate the circuit to switch, an output filter capacitor with some ESR is required to maintain stability. The Quick-PWM architecture can also respond quickly to line input changes by directly feeding the input voltage signal to the on-time calculator. Other methods must wait for the output voltage to sag or soar before action is taken, and this is often too late.

A practical application of Quick-PWM is found in **Figure 12**. The MAX8632 is an integrated DDR memory power supply. Along with a Quick-PWM buck circuit (VDDQ), the MAX8632 integrates a high-speed linear regulator (VTT) to manage bus transients found in DDR memory systems. The linear regulator offers specific advantages over switchers: linear regulators do not have an inductor to limit current slew-rate, so a very fast current slew rate can service load transients. Slower circuits would require large capacitors to provide load current until the power supply can ramp up the current to service the load.

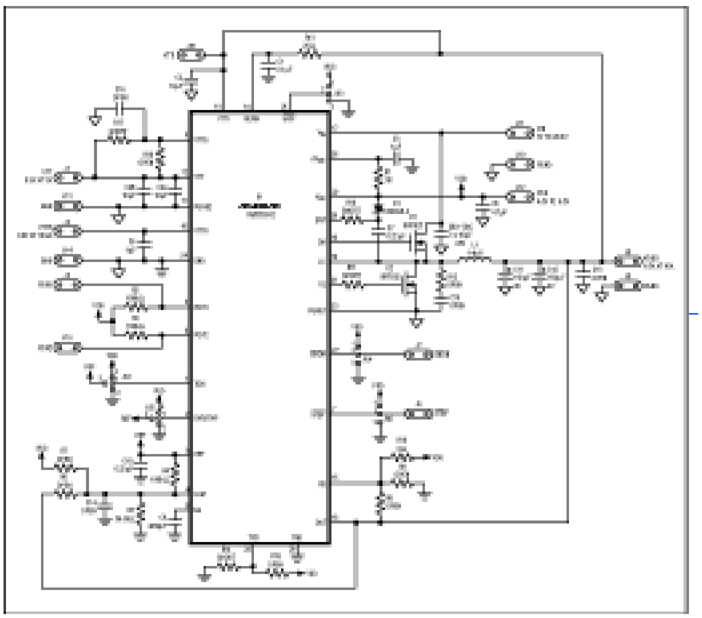

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Figure 12. The MAX8632 uses Maxim's Quick-PWM architecture and a linear regulator to provide a complete DDR power-supply system. The device can be used as a main GPU, or standard core-logic power supply.

Efficiency

One of the largest power-loss factors for switchers is the rectifying diode. The power dissipated is simply the forward voltage drop multiplied by the current going through it. The reverse recovery for silicon diodes can also create loss. These power losses reduce overall efficiency and require thermal management in the form of a heat sink or fan.

To minimize this loss, switching regulators can use Schottky diodes that have a relatively low forward-voltage drop and good reverse recovery. For maximum efficiency, however, you can use a MOSFET switch instead of the diode. This design is known as a 'synchronous rectifier' (see **Figures** 11, 12 and **13**). The synchronous rectifier switch is open when the main switch is closed, and the same is true conversely. To prevent cross-conduction (both top and bottom switches are on simultaneously), the switching scheme must be break-before-make. Because of this, a diode is still required to conduct during the interval between the opening of the main switch and the closing of the synchronous-rectifier switch (dead time). When a MOSFET is used as a synchronous switch, the current normally flows in reverse (source to drain), and this allows the integrated body diode to conduct current during the dead time. When the synchronous rectifier switch closes, the current flows through the MOSFET channel. Because of the very low-channel resistance for power MOSFETs, the standard forward drop of the rectifying diode can be reduced to a few millivolts. Synchronous rectification can provide efficiencies well above 90%

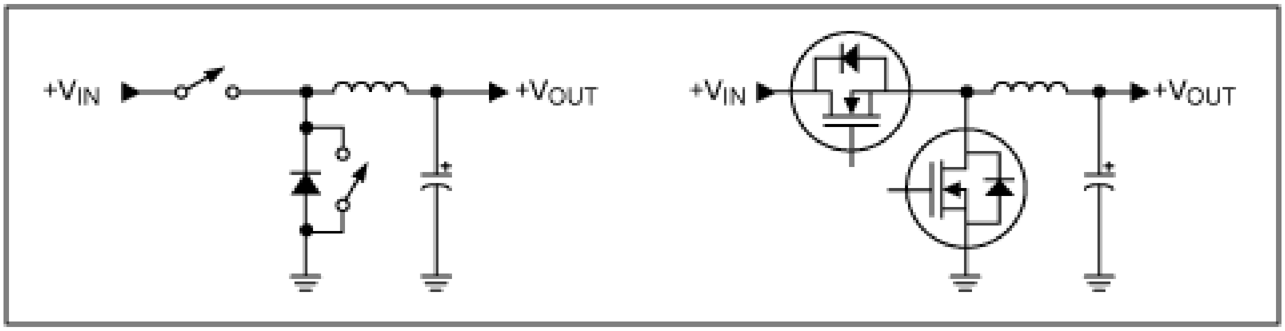


Figure 13. Synchronous rectification for the buck circuit. Notice the integrated MOSFET body diode.

Skip Mode Improves Light Load Efficiency

A feature offered in many modern switching controllers is skip mode. Skip mode allows the regulator to skip cycles when they are not needed, which greatly improves efficiency at light loads. For the standard buck circuit (Figure 1) with a rectifying diode, not initiating a new cycle simply allows the inductor current or inductor energy to discharge to zero. At this point the diode blocks any reverse-inductor current flow and the voltage across the inductor goes to zero. This is called "discontinuous mode" and is shown in **Figure 14**. In skip mode, a new cycle is initiated when the output voltage drops below the regulating threshold. While in skip mode and discontinuous operation, the switching frequency is proportional to the load current. The situation with a synchronous rectifier is, unfortunately, somewhat more complicated. This is because the inductor current can reverse in the MOSFET switch if the gate is left on. The MAX8632 integrates a comparator that senses when the current through the inductor has reversed and opens the switch, allowing the MOSFET's body diode to block the reverse current.

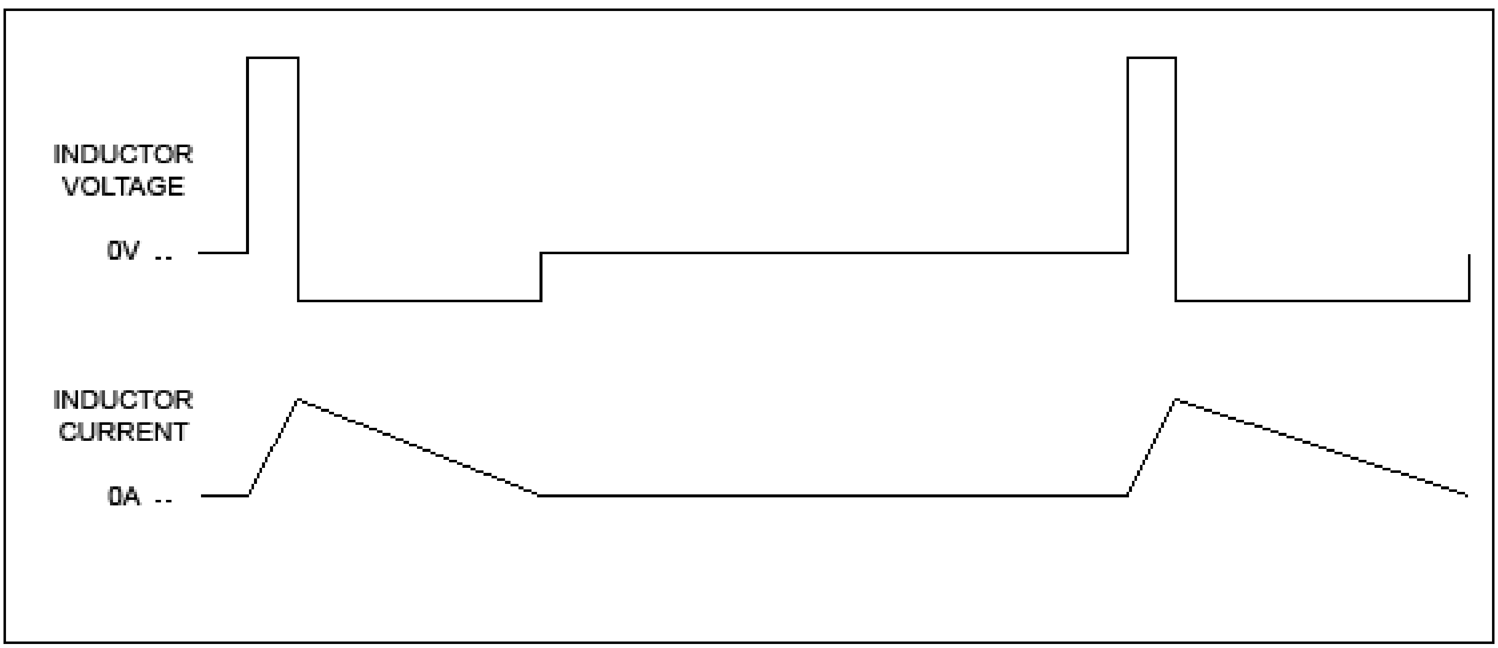


Figure 14. In discontinuous mode the inductor fully discharges and then the inductor voltage rests at zero.

Figure 15 shows that skip mode offers improved light-load efficiencies but at the expense of noise, because the switching frequency is not fixed. The forced-PWM control technique maintains a constant switching frequency, and varies the ratio of charge cycle to discharge cycle as the operating parameters vary. Because the switching frequency is fixed, the noise spectrum is relatively narrow, thereby allowing simple lowpass or notch filter techniques to greatly reduce the peak-to-peak ripple voltage. Because the noise can be placed in a less-sensitive frequency band, PWM is popular with telecom and other applications where noise interference is a concern.

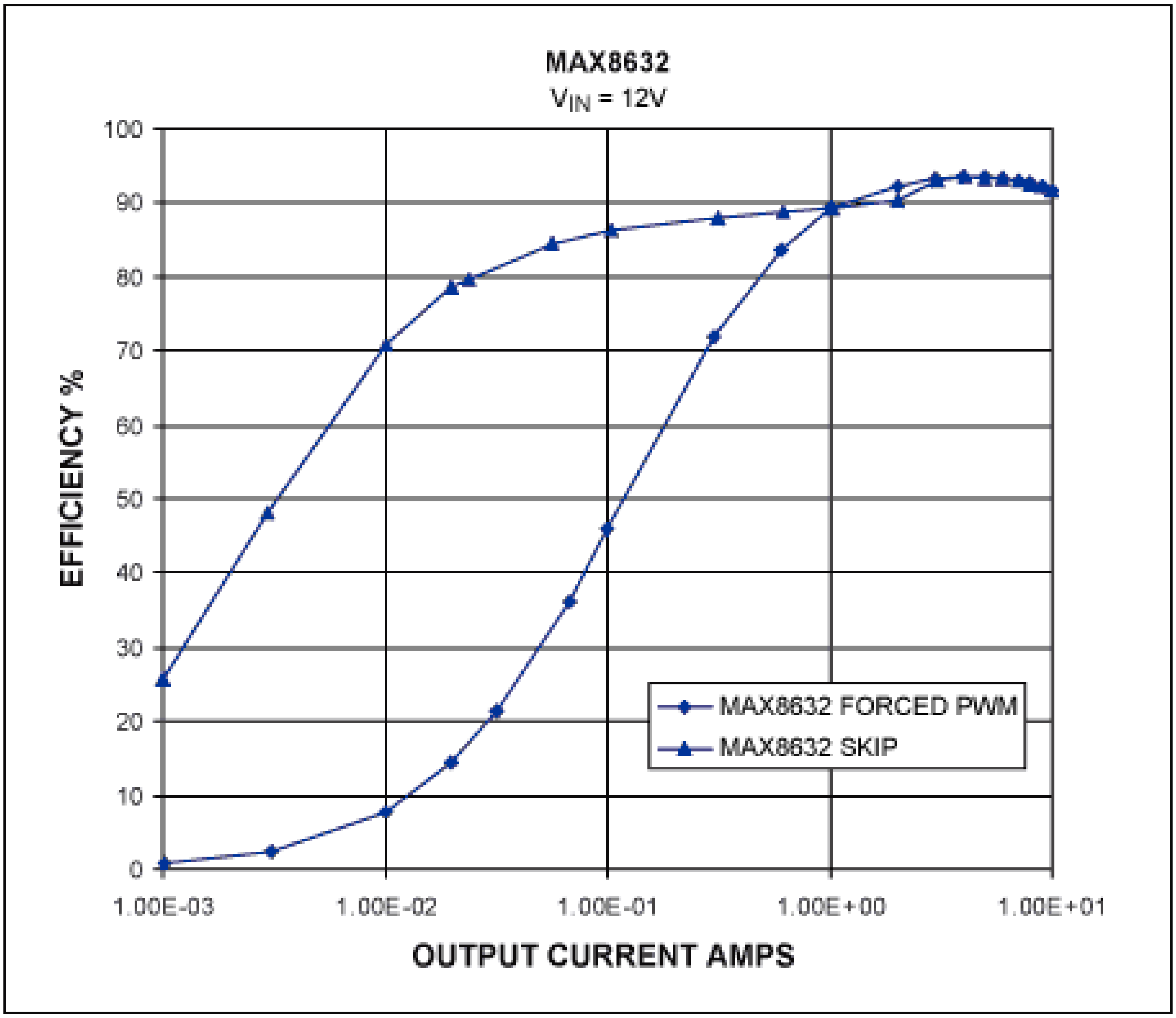


Figure 15. Efficiency with and without skip mode.

Summary

Although switching techniques are more difficult to implement, switching circuits have almost completely replaced linear power supplies in a wide range of portable and stationary designs. This is because switching circuits offer better efficiency, smaller components, and fewer thermal management issues.

MOSFET power switches are now integrated with controllers to form single-chip solutions, like the MAX1945 circuit shown in **Figure 16**. This chip has a metallic slug on the underside that removes heat from the die so the 28-pin TSSOP package can dissipate over 1W, allowing the circuit to supply over 10W to its load. With a 1MHz switching frequency, the output inductor and filter capacitors can be reduced in size, further saving valuable space and component count. As MOSFET power-switch technologies continue to improve, so will switch-mode performance, further reducing cost, size, and thermal management problems.

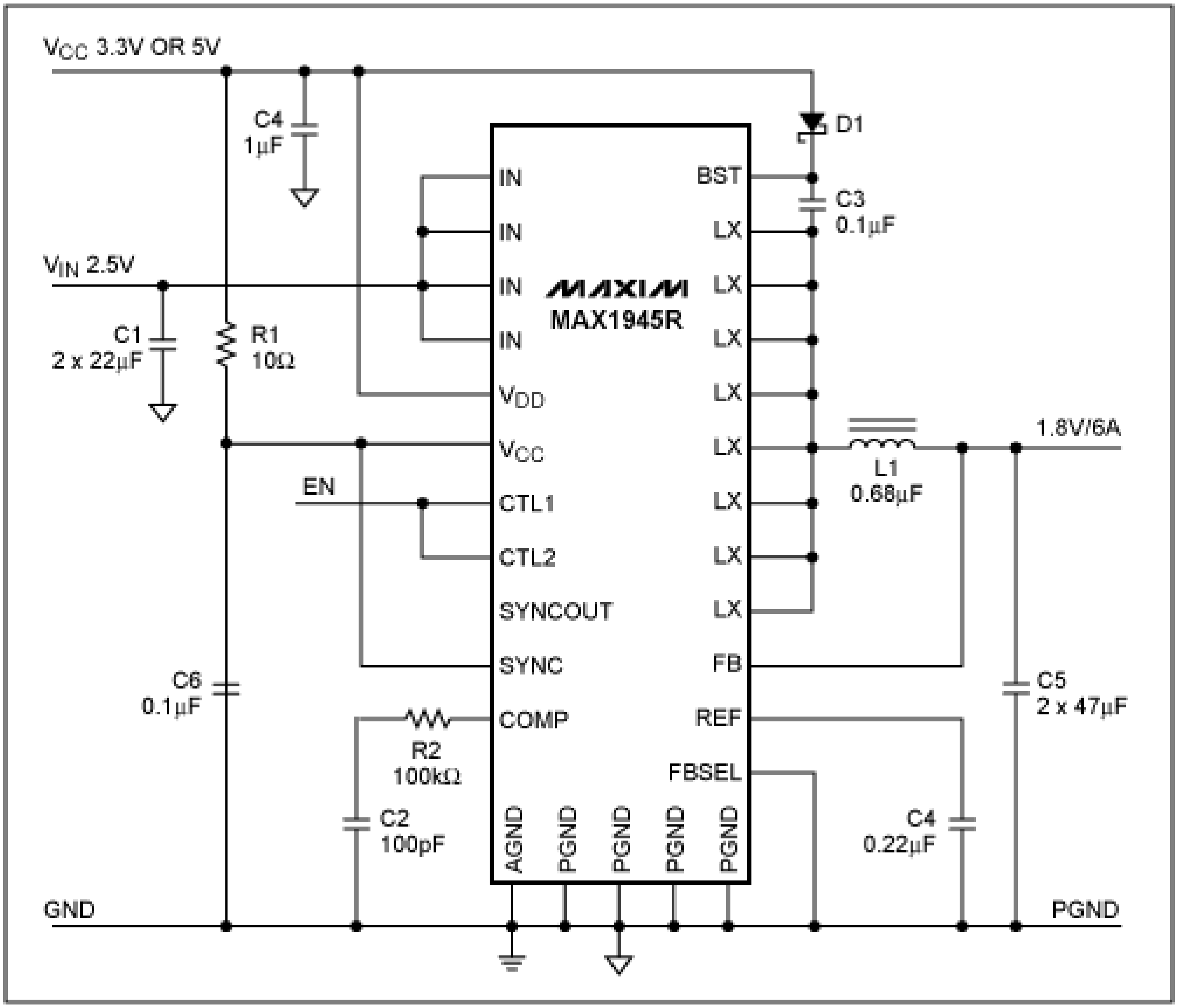


Figure 16. The MAX1945 is a 6A internal switch device with a reduced part count and small footprint to save board space.

Application Note 2031: http://www.maxim-ic.com/an2031

More Information

For technical questions and support: http://www.maxim-ic.com/support
For samples: http://www.maxim-ic.com/samples
Other questions and comments: http://www.maxim-ic.com/contact

Related Parts

MAX1932: QuickView -- Full (PDF) Data Sheet -- Free Samples

MAX668: QuickView -- Full (PDF) Data Sheet -- Free Samples

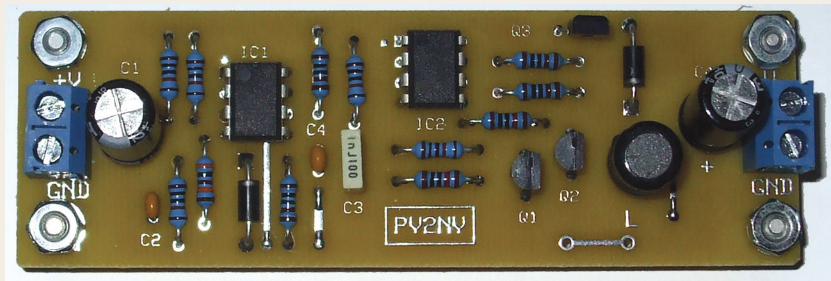
MAX8632: QuickView -- Full (PDF) Data Sheet -- Free Samples

AN2031, AN 2031, APP2031, Appnote2031, Appnote 2031

By Jim Stewart

BUILD AN INVERTING DC-DC CONVERTER

It's often the case that you need +V and -V when all you have is +V. For example, you need +12V and -12V, but all that's available is +12V. It would be really handy to have a "black box" that would give you -V out when you put +V in, and work over a range of voltages without adjustment. In this project, we will build just such a thing. It's a voltage mirror that can supply over 100 mA without a significant drop in voltage. The circuit uses a switched inductor as described in the sidebar.

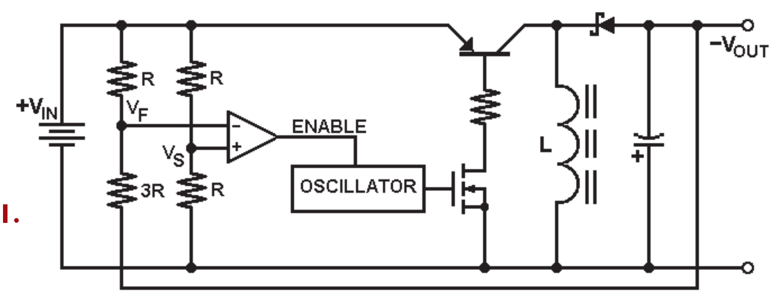


How It Works

Figure 1 shows a block diagram of the circuit. A square wave oscillator drives a MOSFET that, in turn, drives a PNP switching transistor. The oscillator is enabled/disabled by the output of the comparator. A comparator is just an op-amp that compares a set point voltage (V_S) to a feedback voltage (V_F). Depending on which voltage is higher, the comparator's output goes either high (+V) or low (ground).

When V_{OUT} is the correct voltage, V_S) exceeds V_F by a few micro volts and the oscillator is disabled. The PNP switch is off. When V_S) is less than V_F, the oscillator is enabled and the PNP transistor switches on and off.

V_S) is half the input voltage (V_S) = V/2) as set by the R-R voltage divider. V_F is set by the R-3R divider, and becomes V/2 when the output voltage equals -V. The diode blocks the +V from the output while charging the inductor. It then provides a current path for the discharging inductor to transfer charge to the output capacitor.



■ FIGURE 1.

Because the diode turns on and off to control the inductor current, it's called a *commutating* diode. Commutating is an old term, often used with DC motors. It means to switch current from one path to another.

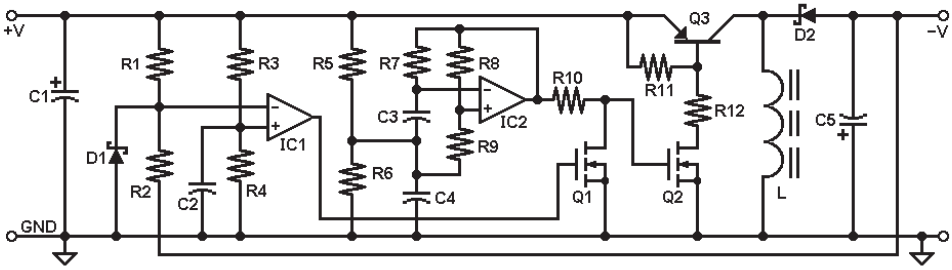
This type of circuit is often referred to as a *buck-boost* DC-to-DC converter. The term *buck* refers to the inductor opposing current through it while it charges. The term *boost* refers to the ability of the inductor to increase output voltage when it discharges. In this case, we have an *inverting* converter since a positive input voltage produces a negative output voltage.

The oscillator causes the PNP transistor switch to open and close at a fixed frequency. When the switch closes (transistor ON), the inductor charges up. When the switch opens (transistor OFF), current from the inductor charges up the voltage on the capacitor.

The switching continues until the voltage on the capacitor becomes V_{OUT} = $-V_{IN}$. At that point, the comparator disables the oscillator and the PNP stays OFF. When voltage across the capacitor drops, the comparator enables the oscillator and the inductor is pumped up again.

The Circuit

Figure 2 shows the schematic. IC1 and IC2 are CMOS op-amps. IC1 is the comparator. IC2 forms a square wave oscillator with frequency given by $f = 1/2.2(R7 \times C3)$. IC2 drives Q2 through R10. IC1 controls the gate of Q1. When Q1 is off, the square wave is applied to the gate of Q2, which toggles Q3 on and off to pump up the inductor.



■ **FIGURE 2.**

When Q1 is on, the gate of Q2 is grounded and Q3 stays off. Since there's a single input voltage, each op-amp uses a resistor divider to "split the rail" to create a signal ground.

That allows the negative input to see plus and minus voltages with respect to the positive input. C2 and C4 bypass the signal grounds to the supply voltage ground.

C1 and C5 are low ESR (Equivalent Series Resistance) aluminum electrolytic capacitors. Tantalum capacitors might be a bit better, but are more expensive.

ESR is a measure of how much power the capacitor will dissipate. $P_{DISS} = I^2_{RC} \times ESR$, where I_{RC} is the AC "ripple current" in the capacitor. The capacitors chosen are rated for a maximum ripple current of 840 mA.

D1 and D2 are Schottky diodes. Schottky diodes can go from conducting to non-conducting very quickly, allowing a high switching frequency. They also have a low voltage drop when conducting.

In a switched inductor circuit, much of the power loss can be due to the voltage drop across the commutating diode. D1 is there in case the feedback fails and V_{OUT} becomes too negative.

Q3 is a ZTX550 PNP transistor, chosen because its specifications suit this application well:

- Maximum power dissipation: P_{MAX} = 1 watt @ 25 °C
- Maximum continuous collector current: I_C = one amp
- Maximum collector-base voltage: V_{CBO} = 60 volts
- Maximum saturation voltage: V_{CE} = 0.25 volts @ I_C = 150 mA
- Transition frequency: f_T = 150 MHz minimum
- Package: E -Line (slightly smaller than TO-92)

SWITCHED INDUCTORS

Like capacitors, inductors can be charged up. **Figure S1** shows a DC voltage applied to an inductor via a switch. Current (I_{IN}) will increase according to $I = (V/L) \times \Delta t$ where Δt is the length of time the inductor is charging. At maximum current (I_M), the inductor stores energy (W) given by $W = (L\, I^2_M)/2$. The energy is stored in the magnetic field around the inductor.

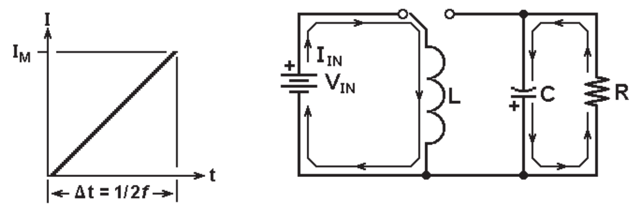


FIGURE S1. Charging an inductor.

Current In Inductor

When the switch is moved from battery to load, the inductor discharges as shown in **Figure S2**.

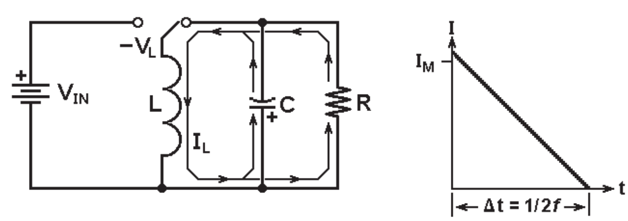


FIGURE S2. Discharging an inductor.

Four important things then happen:

First, the *magnitude* of the inductor current (I_L) just *after* the switch is moved to the load is equal to the current (I_M) just *before* the switch is moved. That's because energy is conserved.

Second, the *direction* of current just after the switch is moved is the same as the direction just before the switch is moved. Again, it's conservation of energy.

Third is the key point: To keep the current I_L flowing in the same direction, the voltage across the inductor will *switch polarity* as the discharging inductor acts like a generator.

Fourth, the voltage across it will become whatever value is necessary to keep the current flowing until all the energy is transferred to the capacitor or dissipated in the load resistance.

PARAMETERS

How are L, *f*, and R_{LOAD} related to each other in this circuit? We start with the requirement that the input power (P_{IN}) be greater than the output power (P_{OUT}). It's *greater than* instead of *equal to* because efficiency is not 100%; some power is lost. So, we need to find expressions for P_{IN} and P_{OUT}. We then let P_{IN} > P_{OUT} and see what we get. We do it for one oscillator cycle.

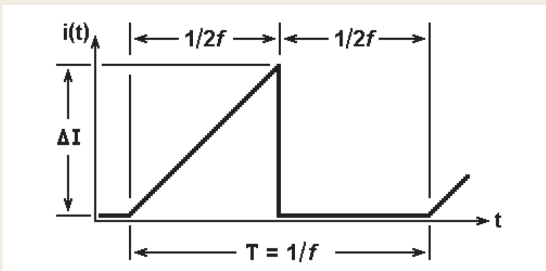


FIGURE P1. I vs. t for one cycle.

Step 1: Find P_{OUT}
For the entire cycle (T = 1/*f*), we have
Equation 1: $P_{OUT} = V^2_{OUT} / R_{LOAD}$

Step 2: Find P_{IN}
$P_{IN} = V_{IN} \times I_{IN}$ where $I_{IN} = I_{DC}$ = average current (I_{AVG}) during the entire cycle
For the first half-cycle, $I_{AVG} = \Delta I/2$ while for second half-cycle $I_{AVG} = 0$
So, for the entire cycle $I_{AVG} = (\Delta I/2 + 0) \div 2 = DI/4$
In a charging inductor, $V = L \times (\Delta I/\Delta t)$, so $\Delta I = (V_{IN} / L) \times \Delta t$
The inductor charges during $\Delta t = T/2 = 1/2f$
So, $\Delta I = V_{IN} / 2fL$
And for the entire cycle $I_{AVG} = \Delta I/4 = V_{IN} / 8fL$
So, $P_{IN} = V_{IN} \times I_{AVG} = V_{IN} \times (V_{IN} / 8fL)$ to give
Equation 2: $P_{IN} = V^2_{IN} / 8fL$

Step 3: Let $P_{IN} > P_{OUT}$ and solve
$P_{IN} > P_{OUT}$ gives $V^2_{IN} / 8fL > V^2_{OUT} / R_{LOAD}$
Invert both sides to get $1/P_{IN} < 1/P_{OUT}$
So, $8fL / V^2_{IN} < R_{LOAD} / V^2_{OUT}$
Re-arrange terms to get $8fL < (V_{IN} / V_{OUT})^2 \times R_{LOAD}$ which gives
Equation 3: $fL < (V_{IN} / V_{OUT})^2 \times (R_{LOAD} / 8)$

And if, $V_{OUT} = V_{IN}$ we get
Equation 4: $fL < R_{LOAD} / 8$

Design Parameters

This design has five important parameters:

- Input voltage: V_{IN}
- Output voltage: V_{OUT}
- Load resistance: R_{LOAD}
- Oscillation frequency: *f*
- Inductance value: L

The parameters are related to each other by the equation:

$f L < (V_{IN} / V_{OUT})2 \times (R_{LOAD} / 8)$

(To see where the equation comes from, see the **sidebar**.)

We want the magnitudes of V_{IN} and V_{OUT} to be equal, so the equation becomes:

$f L < (R_{LOAD} / 8)$

For our circuit, L = 220 mH, *f* = 45 kHz, and R_{LOAD} = 100 Ω. To see if those values work, we need to verify that *f* L is less than R_{LOAD} / 8.

$f L = (45 \times 10^{3}) \times (220 \times 10^{-6}) = 9900 \times 10^{-3} = 9.9\ \Omega$

$R_{LOAD} / 8 = 100/8 = 12.5\ \Omega$

Since 9.9 is less than 12.5, the circuit should work. Note that increasing the load resistor makes *f* L even less than R_{LOAD} / 8.

Note: The above calculation just lets us verify that the values we use are in the ballpark. It's not an exact description.

I replaced the 220 µH inductor with one that was 330 µH and it worked just fine. Also, the value of an inductor usually drops as the DC current through it increases.

Performance

1. Mirror Effect: A 1 kΩ load was connected to the output and V_{OUT} was measured for various values of V_{IN}. The results are shown in **Table 1**. Note that the differences are more or less constant. That means the circuit has a fixed offset of about 0.11 volts. That could be "zeroed out" by adjusting R1 or R2.

2. Efficiency (η): V_{IN} was set to +12V and a 110 Ω load was connected to the output. Input current was measured to obtain input power $P_{IN} = V_{IN} \times I_{IN}$. Output power was V^2_{OUT} / 110 Ω. The result was:

I_{IN} = 150 mA
P_{IN} = 12V x 150 mA = 1.80 watts
$P_{OUT} = (11.86\ V)^2 / 110\ \Omega$ = 1.28 watts

$\eta = (P_{OUT} / P_{IN}) \times 100\%$ = (1.28W / 1.80W) x 100% = 71.1% efficiency

3. Load Regulation: With 12 volts in, V_{OUT} was 11.87 volts with a 1 kΩ load and 11.86 volts with a 110 Ω load. That gives a load regulation of:

LOAD-REG = [(11.87 - 11.86) / 11.87] x 100 % = 0.08%

A load regulation of 0.08% is not bad for such a simple circuit.

4. Ripple Voltage: With V_{IN} = 12 volts and a 110 Ω load on the output, ripple voltage was 50 mV_{PP} as measured with an oscilloscope.

V_{IN}	V_{OUT}	Difference
6.0V	- 5.90V	0.10V
7.0V	- 6.89V	0.11V
8.0V	- 7.89V	0.11V
9.0V	- 8.89V	0.11V
10.0V	- 9.88V	0.12V
12.0V	- 11.87V	0.13V
12.0V	- 11.86V across 110 Ω	0.14V
15.0V	- 14.86V	0.14V

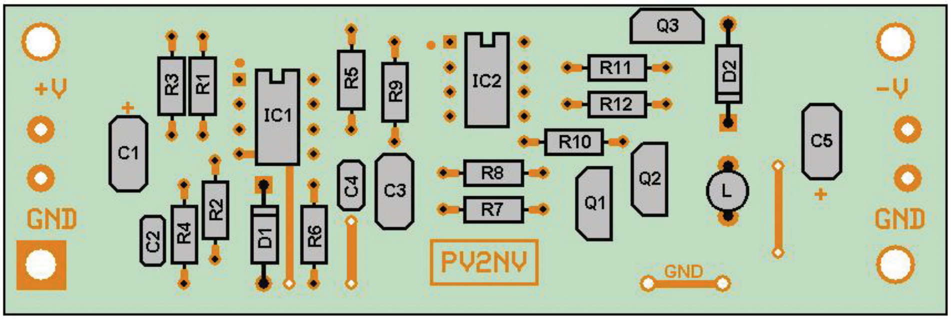
Table 1: V_{OUT} vs. V_{IN}

Construction

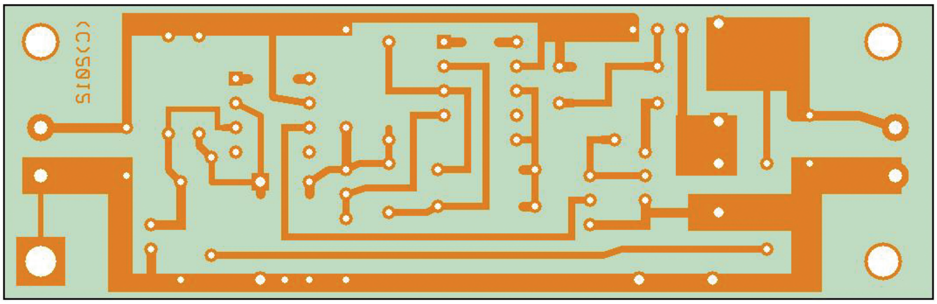
The printed circuit board (PV2NV) component layout is shown in **Figure 3**. The optional terminal blocks are not shown. The solder side is shown in **Figure 4**. An ExpressPCB file (PV2NV.PCB) is available at the article link.

There is a mounting hole at each corner of the board that's big enough for a #4 screw. The mounting hole with the square outline is connected to circuit ground by a thin strip of copper, so mounting the board to a metal chassis will connect ground to the chassis. If that's not desirable, you can cut the strip.

Board dimensions are 3.8 inches by 1.25 inches.



■ FIGURE 3.

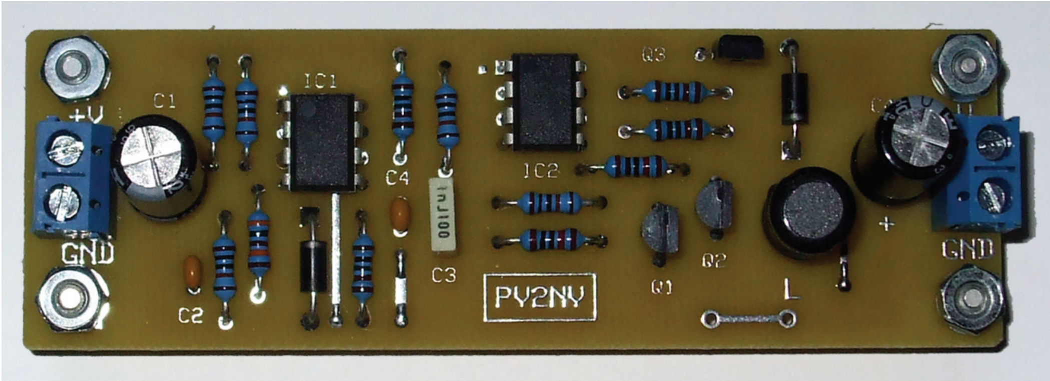


■ FIGURE 4.

Procedure

If you have a lead forming tool (**Figure 5**), bend the resistor and diode leads to a 0.4 inch spacing. Such tools (Electronix Express part #0664ST010; Jameco part #106884) are inexpensive and handy to have. (When inserting components into the board, bend the leads slightly on the solder side to secure them in place.)

■ **FIGURE 5.**



■ **FIGURE 6.**

Here is a suggested sequence for mounting the components:

- Insert R3, R1, R5, R9, and solder. Clip off excess leads.
- Insert R4, R2, R6, and solder. Clip off excess leads.
- Insert R11, R12, R10, R8, R7, and solder. Clip off excess leads.
- Insert D1, D2, Q1, Q2, Q3, and solder. Clip off excess leads.
- Insert IC1 then IC2. Verify pin 1 is in the proper hole and then solder.
- Insert the inductor and capacitors. Note polarity of C1 and C5. Solder and clip off excess leads.
- If using terminal blocks (optional), mount and solder them.
- Clean off the solder side with some rubbing alcohol and a scrub brush. Give the assembled board a good visual inspection and fix any problems you find.

Assembly is now complete. A photograph of the assembled board is shown in **Figure 6**.

PARTS LIST

ITEM	DESCRIPTION
R1:	100 kΩ, 1/4W, 1%
R2:	301 kΩ, 1/4W, 1%
R3:	100 kΩ, 1/4W, 1%
R4:	100 kΩ, 1/4W, 1%
R5:	1 kΩ, 1/4W, 1%
R6:	1 kΩ, 1/4W, 1%
R7:	10 kΩ, 1/4W, 1%
R8:	100 kΩ, 1/4W, 1%
R9:	100 kΩ, 1/4W, 1%
R10:	10 kΩ, 1/4W, 1%
R11:	1 kΩ, 1/4W, 1%
R12:	1 kΩ, 1/4W, 1%
L:	220 µH Digi-Key part #811-1316-ND or equivalent
C1:	220 µF, 35V Digi-Key part #493-1578-ND or equivalent
C2:	0.1 µF, 50V, ceramic
C3:	1 nF film, 5% Digi-Key part #399-5871-ND or equivalent
C4:	0.1 µF, 50V, ceramic
C5:	220 µF, 35V Digi-Key part #493-1578-ND or equivalent
Q1:	2N7000
Q2:	2N7000
Q3:	ZTX550 PNP
D1:	Schottky, one amp, 1N5819 or equivalent
D2:	Schottky, one amp, 1N5819 or equivalent
IC1:	CA3140
IC2:	CA3140

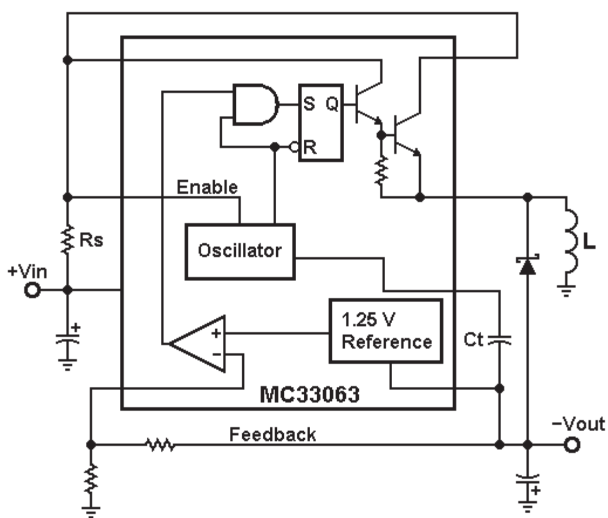
Printed Circuit Board: An ExpressPCB file called PV2NV.PCB is available for download from the article link.

Terminal Blocks (optional): Two-position, 5 mm spacing, Jameco part #2094485 or equivalent.

Testing

Now for the moment of truth: Apply V_{IN} and measure $-V_{OUT}$ with your multimeter. Verify that the magnitudes of V_{IN} and V_{OUT} are equal. If not, remove power and examine the board. Check for the usual suspects:

- Any solder bridges or bad solder joints?
- Are components in the right place?
- Are any components missing?
- Are the cathode and anode of a diode swapped?
- Are the transistors in backwards?
- Are the polarities of the electrolytic caps reversed?
- Do the ICs have pin 1 in the correct hole?



■ **FIGURE 7.**

Wrapping Up

Most buck-boost converters use pulse width modulation (PWM) to control the voltage applied to the inductor. Feedback voltage is compared to a fixed reference voltage and the pulse width is adjusted to obtain the desired output voltage. An example circuit using the MC33063 IC is shown in **Figure 7**.

Note that the output voltage of such a circuit depends only on the reference and does not change with the input voltage. For our circuit, output voltage is a mirror image of input voltage. Since I've never seen a similar circuit, I'm tempted to think there aren't any like it. That, however, is not very likely. **NV**

Four Output Bench Supply

Every workbench needs a power supply — this one provides four different outputs.

Larry Cicchinelli, K3PTO

This project is a four output bench power supply. Three outputs (positive voltage) use identical switching regulator circuits that can be set to be any voltage between 3.3 and 20 V. Each output is independent of the others and is capable of up to 1 A. The fourth output is via a negative regulator capable of about 250 mA. The unit I built has two fixed outputs and two variable outputs. You can also make any combination of them variable within the above range.

The only dependence among the outputs is that they are all driven by a single transformer. One of the features of a switching regulator is that you can essentially trade off between voltage and current. The transformer I used is rated at 25 V and 2 A. As such it is good for 50 W. Assuming that the regulator IC being used has an efficiency of 75%, you will have a total of about 37 W available from all power supply outputs. In practical terms this means you can get more current from the outputs than what the transformer is supplying — as long as you stay within the 37 W and the maximum current per regulator.

The regulator I used is the 3.3 V version of the LM2575. If you examine its data sheet you will see that the only difference among the models is the internal voltage divider. This allows you to design a power supply with a higher output voltage by simply inserting a resistance between the output and the FEEDBACK pin. I selected the 3.3 V version mainly due to its cost relative to the others. With it I can get any voltage from 3.3 V to 20 V from the circuit. You could also use the "adjustable" version, which will then allow you to select any voltage between 1.23 and 20 V.

The regulator is specified for up to 37 V output. Since I have specified 50 V capacitors, I believe you should be able to get up to about 30 V output. If you want to output a higher voltage than 30 V, I recommend that you use higher voltage capacitors. The transformer I am using is rated at 25 V; however, I have measured the loaded output at closer to 30 V ac, so I could probably get up to 25 V from the regulators. You will also need to use the 200 V range of the digital panel meter (DPM).

There is also a high voltage version of the LM2575 that can provide outputs of up to 57 V. I recommend that you use capacitors rated to at least 100 V if you decide to use that version.

A Little Theory

Switching regulators come in essentially three varieties: buck, boost and buck-boost. The regulator in this article is of the buck type — the output voltage is less than the input voltage. The main feature of a switching regulator that differentiates it from a linear regulator is that the switcher oscillates. They generally use a form of pulse width modulation (PWM) in order to regulate the output voltage. The rise and fall times of the oscillator are quite fast and the harmonics can cause interference to communications receivers. This is the reason a spectrum analyzer is one of the pieces of test equipment used to characterize a switching regulator. This is certainly not the case with a linear regulator!

Two of the best tutorials I have found on switching regulators are *Application Note 2031* on the Maxim-IC Web site at **www.maxim-ic.com/appnotes.cfm/an_pk/2031** and at **www.national.com/appinfo/power/files/f5.pdf** from National Semiconductor. Rather than try to repeat much of the material in that note, I suggest that you get a copy and read them for yourself.

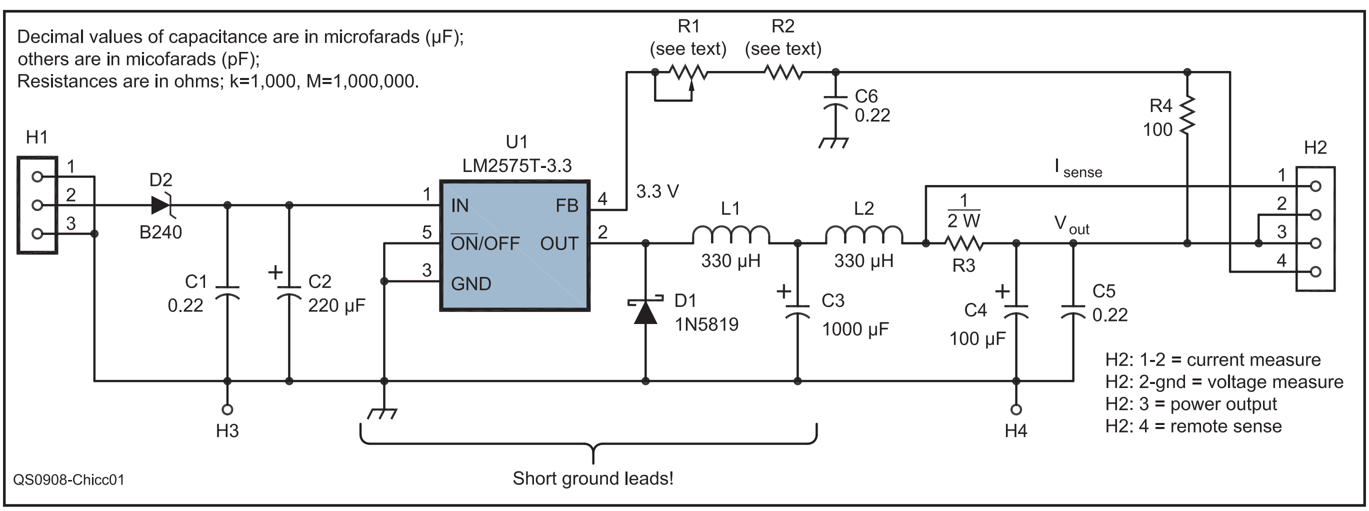


Figure 1 — Schematic diagram of a single positive regulator module.

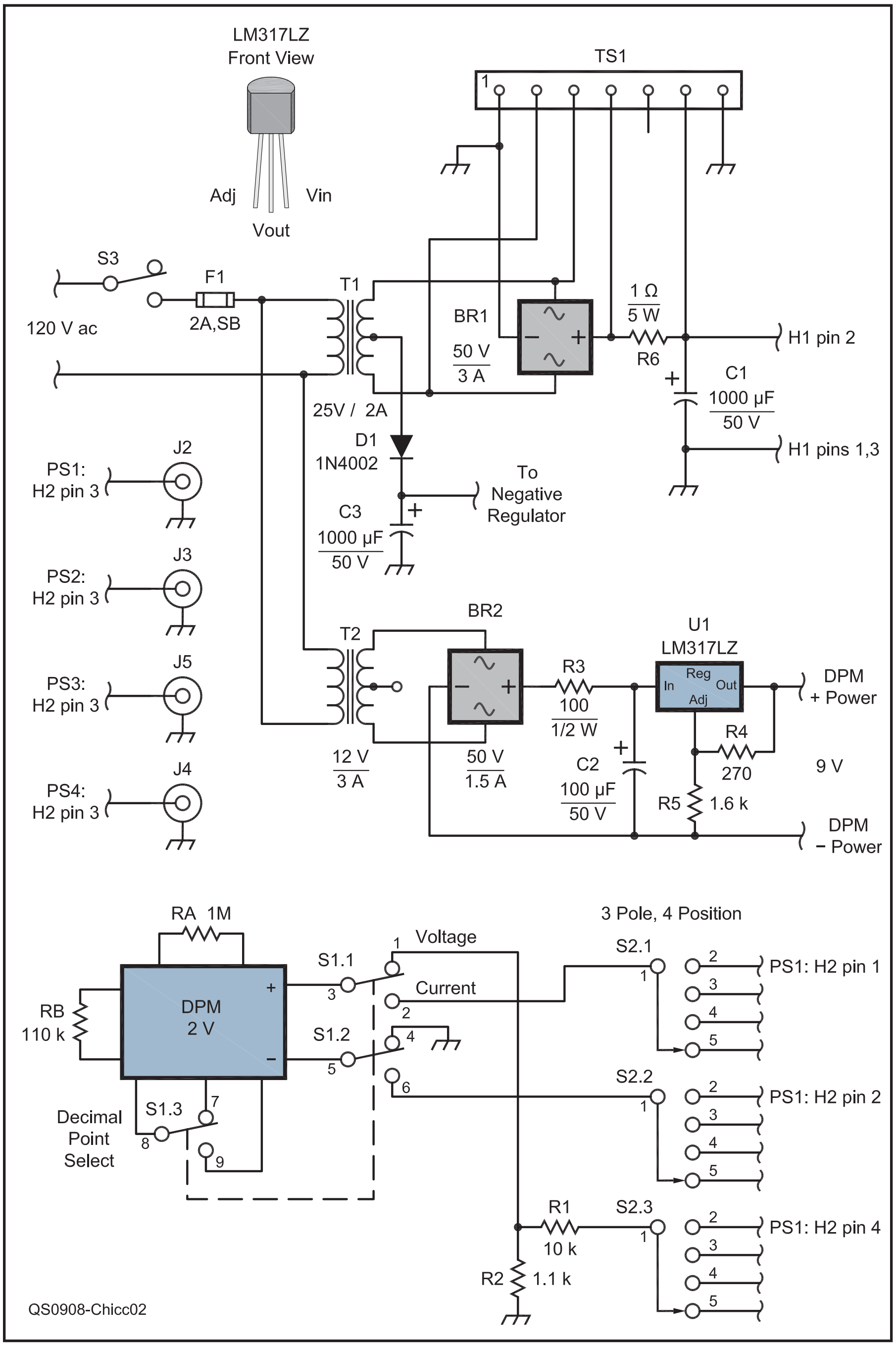


Figure 2 — Chassis schematic showing the interconnection of the modules as well as control and metering details.

A switching regulator will have some amount of high frequency noise on its output at the switching frequency, about 52 kHz for the LM2575. In the circuits described here there is a low pass filter on each output that reduces, but does not completely eliminate, this noise. If your requirement is for fixed voltages, you can add a low drop out series regulator (LDO). A good LDO typically requires only about 100 mV between the input and output voltages, so you can design the switcher to be a little higher than the desired voltage and get the benefits of both types of regulators.

Some Circuit Details

Figure 1 is a schematic of a single positive regulator. There are several variations of the circuit which could be implemented. L2 and C4 are optional. These two components provide a low pass filter that will decrease the 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 if the calculated value is nonstandard and you want to use two fixed resistors.

The formula for the output voltage (with the 3.3 V version) can be calculated as follows. The current (in mA) through the internal voltage divider is

$I = 3.3\ V/2.7\ k\Omega = 1.22\ mA$
$R1 + R2 = (V_{OUT} - 3.3) / 1.22\ \ k\Omega$

transposing terms yields:
$V_{OUT} = [1.22 \times (R1+R2)] + 3.3$

Note that if you make R1=R2=0, the calculation results in an output of 3.3 V. The leakage current of the error amplifier in the regulator is somewhat less than –25 nA, so it can be ignored. Also, since the 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.

If you want to have an accurate, fixed output voltage, I recommend selecting a value for R2 that is lower than the calculated value. Then select a potentiometer for R1 that yields a reasonable adjustment range.

If you decide to use the extra LC filter, you will have to install L1 and L2 such that their phasing dots line up with the dot symbols on the circuit board. I found out the hard way that if the dots are at the same end of the board, the output will have an additional low frequency ripple. When I built my board I just happened to have three circuits assembled correctly. The fourth one had a serious low frequency ripple that I could not get rid of. I eventually replaced every component, one at a time, to find the bad one. When I replaced L2 the output was okay. It was then I noticed the phasing dots. I reversed L2 just to see what would happen and the ripple came back. There can be inductive coupling, even though there is a ground plane on both sides of the board under the inductors.

Remote Sensing

A feature of many power supplies is that of remote sensing. This is used to electronically adjust for the voltage drop in the wires carrying current to the load. I found that, 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. If you are not going to use remote sensing then you should insert a jumper in place of R4. R4 (100 Ω) is there for protection just in case the remote sense connection is missing.

To do remote sensing most effectively you will have to implement the circuit somewhat differently than is indicated in the chassis schematic. I used the same point to pick up the output voltage for both the voltage and current measurements (H2-2). This measures the voltage at the output of the regulator board — not at the load. If you want really accurate readings on the DPM, and have accurate remote sensing, you will have to pick up the voltage measurement from the remote sense input of the board at H2-4 (see Figure 2). This will involve using a three pole, four position rotary switch. This is necessary because you still need to measure the voltage drop across the current sense resistor right at the resistor.

If you do not want to use remote sensing you can simplify the switch wiring to use a two pole switch instead of the three pole listed. In this case you would essentially not use S2.2 and connect S1.2 to the common of S2.3 instead of S2.2.

Strictly speaking even this does not fully implement remote sensing. This circuit does not have a mechanism to adjust for the voltage drop in the ground leg. Most high-end commercial supplies will have both power and ground sense inputs. For this power supply make sure that the ground leads have minimal voltage drop. Measurements inside the chassis have indicated this. I have measured about 100 mV drop at 1 A between the positive output of the regulator board and the chassis connector. There was no measurable voltage drop in the ground circuit. You just have to be sure to use relatively heavy wires for the ground connections.

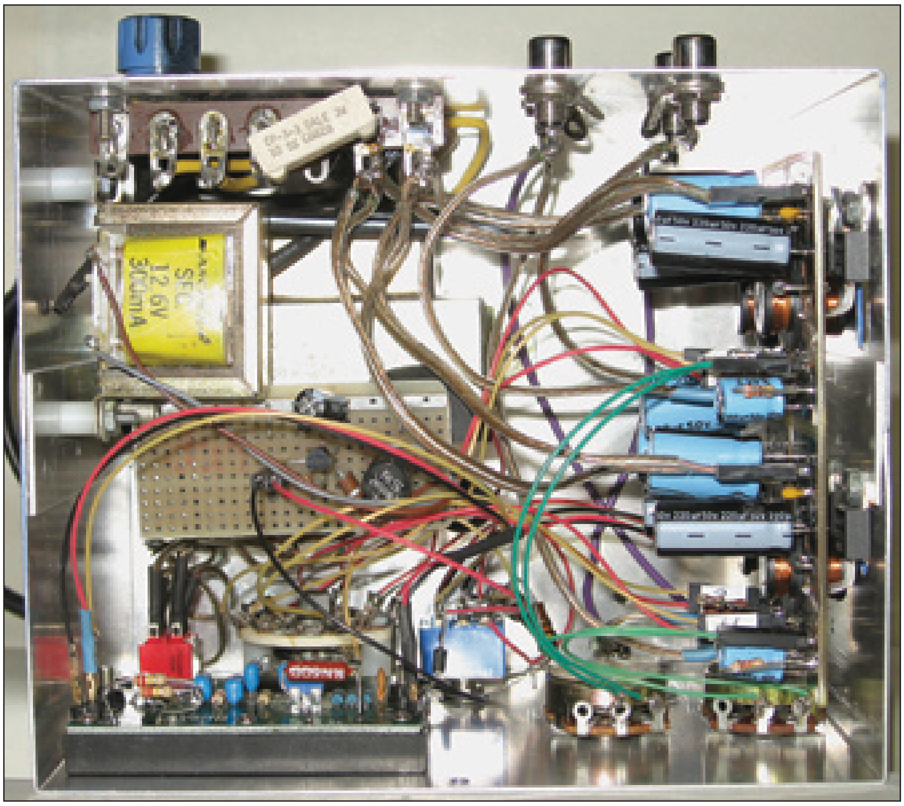


Figure 3 — Underchassis view of the completed power supply. The components used for the main rectifier circuit are mounted on a terminal strip shown at the top left.

Efficiency

Table 1 on the binaries web version shows the efficiency of the positive regulator with various input voltages. Notice that the efficiency is really good at 14 V; however, the circuit is no longer regulating! Optimum efficiency seems to occur at 20 V but there is not a whole lot of variation between 16 and 28 V.

Rectifier Circuit

Figure 2 shows the connections among the parts of the system — regulator boards, DPM and rectifier circuit. The components used for the main rectifier circuit are mounted on a terminal strip (Mouser 158-1008). You can see the terminal strip and R6 at the top left of Figure 3. You can hardly see it, but C1 is mounted underneath the terminal strip. The leads of the bridge rectifier are soldered into the holes that are used to rivet the terminals to the Bakelite. One of the four leads, the negative output, is soldered to a grounded terminal. Since I have had quite a few of these terminal strips for several years I used fine Emory paper to clean their surfaces as well as a small file to clean the holes. This was done in order to insure good solder connections.

Negative Regulator

The negative regulator is of the buck-boost configuration. It converts a positive voltage into a negative one (see Figure 4). This design uses many of the same component values as the positive regulators. I was unable to implement the current measuring circuit within a feedback loop. I tried several configurations but each introduced a significant low frequency noise component to the output voltage. There was also some

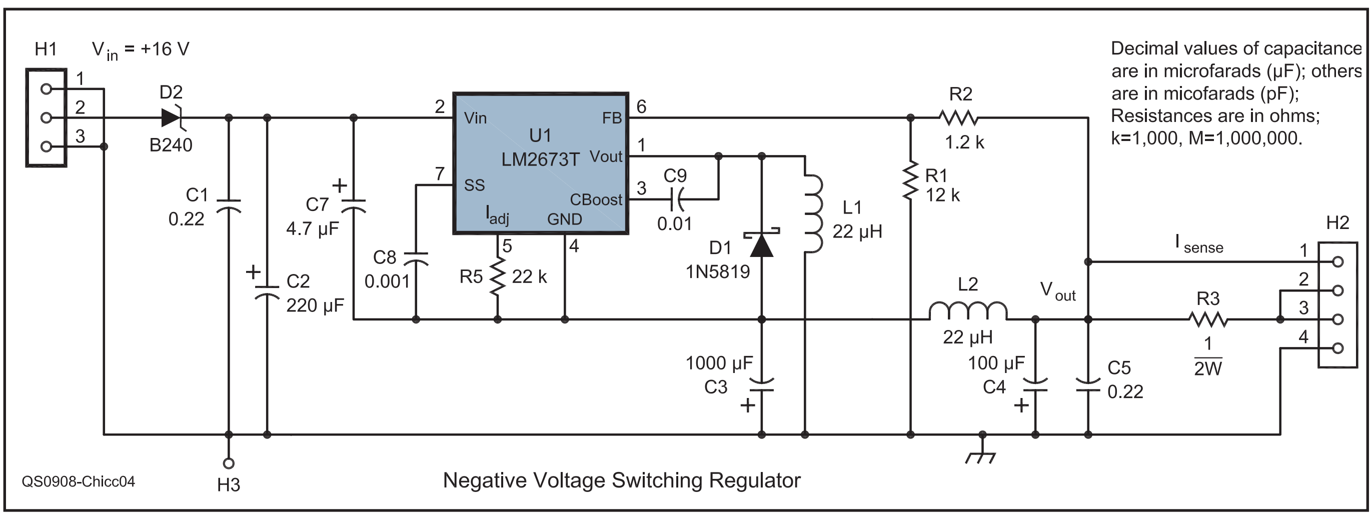


Figure 4 — Schematic diagram of a single negative regulator module. The complete supply can include up to four modules.

Table 1
Power Supply Efficiency Using the Positive Regulator with R_{LOAD} of 50 Ω

Measured Values			*Calculated Values*		
Voltage In	*Current*	*Voltage Out*	*Power In*	*Power Out*	*Efficiency (%)*
14	0.275	13.7	3.85	3.75	98
16	0.25	12.6	4	3.18	79
18	0.23	12.6	4.14	3.18	77
20	0.2	12.6	4	3.18	79
22	0.19	12.6	4.18	3.18	76
24	0.175	12.6	4.2	3.18	76
26	0.165	12.6	4.29	3.18	74
28	0.155	12.6	4.34	3.18	73

Table 2
Power Supply Efficiency Using the Negative Regulator with R_{LOAD} of 50 Ω

Measured Values			*Calculated Values*		
Voltage In	*Current*	*Voltage Out*	*Power In*	*Power Out*	*Efficiency (%)*
5	1.3	11.1	6.50	2.46	38
7	0.77	12.1	5.39	2.93	54
9	0.48	12.2	4.32	2.98	69
11	0.38	12.2	4.18	2.98	71
13	0.32	12.3	4.16	3.03	73
15	0.27	12.3	4.05	3.03	75
16	0.25	12.3	4.00	3.03	76

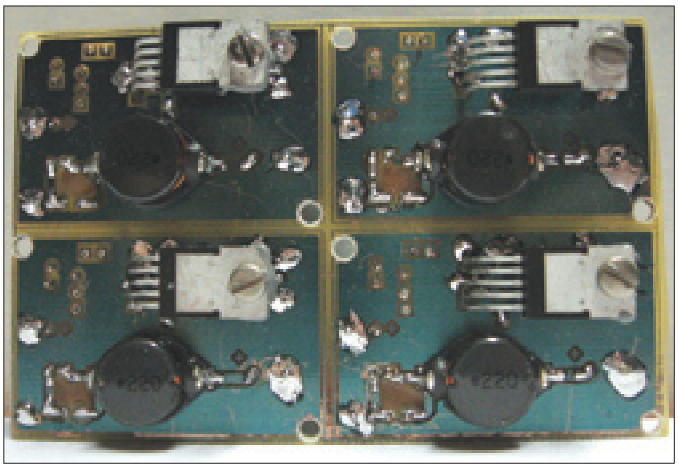


Figure 5 — Back side of the PCB assembly with the ICs folded over and the bolts holding the nuts in place while the epoxy hardens.

50 kHz noise present on the output, so I implemented an additional low pass filter on the output that reduces it considerably.

The negative regulator uses a different rectifier than the positive regulators. This was necessary because the LM2575 has a maximum input to output voltage specification of 40 V. Since the output of the main bridge circuit is 30 V and the output of the negative regulator is –12 V, the difference exceeds the specification. Using the center tap of the transformer yields close to +16 V for the negative regulator. The half wave rectifier circuit is not as efficient as a bridge, but it will suffice for bench top analog circuits. The components can be mounted to the unused terminals of the terminal strip used for the main rectifier.

One surprising parameter of this circuit is that the inductor current is quite high. Even with no load, the formula indicates that there will be about 200 mA flowing through it. With a 100 mA load there will be an increase of about 175 mA. You can find the formula on page 20 of the LM2575 data sheet. These current values may be surprising to you, since the input and output currents are roughly the same with a 16 V input. The cited references show why this happens — the input voltage is applied across the inductor for 50% of the time.

This circuit also has a significant turn-on surge current. I found that the circuit will not turn on if a relatively heavy load is connected and the input voltage is current limited. I discovered this while testing the circuit and it was being driven by one of the positive regulators. There was no problem with it being driven from the rectifier circuit.

Table 2 is a chart showing the efficiency of the negative regulator with various input voltages. As you can easily see, the efficiency is much better with an input voltage of at least 9 V.

The Digital Panel Meter

Another feature of the unit is the digital panel meter (DPM). It can be switched to measure the output voltage (H2 pin 1 to ground) as well as the current drawn (voltage between H2 pins 1 and 2) for each of the positive supplies. Figure 2 shows the circuit I implemented. A three pole, four position rotary switch selects which power supply is being monitored and a three pole, double throw toggle switch selects between voltage and current measurements.

The DPM is a 2000 count unit with a basic range of 200 mV. It does not have a 2 V range. I inserted my own resistors on the DPM board for RA (1 MΩ) and RB (111 kΩ). In order to get a 10:1 voltage ratio the resistor ratio needs to be 9:1. If you have to do this for your DPM, you will want to insure that you maintain the accuracy of the meter. I strongly suggest that you maintain at least a 1 MΩ input resistance so that it does not affect the external voltage divider used for measuring the voltage. I used the calibration potentiometer on the DPM for the final accuracy adjustment. I borrowed a four digit DVM of known accuracy to insure good calibration.

It may be hard to get two resistors with exactly a 9:1 ratio from your "junk box." On the DPM I used two 220 kΩ resistors in parallel to get the required 111 kΩ resistance. By measuring several 220 kΩ resistors I was able to find a combination that was quite close to 111 kΩ. For the voltage measuring divider you can do the same thing using a 10 kΩ input resistor and then a 1 kΩ and 110 Ω in series for the "low" side of the divider. The parts list specifies 1% resistors, in case you don't want to combine resistors as I did.

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 (which could require another power supply anyway). For this system I selected the isolated power supply implementation. I used a series regulator because they are somewhat easier to implement and because the DPM has a very low current requirement. Rather than build another printed circuit board I decided to mount all the components, except the transformer, on a breadboard.

The DPM also has a set of jumpers that allow selection of the decimal point location. As can be seen on Figure 2, I use one pole of the toggle switch to select those inputs.

Some Construction Hints

On the DPM and the regulator boards, I used pin headers for all of the connections that come off the boards. (see the parts list

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Dual Output Power Supply

Need a positive and negative power supply for your next op amp or other project? Here is the answer.

The unique circuit shown in Figure 1 provides very nearly equal positive and negative dc voltages with a common ground from an ac power transformer. Plus and minus outputs from an untapped transformer secondary normally require plus and minus half wave rectifiers, but unless the loads are balanced there will be a net field in the core, which might cause it to saturate. This double bridge design prevents that from happening. D1 to D8 form two bridge rectifiers. All capacitors are electrolytic. If C2 and C3 are considerably larger than C4, then V1 and V2 will be very close to the same at any load.

This circuit can be used for op amps and many other ICs, such as comparators and function generators. It was designed by Robert Dehoney, of IEEE, and verified by extensive testing and measurements of test-bench prototypes.

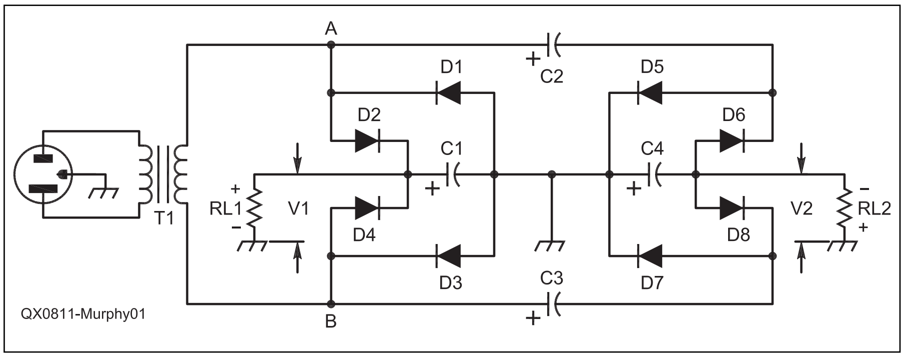


Figure 1 — This schematic diagram shows the double bridge rectifier circuit.

How it Works

In Figure 2A, when point A is more positive than point B by a large enough margin, C1 charges through D2 and D3, C2 charges through D5 and D3, and C4 charges through C2, D5, D8 and C3. During this half cycle, voltage V_{C4} follows V_{C3} since $V_{C4} = V_{D3} + V_{C3} - V_{D8}$.

In Figure 2B, when point B is enough more positive than point A, C1 charges through D1 and D4, C3 charges through D7 and D1, and C4 charges through C3, D7, D6 and C2. During this half cycle, voltage V_{C4} follows V_{C2} since $V_{C4} = V_{D1} + V_{C2} - V_{D6}$.

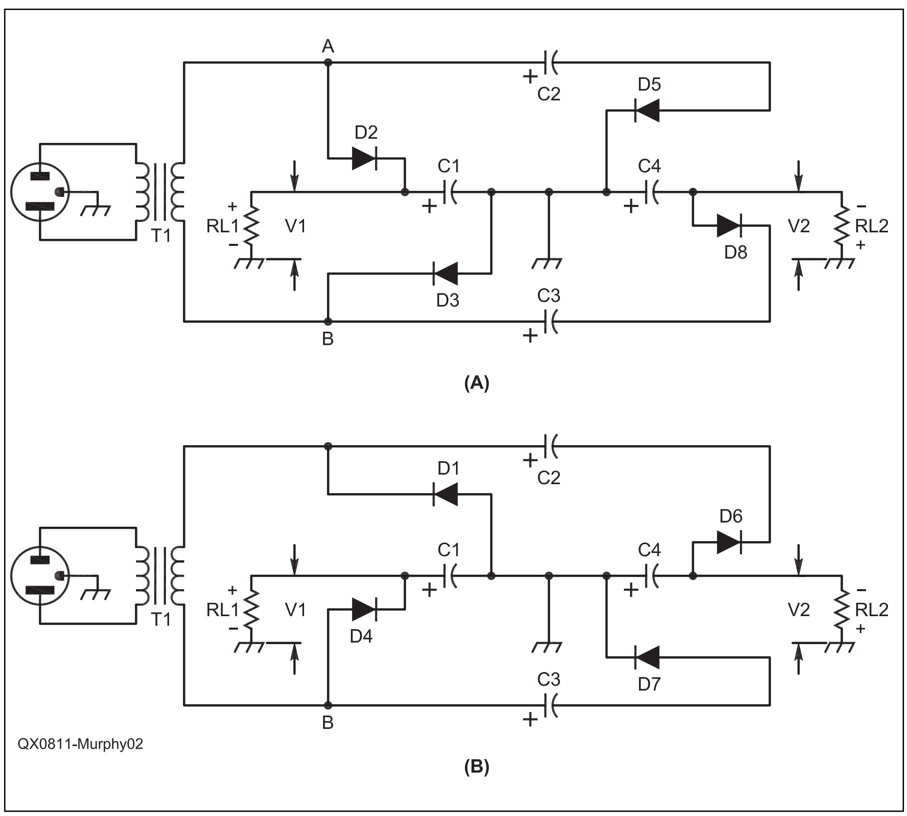


Figure 2 — Part A shows the current path through the circuit when the top of the transformer is positive. Part B shows the current path when the bottom of the transformer is positive.

Design a Dual Output Power Supply

For computerized design using these equations, download *HamCalc* (version 103 or later) and run the "Power Supply — Dual Output" program.[1] If you prefer to do the math yourself, proceed as follows:

[1]George Murphy, VE3ERP, *HamCalc* "Painless Math for Radio Amateurs." This free software is available for download at **www.cq-amateur-radio.com**.

a) Specify desired voltage (E) to RL1 and RL2 and current I1 and I2 through each, where:
E is in volts
RL1 and RL2 are in ohms
I1 and I2 are in amps.

b) Calculate the values of RL1 and R2:

$$RL1 = E / I1 \qquad \text{[Eq 1]}$$

$$RL2 = E / I2 \qquad \text{[Eq 2]}$$

where:
RL1 and RL2 are in ohms
E is in volts dc
I1 and I2 are in amps.

d) Specify the ac line frequency (F) in hertz.

e) Specify allowable peak-to-peak ripple voltage of V1 (across R1) and V2 (across R2).

f) Calculate the values of C1 and C4:

$$C1 = 375 \times I1 \times 10^3 / F / R1 \qquad \text{[Eq 4]}$$

$$C4 = 375 \times I2 \times 10^3 / F / R2 \qquad \text{[Eq 5]}$$

where:
C1 and C4 are in µF
I1 and I2 are in amps
F is in hertz
R1 and R2 are peak-to-peak volts.

g) Calculate the values of C2 and C3:

$$C2 = C3 = 3 \times C4 \text{ (minimum) or } 5 \times C4 \text{ (recommended)} \qquad \text{[Eq 6]}$$

where:
C2 and C3 are in µF.

h) Estimate the required transformer secondary voltage (TE) and current (TI):

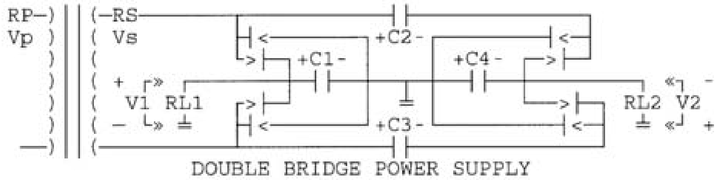
$$TE = 2 + E / 1.41 \qquad \text{[Eq 7]}$$

$$TI = 1.8 \times (I1 + I2) \qquad \text{[Eq 8]}$$

where:
TE is in volts
TI is in amps.

Design Notes

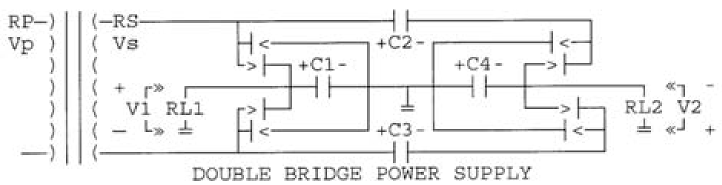
1. Rectifier diodes should have a rating not less than 1.4 × TE volts at TI amperes.
2. All capacitors are electrolytic.
3. C1 can be any value not less than the calculated value of C1.




```
Line....................................   120 v. 60 Hz
Transformer primary resistance............     7.60 Ω
Transformer seconday open circuit voltage..   28.70 volts
Tranformer secondary winding resistance....    0.70 Ω
Positive power supply input capacitor C1...  1470.00 µF
Negative power supply input capacitor C4...  1470.00 µF
Blocking capacitors C2 and C3.............  5000.00 µF
Pos. supply load resistance RL1...........    35.00 Ω
Neg. supply load resistance RL2...........    35.00 Ω
 V1max=  33.457v.  V1min=  29.793v.  V1=  31.635v.  IL1=  0.904 amp
 V2max= -32.261v.  V2min= -28.888v.  V2= -30.498v.  IL2= -0.871 amp
V1 ripple=  3.66v. :   V2 ripple = 3.37v.
Transformer minimum rating  86.4 VA
```

QX0811-Murphy03

Figure 3 — Here is the *HamCalc* printout for prototype number 1.




```
Line....................................   120 v. 60 Hz
Transformer primary resistance............     6.20 Ω
Transformer seconday open circuit voltage..   29.00 volts
Tranformer secondary winding resistance....    0.45 Ω
Positive power supply input capacitor C1...  1470.00 µF
Negative power supply input capacitor C4...  1470.00 µF
Blocking capacitors C2 and C3.............  5000.00 µF
Pos. supply load resistance RL1...........    35.00 Ω
Neg. supply load resistance RL2...........    35.00 Ω
 V1max=  35.326v.  V1min=  31.260v.  V1=  33.258v.  IL1=  0.950 amp
 V2max= -33.863v.  V2min= -30.376v.  V2= -32.117v.  IL2= -0.918 amp
V1 ripple=  4.07v. :   V2 ripple = 3.49v.
Transformer minimum rating  96.0 VA
```

QX0811-Murphy04

Figure 4 — Here is the *HamCalc* printout for prototype number 2.

Table 1
Prototype Bench Test Values

PROTOTYPE	TRANSFORMER - 120 V. 60 Hz					C1 & C4 µF	C2 & C3 µF	CALCULATED VALUES				MEASURED VALUES			
	PRIMARY WINDING OHMS	SECONDARY WINDING OHMS	SECONDARY NO-LOAD VOLTS	LOAD				V1		V2		V1		V2	
				RL1 OHMS	RL2 OHMS			Volts	Ripple volts	Volts	Ripple volts	Volts	Ripple volts	Volts	Ripple volts
# 1	7.60	0.70	28.70	35	35	1470	5000	31.7	3.7	-30.5	3.4	32.2	2.7	-30.9	2.9
# 2	6.20	0.45	29.00					33.3	4.1	-32.1	3.5	32.8	2.5	-31.5	2.6
# 3	40.50	1.00	13.76	150	150	1000		16.4	0.70	-16.0	0.69	16.1	1.7	15.7	1.7

4. C4 can be any value not less than the calculated value of C4.

5. C2 and C3 can be any value not less than 3 times the value selected for C4. (We recommend 5 or more times the value of C4, for minimum ripple.)

6. Select a transformer with a secondary voltage somewhat higher than TE and a current rating not less than TI.

7. When a suitable transformer has been found, you may verify the design and predict the actual V1 and V2 voltages using the *HamCalc* "Power Supply - Double Bridge" program or an equivalent program.

Prototype Bench Test Results

Three prototypes were built and tested using salvaged junk box transformers and capacitors.

Figures 3 and 4 are identical except for the transformers. Unfortunately, there were no capacitors on hand to allow C2 and C3 to be 5 times the value of C4, so the V2 value is less than optimum. Figure 5 indicates a situation in which the capacitor values have a 5:1 ratio, and where V2 is within a few percent of V1.

Table 1 shows the component values of each prototype, the output values predicted by the *HamCalc* program, and the actual measured output values, assuming that the values of all capacitors are exactly as marked on the capacitor which, of course, never happens in the real world. Figure 6 shows how an oscilloscope might show the output in a perfect world.

This design is very flexible, limited only by your tolerance for precision and the contents of your junk box!

George Murphy is a retired industrial designer and professional musician with no vocational RF engineering experience. Licensed as VE3ERP in 1960, his Amateur Radio hobby has led to his writing many articles for international Amateur Radio publications since 1985 and worldwide distribution of his HamCalc software since 1993. George is an ARRL member.

Robert Dehoney is a retired professional electrical engineer who worked for a defense company developing HF, VHF and UHF systems, enjoys analyzing and constructing useful circuits and has authored papers for RF Design Magazine.

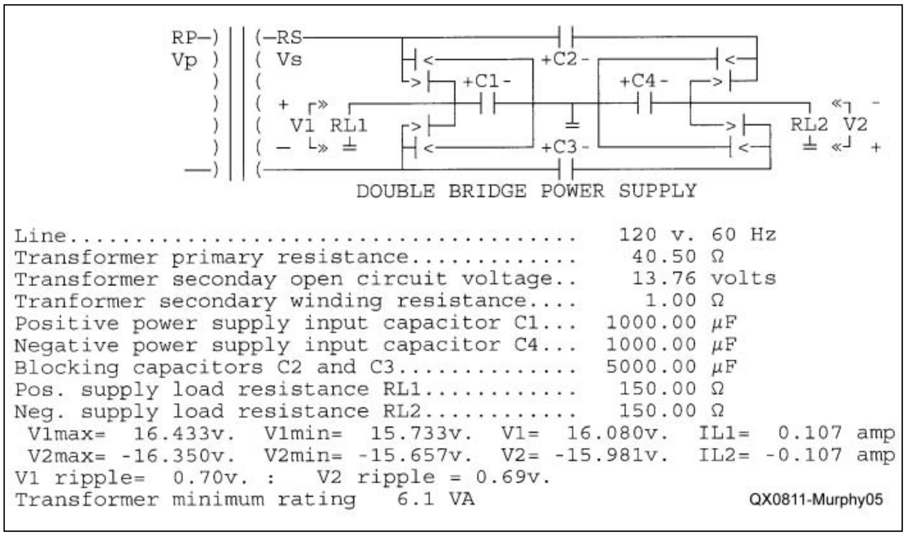


Figure 5 — This *HamCalc* printout is for prototype number 3.

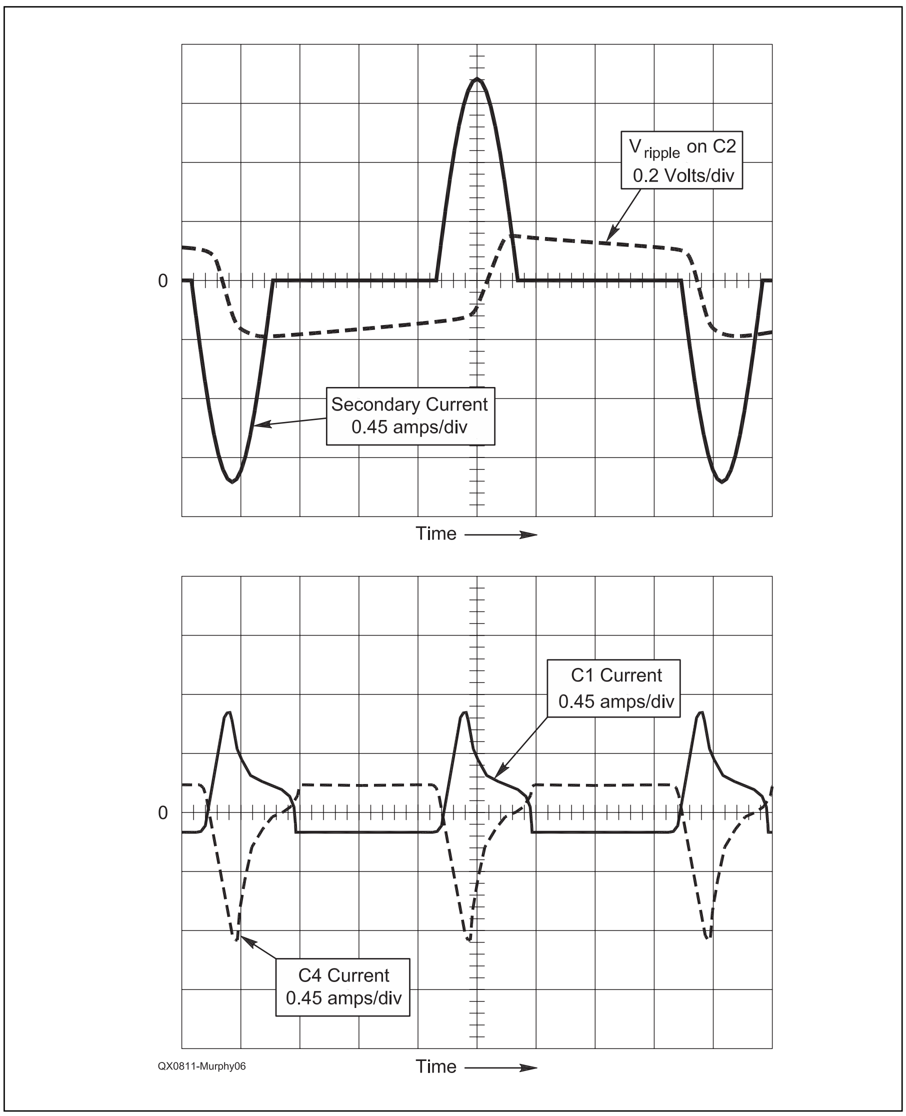


Figure 6 — Part A shows the transformer secondary current and the ripple voltage on C2. During the positive current pulse, the voltage on C2 increases a bit, and during the negative current pulse the voltage decreases. This ripple is riding on the 40 or so volts dc across C2. Part B shows the current into and out of C1 and C4. The positive going pulse is across C1. You can see the capacitor charging for about 25% of the time and discharging for about 75% of the time. In both cases, RL1 = RL2 = 220 Ω.

QEX

Four Output Bench Supply

Every workbench needs a power supply — this one provides four different outputs.

Larry Cicchinelli, K3PTO

This project is a four output bench power supply. Three outputs (positive voltage) use identical switching regulator circuits that can be set to be any voltage between 3.3 and 20 V. Each output is independent of the others and is capable of up to 1 A. The fourth output is via a negative regulator capable of about 250 mA. The unit I built has two fixed outputs and two variable outputs. You can also make any combination of them variable within the above range.

The only dependency among the outputs is that they are all driven by a single transformer. One of the features of a switching regulator is that you can essentially trade off between voltage and current. The transformer I used is rated at 25 V and 2 A. As such it is good for 50 W. Assuming that the regulator IC being used has an efficiency of 75%, you will have a total of about 37 W available from all power supply outputs. In practical terms this means you can get more current from the outputs than what the transformer is supplying — as long as you stay within the 37 W and the maximum current per regulator.

The regulator I used for the positive supplies is the 3.3 V version of the LM2575. If you examine its data sheet you will see that the only difference among the models is the internal voltage divider. This allows you to design a power supply with a higher output voltage by simply inserting a resistance between the output and the FEEDBACK pin. I selected the 3.3 V version mainly due to its cost relative to the others. With it I can get any voltage from 3.3 V to 20 V from the circuit. You could also use the "adjustable" version, which will then allow you to select any voltage between 1.23 and 20 V.

The regulator is specified for up to 37 V output. Since I have specified 50 V capacitors, I believe you should be able to get up to about 30 V output. If you want to output a higher voltage than 30 V, I recommend that you use higher voltage capacitors. The transformer I am using is rated at 25 V; however, I have measured the loaded output at closer to 30 V ac, so I could probably get up to 25 V from the regulators. You will also need to use the 200 V range of the digital panel meter (DPM).

There is also a high voltage version of the LM2575 that can provide outputs of up to 57 V. I recommend that you use capacitors rated to at least 100 V if you decide to use that version.

A Little Theory

Switching regulators come in essentially three varieties: buck, boost and buck-boost. The positive regulator in this article is of the buck type — the output voltage is less than the input voltage. The main feature of a switching regulator that differentiates it from a linear regulator is that the switcher oscillates. They generally use a form of pulse width modulation (PWM) in order to regulate the output voltage. The rise and fall times of the oscillator are quite fast and the harmonics can cause interference to communications receivers. This is the reason a spectrum analyzer is one of the pieces of test equipment used to characterize a switching regulator. This is certainly not the case with a linear regulator!

Two of the best tutorials I have found on switching regulators are *Application Note 2031* on the Maxim-IC Web site at **www.maxim-ic.com/appnotes.cfm/an_pk/2031** and at **www.national.com/appinfo/power/**

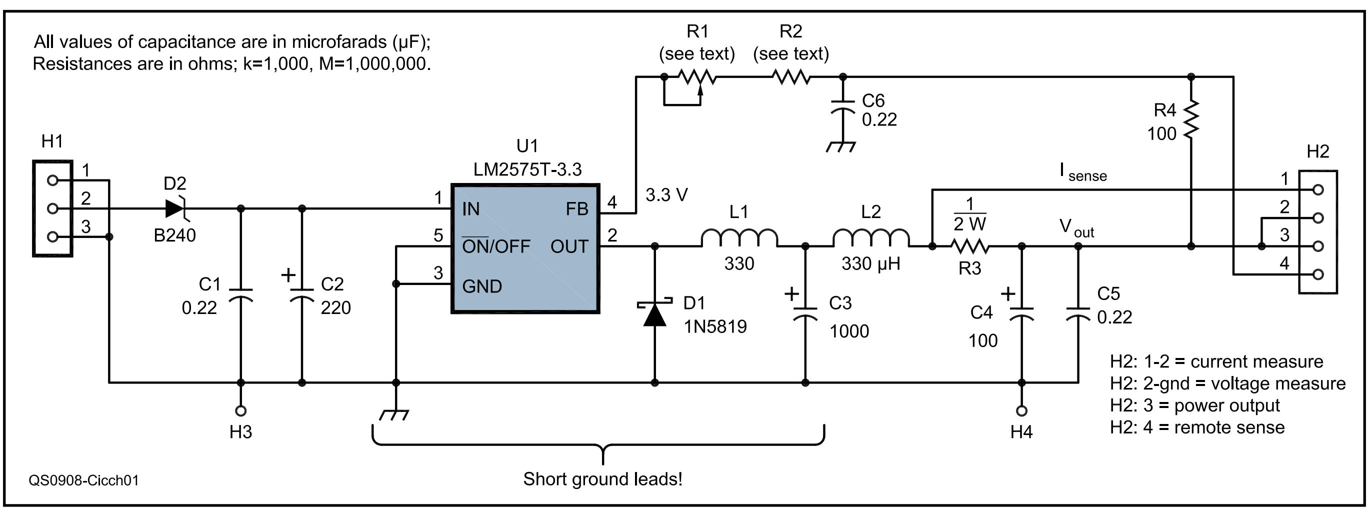


Figure 1 — Schematic diagram of a single positive regulator module.

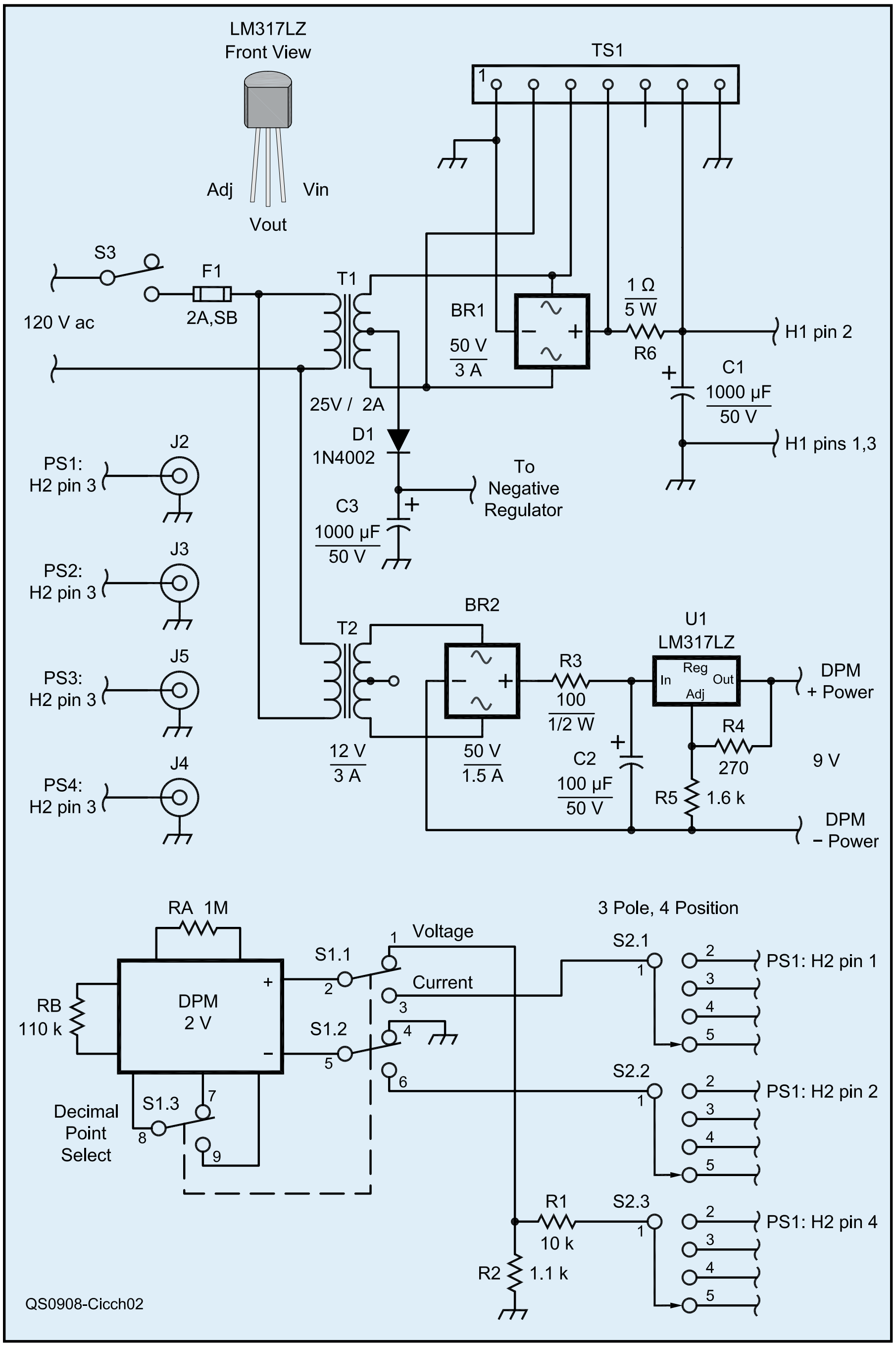


Figure 2 — Chassis schematic showing the interconnection of the modules as well as control and metering details.

files/f5.pdf from National Semiconductor. Rather than try to repeat much of the material in that note, I suggest that you get a copy and read them for yourself.

A switching regulator will have some amount of high frequency noise on its output at the switching frequency, about 52 kHz for the LM2575. In the circuits described here there is a low pass filter on each output that reduces, but does not completely eliminate, this noise. If your requirement is for fixed voltages, you can add a low drop out series regulator (LDO). A good LDO typically requires only about 100 mV between the input and output voltages, so you can design the switcher to be a little higher than the desired voltage and get the benefits of both types of regulators.

Some Circuit Details

Figure 1 is a schematic of a single positive regulator. There are several variations of the circuit which could be implemented. L2 and C4 are optional. These two components provide a low pass filter that will decrease the 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 if the calculated value is nonstandard and you want to use two fixed resistors.

The formula for the output voltage (with the 3.3 V version) can be calculated as follows. The current (in mA) through the internal voltage divider is

$$I = 3.3\ \text{V}/2.7\ \text{k}\Omega = 1.22\ \text{mA}$$

$$R1 + R2 = (V_{OUT} - 3.3) / 1.22\ \text{k}\Omega$$

transposing terms yields:

$$V_{OUT} = [1.22 \times (R1+R2)] + 3.3$$

Note that if you make R1=R2=0, the calculation results in an output of 3.3 V. The leakage current of the error amplifier in the regulator is somewhat less than –25 nA, so it can be ignored. Also, since the 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.

If you want to have an accurate, fixed output voltage, I recommend selecting a value for R2 that is lower than the calculated value. Then select a potentiometer for R1 that yields a reasonable adjustment range.

If you decide to use the extra LC filter, you will have to install L1 and L2 such that their phasing dots line up with the dot symbols on the circuit board. I found out the hard way that if the dots are at the same end of the board, the output will have an additional low frequency ripple. When I built my board I just happened to have three circuits assembled correctly. The fourth one had a serious low frequency ripple that I could not get rid of. I eventually replaced every component, one at a time, to find the bad one. When I replaced L2 the output was okay. It was then I noticed the phasing dots. I reversed L2 just to see what would happen and the ripple came back. There can be inductive coupling, even though there is a ground plane on both sides of the board under the inductors.

Remote Sensing

A feature of many power supplies is that of remote sensing. This is used to electronically adjust for the voltage drop in the wires carrying current to the load. I found that, 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. If you are not going to use remote sensing then you should insert a jumper in place of R4. R4 (100 Ω) is there for protection just in case the remote sense connection is missing.

The schematic shows the connections necessary for remote sensing. It will only work, however, if you run a separate wire from H2-4 to the load. The added accuracy is the result of the load current flowing to the load via H2-3 and essentially no current flowing via H2-4. The connection to H2-2 is still needed in order to measure the voltage drop across the current sense resistor accurately.

If you do not want to use remote sensing you can simplify the switch wiring to use a two pole switch instead of the three pole listed. In this case you would essentially not use S2.2 and connect S1.2 to the common of S2.3 instead of S2.2.

Strictly speaking even this does not fully implement remote sensing. This circuit does not have a mechanism to adjust for the voltage drop in the ground leg. Most high-end commercial supplies will have both power and ground sense inputs. For this power supply make sure that the ground leads have minimal voltage drop. Measurements inside the chassis have indicated this. I have measured about 100 mV drop at 1 A between the positive output of the regulator board and the chassis connector. There was no measurable voltage drop in the ground circuit. You just have to be sure to use relatively heavy wires for the ground connections.

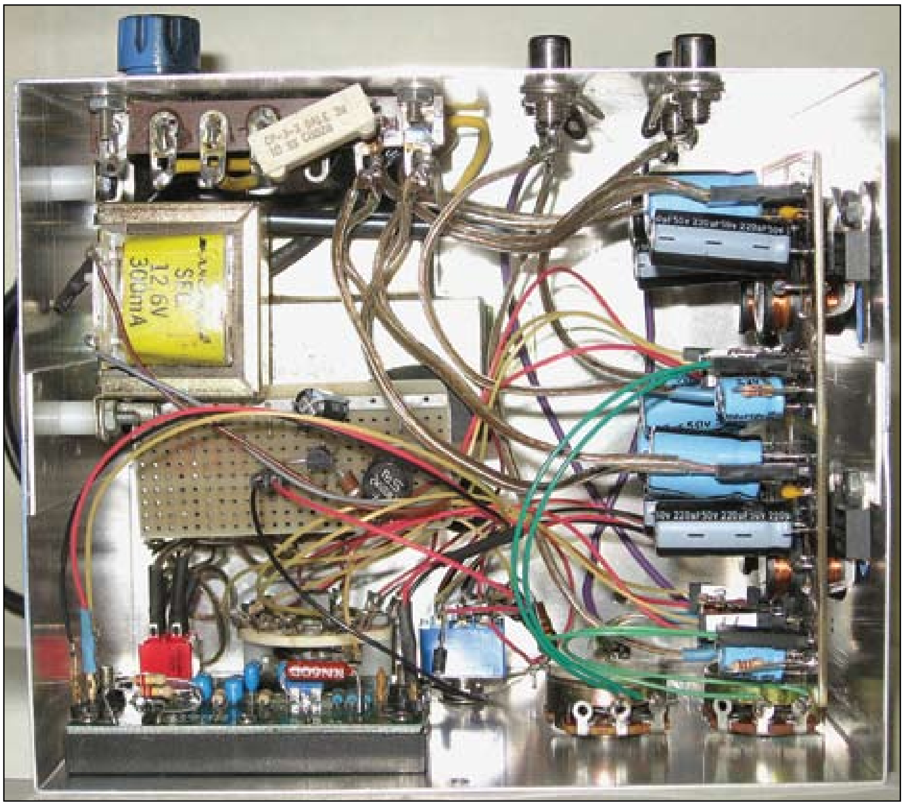


Figure 3 — Underchassis view of the completed power supply. The components used for the main rectifier circuit are mounted on a terminal strip shown at the top left.

Efficiency

Table 1 on the binaries Web version shows the efficiency of the positive regulator with various input voltages. Notice that the efficiency is really good at 14 V; however, the circuit is no longer regulating! Optimum efficiency seems to occur at 20 V but there is not a whole lot of variation between 16 and 28 V.

Rectifier Circuit

Figure 2 shows the connections among the parts of the system — regulator boards, DPM and rectifier circuit. The components used for the main rectifier circuit are mounted on a terminal strip (Mouser 158-1008). You can see the terminal strip and R6 at the top left of Figure 3. You can hardly see it, but C1 is mounted underneath the terminal strip. The leads of the bridge rectifier are soldered into the holes that are used to rivet the terminals to the Bakelite. One of the four leads, the negative output, is soldered to a grounded terminal. Since I have had quite a few of these terminal strips for several years I used fine Emory paper to clean their surfaces as well as a small file to clean the holes. This was done in order to insure good solder connections.

Negative Regulator

The negative regulator is generally similar to the positive regulator. Its description and schematic are on the binaries Web site.

The Digital Panel Meter

Another feature of the unit is the digital panel meter (DPM). It can be switched to measure the output voltage (H2 pin 1 to ground) as well as the current drawn (voltage between H2 pins 1 and 2) for each of the supplies. Figure 2 shows the circuit I implemented. A three pole, four position rotary switch selects which power supply is being monitored and a three pole, double throw toggle switch selects between voltage and current measurements.

The DPM is a 2000 count unit with a basic range of 200 mV. It does not have a 2 V range. I inserted my own resistors on the DPM board for RA (1 MΩ) and RB (111 kΩ). In order to get a 10:1 voltage ratio the resistor ratio needs to be 9:1. If you have to do this for your DPM, you will want to insure that you maintain the accuracy of the meter. I strongly suggest that you maintain at least a 1 MΩ input resistance so that it does not affect the external voltage divider used for measuring the voltage. I used the calibration potentiometer on the DPM for the final accuracy adjustment. I borrowed a four digit DVM of known accuracy to insure good calibration.

It may be hard to get two resistors with exactly a 9:1 ratio from your "junk box." On the DPM I used two 220 kΩ resistors in parallel to get the required 111 kΩ resistance. By measuring several 220 kΩ resistors I was able to find a combination that was quite close to 111 kΩ. For the voltage measuring divider you can do the same thing using a 10 kΩ input resistor and then a 1 kΩ and 110 Ω in series for the "low" side of the divider. The parts list specifies 1% resistors, in case you don't want to combine resistors as I did.

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 (which could require another power supply anyway). For this system I selected the isolated power supply implementation. I used a series regulator because they are somewhat easier to implement and because the DPM has a very low current requirement. Rather than build another printed circuit board I decided to mount all the components, except the transformer, on a breadboard.

The DPM also has a set of jumpers that allow selection of the decimal point location. As can be seen on Figure 2, I use one pole of the toggle switch to select those inputs.

Some Construction Hints

On the DPM and the regulator boards, I used pin headers for all of the connections that come off the boards. (see the parts list for details). This allowed me to assemble the subsystems without having to consider any attached wires. I would then determine the appropriate wire lengths and install the mating connectors on the wires and simply push them onto the pins. This connection method costs about 20 cents per connection.

I have used this method for quite a few

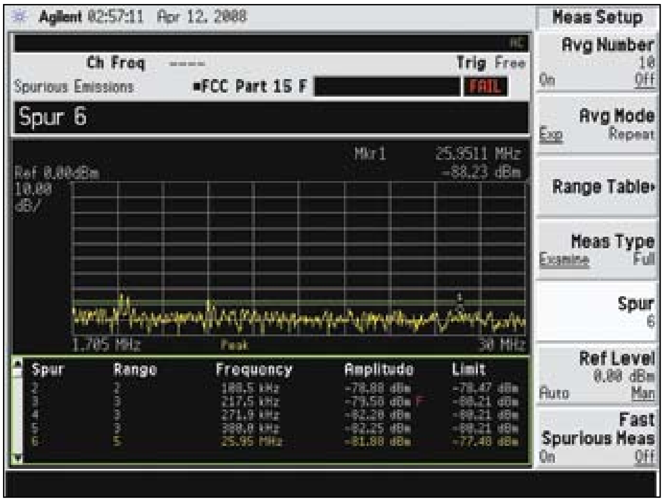


Figure 4 — Measured conducted noise spectrum amplitude throughout the range of 1 to 30 MHz. These show a very quiet output.

projects and have found it to be very useful. It allows me to disconnect all or part of a circuit for debugging as well as for repair. I typically assemble the connectors under a three power magnifying lens and use a pair of 4 inch needle nose pliers for crimping. A crimping tool would make the job easier but they can be expensive.

I have a supply of eight conductor telephone cable that I use for many of my projects. I cut the cable to an appropriate length and then pull the wires out of the sheath. This gives me wires of eight different colors, making them much easier to trace.

The printed circuit board for the regulators contains four identical circuits. At first I thought I might separate the boards for mounting in the enclosure. I decided to keep them as a single board, however. This made it easier to mount the board and it also gave me an idea as to how to heat sink the regulator ICs. I mounted them on the bottom side instead of on the top as I had originally planned. I then folded over the ICs so that the flat side was parallel to the PCB and farthest from the board. I managed to fold the ICs identically so that I was able to use their mounting holes to fasten them to the side of my enclosure. This not only is a convenient method of mounting the board it also gives the ICs a good heatsink and ground. Figure 3 shows the completed assembly bolted to the side of my enclosure.

One problem I encountered with the above method was that of mounting the nuts and bolts required by the ICs. I solved this problem by using a small amount of epoxy to attach the nuts to the "front" side of the regulators. I used a short bolt to hold the nut in place while the epoxy hardened being careful to avoid getting any epoxy on the threads. Figure 5 on the binaries version shows the back side of the PCB assembly with the ICs folded over and the bolts holding the nuts in place while the epoxy hardens.

Another issue I had to solve was the length of the bolts. I had to insure that they were short enough so that they would not touch the PCB once they were used to bolt the ICs to the chassis. Since I did not have bolts of the proper length I had to cut them to length. I used a pair of bolt cutters with threaded holes for the 10-32 bolts. A word of caution here — to ensure that I could still thread a nut onto the bolt after cutting it, I threaded a nut onto each bolt beforehand, so that after I cut it I would have to remove the nut. This method helps to insure that the bolt will still allow a nut to be threaded onto it.

I then measured the distance between the mounting holes of the four regulator ICs and drilled holes in the side of my enclosure accordingly. This method proved relatively easy to implement and made for a very simple mounting procedure. I suggest making a drilling template out of card stock (an index card, for example) to ensure correct spacing, before you drill any holes.

Since the negative regulator was an addition to the "completed" system its installation was different from that of the positive regulators. The circuit was built on a prototyping breadboard and bolted to the bottom of the enclosure. I removed the connections to my regulator #3 and rerouted them to the negative regulator. I have since designed a printed circuit board, but have not replaced the one in my system.

For those who may already have a positive voltage power supply, the negative regulator can be constructed as a separate project and simply connected to your existing supply.

A caution regarding the circuit boards. My source is FAR Circuits, which makes 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.

The QST Binaries Web site has the artwork for the board as well as the Gerbers and a drill file.[1] These can be used to make your own boards if you want. The schematic capture software I use is *DipTrace*. The schematic and PCB files are also on the binaries site.

RFI

I have access to a really good spectrum analyzer at my employers, as well as someone who knows how to use it (thank you, Matt!). 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. I put an 8 Ω, 20 W resistor on the #2 regulator output and adjusted the voltage to +8 V. We then made a series of measurements. The spectrum analyzer has a mode in which it does FCC Part 15 (RFI) tests automatically — very convenient! Even with the antenna so close, the only interference that would have failed the test occurs around 88 MHz). The horizontal green line shows the FCC limit of about –62 dBm. Figure 4 shows conducted noise — the analyzer probe was connected directly to the output through a 0.22 μF capacitor. Figure 4 shows the noise throughout the range of 1 to 30 MHz. This shows a very quiet output.

Parts

The only critical parts are the resistors that form the two voltage dividers. Even their values can be changed, within reason, as long as the ratios are maintained. Although the value of C3 is not very critical, it should be a low ESR type.

The parts list (also on the binaries site) provides the sources from which I obtained my parts. Since I have built quite a few projects over the years I have developed a fairly good supply of components. I have a spreadsheet I keep updated with everything I purchase so that I can use the same parts in new projects. Many of the parts can be obtained from several sources so you may want to do a little shopping around. I try to minimize costs by getting parts from as few sources as possible in order to keep shipping costs down.

Photos by the author.

Larry Cicchinelli, K3PTO, is an ARRL member who has been licensed as K3PTO since 1961. He holds an Advanced class license. Larry earned a BSEE from the Drexel Institute of Technology in 1969 and an MSES from The Pennsylvania State University in 1981. He was employed at Ford Motor Company for 33 years until 2000, responsible for the design and fabrication of test equipment first for ICs and then for automotive electronics. He is now Technical Support Manager for Rabbit Brand at Digi International. He has had articles published in QEX, Circuit Cellar *and* Nuts & Volts *magazines. You can reach Larry at 119 River Run Cir, Sacramento, CA 95833 or at* **k3pto@arrl.net**.

[1]**www.arrl.org/files/qst-binaries/**

National Semiconductor

February 29, 2008

LM2673 SIMPLE SWITCHER® 3A Step-Down Voltage Regulator with Adjustable Current Limit

General Description

The LM2673 series of regulators are monolithic integrated circuits which provide all of the active functions for a step-down (buck) switching regulator capable of driving up to 3A loads with excellent line and load regulation characteristics. High efficiency (>90%) is obtained through the use of a low ON-resistance DMOS power switch. The series consists of fixed output voltages of 3.3V, 5V and 12V and an adjustable output version.

The SIMPLE SWITCHER concept provides for a complete design using a minimum number of external components. A high fixed frequency oscillator (260KHz) allows the use of physically smaller sized components. A family of standard inductors for use with the LM2673 are available from several manufacturers to greatly simplify the design process.

Other features include the ability to reduce the input surge current at power-ON by adding a softstart timing capacitor to gradually turn on the regulator. The LM2673 series also has built in thermal shutdown and resistor programmable current limit of the power MOSFET switch to protect the device and load circuitry under fault conditions. The output voltage is guaranteed to a ±2% tolerance. The clock frequency is controlled to within a ±11% tolerance.

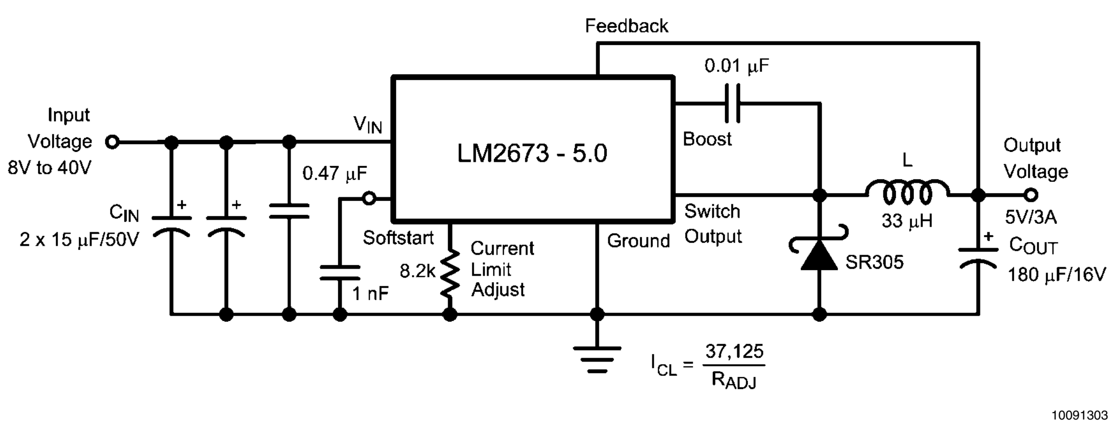
Features

- Efficiency up to 94%
- Simple and easy to design with (using off-the-shelf external components)
- Resistor programmable peak current limit over a range of 2A to 5A.
- 150 mΩ DMOS output switch
- 3.3V, 5V and 12V fixed output and adjustable (1.2V to 37V) versions
- ±2%maximum output tolerance over full line and load conditions
- Wide input voltage range: 8V to 40V
- 260 KHz fixed frequency internal oscillator
- Softstart capability
- −40 to +125°C operating junction temperature range

Applications

- Simple to design, high efficiency (>90%) step-down switching regulators
- Efficient system pre-regulator for linear voltage regulators
- Battery chargers

Typical Application



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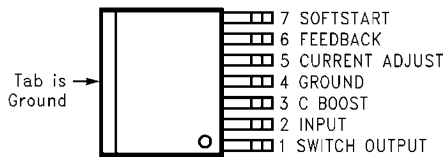
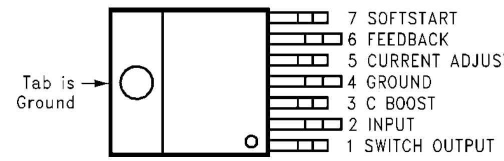
Connection Diagrams and Ordering Information

TO-263 Package
Top View



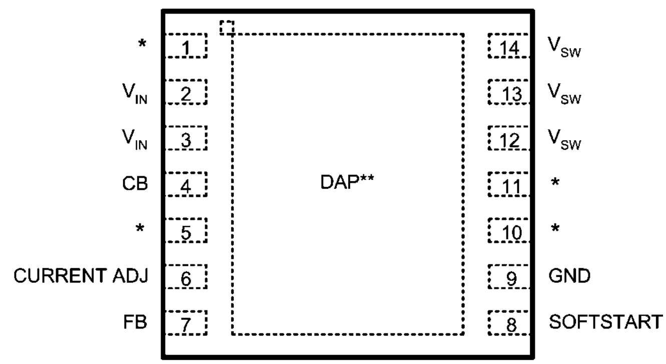
Order Number
LM2673S-3.3, LM2673S-5.0,
LM2673S-12 or LM2673S-ADJ
See NSC Package Number TS7B

TO-220 Package
Top View



Order Number
LM2673T-3.3, LM2673T-5.0,
LM2673T-12 or LM2673T-ADJ
See NSC Package Number TA07B

Top View



*No Connections

**Connect to Pin 9 on PCB

10091335

LLP-14
See NS package Number SRC14A

Ordering Information for LLP Package

Output Voltage	Order Information	Package Marking	Supplied As
12	LM2673SD-12	S0002SB	250 Units on Tape and Reel
12	LM2673SDX-12	S0002SB	2500 Units on Tape and Reel
3.3	LM2673SD-3.3	S0002TB	250 Units on Tape and Reel
3.3	LM2673SDX-3.3	S0002TB	2500 Units on Tape and Reel
5.0	LM2673SD-5.0	S0002UB	250 Units on Tape and Reel
5.0	LM2673SDX-5.0	S0002UB	2500 Units on Tape and Reel
ADJ	LM2673SD-ADJ	S0002VB	250 Units on Tape and Reel
ADJ	LM2673SDX-ADJ	S0002VB	2500 Units on Tape and Reel

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/ Distributors for availability and specifications.

Input Supply Voltage	45V
Softstart Pin Voltage	−0.1V to 6V
Switch Voltage to Ground (Note 13)	−1V to V_{IN}
Boost Pin Voltage	V_{SW} + 8V
Feedback Pin Voltage	−0.3V to 14V
Power Dissipation	Internally Limited
ESD (Note 2)	2 kV
Storage Temperature Range	−65°C to 150°C
Soldering Temperature	
Wave	4 sec, 260°C
Infrared	10 sec, 240°C
Vapor Phase	75 sec, 219°C

Operating Ratings

Supply Voltage	8V to 40V
Junction Temperature Range (T_J)	−40°C to 125°C

Electrical Characteristics

Limits appearing in **bold type face** apply over the entire junction temperature range of operation, −40°C to 125°C. Specifications appearing in normal type apply for $T_A = T_J = 25°C$. R_{ADJ} = 8.2KΩ

LM2673-3.3

Symbol	Parameter	Conditions	Typical (Note 3)	Min (Note 4)	Max (Note 4)	Units
V_{OUT}	Output Voltage	V_{IN} = 8V to 40V, 100mA ≤ I_{OUT} ≤ 3A	3.3	3.234/**3.201**	3.366/**3.399**	V
η	Efficiency	V_{IN} = 12V, I_{LOAD} = 3A	86			%

LM2673-5.0

Symbol	Parameter	Conditions	Typical (Note 3)	Min (Note 4)	Max (Note 4)	Units
V_{OUT}	Output Voltage	V_{IN} = 8V to 40V, 100mA ≤ I_{OUT} ≤ 3A	5.0	4.900/**4.850**	5.100/**5.150**	V
η	Efficiency	V_{IN} = 12V, I_{LOAD} = 3A	88			%

LM2673-12

Symbol	Parameter	Conditions	Typical (Note 3)	Min (Note 4)	Max (Note 4)	Units
V_{OUT}	Output Voltage	V_{IN} = 15V to 40V, 100mA ≤ I_{OUT} ≤ 3A	12	11.76/**11.64**	12.24/**12.36**	V
η	Efficiency	V_{IN} = 24V, I_{LOAD} = 3A	94			%

LM2673-ADJ

Symbol	Parameter	Conditions	Typ (Note 3)	Min (Note 4)	Max (Note 4)	Units
V_{FB}	Feedback Voltage	V_{IN} = 8V to 40V, 100mA ≤ I_{OUT} ≤ 3A V_{OUT} Programmed for 5V	1.21	1.186/**1.174**	1.234/**1.246**	V
η	Efficiency	V_{IN} = 12V, I_{LOAD} = 3A	88			%

All Output Voltage Versions Electrical Characteristics

Limits appearing in **bold type face** apply over the entire junction temperature range of operation, −40°C to 125°C.

Specifications appearing in normal type apply for $T_A = T_J = 25°C$. Unless otherwise specified, $R_{ADJ} = 8.2K\Omega$, V_{IN}=12V for the 3.3V, 5V and Adjustable versions and V_{IN}=24V for the 12V version.

Symbol	Parameter	Conditions	Typ	Min	Max	Units
DEVICE PARAMETERS						
I_Q	Quiescent Current	$V_{FEEDBACK}$ = 8V For 3.3V, 5.0V, and ADJ Versions $V_{FEEDBACK}$ = 15V For 12V Versions	4.2		6	mA
V_{ADJ}	Current Limit Adjust Voltage		1.21	1.181/**1.169**	1.229/**1.246**	V
I_{CL}	Current Limit	R_{ADJ} = 8.2KΩ, (Note 5)	4.5	3.8/**3.6**	5.25/**5.4**	A
I_L	Output Leakage Current	V_{IN} = 40V, Softstart Pin = 0V V_{SWITCH} = 0V V_{SWITCH} = −1V	1.0 6		1.5 15	mA mA
$R_{DS(ON)}$	Switch On-Resistance	I_{SWITCH} = 3A	0.15		0.17/**0.29**	Ω
f_O	Oscillator Frequency	Measured at Switch Pin	260	**225**	**280**	kHz
D	Duty Cycle	Maximum Duty Cycle Minimum Duty Cycle	91 0			% %
I_{BIAS}	Feedback Bias Current	$V_{FEEDBACK}$ = 1.3V ADJ Version Only	85			nA
V_{SFST}	Softstart Threshold Voltage		0.63	**0.53**	**0.74**	V
I_{SFST}	Softstart Pin Current	Softstart Pin = 0V	3.7		**6.9**	μA
θ_{JA}	Thermal Resistance	T Package, Junction to Ambient (Note 6)	65			
θ_{JA}		T Package, Junction to Ambient (Note 7)	45			
θ_{JC}		T Package, Junction to Case	2			
θ_{JA}		S Package, Junction to Ambient (Note 8)	56			°C/W
θ_{JA}		S Package, Junction to Ambient (Note 9)	35			
θ_{JA}		S Package, Junction to Ambient (Note 10)	26			
θ_{JC}		S Package, Junction to Case	2			++
θ_{JA}		SD Package, Junction to Ambient (Note 11)	55			°C/W
θ_{JA}		SD Package, Junction to Ambient (Note 12)	29			

Note 1: Absolute Maximum Ratings are limits beyond which damage to the device may occur. Operating Ratings indicate conditions under which of the device is guaranteed. Operating Ratings do not imply guaranteed performance limits. For guaranteed performance limits and associated test condition, see the electrical Characteristics tables.

Note 2: ESD was applied using the human-body model, a 100pF capacitor discharged through a 1.5 kΩ resistor into each pin.

Note 3: Typical values are determined with $T_A = T_J = 25°C$ and represent the most likely norm.

Note 4: All limits are guaranteed at room temperature (standard type face) and at **temperature extremes (bold type face)**. All room temperature limits are 100% tested during production with $T_A = T_J = 25°C$. All limits at temperature extremes are guaranteed via correlation using standard standard Quality Control (SQC) methods. All limits are used to calculate Average Outgoing Quality Level (AOQL).

Note 5: The peak switch current limit is determined by the following relationship: I_{CL}=37,125/ R_{ADJ}.

Note 6: Junction to ambient thermal resistance (no external heat sink) for the 7 lead TO-220 package mounted vertically, with ½ inch leads in a socket, or on a PC board with minimum copper area.

Note 7: Junction to ambient thermal resistance (no external heat sink) for the 7 lead TO-220 package mounted vertically, with ½ inch leads soldered to a PC board containing approximately 4 square inches of (1 oz.) copper area surrounding the leads.

Note 8: Junction to ambient thermal resistance for the 7 lead TO-263 mounted horizontally against a PC board area of 0.136 square inches (the same size as the TO-263 package) of 1 oz. (0.0014 in. thick) copper.

Note 9: Junction to ambient thermal resistance for the 7 lead TO-263 mounted horizontally against a PC board area of 0.4896 square inches (3.6 times the area of the TO-263 package) of 1 oz. (0.0014 in. thick) copper.

Note 10: Junction to ambient thermal resistance for the 7 lead TO-263 mounted horizontally against a PC board copper area of 1.0064 square inches (7.4 times the area of the TO-263 package) of 1 oz. (0.0014 in. thick) copper. Additional copper area will reduce thermal resistance further. See the thermal model in Switchers Made Simple® software.

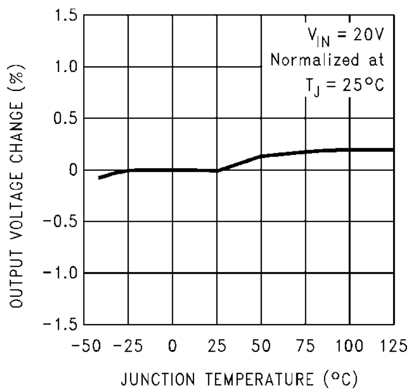
Note 11: Junction to ambient thermal resistance for the 14-lead LLP mounted on a PC board copper area equal to the die attach paddle.

Note 12: Junction to ambient thermal resistance for the 14-lead LLP mounted on a PC board copper area using 12 vias to a second layer of copper equal to die attach paddle. Additional copper area will reduce thermal resistance further. For layout recommendations, refer to Application Note AN-1187.

Note 13: The absolute maximum specification of the 'Switch Voltage to Ground' applies to DC voltage. An extended negative voltage limit of -8V applies to a pulse of up to 20 ns, -6V of 60 ns and -3V of up to 100 ns.

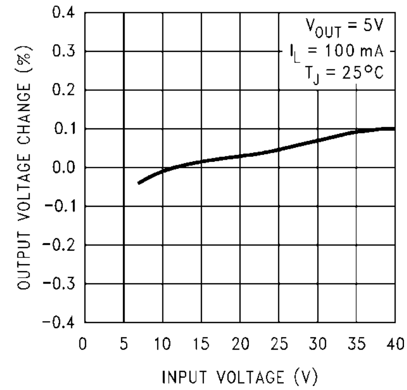
Typical Performance Characteristics

Normalized Output Voltage



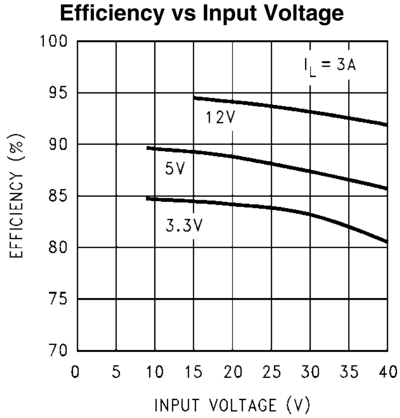
10091304

Line Regulation



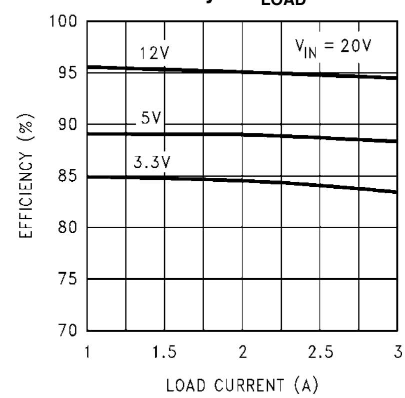
10091305

Efficiency vs Input Voltage



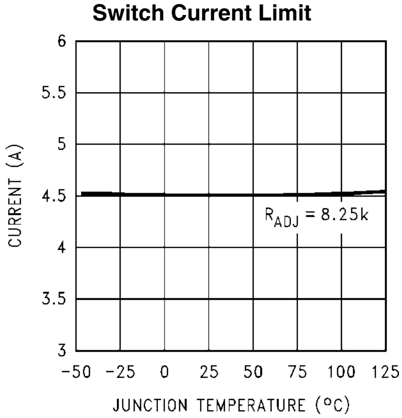
10091306

Efficiency vs I_{LOAD}



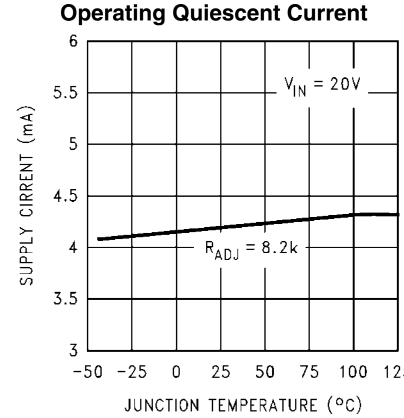
10091307

Switch Current Limit



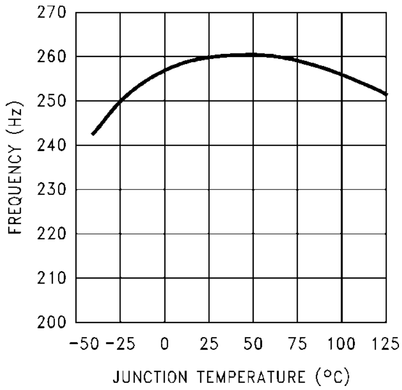
10091308

Operating Quiescent Current



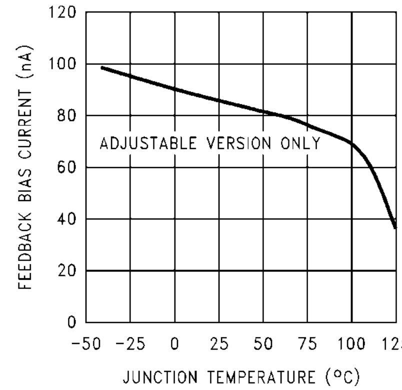
10091309

Switching Frequency



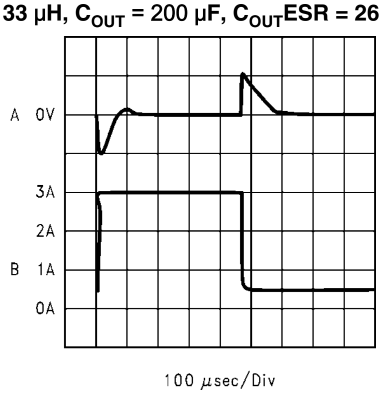
10091312

Feedback Pin Bias Current



10091313

Load Transient Response for Continuous Mode
V_{IN} = 20V, V_{OUT} = 5V
L = 33 µH, C_{OUT} = 200 µF, C_{OUT}ESR = 26 mΩ



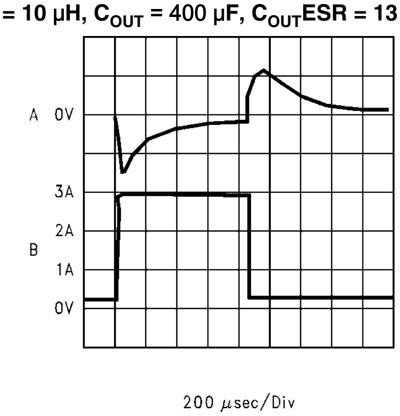
10091317

A: Output Voltage, 100 mV//div, AC-Coupled.

B: Load Current: 500 mA to 3A Load Pulse

Horizontal Time Base: 100 µs/div

Load Transient Response for Discontinuous Mode
V_{IN} = 20V, V_{OUT} = 5V,
L = 10 µH, C_{OUT} = 400 µF, C_{OUT}ESR = 13 mΩ



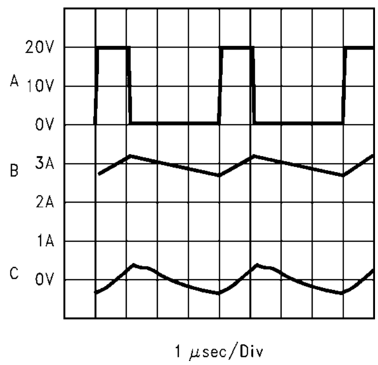
10091318

A: Output Voltage, 100 mV/div, AC-Coupled.

B: Load Current: 200 mA to 3A Load Pulse

Horizontal Time Base: 200 µs/div

Continuous Mode Switching Waveforms
V_{IN} = 20V, V_{OUT} = 5V, I_{LOAD} = 3A
L = 33 µH, C_{OUT} = 200 µF, C_{OUT}ESR = 26 mΩ

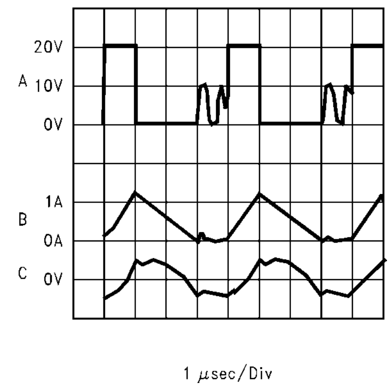


10091315

A: V_{SW} Pin Voltage, 10 V/div.
B: Inductor Current, 1 A/div
C: Output Ripple Voltage, 20 mV/div AC-Coupled

Horizontal Time Base: 1 µs/div

Discontinuous Mode Switching Waveforms
V_{IN} = 20V, V_{OUT} = 5V, I_{LOAD} = 500 mA
L = 10 µH, C_{OUT} = 400 µF, C_{OUT}ESR = 13 mΩ

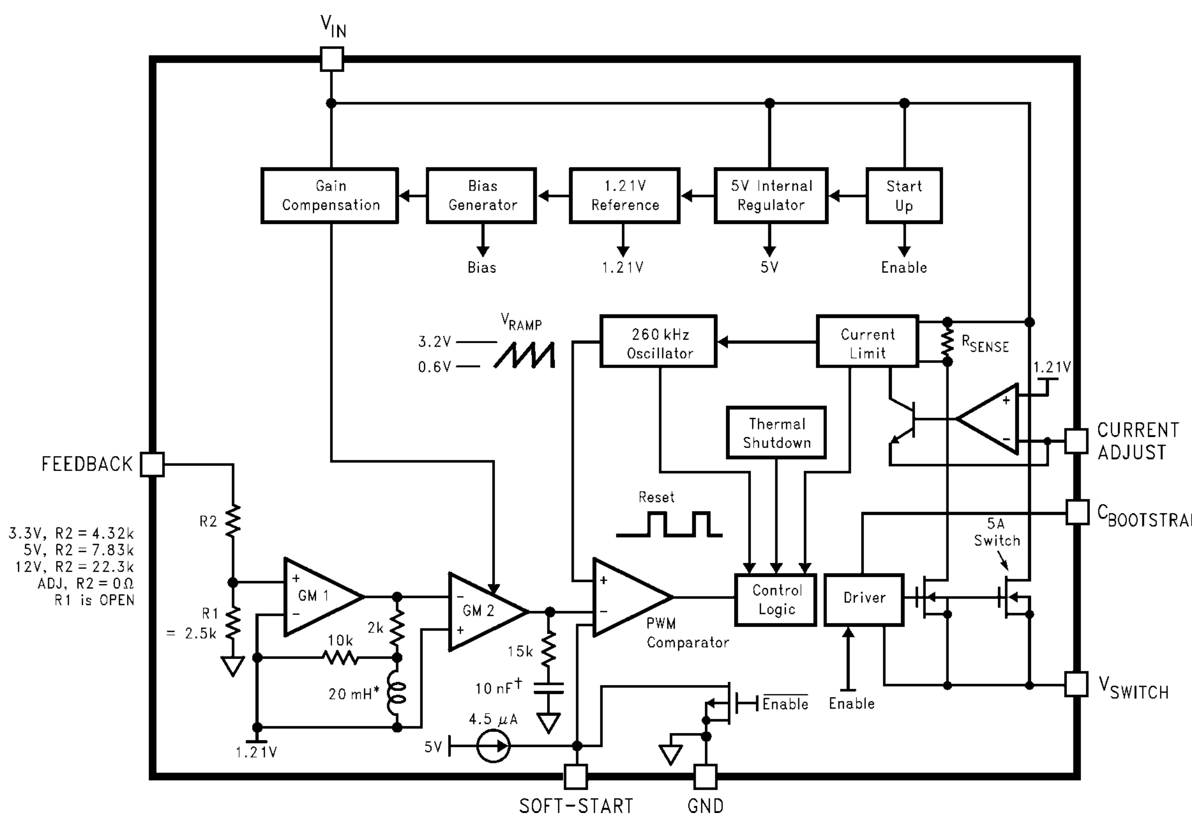


10091316

A: V_{SW} Pin Voltage, 10 V/div.
B: Inductor Current, 1 A/div
C: Output Ripple Voltage, 20 mV/div AC-Coupled

Horizontal Time Base: 1 µs//iv

Block Diagram



10091314

* Active Inductor Patent Number 5,514,947
† Active Capacitor Patent Number 5,382,918

Application Hints

The LM2673 provides all of the active functions required for a step-down (buck) switching regulator. The internal power switch is a DMOS power MOSFET to provide power supply designs with high current capability, up to 3A, and highly efficient operation.

The LM2673 is part of the SIMPLE SWITCHER family of power converters. A complete design uses a minimum number of external components, which have been pre-determined from a variety of manufacturers. Using either this data sheet or a design software program called ***LM267X Made Simple*** (version 2.0) a complete switching power supply can be designed quickly. The software is provided free of charge and can be downloaded from National Semiconductor's Internet site located at http://www.national.com.

SWITCH OUTPUT

This is the output of a power MOSFET switch connected directly to the input voltage. The switch provides energy to an inductor, an output capacitor and the load circuitry under control of an internal pulse-width-modulator (PWM). The PWM controller is internally clocked by a fixed 260KHz oscillator. In a standard step-down application the duty cycle (Time ON/ Time OFF) of the power switch is proportional to the ratio of the power supply output voltage to the input voltage. The voltage on pin 1 switches between Vin (switch ON) and below ground by the voltage drop of the external Schottky diode (switch OFF).

INPUT

The input voltage for the power supply is connected to pin 2. In addition to providing energy to the load the input voltage also provides bias for the internal circuitry of the LM2673. For guaranteed performance the input voltage must be in the range of 8V to 40V. For best performance of the power supply the input pin should always be bypassed with an input capacitor located close to pin 2.

C BOOST

A capacitor must be connected from pin 3 to the switch output, pin 1. This capacitor boosts the gate drive to the internal MOSFET above Vin to fully turn it ON. This minimizes conduction losses in the power switch to maintain high efficiency. The recommended value for C Boost is 0.01µF.

GROUND

This is the ground reference connection for all components in the power supply. In fast-switching, high-current applications such as those implemented with the LM2673, it is recommended that a broad ground plane be used to minimize signal coupling throughout the circuit

CURRENT ADJUST

A key feature of the LM2673 is the ability to tailor the peak switch current limit to a level required by a particular application. This alleviates the need to use external components that must be physically sized to accommodate current levels (under shorted output conditions for example) that may be much higher than the normal circuit operating current requirements.

A resistor connected from pin 5 to ground establishes a current ($I_{(pin\ 5)} = 1.2V / R_{ADJ}$) that sets the peak current through the power switch. The maximum switch current is fixed at a level of 37,125 / R_{ADJ}.

FEEDBACK

This is the input to a two-stage high gain amplifier, which drives the PWM controller. It is necessary to connect pin 6 to the actual output of the power supply to set the dc output voltage. For the fixed output devices (3.3V, 5V and 12V outputs), a direct wire connection to the output is all that is required as internal gain setting resistors are provided inside the LM2673. For the adjustable output version two external resistors are required to set the dc output voltage. For stable operation of the power supply it is important to prevent coupling of any inductor flux to the feedback input.

SOFTSTART

A capacitor connected from pin 7 to ground allows for a slow turn-on of the switching regulator. The capacitor sets a time delay to gradually increase the duty cycle of the internal power switch. This can significantly reduce the amount of surge current required from the input supply during an abrupt application of the input voltage. If softstart is not required this pin should be left open circuited. Please see the C_{SS} softstart capacitor section for further information regarding softstart capacitor values.

DAP (LLP PACKAGE)

The Die Attach Pad (DAP) can and should be connected to PCB Ground plane/island. For CAD and assembly guidelines refer to Application Note AN-1187 at http://power.national.com.

DESIGN CONSIDERATIONS

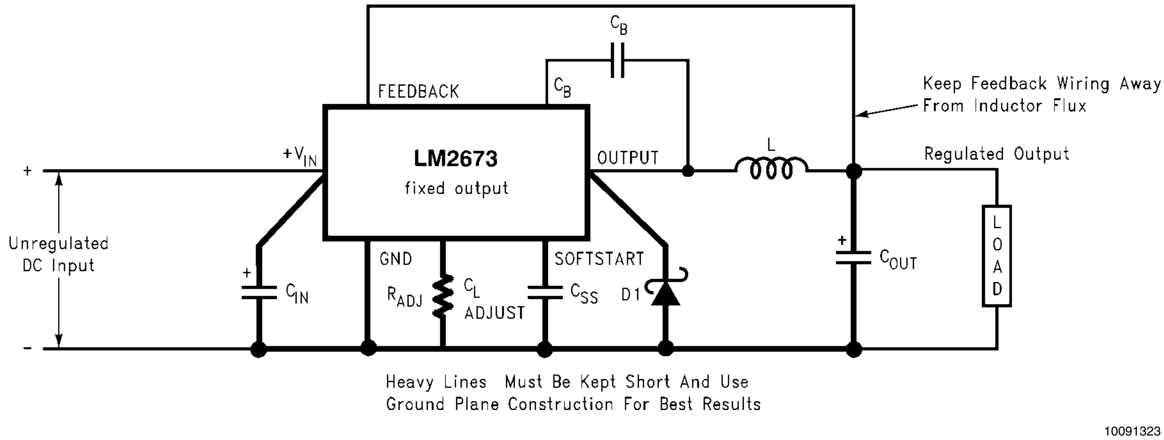


FIGURE 1. Basic circuit for fixed output voltage applications.

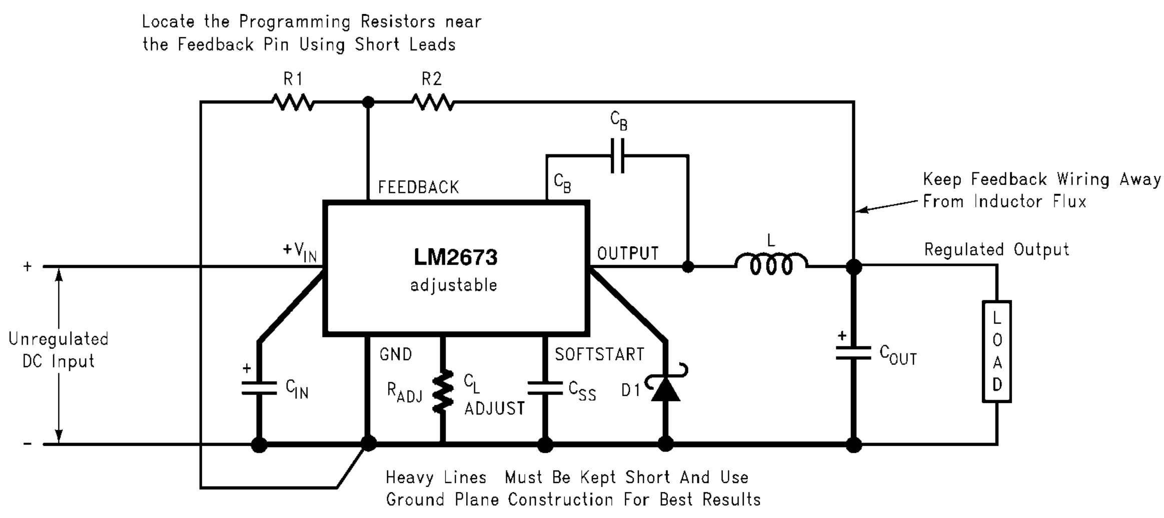


FIGURE 2. Basic circuit for adjustable output voltage applications

Power supply design using the LM2673 is greatly simplified by using recommended external components. A wide range of inductors, capacitors and Schottky diodes from several manufacturers have been evaluated for use in designs that cover the full range of capabilities (input voltage, output voltage and load current) of the LM2673. A simple design procedure using nomographs and component tables provided in this data sheet leads to a working design with very little effort. Alternatively, the design software, ***LM267X Made Simple*** (version 6.0), can also be used to provide instant component selection, circuit performance calculations for evaluation, a bill of materials component list and a circuit schematic.

The individual components from the various manufacturers called out for use are still just a small sample of the vast array of components available in the industry. While these components are recommended, they are not exclusively the only components for use in a design. After a close comparison of component specifications, equivalent devices from other manufacturers could be substituted for use in an application.

Important considerations for each external component and an explanation of how the nomographs and selection tables were developed follows.

INDUCTOR

The inductor is the key component in a switching regulator. For efficiency the inductor stores energy during the switch ON time and then transfers energy to the load while the switch is OFF.

Nomographs are used to select the inductance value required for a given set of operating conditions. The nomographs assume that the circuit is operating in continuous mode (the current flowing through the inductor never falls to zero). The magnitude of inductance is selected to maintain a maximum ripple current of 30% of the maximum load current. If the ripple current exceeds this 30% limit the next larger value is selected.

The inductors offered have been specifically manufactured to provide proper operation under all operating conditions of input and output voltage and load current. Several part types are offered for a given amount of inductance. Both surface mount and through-hole devices are available. The inductors from each of the three manufacturers have unique characteristics.

Renco: ferrite stick core inductors; benefits are typically lowest cost and can withstand ripple and transient peak currents above the rated value. These inductors have an external magnetic field, which may generate EMI.

Pulse Engineering: powdered iron toroid core inductors; these also can withstand higher than rated currents and, being toroid inductors, will have low EMI.

Coilcraft: ferrite drum core inductors; these are the smallest physical size inductors and are available only as surface mount components. These inductors also generate EMI but less than stick inductors.

OUTPUT CAPACITOR

The output capacitor acts to smooth the dc output voltage and also provides energy storage. Selection of an output capacitor, with an associated equivalent series resistance (ESR), impacts both the amount of output ripple voltage and stability of the control loop.

The output ripple voltage of the power supply is the product of the capacitor ESR and the inductor ripple current. The capacitor types recommended in the tables were selected for having low ESR ratings.

In addition, both surface mount tantalum capacitors and through-hole aluminum electrolytic capacitors are offered as solutions.

Impacting frequency stability of the overall control loop, the output capacitance, in conjunction with the inductor, creates a double pole inside the feedback loop. In addition the capacitance and the ESR value create a zero. These frequency response effects together with the internal frequency compensation circuitry of the LM2673 modify the gain and phase shift of the closed loop system.

As a general rule for stable switching regulator circuits it is desired to have the unity gain bandwidth of the circuit to be limited to no more than one-sixth of the controller switching frequency. With the fixed 260KHz switching frequency of the LM2673, the output capacitor is selected to provide a unity gain bandwidth of 40KHz maximum. Each recommended capacitor value has been chosen to achieve this result.

In some cases multiple capacitors are required either to reduce the ESR of the output capacitor, to minimize output ripple (a ripple voltage of 1% of Vout or less is the assumed performance condition), or to increase the output capacitance to reduce the closed loop unity gain bandwidth (to less than 40KHz). When parallel combinations of capacitors are required it has been assumed that each capacitor is the exact same part type.

The RMS current and working voltage (WV) ratings of the output capacitor are also important considerations. In a typical step-down switching regulator, the inductor ripple current (set to be no more than 30% of the maximum load current by the inductor selection) is the current that flows through the output capacitor. The capacitor RMS current rating must be greater than this ripple current. The voltage rating of the output capacitor should be greater than 1.3 times the maximum output voltage of the power supply. If operation of the system at elevated temperatures is required, the capacitor voltage rating may be de-rated to less than the nominal room temperature rating. Careful inspection of the manufacturer's specification for de-rating of working voltage with temperature is important.

INPUT CAPACITOR

Fast changing currents in high current switching regulators place a significant dynamic load on the unregulated power source. An input capacitor helps to provide additional current to the power supply as well as smooth out input voltage variations.

Like the output capacitor, the key specifications for the input capacitor are RMS current rating and working voltage. The RMS current flowing through the input capacitor is equal to one-half of the maximum dc load current so the capacitor should be rated to handle this. Paralleling multiple capacitors proportionally increases the current rating of the total capacitance. The voltage rating should also be selected to be 1.3 times the maximum input voltage. Depending on the unregulated input power source, under light load conditions the maximum input voltage could be significantly higher than normal operation and should be considered when selecting an input capacitor.

The input capacitor should be placed very close to the input pin of the LM2673. Due to relative high current operation with fast transient changes, the series inductance of input connecting wires or PCB traces can create ringing signals at the input terminal which could possibly propagate to the output or other parts of the circuitry. It may be necessary in some designs to add a small valued (0.1µF to 0.47µF) ceramic type capacitor in parallel with the input capacitor to prevent or minimize any ringing.

CATCH DIODE

When the power switch in the LM2673 turns OFF, the current through the inductor continues to flow. The path for this current is through the diode connected between the switch output and ground. This forward biased diode clamps the switch output to a voltage less than ground. This negative voltage must be greater than −1V so a low voltage drop (particularly at high current levels) Schottky diode is recommended. Total efficiency of the entire power supply is significantly impacted by the power lost in the output catch diode. The average current through the catch diode is dependent on the switch duty cycle (D) and is equal to the load current times (1-D). Use of a diode rated for much higher current than is required by the actual application helps to minimize the voltage drop and power loss in the diode.

During the switch ON time the diode will be reversed biased by the input voltage. The reverse voltage rating of the diode should be at least 1.3 times greater than the maximum input voltage.

BOOST CAPACITOR

The boost capacitor creates a voltage used to overdrive the gate of the internal power MOSFET. This improves efficiency by minimizing the on resistance of the switch and associated power loss. For all applications it is recommended to use a 0.01µF/50V ceramic capacitor.

R_{ADJ}, ADJUSTABLE CURRENT LIMIT

A key feature of the LM2673 is the ability to control the peak switch current. Without this feature the peak switch current would be internally set to 5A or higher to accommodate 3A load current designs. This requires that both the inductor (which could saturate with excessively high currents) and the catch diode be able to safely handle up to 5A which would be conducted under load fault conditions.

If an application only requires a load current of 2A or so the peak switch current can be set to a limit just over the maximum load current with the addition of a single programming resistor. This allows the use of less powerful and more cost effective inductors and diodes.

The peak switch current is equal to a factor of 37,125 divided by R_{ADJ}. A resistance of 8.2KΩ sets the current limit to typically 4.5A. For predictable control of the current limit it is recommended to keep the peak switch current greater than 1A. For lower current applications 500mA and 1A switching regulators, the LM2674 and LM2672, are available.

When the power switch reaches the current limit threshold it is immediately turned OFF and the internal switching frequency is reduced. This extends the OFF time of the switch to prevent a steady state high current condition. As the switch current falls below the current limit threshold, the switch will turn back ON. If a load fault continues, the switch will again exceed the threshold and switch back OFF. This will result in a low duty cycle pulsing of the power switch to minimize the overall fault condition power dissipation.

Css SOFTSTART CAPACITOR

This optional capacitor controls the rate at which the LM2673 starts up at power on. The capacitor is charged linearly by an internal current source. This voltage ramp gradually increases the duty cycle of the power switch until it reaches the normal operating duty cycle defined primarily by the ratio of the output voltage to the input voltage. The softstart turn-on time is programmable by the selection of Css.

The formula for selecting a softstart capacitor is:

$$C_{SS} \cong (I_{SST} \cdot t_{SS}) / [V_{SST} + 2.6V \cdot (\frac{V_{OUT} + V_{SCHOTTKY}}{V_{IN}})]$$

Where:

I_{SST} = Softstart Current, 3.7µA typical

t_{SS} = Softstart time, from design requirements

V_{SST} = Softstart Threshold Voltage, 0.63V typical

V_{OUT} = Output Voltage, from design requirements

$V_{SCHOTTKY}$ = Schottky Diode Voltage Drop, typically 0.5V

V_{IN} = Maximum Input Voltage, from design requirements

If this feature is not desired, leave the Softstart pin (pin 7) open circuited

With certain softstart capacitor values and operating conditions, the LM2673 can exhibit an overshoot on the output voltage during turn on. Especially when starting up into no load or low load, the softstart function may not be effective in preventing a larger voltage overshoot on the output. With larger loads or lower input voltages during startup this effect is minimized. In particular, avoid using softstart capacitors between 0.033µF and 1µF.

ADDITIONAL APPLICATION INFORMATION

When the output voltage is greater than approximately 6V, and the duty cycle at minimum input voltage is greater than approximately 50%, the designer should exercise caution in selection of the output filter components. When an application designed to these specific operating conditions is subjected to a current limit fault condition, it may be possible to observe a large hysteresis in the current limit. This can affect the output voltage of the device until the load current is reduced sufficiently to allow the current limit protection circuit to reset itself.

Under current limiting conditions, the LM267x is designed to respond in the following manner:

1. At the moment when the inductor current reaches the current limit threshold, the ON-pulse is immediately terminated. This happens for any application condition.
2. However, the current limit block is also designed to momentarily reduce the duty cycle to below 50% to avoid subharmonic oscillations, which could cause the inductor to saturate.
3. Thereafter, once the inductor current falls below the current limit threshold, there is a small relaxation time during which the duty cycle progressively rises back above 50% to the value required to achieve regulation.

If the output capacitance is sufficiently 'large', it may be possible that as the output tries to recover, the output capacitor charging current is large enough to repeatedly re-trigger the current limit circuit before the output has fully settled. This condition is exacerbated with higher output voltage settings because the energy requirement of the output capacitor varies as the square of the output voltage (½CV^2), thus requiring an increased charging current.

A simple test to determine if this condition might exist for a suspect application is to apply a short circuit across the output of the converter, and then remove the shorted output condition. In an application with properly selected external components, the output will recover smoothly.

Practical values of external components that have been experimentally found to work well under these specific operating conditions are C_{OUT} = 47µF, L = 22µH. It should be noted that even with these components, for a device's current limit of I_{CLIM}, the maximum load current under which the possibility of the large current limit hysteresis can be minimized is $I_{CLIM}/2$. For example, if the input is 24V and the set output voltage is 18V, then for a desired maximum current of 1.5A, the current limit of the chosen switcher must be confirmed to be at least 3A.

SIMPLE DESIGN PROCEDURE

Using the nomographs and tables in this data sheet (or use the available design software at http://www.national.com) a complete step-down regulator can be designed in a few simple steps.

Step 1: Define the power supply operating conditions:

Required output voltage

Maximum DC input voltage

Maximum output load current

Step 2: Set the output voltage by selecting a fixed output LM2673 (3.3V, 5V or 12V applications) or determine the required feedback resistors for use with the adjustable LM2673 –ADJ

Step 3: Determine the inductor required by using one of the four nomographs, *Figure 3* through *Figure 6*. Table 1 provides a specific manufacturer and part number for the inductor.

Step 4: Using Table 3 (fixed output voltage) or Table 6 (adjustable output voltage), determine the output capacitance required for stable operation. Table 2 provides the specific capacitor type from the manufacturer of choice.

Step 5: Determine an input capacitor from Table 4 for fixed output voltage applications. Use Table 2 to find the specific capacitor type. For adjustable output circuits select a capacitor from Table 2 with a sufficient working voltage (WV) rating greater than Vin max, and an rms current rating greater than one-half the maximum load current (2 or more capacitors in parallel may be required).

Step 6: Select a diode from Table 5. The current rating of the diode must be greater than I load max and the Reverse Voltage rating must be greater than Vin max.

Step 7: Include a 0.01µF/50V capacitor for Cboost in the design and then determine the value of a softstart capacitor if desired.

Step 8: Define a value for R_{ADJ} to set the peak switch current limit to be at least 20% greater than Iout max to allow for at least 30% inductor ripple current (±15% of Iout). For designs that must operate over the full temperature range the switch current limit should be set to at least 50% greater than Iout max (1.5 x I_{out} max).

FIXED OUTPUT VOLTAGE DESIGN EXAMPLE

A system logic power supply bus of 3.3V is to be generated from a wall adapter which provides an unregulated DC voltage of 13V to 16V. The maximum load current is 2.5A. A softstart delay time of 50mS is desired. Through-hole components are preferred.

Step 1: Operating conditions are:

Vout = 3.3V

Vin max = 16V

Iload max = 2.5A

Step 2: Select an LM2673T-3.3. The output voltage will have a tolerance of

±2% at room temperature and ±3% over the full operating temperature range.

Step 3: Use the nomograph for the 3.3V device ,*Figure 3*. The intersection of the 16V horizontal line (V_{in} max) and the 2.5A vertical line (I_{load} max) indicates that L33, a 22µH inductor, is required.

From Table 1, L33 in a through-hole component is available from Renco with part number RL-1283-22-43 or part number PE-53933 from Pulse Engineering.

Step 4: Use Table 3 to determine an output capacitor. With a 3.3V output and a 33µH inductor there are four through-hole output capacitor solutions with the number of same type capacitors to be paralleled and an identifying capacitor code given. Table 2 provides the actual capacitor characteristics. Any of the following choices will work in the circuit:

1 x 220µF/10V Sanyo OS-CON (code C5)

1 x 1000µF/35V Sanyo MV-GX (code C10)

1 x 2200µF/10V Nichicon PL (code C5)

1 x 1000µF/35V Panasonic HFQ (code C7)

Step 5: Use Table 4 to select an input capacitor. With 3.3V output and 22µH there are three through-hole solutions. These capacitors provide a sufficient voltage rating and an rms current rating greater than 1.25A (1/2 I_{load} max). Again using Table 2 for specific component characteristics the following choices are suitable:

1 x 1000µF/63V Sanyo MV-GX (code C14)

1 x 820µF/63V Nichicon PL (code C24)

1 x 560µF/50V Panasonic HFQ (code C13)

Step 6: From Table 5 a 3A or more Schottky diode must be selected. The 20V rated diodes are sufficient for the application and for through-hole components two part types are suitable:

1N5820

SR302

Step 7: A 0.01µF capacitor will be used for Cboost. For the 50mS softstart delay the following parameters are to be used:

I_{SST}: 3.7µA

t_{SS}: 50mS

V_{SST}: 0.63V

V_{OUT}: 3.3V

$V_{SCHOTTKY}$: 0.5V

V_{IN}: 16V

Using Vin max ensures that the softstart delay time will be at least the desired 50mS.

Using the formula for Css a value of 0.148µF is determined to be required. Use of a standard value 0.22µF capacitor will produce more than sufficient softstart delay.

Step 8: Determine a value for R_{ADJ} to provide a peak switch current limit of at least 2.5A plus 50% or 3.75A.

$$R_{ADJ} = \frac{37,125}{3.75A} = 9.9\ k\Omega$$

Use a value of 10KΩ.

ADJUSTABLE OUTPUT DESIGN EXAMPLE

In this example it is desired to convert the voltage from a two battery automotive power supply (voltage range of 20V to 28V, typical in large truck applications) to the 14.8VDC alternator supply typically used to power electronic equipment from single battery 12V vehicle systems. The load current required is 2A maximum. It is also desired to implement the power supply with all surface mount components. Softstart is not required.

Step 1: Operating conditions are:

Vout = 14.8V

Vin max = 28V

Iload max = 2A

Step 2: Select an LM2673S-ADJ. To set the output voltage to 14.9V two resistors need to be chosen (R1 and R2 in *Figure 2*). For the adjustable device the output voltage is set by the following relationship:

$$V_{OUT} = V_{FB}\left(1 + \frac{R_2}{R_1}\right)$$

Where V_{FB} is the feedback voltage of typically 1.21V.

A recommended value to use for R1 is 1K. In this example then R2 is determined to be:

$$R_2 = R_1\left(\frac{V_{OUT}}{V_{FB}} - 1\right) = 1\ k\Omega\left(\frac{14.8V}{1.21V} - 1\right)$$

R2 = 11.23KΩ

The closest standard 1% tolerance value to use is 11.3KΩ

This will set the nominal output voltage to 14.88V which is within 0.5% of the target value.

Step 3: To use the nomograph for the adjustable device, *Figure 6*, requires a calculation of the inductor Volt•microsecond constant (E•T expressed in V•µS) from the following formula:

$$E \cdot T = (V_{IN(MAX)} - V_{OUT} - V_{SAT}) \cdot \frac{V_{OUT} + V_D}{V_{IN(MAX)} - V_{SAT} + V_D} \cdot \frac{1000}{260}\ (V \cdot \mu s)$$

where V_{SAT} is the voltage drop across the internal power switch which is $R_{ds(ON)}$ times I_{load}. In this example this would be typically 0.15Ω x 2A or 0.3V and V_D is the voltage drop across the forward bisased Schottky diode, typically 0.5V. The switching frequency of 260KHz is the nominal value to use to estimate the ON time of the switch during which energy is stored in the inductor.

For this example E•T is found to be:

$$E \cdot T = (28 - 14.8 - 0.3) \cdot \frac{14.8 + 0.5}{28 - 0.3 + 0.5} \cdot \frac{1000}{260}\ (V \cdot \mu s)$$

$$E \cdot T = (12.9V) \cdot \frac{15.3}{28.2} \cdot 3.85\ (V \cdot \mu s) = 26.9\ (V \cdot \mu s)$$

Using *Figure 6*, the intersection of 27V•µS horizontally and the 2A vertical line (I_{load} max) indicates that L38 , a 68µH inductor, should be used.

From Table 1, L38 in a surface mount component is available from Pulse Engineering with part number PE-54038S.

Step 4: Use Table 6 to determine an output capacitor. With a 14.8V output the 12.5 to 15V row is used and with a 68µH inductor there are three surface mount output capacitor solutions. Table 2 provides the actual capacitor characteristics based on the C Code number. Any of the following choices can be used:

1 x 33µF/20V AVX TPS (code C6)

1 x 47µF/20V Sprague 594 (code C8)

1 x 47µF/20V Kemet T495 (code C8)

Important Note: When using the adjustable device in low voltage applications (less than 3V output), if the nomograph, Figure 6, selects an inductance of 22µH or less, Table 6 does not provide an output capacitor solution. With these conditions the number of output capacitors required for stable operation becomes impractical. It is recommended to use either a 33µH or 47µH inductor and the output capacitors from Table 6.

Step 5: An input capacitor for this example will require at least a 35V WV rating with an rms current rating of 1A (1/2 Iout max). From Table 2 it can be seen that C12, a 33µF/35V capacitor from Sprague, has the required voltage/current rating of the surface mount components.

Step 6: From Table 5 a 3A Schottky diode must be selected. For surface mount diodes with a margin of safety on the voltage rating one of five diodes can be used:

SK34

30BQ040

30WQ04F

MBRS340

MBRD340

Step 7: A 0.01µF capacitor will be used for Cboost.

The softstart pin will be left open circuited.

Step 8: Determine a value for R_{ADJ} to provide a peak switch current limit of at least 2A plus 50% or 3A.

$$R_{ADJ} = \frac{37,125}{3A} = 12.375\,k\Omega$$

Use a value of 12.4KΩ.

LLP PACKAGE DEVICES

The LM2673 is offered in the 14 lead LLP surface mount package to allow for a significantly decreased footprint with equivalent power dissipation compared to the TO-263. For details on mounting and soldering specifications, refer to Application Note AN-1187.

Inductor Selection Guides

For Continuous Mode Operation

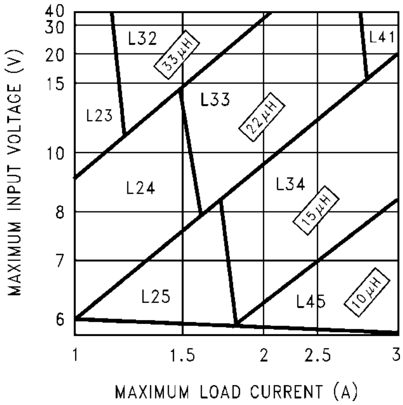


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FIGURE 3. LM2673-3.3

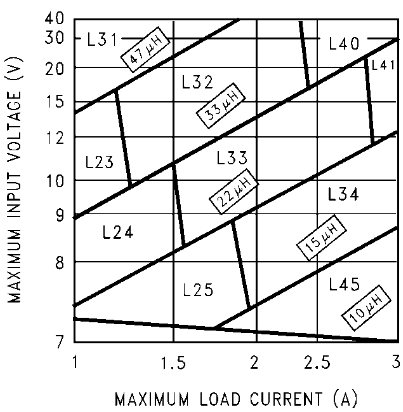

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FIGURE 4. LM2673-5.0

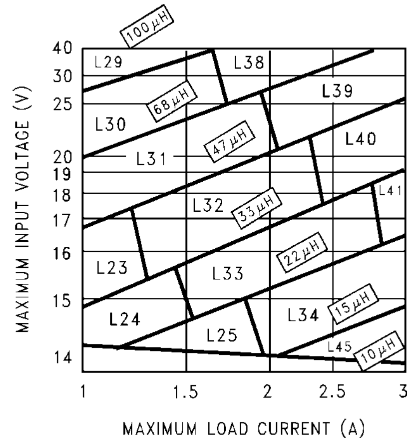

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FIGURE 5. LM2673-12

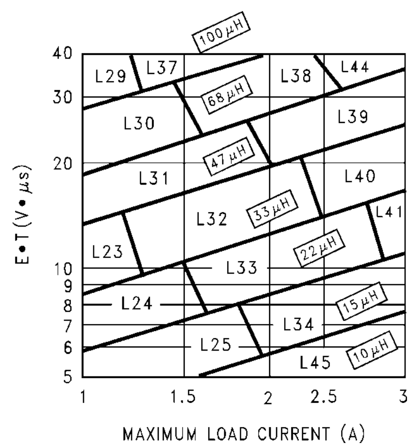

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FIGURE 6. LM2673-ADJ

Table 1. Inductor Manufacturer Part Numbers

Inductor Reference Number	Inductance (µH)	Current (A)	Renco		Pulse Engineering		Coilcraft
			Through Hole	Surface Mount	Through Hole	Surface Mount	Surface Mount
L23	33	1.35	RL-5471-7	RL1500-33	PE-53823	PE-53823S	DO3316-333
L24	22	1.65	RL-1283-22-43	RL1500-22	PE-53824	PE-53824S	DO3316-223
L25	15	2.00	RL-1283-15-43	RL1500-15	PE-53825	PE-53825S	DO3316-153
L29	100	1.41	RL-5471-4	RL-6050-100	PE-53829	PE-53829S	DO5022P-104
L30	68	1.71	RL-5471-5	RL6050-68	PE-53830	PE-53830S	DO5022P-683
L31	47	2.06	RL-5471-6	RL6050-47	PE-53831	PE-53831S	DO5022P-473
L32	33	2.46	RL-5471-7	RL6050-33	PE-53932	PE-53932S	DO5022P-333
L33	22	3.02	RL-1283-22-43	RL6050-22	PE-53933	PE-53933S	DO5022P-223
L34	15	3.65	RL-1283-15-43	—	PE-53934	PE-53934S	DO5022P-153
L38	68	2.97	RL-5472-2	—	PE-54038	PE-54038S	—
L39	47	3.57	RL-5472-3	—	PE-54039	PE-54039S	—
L40	33	4.26	RL-1283-33-43	—	PE-54040	PE-54040S	—
L41	22	5.22	RL-1283-22-43	—	PE-54041	P0841	—
L44	68	3.45	RL-5473-3	—	PE-54044	—	—
L45	10	4.47	RL-1283-10-43	—	—	P0845	DO5022P-103HC

Inductor Manufacturer Contact Numbers

Coilcraft	Phone	(800) 322-2645
	FAX	(708) 639-1469
Coilcraft, Europe	Phone	+44 1236 730 595
	FAX	+44 1236 730 627
Pulse Engineering	Phone	(619) 674-8100
	FAX	(619) 674-8262
Pulse Engineering, Europe	Phone	+353 93 24 107
	FAX	+353 93 24 459
Renco Electronics	Phone	(800) 645-5828
	FAX	(516) 586-5562

Capacitor Selection Guides

Table 2. Input and Output Capacitor Codes

Capacitor Reference Code	Surface Mount								
	AVX TPS Series			Sprague 594D Series			Kemet T495 Series		
	C (µF)	WV (V)	Irms (A)	C (µF)	WV (V)	Irms (A)	C (µF)	WV (V)	Irms (A)
C1	330	6.3	1.15	120	6.3	1.1	100	6.3	0.82
C2	100	10	1.1	220	6.3	1.4	220	6.3	1.1
C3	220	10	1.15	68	10	1.05	330	6.3	1.1
C4	47	16	0.89	150	10	1.35	100	10	1.1
C5	100	16	1.15	47	16	1	150	10	1.1
C6	33	20	0.77	100	16	1.3	220	10	1.1
C7	68	20	0.94	180	16	1.95	33	20	0.78
C8	22	25	0.77	47	20	1.15	47	20	0.94
C9	10	35	0.63	33	25	1.05	68	20	0.94
C10	22	35	0.66	68	25	1.6	10	35	0.63
C11				15	35	0.75	22	35	0.63
C12				33	35	1	4.7	50	0.66
C13				15	50	0.9			

Input and Output Capacitor Codes (continued)

Capacitor Reference Code	Through Hole											
	Sanyo OS-CON SA Series			Sanyo MV-GX Series			Nichicon PL Series			Panasonic HFQ Series		
	C (µF)	WV (V)	Irms (A)	C (µF)	WV (V)	Irms (A)	C (µF)	WV (V)	Irms (A)	C (µF)	WV (V)	Irms (A)
C1	47	6.3	1	1000	6.3	0.8	680	10	0.8	82	35	0.4
C2	150	6.3	1.95	270	16	0.6	820	10	0.98	120	35	0.44
C3	330	6.3	2.45	470	16	0.75	1000	10	1.06	220	35	0.76
C4	100	10	1.87	560	16	0.95	1200	10	1.28	330	35	1.01
C5	220	10	2.36	820	16	1.25	2200	10	1.71	560	35	1.4
C6	33	16	0.96	1000	16	1.3	3300	10	2.18	820	35	1.62
C7	100	16	1.92	150	35	0.65	3900	10	2.36	1000	35	1.73
C8	150	16	2.28	470	35	1.3	6800	10	2.68	2200	35	2.8
C9	100	20	2.25	680	35	1.4	180	16	0.41	56	50	0.36
C10	47	25	2.09	1000	35	1.7	270	16	0.55	100	50	0.5
C11				220	63	0.76	470	16	0.77	220	50	0.92
C12				470	63	1.2	680	16	1.02	470	50	1.44
C13				680	63	1.5	820	16	1.22	560	50	1.68
C14				1000	63	1.75	1800	16	1.88	1200	50	2.22
C15							220	25	0.63	330	63	1.42
C16							220	35	0.79	1500	63	2.51
C17							560	35	1.43			
C18							2200	35	2.68			
C19							150	50	0.82			
C20							220	50	1.04			
C21							330	50	1.3			
C22							100	63	0.75			
C23							390	63	1.62			
C24							820	63	2.22			
C25							1200	63	2.51			

Capacitor Manufacturer Contact Numbers

Nichicon	Phone	(847) 843-7500
	FAX	(847) 843-2798
Panasonic	Phone	(714) 373-7857
	FAX	(714) 373-7102
AVX	Phone	(845) 448-9411
	FAX	(845) 448-1943
Sprague/Vishay	Phone	(207) 324-4140
	FAX	(207) 324-7223
Sanyo	Phone	(619) 661-6322
	FAX	(619) 661-1055
Kemet	Phone	(864) 963-6300
	FAX	(864) 963-6521

Table 3. Output Capacitors for Fixed Output Voltage Application

Output Voltage (V)	Inductance (µH)	Surface Mount					
		AVX TPS Series		Sprague 594D Series		Kemet T495 Series	
		No.	C Code	No.	C Code	No.	C Code
3.3	10	4	C2	3	C1	4	C4
	15	4	C2	3	C1	4	C4
	22	3	C2	2	C7	3	C4
	33	2	C2	2	C6	2	C4
5	10	4	C2	4	C6	4	C4
	15	3	C2	2	C7	3	C4
	22	3	C2	2	C7	3	C4
	33	2	C2	2	C3	2	C4
	47	2	C2	1	C7	2	C4
12	10	4	C5	3	C6	5	C9
	15	3	C5	2	C7	4	C8
	22	2	C5	2	C6	3	C8
	33	2	C5	1	C7	2	C8
	47	2	C4	1	C6	2	C8
	68	1	C5	1	C5	2	C7
	100	1	C4	1	C5	1	C8

Output Voltage (V)	Inductance (µH)	Through Hole							
		Sanyo OS-CON SA Series		Sanyo MV-GX Series		Nichicon PL Series		Panasonic HFQ Series	
		No.	C Code	No.	C Code	No.	C Code	No.	C Code
3.3	10	1	C3	1	C10	1	C6	2	C6
	15	1	C3	1	C10	1	C6	2	C5
	22	1	C5	1	C10	1	C5	1	C7
	33	1	C2	1	C10	1	C13	1	C5
5	10	2	C4	1	C10	1	C6	2	C5
	15	1	C5	1	C10	1	C5	1	C6
	22	1	C5	1	C5	1	C5	1	C5
	33	1	C4	1	C5	1	C13	1	C5
	47	1	C4	1	C4	1	C13	2	C3
12	10	2	C7	2	C5	1	C18	2	C5
	15	1	C8	1	C5	1	C17	1	C5
	22	1	C7	1	C5	1	C13	1	C5
	33	1	C7	1	C3	1	C11	1	C4
	47	1	C7	1	C3	1	C10	1	C3
	68	1	C7	1	C2	1	C10	1	C3
	100	1	C7	1	C2	1	C9	1	C1

No. represents the number of identical capacitor types to be connected in parallel
C Code indicates the Capacitor Reference number in Table 2 for identifying the specific component from the manufacturer.

Table 4. Input Capacitors for Fixed Output Voltage Application

(Assumes worst case maximum input voltage and load current for a given inductance value)

Output Voltage (V)	Inductance (μH)	Surface Mount					
		AVX TPS Series		Sprague 594D Series		Kemet T495 Series	
		No.	C Code	No.	C Code	No.	C Code
3.3	10	2	C5	1	C7	2	C8
	15	3	C9	1	C10	3	C10
	22	*	*	2	C13	3	C12
	33	*	*	2	C13	2	C12
5	10	2	C5	1	C7	2	C8
	15	2	C5	1	C7	2	C8
	22	3	C10	2	C12	3	C11
	33	*	*	2	C13	3	C12
	47	*	*	1	C13	2	C12
12	10	2	C7	2	C10	2	C7
	15	2	C7	2	C10	2	C7
	22	3	C10	2	C12	3	C10
	33	3	C10	2	C12	3	C10
	47	*	*	2	C13	3	C12
	68	*	*	2	C13	2	C12
	100	*	*	1	C13	2	C12

Output Voltage (V)	Inductance (μH)	Through Hole							
		Sanyo OS-CON SA Series		Sanyo MV-GX Series		Nichicon PL Series		Panasonic HFQ Series	
		No.	C Code	No.	C Code	No.	C Code	No.	C Code
3.3	10	1	C7	2	C4	1	C5	1	C6
	15	1	C10	1	C10	1	C18	1	C6
	22	*	*	1	C14	1	C24	1	C13
	33	*	*	1	C12	1	C20	1	C12
5	10	1	C7	2	C4	1	C14	1	C6
	15	1	C7	2	C4	1	C14	1	C6
	22	*	*	1	C10	1	C18	1	C13
	33	*	*	1	C14	1	C23	1	C13
	47	*	*	1	C12	1	C20	1	C12
12	10	1	C9	1	C10	1	C18	1	C6
	15	1	C10	1	C10	1	C18	1	C6
	22	1	C10	1	C10	1	C18	1	C6
	33	*	*	1	C10	1	C18	1	C6
	47	*	*	1	C13	1	C23	1	C13
	68	*	*	1	C12	1	C21	1	C12
	100	*	*	1	C11	1	C22	1	C11

* Check voltage rating of capacitors to be greater than application input voltage.
No. represents the number of identical capacitor types to be connected in parallel
C Code indicates the Capacitor Reference number in Table 2 for identifying the specific component from the manufacturer.

Table 5. Schottky Diode Selection Table

Reverse Voltage (V)	Surface Mount		Through Hole	
	3A	**5A or More**	**3A**	**5A or More**
20V	SK32		1N5820 SR302	
30V	SK33 30WQ03F	MBRD835L	1N5821 31DQ03	
40V	SK34 30BQ040 30WQ04F MBRS340 MBRD340	MBRB1545CT 6TQ045S	1N5822 MBR340 31DQ04 SR403	MBR745 80SQ045 6TQ045
50V or More	SK35 30WQ05F		MBR350 31DQ05 SR305	

Diode Manufacturer Contact Numbers

International Rectifier	Phone	(310) 322-3331
	FAX	(310) 322-3332
Motorola	Phone	(800) 521-6274
	FAX	(602) 244-6609
General Semiconductor	Phone	(516) 847-3000
	FAX	(516) 847-3236
Diodes, Inc.	Phone	(805) 446-4800
	FAX	(805) 446-4850

Table 6. Output Capacitors for Adjustable Output Voltage Applications

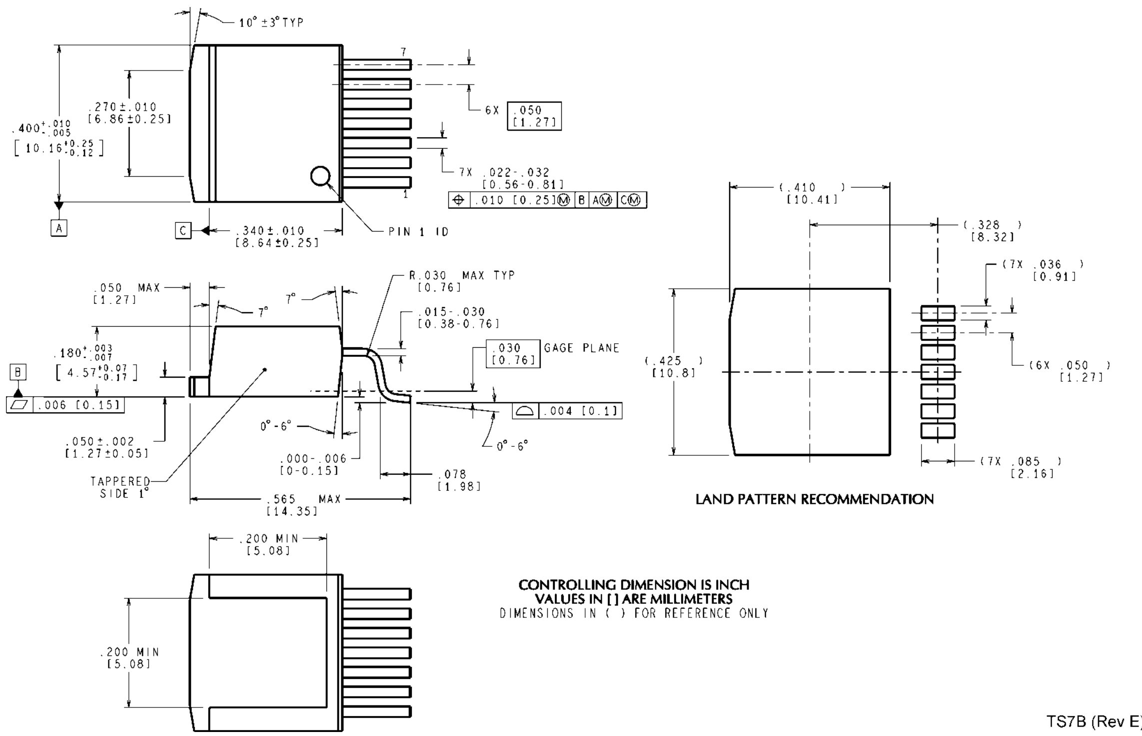
Output Voltage (V)	Inductance (μH)	Surface Mount					
		AVX TPS Series		Sprague 594D Series		Kemet T495 Series	
		No.	C Code	No.	C Code	No.	C Code
1.21 to 2.50	33*	7	C1	6	C2	7	C3
	47*	5	C1	4	C2	5	C3
2.5 to 3.75	33*	4	C1	3	C2	4	C3
	47*	3	C1	2	C2	3	C3
3.75 to 5	22	4	C1	3	C2	4	C3
	33	3	C1	2	C2	3	C3
	47	2	C1	2	C2	2	C3
5 to 6.25	22	3	C2	3	C3	3	C4
	33	2	C2	2	C3	2	C4
	47	2	C2	2	C3	2	C4
	68	1	C2	1	C3	1	C4
6.25 to 7.5	22	3	C2	1	C4	3	C4
	33	2	C2	1	C3	2	C4
	47	1	C3	1	C4	1	C6
	68	1	C2	1	C3	1	C4
7.5 to 10	33	2	C5	1	C6	2	C8
	47	1	C5	1	C6	2	C8
	68	1	C5	1	C6	1	C8
	100	1	C4	1	C5	1	C8
10 to 12.5	33	1	C5	1	C6	2	C8
	47	1	C5	1	C6	2	C8
	68	1	C5	1	C6	1	C8
	100	1	C5	1	C6	1	C8
12.5 to 15	33	1	C6	1	C8	1	C8
	47	1	C6	1	C8	1	C8
	68	1	C6	1	C8	1	C8
	100	1	C6	1	C8	1	C8
15 to 20	33	1	C8	1	C10	2	C10
	47	1	C8	1	C9	2	C10
	68	1	C8	1	C9	2	C10
	100	1	C8	1	C9	1	C10
20 to 30	33	2	C9	2	C11	2	C11
	47	1	C10	1	C12	1	C11
	68	1	C9	1	C12	1	C11
	100	1	C9	1	C12	1	C11
30 to 37	10	No Values Available		4	C13	8	C12
	15			3	C13	5	C12
	22			2	C13	4	C12
	33			1	C13	3	C12
	47			1	C13	2	C12
	68			1	C13	2	C12

Output Capacitors for Adjustable Output Voltage Applications (continued)

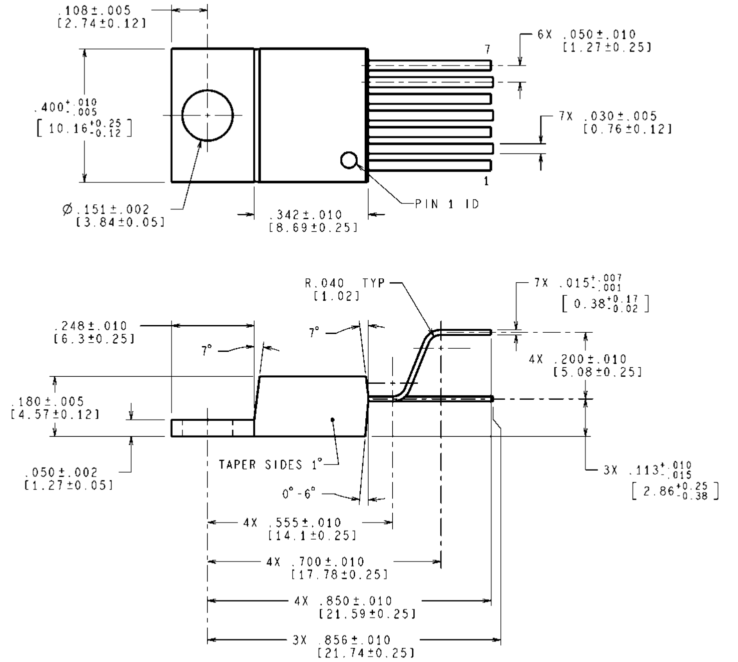
Output Voltage (V)	Inductance (µH)	Through Hole							
		Sanyo OS-CON SA Series		Sanyo MV-GX Series		Nichicon PL Series		Panasonic HFQ Series	
		No.	C Code	No.	C Code	No.	C Code	No.	C Code
1.21 to 2.50	33*	2	C3	5	C1	5	C3	3	C
	47*	2	C2	4	C1	3	C3	2	C5
2.5 to 3.75	33*	1	C3	3	C1	3	C1	2	C5
	47*	1	C2	2	C1	2	C3	1	C5
3.75 to 5	22	1	C3	3	C1	3	C1	2	C5
	33	1	C2	2	C1	2	C1	1	C5
	47	1	C2	2	C1	1	C3	1	C5
5 to 6.25	22	1	C5	2	C6	2	C3	2	C5
	33	1	C4	1	C6	2	C1	1	C5
	47	1	C4	1	C6	1	C3	1	C5
	68	1	C4	1	C6	1	C1	1	C5
6.25 to 7.5	22	1	C5	1	C6	2	C1	1	C5
	33	1	C4	1	C6	1	C3	1	C5
	47	1	C4	1	C6	1	C1	1	C5
	68	1	C4	1	C2	1	C1	1	C5
7.5 to 10	33	1	C7	1	C6	1	C14	1	C5
	47	1	C7	1	C6	1	C14	1	C5
	68	1	C7	1	C2	1	C14	1	C2
	100	1	C7	1	C2	1	C14	1	C2
10 to 12.5	33	1	C7	1	C6	1	C14	1	C5
	47	1	C7	1	C2	1	C14	1	C5
	68	1	C7	1	C2	1	C9	1	C2
	100	1	C7	1	C2	1	C9	1	C2
12.5 to 15	33	1	C9	1	C10	1	C15	1	C2
	47	1	C9	1	C10	1	C15	1	C2
	68	1	C9	1	C10	1	C15	1	C2
	100	1	C9	1	C10	1	C15	1	C2
15 to 20	33	1	C10	1	C7	1	C15	1	C2
	47	1	C10	1	C7	1	C15	1	C2
	68	1	C10	1	C7	1	C15	1	C2
	100	1	C10	1	C7	1	C15	1	C2
20 to 30	33	No Values Available		1	C7	1	C16	1	C2
	47			1	C7	1	C16	1	C2
	68			1	C7	1	C16	1	C2
	100			1	C7	1	C16	1	C2
30 to 37	10	No Values Available		1	C12	1	C20	1	C10
	15			1	C11	1	C20	1	C11
	22			1	C11	1	C20	1	C10
	33			1	C11	1	C20	1	C10
	47			1	C11	1	C20	1	C10
	68			1	C11	1	C20	1	C10

* Set to a higher value for a practical design solution. See Applications Hints section
No. represents the number of identical capacitor types to be connected in parallel
C Code indicates the Capacitor Reference number in Table 2 for identifying the specific component from the manufacturer.

Physical Dimensions inches (millimeters) unless otherwise noted



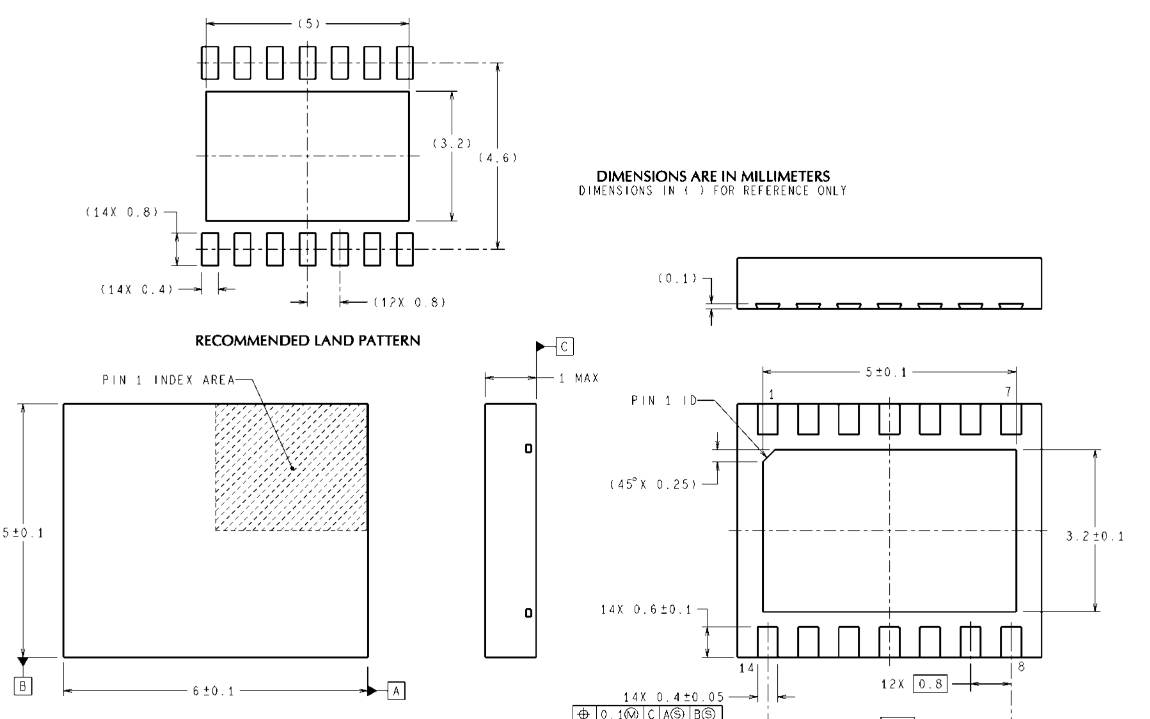
TO-263 Surface Mount Power Package
Order Number LM2673S-3.3, LM2673S-5.0,
LM2673S-12 or LM2673S-ADJ
NS Package Number TS7B



CONTROLING DIMENSIONS IS INCH
VALUES IN [ ] ARE MILLIMETERS

TA07B (Rev E)

TO-220 Power Package
Order Number LM2673T-3.3, LM2673T-5.0,
LM2673T-12 or LM2673T-ADJ
NS Package Number TA07B



SRC14A (Rev A)

14-Lead LLP Package
NS Package Number SRC14A

Notes

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Data Converters	www.national.com/adc	Distributors	www.national.com/contacts
Displays	www.national.com/displays	Green Compliance	www.national.com/quality/green
Ethernet	www.national.com/ethernet	Packaging	www.national.com/packaging
Interface	www.national.com/interface	Quality and Reliability	www.national.com/quality
LVDS	www.national.com/lvds	Reference Designs	www.national.com/refdesigns
Power Management	www.national.com/power	Feedback	www.national.com/feedback
Switching Regulators	www.national.com/switchers		
LDOs	www.national.com/ldo		
LED Lighting	www.national.com/led		
PowerWise	www.national.com/powerwise		
Serial Digital Interface (SDI)	www.national.com/sdi		
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German Tel: +49 (0) 180 5010 771
English Tel: +44 (0) 870 850 4288

National Semiconductor Asia Pacific Technical Support Center
Email: ap.support@nsc.com

National Semiconductor Japan Technical Support Center
Email: jpn.feedback@nsc.com

THE MICRO M+

The Micro M+ is an ideal photovoltaic (PV) controller for use at home or in the field. It's an easy-to-build, one-evening project even a beginner can master. This project was designed by Mike Bryce, WB8VGE. An earlier charge controller called the "Micro M" proved to be a very popular project.[1]

Hams really do like to operate their rigs from solar power. Many have found solar power to be very addictive. I had dozens of requests for information on how to increase the current capacity of the original "Micro M" controller. The Micro M would handle up to 2 A of current. I wanted to improve the performance of the Micro M while I was at it. Because the Micro M switched the negative lead of the solar panel on and off, that lead had to be insolated from the system ground. While that's not a problem with portable use, it may cause trouble with a home station where all the grounds should be connected. Here's what I wanted to do:

- Reduce the standby current at night
- Increase current handling capacity to 4 A
- Change the charging scheme to high (positive) side switching
- Improve the charging algorithm
- Keep the size as small as possible, but large enough to easily construct.

I called the end result the Micro M+. You can assemble one in about an hour. Everything mounts on one double-sided PC board. It's small enough to mount inside your rig yet large enough so you won't misplace it! You can stuff four of them in your shirt pocket! And, you need not worry about RFI being generated by the Micro M+. It's completely silent and makes absolutely no RFI!

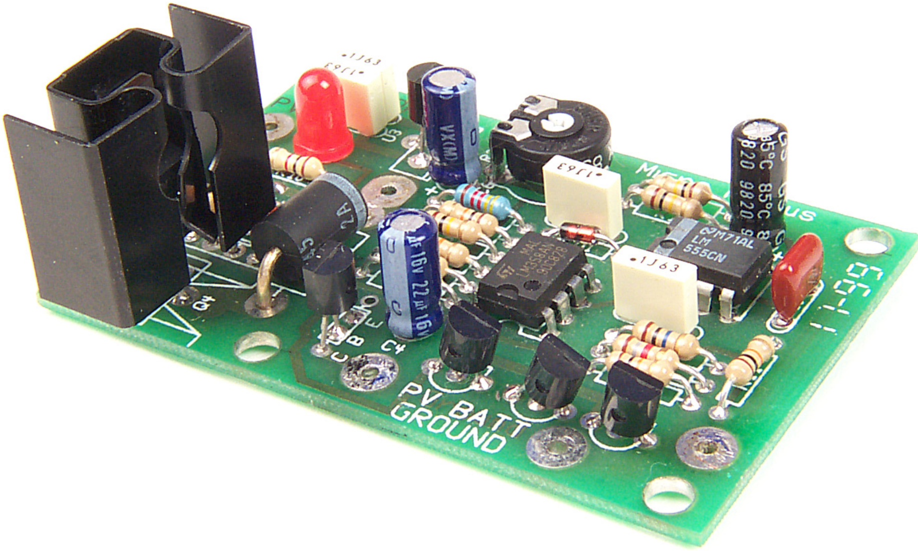


Fig 17.50 — This photo shows the Micro M+ charge controller circuit board. Leads solder to the board and connect to a solar panel and to the battery being charged.

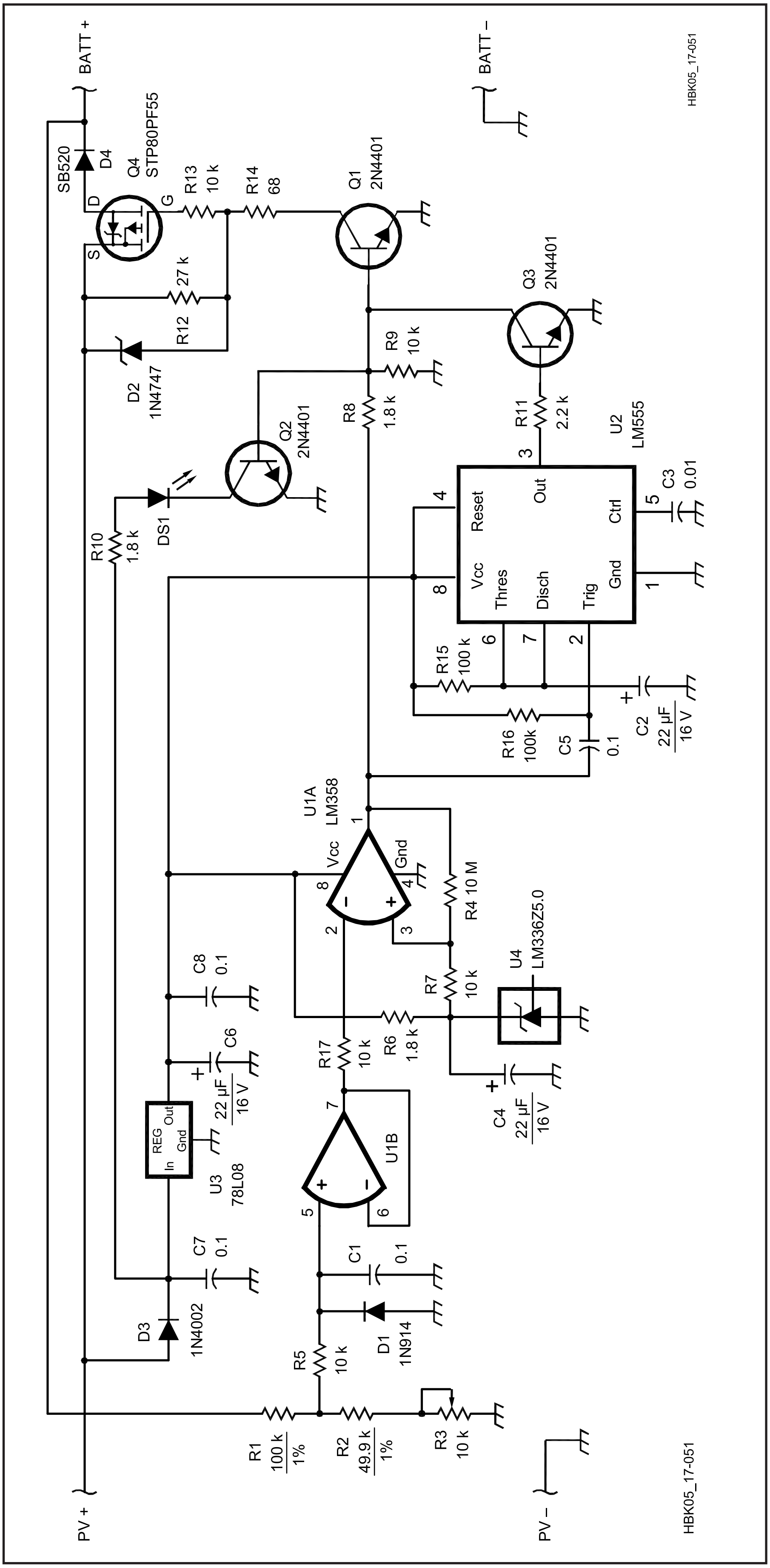
The Micro M+ will handle up to 4 A of current from a solar panel. That's equal to a 75-W solar panel.[2] I've reduced the standby current to less than one milliamp. I've also introduced a new charging algorithm to the Micro M+. All the current switching is done on the positive side. Now, you can connect the photovoltaic array, battery and load grounds together.

A complete kit of parts is available as well as just the PC board.[3] The Micro M+ is easy to build, making it a perfect first time project.

HOW IT WORKS

Fig 17.50 shows the complete Micro M+. **Fig 17.51** shows the schematic diagram. Let's begin with the current handling part of the Micro M+. Current from the solar panel is controlled by a power MOSFET. Instead of using a common N-channel MOSFET, however, the Micro M+ uses an STMicroelectronics STP80PF55 P-channel MOSFET. This P-channel FET has a current rating of 80 A with an RDS_{on} of 0.016 Ω. It comes in a TO-220 case. Current from the solar panel is routed directly to the MOSFET source lead.

N-channel power MOSFETs have very low RDS_{on} and even lower prices. To switch current on and off in a high-side application, though, the gate of an N-channel MOSFET must be at least 10 V higher



than the rail it is switching. In a typical 12-V system, the gate voltage must be at least 22 V to ensure the MOSFET is turned completely on. If the gate voltage is less than that required to fully enhance the MOSFET, it will be almost on and somewhat off (the MOSFET is operating in its linear region). Hence, the device will likely be destroyed at high current levels.

Normally, to produce this higher gate voltage, some sort of oscillator is used to charge a capacitor via a voltage doubler. This charge pump generates harmonics that may ride on the dc flowing into the battery under charge. Normally, this would not cause any problem, and in most cases, a filter or two on the dc bus will eliminate most of the harmonics generated. Even the best filter won't get rid of all the harmonics, however. To compound the problem, long wire runs to and from the solar panels and batteries act like antennas.

The P-channel MOSFET eliminates the need for a charge pump altogether. To turn on a P-channel MOSFET, all we have to do is pull the gate lead to ground! Since the Micro M+ does not have a charge pump, it generates NO RFI!

Now, you may be wondering if the P-channel MOSFET is so great, why have you not seen them in applications like this before? The answer is twofold. First, the RDS_{on} of a P-channel MOSFET has always been much higher than its N-channel cousin. Several years ago, a P-channel MOSFET with an RDS_{on} of 0.12 Ω was

Fig 17.51 — The schematic diagram of the Micro M+ charge controller.
C1, C5, C7, C8 — 0.1 µF
C2, C4, C6 — 22 µF, 16 V electrolytic
C3 — 0.01 µF
D1 — 1N914, small signal silicon switching diode
D2 — 1N4747, 20-V, 1-W Zener
D3 — 1N4002, silicon rectifier diode
D4 — SB520 20V, 5A Schottky diode (Mouser 512-SB520)
DS1 — LED, junkbox variety
Q1, Q2, Q3 — 2N4401 NPN small-signal transistor (2N2222 or 2N3904 will also work.)
Q4 — STP80PF55 P-channel MOSFET in TO-220 case (Mouser 511-STP80PF55). You will also need a small clip-on heat sink for this case.
R1 — 100 kΩ, 1%
R2 — 49.9 kΩ, 1%
R3 — 10 kΩ trimmer
U1 — LM358AN, Dual op-amp
U2 — LM555AN timer
U3 — LM78L08, 8-V regulator
U4 — LM336Z-5.0, 5.0-V Zener diode in TO-92 case. The adjust terminal allows control of the temperature coefficient and voltage over a range. The adjust terminal is not used for the Micro M+.

considered very low. At that time an N channel MOSFET had an RDS_{on} of 0.009 Ω. Suppose you want to control 10 A of current from your solar panel. Using the N-channel MOSFET above we find the MOSFET will dissipate less than a watt of power. On the other hand, the P-channel MOSFET will dissipate 12 W of power! Current generated by our solar panels is way too precious (and expensive) to have 12 W go up as heat from the charge controller.

The second factor was price. The P-channel MOSFET I described above would have easily sold for $19 each. The N-channel device would have been a few dollars.

More recently, the RDS_{on} of a typical P-channel MOSFET has fallen to 0.028 Ω. The price, while still a bit expensive, has dropped to about $8 each.

With the P channel MOSFET controlling the current, diode D4 — an SB520 Schottky — prevents battery current from flowing into the solar panel at night. This diode also provides reverse polarity protection to the battery in the event you connect the solar panel backwards. This protects the expensive P channel MOSFET.

Zener diode D2, a 1N4747, protects the gate from damage due to spikes on the solar panel line. Resistor R12 pulls the gate up, ensuring the power MOSFET is off when it is supposed to be.

THE MICRO M+ LIKES TO SLEEP

The Micro M+ never draws current from the battery. The solar panel provides all the power the Micro M+ needs, which means the Micro M+ goes to sleep at night. When the sun rises, the Micro M+ will start up again. As soon as the solar panel is producing enough current and voltage to start charging the battery, the Micro M+ will pass current into the battery.

To reduce the amount of stand-by current, diode D3 passes current from the solar panel to U3, the voltage regulator. U3, an LM78L08 regulator, provides a steady +8 V to the Micro M+ controller. Bypass capacitors, C6, C7 and C8 are used to keep everything happy. As long as there is power being produced by the solar panel, the Micro M+ will be awake. At sun down, the Micro M+ will go to sleep. Sleep current is on the order of less than 1 mA!

BATTERY SENSING

The battery terminal voltage is divided down to a more usable level by resistors, R1, R2 and R3. Resistor R3, a 10 kΩ trimmer, sets the state-of-charge for the Micro M+. A filter consisting of R5 and C1 helps keep the input clean from noise picked up by the wires to and from the solar panel. Diode D1 protects the op-amp input in case the battery sense line was connected backwards.

An LM358 dual op-amp is used in the Micro M+. One section (U1B) buffers the divided battery voltage before passing it along to the voltage comparator, U1A. Here the battery sense voltage is compared to the reference voltage supplied by U4. U4 is an LM336Z-5.0 precision diode. To prevent U1A from oscillating, a 10-MΩ resistor is used to eliminate any hysteresis.

As long as the voltage of the battery under charge is below the reference point, the output of U1A will be high. This saturates transistors Q1 and Q2. Q2 conducts and lights LED DS1, the CHARGING LED. Q1, also fully saturated, pulls the gate of the P channel MOSFET to ground. This effectively turns on the FET, and current flows from the solar panel into the battery via D4.

As the battery begins to take up the charge, its terminal voltage will increase. When the battery reaches the state-of-charge set point, the output of U1A goes low. With Q1 and Q2 now off, the P channel MOSFET is turned off, stopping all current into the battery. With Q2 off, the CHARGING LED goes dark.

Since we have eliminated any hysteresis in U1A, as soon as the current stops, the output of U1A pops back up high again. Why? Because the battery terminal voltage will fall back down as the charging current is removed. If left like this, the Micro M+ would sit and oscillate at the state-of-charge set point.

To prevent that from happening, the output of U1A is monitored by U2, an LM555 timer chip. As soon as the output of U1A goes low, this low trips U2. The output of U2 goes high, fully saturating transistor Q3. With Q3 turned on, it pulls the base of Q1 and Q2 low. Since both Q1 and Q2 are now deprived of base current, they remain off.

With the values shown for R15 and C2, charging current is stopped for about four seconds after the state-of-charge has been reached.

After the four second delay, Q1 and Q2 are allowed to have base drive from U1A. This lights up the charging LED and allows Q4 to pass current once more to the battery.

As soon as the battery hits the state-of-charge once more, the process is repeated. As the battery becomes fully charged, the "on" time will shorten up while the "off" time will always remain the same four seconds. In effect, a pulse of current will be sent to the battery that will shorten over time. I call this charging algorithm "Pulse Time Modulation."

As a side benefit of the pulse time modulation, the Micro M+ won't go nuts if you put a large solar panel onto a small battery. The charging algorithm will always keep the off time at four seconds allowing the battery time to rest before being hit by higher current than normal for its capacity.

BUILDING YOUR OWN MICRO M+

There's nothing special about the circuit. The use of a PC board makes the assembly of the Micro M+ quick and easy. It also makes it much easier if you need to troubleshoot the circuit. The entire circuit can be built on a piece of perf board.

The power MOSFET must be protected against static discharges. A dash of common sense and standard MOSFET handling procedures will work best. Don't handle the MOSFET until you need to install it in the circuit. A wrist strap is a good idea to prevent static damage. Once installed in the PC board, the device is quite robust.

A small clip-on heat sink is used for the power MOSFET. If you desire, the MOSFET could be mounted to a metal chassis. If you do this, make sure you electrically insulate the MOSFET tab from the chassis.

If you plan to use the Micro M+ outside, then consider soldering the IC directly onto the board. I've found that cheap solder-plated IC sockets corrode. If you want to use an IC socket, use one with gold plated contacts.

Feel free to substitute part values. There's nothing really critical. I do suggest you stick with 1% resistors for both R1 and R2. This isn't so important for their closer tolerance but for the 50-PPM temperature compensation they have. You can use standard off-the-shelf parts for either or both R1 and R2, but the entire circuit should then be located in an environment with a stable temperature.

ADJUSTMENTS

You'll need a good digital voltmeter and a variable power supply. Set the power supply to 14.3 V. Connect the Micro M+ battery negative lead to the power supply negative lead. Connect the Micro M+ PV positive and battery positive leads to the power supply positive lead. The charging LED should be on. If not, adjust trimmer R3 until it comes on. Check for +8 V at the VCC pins of the LM358 and the LM555. You should also see +5 V from the LM336Z5.0 diode.

Quickly move the trimmer from one end of its travel to the other. At one point the LED will go dark. This is the switch point. To verify that the "off pulse" is working, as soon as the LED goes dark quickly reverse the direction of the trimmer. The LED should remain off for several seconds and then come back on. If everything seems to be working, it's time to set the state-of-charge trimmer.

Now, slowly adjust the trimmer until the LED goes dark. You might want to try this adjustment more than once as the closer you get the comparator to switch at exactly 14.3 V, the more accurate the Micro M+ will be. Here's a hint I've learned after adjusting hundreds of Micro M+ controllers. Set the power supply to slightly above the cut-off voltage that you want. If you want 14.3 V, then set the supply to 14.5 V. I've found that in the time it takes to react to the LED going dark, you overshoot the cut-off point. Setting the supply higher takes this into account and usually you can get the trimmer set to exactly what you need in one try. That's all you need to do. Disconnect the supply from the Micro M+ and you're ready for the solar panel.

ODDS AND ENDS

The 14.3-V terminal voltage will be correct for just about all sealed and flooded-cell lead-acid batteries. You can change the state-of-charge set point if you want to recharge NiCds or captive sealed lead-acid batteries.

Keep the current from the solar panel within reason for the size of the battery you're going to be using. If you have a 7-amp hour battery, then don't use a 75-W solar panel. You'll get much better results and smoother operation with a smaller panel.

The tab of the power MOSFET is electrically hot. If you plan on using the Micro M+ without a protective case, make sure you insulate the tab from the heatsink. A misplaced wire touching the heatsink could cause real damage to both the Micro M+ and your equipment. A small plastic box from RadioShack works great.

MORE CURRENT?

Well yes, you can get the Micro M+ to handle more current. You must increase the capacity of the blocking diode and mount the power MOSFET on a larger heat sink. I've used an MBR2025 diode and a large heatsink for the MOSFET and can easily control 12 A of current.

BATTERY CHARGING WITHOUT A SOLAR PANEL?

Yes, it's possible. The trick is to use a power supply for which you can limit the output current. A discharged lead-acid battery will draw all the current it can from the charging source. In a solar panel setup, if the panel produces 3 A, that's all it will do. With an ac-powered supply, the current can be excessive. To use the Micro M+ with an ac-powered supply, set the voltage to 15.5 V. Then limit the current to 2 or 3 A.

No matter if you're camping in the outback, or storing photons just in case of an emergency, the Micro M+ will provide your battery with the fullest charge. The Micro M+ is simple to use and completely silent. Just like the sun!

Notes

[1]The Micro M, September 1996 *QST*, p 41.

[2]A 75-W module produces 4.4 A at 17 V. The Micro M+ can easily handle the extra 400 ma.

[3]PC boards, partial kits and full kits are available from Sunlight Energy Systems, **www.seslogic.com**.

Obsolete 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.

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.

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.

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.

April 2007

National Semiconductor

LM1575/LM2575/LM2575HV SIMPLE SWITCHER® 1A Step-Down Voltage Regulator

General Description

The LM2575 series of regulators are monolithic integrated circuits that provide all the active functions for a step-down (buck) switching regulator, capable of driving a 1A load with excellent line and load regulation. These devices are available in fixed output voltages of 3.3V, 5V, 12V, 15V, and an adjustable output version.

Requiring a minimum number of external components, these regulators are simple to use and include internal frequency compensation and a fixed-frequency oscillator.

The LM2575 series offers a high-efficiency replacement for popular three-terminal linear regulators. It substantially reduces the size of the heat sink, and in many cases no heat sink is required.

A standard series of inductors optimized for use with the LM2575 are available from several different manufacturers. This feature greatly simplifies the design of switch-mode power supplies.

Other features include a guaranteed ±4% tolerance on output voltage within specified input voltages and output load conditions, and ±10% on the oscillator frequency. External shutdown is included, featuring 50 μA (typical) standby current. The output switch includes cycle-by-cycle current limiting, as well as thermal shutdown for full protection under fault conditions.

Features

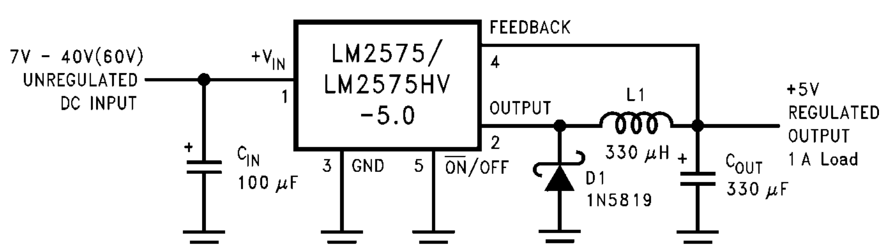
- 3.3V, 5V, 12V, 15V, and adjustable output versions
- Adjustable version output voltage range, 1.23V to 37V (57V for HV version) ±4% max over line and load conditions
- Guaranteed 1A output current
- Wide input voltage range, 40V up to 60V for HV version
- Requires only 4 external components
- 52 kHz fixed frequency internal oscillator
- TTL shutdown capability, low power standby mode
- High efficiency
- Uses readily available standard inductors
- Thermal shutdown and current limit protection
- P+ Product Enhancement tested

Applications

- Simple high-efficiency step-down (buck) regulator
- Efficient pre-regulator for linear regulators
- On-card switching regulators
- Positive to negative converter (Buck-Boost)

Typical Application

(Fixed Output Voltage Versions)



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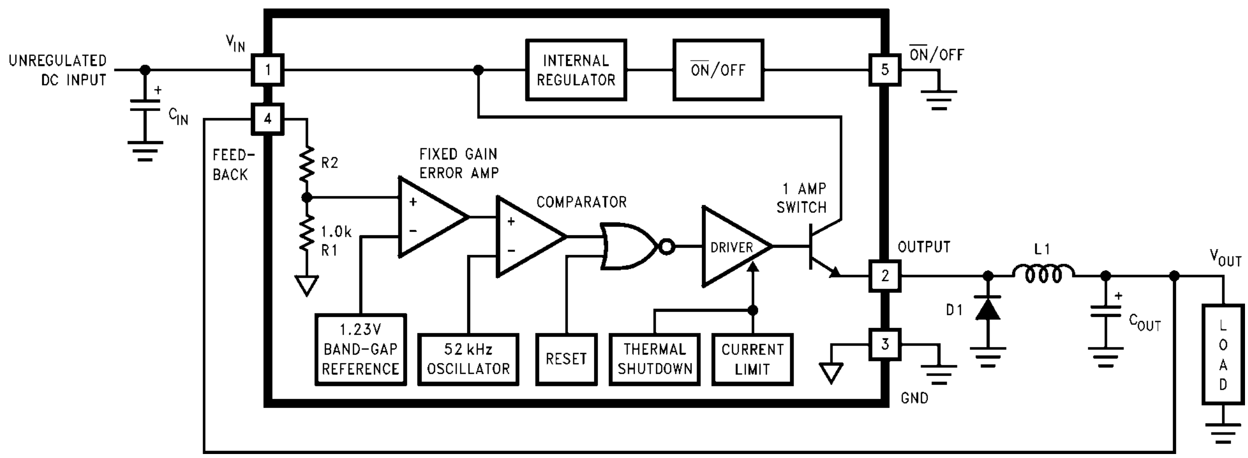
Note: Pin numbers are for the TO-220 package.

SIMPLE SWITCHER® is a registered trademark of National Semiconductor Corporation

LM1575/LM2575/LM2575HV Series SIMPLE SWITCHER 1A Step-Down Voltage Regulator

Block Diagram and Typical Application



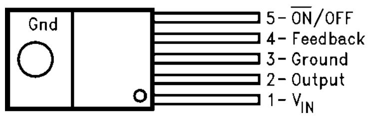
3.3V, R2 = 1.7k
5V, R2 = 3.1k
12V, R2 = 8.84k
15V, R2 = 11.3k
For ADJ. Version
R1 = Open, R2 = 0Ω
Note: Pin numbers are for the TO-220 package.

FIGURE 1.

Connection Diagrams

(XX indicates output voltage option. See Ordering Information table for complete part number.)

Straight Leads
5–Lead TO-220 (T)



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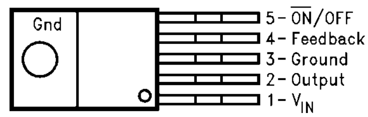
Top View
LM2575T-XX or LM2575HVT-XX
See NS Package Number T05A

Bent, Staggered Leads
5-Lead TO-220 (T)



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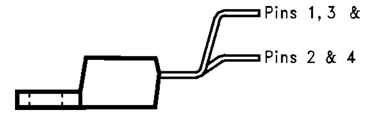
Top View



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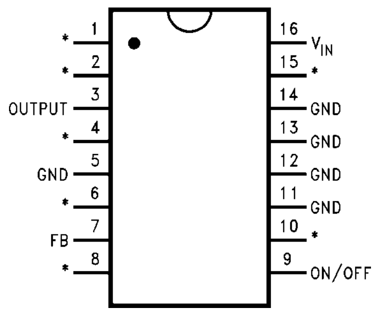
Side View
LM2575T-XX Flow LB03 or
LM2575HVT-XX Flow LB03
See NS Package Number T05D

16–Lead DIP (N or J)



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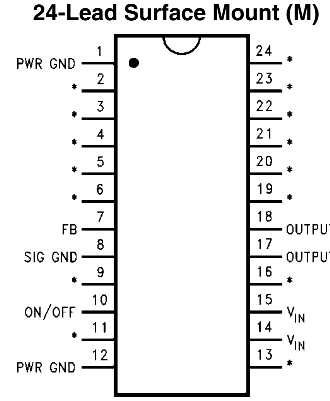
*No Internal Connection

Top View
LM2575N-XX or LM2575HVN-XX
See NS Package Number N16A
LM1575J-XX-QML
See NS Package Number J16A

24-Lead Surface Mount (M)



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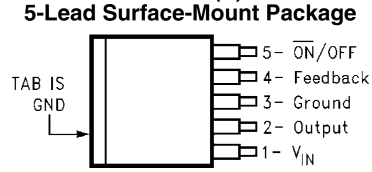
*No Internal Connection

Top View
LM2575M-XX or LM2575HVM-XX
See NS Package Number M24B

TO-263(S)
5-Lead Surface-Mount Package



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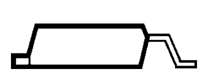
Top View



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Side View
LM2575S-XX or LM2575HVS-XX
See NS Package Number TS5B

Ordering Information

Package Type	NSC Package Number	Standard Voltage Rating (40V)	High Voltage Rating (60V)	Temperature Range
5-Lead TO-220 Straight Leads	T05A	LM2575T-3.3 LM2575T-5.0 LM2575T-12 LM2575T-15 LM2575T-ADJ	LM2575HVT-3.3 LM2575HVT-5.0 LM2575HVT-12 LM2575HVT-15 LM2575HVT-ADJ	
5-Lead TO-220 Bent and Staggered Leads	T05D	LM2575T-3.3 Flow LB03 LM2575T-5.0 Flow LB03 LM2575T-12 Flow LB03 LM2575T-15 Flow LB03 LM2575T-ADJ Flow LB03	LM2575HVT-3.3 Flow LB03 LM2575HVT-5.0 Flow LB03 LM2575HVT-12 Flow LB03 LM2575HVT-15 Flow LB03 LM2575HVT-ADJ Flow LB03	
16-Pin Molded DIP	N16A	LM2575N-5.0 LM2575N-12 LM2575N-15 LM2575N-ADJ	LM2575HVN-5.0 LM2575HVN-12 LM2575HVN-15 LM2575HVN-ADJ	$-40°C \le T_J \le +125°C$
24-Pin Surface Mount	M24B	LM2575M-5.0 LM2575M-12 LM2575M-15 LM2575M-ADJ	LM2575HVM-5.0 LM2575HVM-12 LM2575HVM-15 LM2575HVM-ADJ	
5-Lead TO-263 Surface Mount	TS5B	LM2575S-3.3 LM2575S-5.0 LM2575S-12 LM2575S-15 LM2575S-ADJ	LM2575HVS-3.3 LM2575HVS-5.0 LM2575HVS-12 LM2575HVS-15 LM2575HVS-ADJ	
16-Pin Ceramic DIP	J16A	LM1575J-3.3-QML LM1575J-5.0-QML LM1575J-12-QML LM1575J-15-QML LM1575J-ADJ-QML		$-55°C \le T_J \le +150°C$

Absolute Maximum Ratings (Note 1)

If Military/Aerospace specified devices are required, please contact the National Semiconductor Sales Office/ Distributors for availability and specifications.

Maximum Supply Voltage	
LM1575/LM2575	45V
LM2575HV	63V
$\overline{ON}$ /OFF Pin Input Voltage	$-0.3V \le V \le +V_{IN}$
Output Voltage to Ground	
(Steady State)	−1V
Power Dissipation	Internally Limited
Storage Temperature Range	−65°C to +150°C
Maximum Junction Temperature	150°C
Minimum ESD Rating	
(C = 100 pF, R = 1.5 kΩ)	2 kV
Lead Temperature	
(Soldering, 10 sec.)	260°C

Operating Ratings

Temperature Range	
LM1575	$-55°C \le T_J \le +150°C$
LM2575/LM2575HV	$-40°C \le T_J \le +125°C$
Supply Voltage	
LM1575/LM2575	40V
LM2575HV	60V

LM1575-3.3, LM2575-3.3, LM2575HV-3.3 Electrical Characteristics

Specifications with standard type face are for T_J = 25°C, and those with **boldface type** apply over **full Operating Temperature Range** .

Symbol	Parameter	Conditions	Typ	LM1575-3.3 Limit (Note 2)	LM2575-3.3 LM2575HV-3.3 Limit (Note 3)	Units (Limits)
SYSTEM PARAMETERS (Note 4) Test Circuit *Figure 2*						
V_{OUT}	Output Voltage	V_{IN} = 12V, I_{LOAD} = 0.2A Circuit of *Figure 2*	3.3	 3.267 3.333	 3.234 3.366	V V(Min) V(Max)
V_{OUT}	Output Voltage LM1575/LM2575	$4.75V \le V_{IN} \le 40V$, $0.2A \le I_{LOAD} \le 1A$ Circuit of *Figure 2*	3.3	 3.200/**3.168** 3.400/**3.432**	 3.168/**3.135** 3.432/**3.465**	V V(Min) V(Max)
V_{OUT}	Output Voltage LM2575HV	$4.75V \le V_{IN} \le 60V$, $0.2A \le I_{LOAD} \le 1A$ Circuit of *Figure 2*	3.3	 3.200/**3.168** 3.416/**3.450**	 3.168/**3.135** 3.450/**3.482**	V V(Min) V(Max)
η	Efficiency	V_{IN} = 12V, I_{LOAD} = 1A	75			%

LM1575-5.0, LM2575-5.0, LM2575HV-5.0 Electrical Characteristics

Specifications with standard type face are for T_J = 25°C, and those with **boldface type** apply over **full Operating Temperature Range**.

Symbol	Parameter	Conditions	Typ	LM1575-5.0 Limit (Note 2)	LM2575-5.0 LM2575HV-5.0 Limit (Note 3)	Units (Limits)
SYSTEM PARAMETERS (Note 4) Test Circuit *Figure 2*						
V_{OUT}	Output Voltage	V_{IN} = 12V, I_{LOAD} = 0.2A Circuit of *Figure 2*	5.0	 4.950 5.050	 4.900 5.100	V V(Min) V(Max)
V_{OUT}	Output Voltage LM1575/LM2575	$0.2A \le I_{LOAD} \le 1A$, $8V \le V_{IN} \le 40V$ Circuit of *Figure 2*	5.0	 4.850/**4.800** 5.150/**5.200**	 4.800/**4.750** 5.200/**5.250**	V V(Min) V(Max)

Symbol	Parameter	Conditions	Typ	LM1575-5.0	LM2575-5.0 LM2575HV-5.0	Units (Limits)
				Limit (Note 2)	Limit (Note 3)	
V_{OUT}	Output Voltage LM2575HV	$0.2A \le I_{LOAD} \le 1A$, $8V \le V_{IN} \le 60V$ Circuit of *Figure 2*	5.0			V
				4.850/**4.800**	4.800/**4.750**	V(Min)
				5.175/**5.225**	5.225/**5.275**	V(Max)
η	Efficiency	$V_{IN} = 12V$, $I_{LOAD} = 1A$	77			%

LM1575-12, LM2575-12, LM2575HV-12 Electrical Characteristics

Specifications with standard type face are for $T_J = 25°C$, and those with **boldface type** apply over **full Operating Temperature Range** .

Symbol	Parameter	Conditions	Typ	LM1575-12	LM2575-12 LM2575HV-12	Units (Limits)
				Limit (Note 2)	Limit (Note 3)	
SYSTEM PARAMETERS (Note 4) Test Circuit *Figure 2*						
V_{OUT}	Output Voltage	$V_{IN} = 25V$, $I_{LOAD} = 0.2A$ Circuit of *Figure 2*	12			V
				11.88	11.76	V(Min)
				12.12	12.24	V(Max)
V_{OUT}	Output Voltage LM1575/LM2575	$0.2A \le I_{LOAD} \le 1A$, $15V \le V_{IN} \le 40V$ Circuit of *Figure 2*	12			V
				11.64/**11.52**	11.52/**11.40**	V(Min)
				12.36/**12.48**	12.48/**12.60**	V(Max)
V_{OUT}	Output Voltage LM2575HV	$0.2A \le I_{LOAD} \le 1A$, $15V \le V_{IN} \le 60V$ Circuit of *Figure 2*	12			V
				11.64/**11.52**	11.52/**11.40**	V(Min)
				12.42/**12.54**	12.54/**12.66**	V(Max)
η	Efficiency	$V_{IN} = 15V$, $I_{LOAD} = 1A$	88			%

LM1575-15, LM2575-15, LM2575HV-15 Electrical Characteristics

Specifications with standard type face are for $T_J = 25°C$, and those with **boldface type** apply over **full Operating Temperature Range** .

Symbol	Parameter	Conditions	Typ	LM1575-15	LM2575-15 LM2575HV-15	Units (Limits)
				Limit (Note 2)	Limit (Note 3)	
SYSTEM PARAMETERS (Note 4) Test Circuit *Figure 2*						
V_{OUT}	Output Voltage	$V_{IN} = 30V$, $I_{LOAD} = 0.2A$ Circuit of *Figure 2*	15			V
				14.85	14.70	V(Min)
				15.15	15.30	V(Max)
V_{OUT}	Output Voltage LM1575/LM2575	$0.2A \le I_{LOAD} \le 1A$, $18V \le V_{IN} \le 40V$ Circuit of *Figure 2*	15			V
				14.55/**14.40**	14.40/**14.25**	V(Min)
				15.45/**15.60**	15.60/**15.75**	V(Max)
V_{OUT}	Output Voltage LM2575HV	$0.2A \le I_{LOAD} \le 1A$, $18V \le V_{IN} \le 60V$ Circuit of *Figure 2*	15			V
				14.55/**14.40**	14.40/**14.25**	V(Min)
				15.525/**15.675**	15.68/**15.83**	V(Max)
η	Efficiency	$V_{IN} = 18V$, $I_{LOAD} = 1A$	88			%

LM1575-ADJ, LM2575-ADJ, LM2575HV-ADJ Electrical Characteristics

Specifications with standard type face are for T_J= 25°C, and those with **boldface type** apply over **full Operating Temperature Range**.

Symbol	Parameter	Conditions	Typ	LM1575-ADJ	LM2575-ADJ LM2575HV-ADJ	Units (Limits)
				Limit (Note 2)	Limit (Note 3)	
SYSTEM PARAMETERS (Note 4) Test Circuit *Figure 2*						
V_{OUT}	Feedback Voltage	V_{IN} = 12V, I_{LOAD} = 0.2A	1.230			V
		V_{OUT} = 5V		1.217	1.217	V(Min)
		Circuit of *Figure 2*		1.243	1.243	V(Max)
V_{OUT}	Feedback Voltage	0.2A ≤ I_{LOAD} ≤ 1A,	1.230			V
	LM1575/LM2575	8V ≤ V_{IN} ≤ 40V		1.205/**1.193**	1.193/**1.180**	V(Min)
		V_{OUT} = 5V, Circuit of *Figure 2*		1.255/**1.267**	1.267/**1.280**	V(Max)
V_{OUT}	Feedback Voltage	0.2A ≤ I_{LOAD} ≤ 1A,	1.230			V
	LM2575HV	8V ≤ V_{IN} ≤ 60V		1.205/**1.193**	1.193/**1.180**	V(Min)
		V_{OUT} = 5V, Circuit of *Figure 2*		1.261/**1.273**	1.273/**1.286**	V(Max)
η	Efficiency	V_{IN} = 12V, I_{LOAD} = 1A, V_{OUT} = 5V	77			%

All Output Voltage Versions Electrical Characteristics

Specifications with standard type face are for T_J = 25°C, and those with **boldface type** apply over **full Operating Temperature Range**. Unless otherwise specified, V_{IN} = 12V for the 3.3V, 5V, and Adjustable version, V_{IN} = 25V for the 12V version, and V_{IN} = 30V for the 15V version. I_{LOAD} = 200 mA.

Symbol	Parameter	Conditions	Typ	LM1575-XX	LM2575-XX LM2575HV-XX	Units (Limits)
				Limit (Note 2)	Limit (Note 3)	
DEVICE PARAMETERS						
I_b	Feedback Bias Current	V_{OUT} = 5V (Adjustable Version Only)	50	100/**500**	100/**500**	nA
f_O	Oscillator Frequency	(Note 13)	52			kHz
				47/**43**	47/**42**	kHz(Min)
				58/**62**	58/**63**	kHz(Max)
V_{SAT}	Saturation Voltage	I_{OUT} = 1A (Note 5)	0.9			V
				1.2/**1.4**	1.2/**1.4**	V(Max)
DC	Max Duty Cycle (ON)	(Note 6)	98			%
				93	93	%(Min)
I_{CL}	Current Limit	Peak Current (Notes 5, 13)	2.2			A
				1.7/**1.3**	1.7/**1.3**	A(Min)
				3.0/**3.2**	3.0/**3.2**	A(Max)
I_L	Output Leakage Current	(Notes 7, 8) Output = 0V		2	2	mA(Max)
		Output = −1V	7.5			mA
		Output = −1V		30	30	mA(Max)
I_Q	Quiescent Current	(Note 7)	5			mA
				10/**12**	10	mA(Max)
I_{STBY}	Standby Quiescent Current	$\overline{ON}$ /OFF Pin = 5V (OFF)	50			μA
				200/**500**	200	μA(Max)

Symbol	Parameter	Conditions	Typ	LM1575-XX Limit (Note 2)	LM2575-XX LM2575HV-XX Limit (Note 3)	Units (Limits)
θ_{JA}	Thermal Resistance	T Package, Junction to Ambient (Note 9)	65			
θ_{JA}		T Package, Junction to Ambient (Note 10)	45			°C/W
θ_{JC}		T Package, Junction to Case	2			
θ_{JA}		N Package, Junction to Ambient (Note 11)	85			
θ_{JA}		M Package, Junction to Ambient (Note 11)	100			
θ_{JA}		S Package, Junction to Ambient (Note 12)	37			
$\overline{ON}$ /OFF CONTROL Test Circuit *Figure 2*						
V_{IH}	$\overline{ON}$ /OFF Pin Logic Input Level	V_{OUT} = 0V	1.4	2.2/**2.4**	2.2/**2.4**	V(Min)
V_{IL}		V_{OUT} = Nominal Output Voltage	1.2	1.0/**0.8**	1.0/**0.8**	V(Max)
I_{IH}	$\overline{ON}$ /OFF Pin Input Current	$\overline{ON}$ /OFF Pin = 5V (OFF)	12			µA
				30	30	µA(Max)
I_{IL}		$\overline{ON}$ /OFF Pin = 0V (ON)	0			µA
				10	10	µA(Max)

Note 1: Absolute Maximum Ratings indicate limits beyond which damage to the device may occur. Operating Ratings indicate conditions for which the device is intended to be functional, but do not guarantee specific performance limits. For guaranteed specifications and test conditions, see the Electrical Characteristics.

Note 2: All limits guaranteed at room temperature (standard type face) and at **temperature extremes (bold type face)**. All limits are used to calculate Average Outgoing Quality Level, and all are 100% production tested.

Note 3: All limits guaranteed at room temperature (standard type face) and at **temperature extremes (bold type face)**. All room temperature limits are 100% production tested. All limits at **temperature extremes** are guaranteed via correlation using standard Statistical Quality Control (SQC) methods.

Note 4: External components such as the catch diode, inductor, input and output capacitors can affect switching regulator system performance. When the LM1575/ LM2575 is used as shown in the *Figure 2* test circuit, system performance will be as shown in system parameters section of Electrical Characteristics.

Note 5: Output (pin 2) sourcing current. No diode, inductor or capacitor connected to output pin.

Note 6: Feedback (pin 4) removed from output and connected to 0V.

Note 7: Feedback (pin 4) removed from output and connected to +12V for the Adjustable, 3.3V, and 5V versions, and +25V for the 12V and 15V versions, to force the output transistor OFF.

Note 8: V_{IN} = 40V (60V for the high voltage version).

Note 9: Junction to ambient thermal resistance (no external heat sink) for the 5 lead TO-220 package mounted vertically, with ½ inch leads in a socket, or on a PC board with minimum copper area.

Note 10: Junction to ambient thermal resistance (no external heat sink) for the 5 lead TO-220 package mounted vertically, with ½ inch leads soldered to a PC board containing approximately 4 square inches of copper area surrounding the leads.

Note 11: Junction to ambient thermal resistance with approximately 1 square inch of pc board copper surrounding the leads. Additional copper area will lower thermal resistance further. See thermal model in Switchers made Simple software.

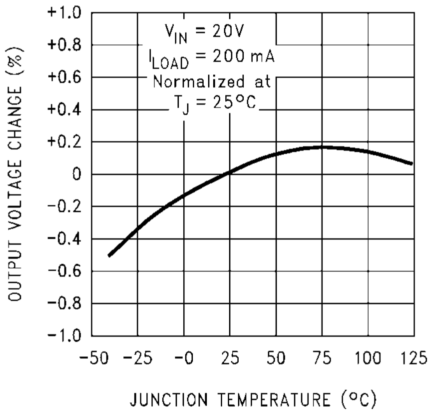
Note 12: If the TO-263 package is used, the thermal resistance can be reduced by increasing the PC board copper area thermally connected to the package: Using 0.5 square inches of copper area, θ_{JA} is 50°C/W; with 1 square inch of copper area, θ_{JA} is 37°C/W; and with 1.6 or more square inches of copper area, θ_{JA} is 32°C/W.

Note 13: The oscillator frequency reduces to approximately 18 kHz in the event of an output short or an overload which causes the regulated output voltage to drop approximately 40% from the nominal output voltage. This self protection feature lowers the average power dissipation of the IC by lowering the minimum duty cycle from 5% down to approximately 2%.

Note 14: Refer to RETS LM1575J for current revision of military RETS/SMD.

Typical Performance Characteristics (Circuit of *Figure 2*)

Normalized Output Voltage



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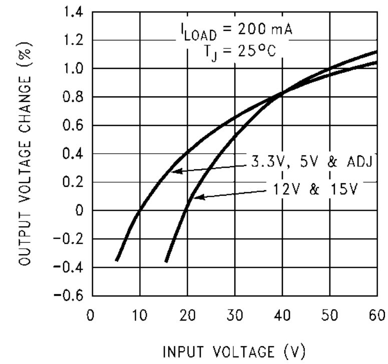
Line Regulation



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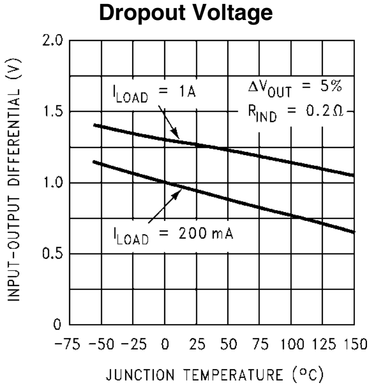
Dropout Voltage



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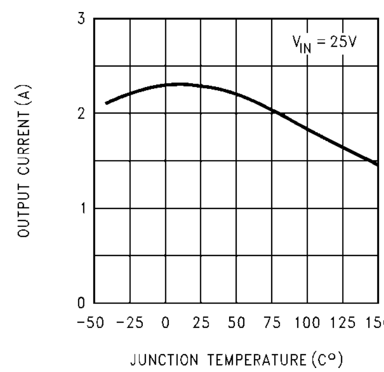
Current Limit



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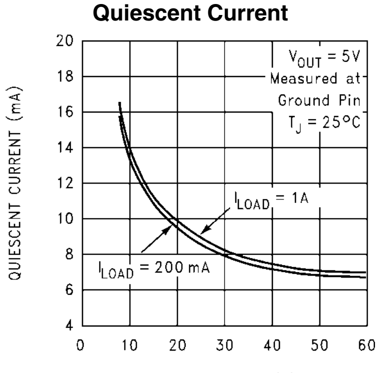
Quiescent Current



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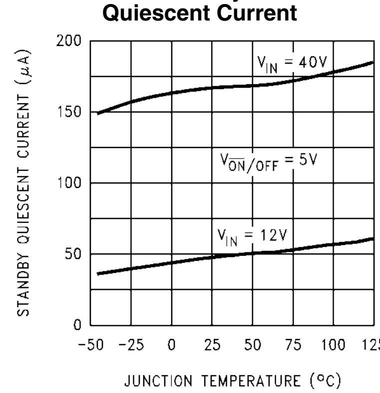
Standby Quiescent Current



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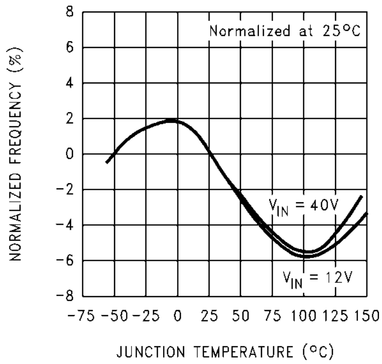
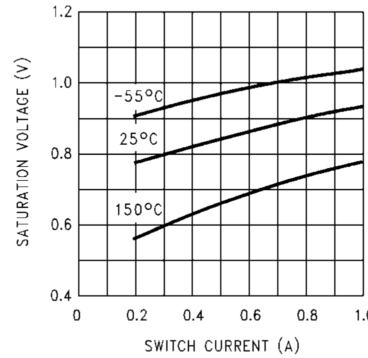
Oscillator Frequency



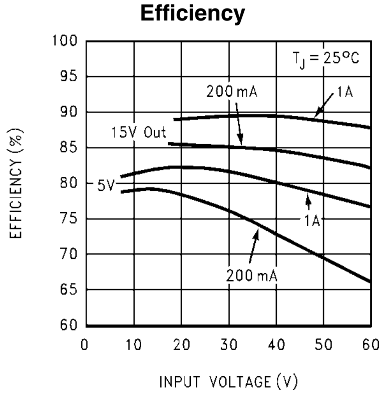
1147538

Switch Saturation Voltage



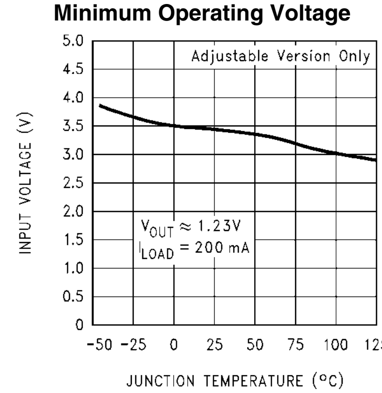
1147539

Efficiency



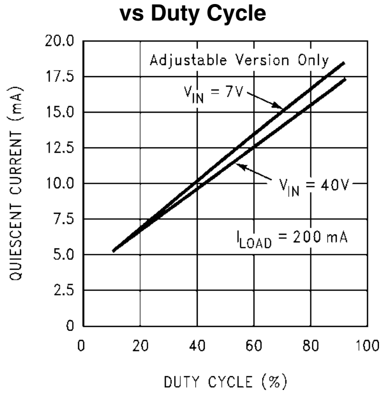
1147540

Minimum Operating Voltage



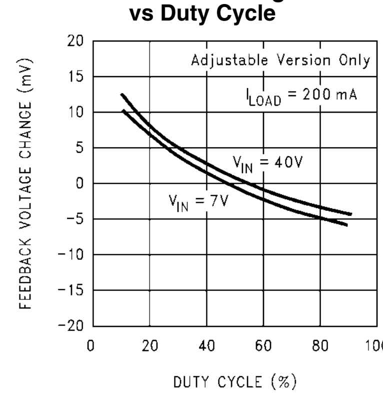
1147541

Quiescent Current vs Duty Cycle



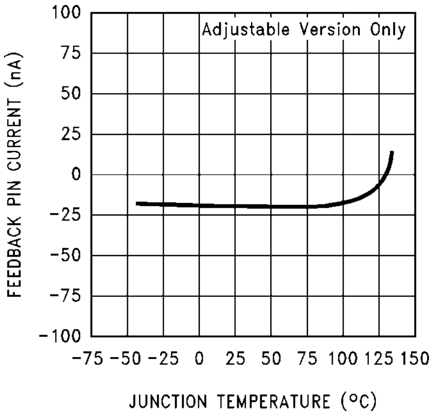
1147542

Feedback Voltage vs Duty Cycle



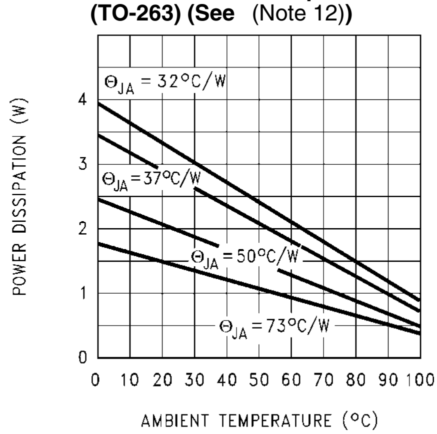
1147543

Feedback Pin Current



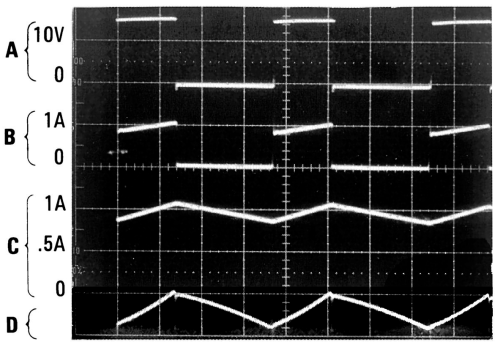
1147505

Maximum Power Dissipation (TO-263) (See (Note 12))



1147528

Switching Waveforms



1147506

V_{OUT} = 5V

A: Output Pin Voltage, 10V/div

B: Output Pin Current, 1A/div

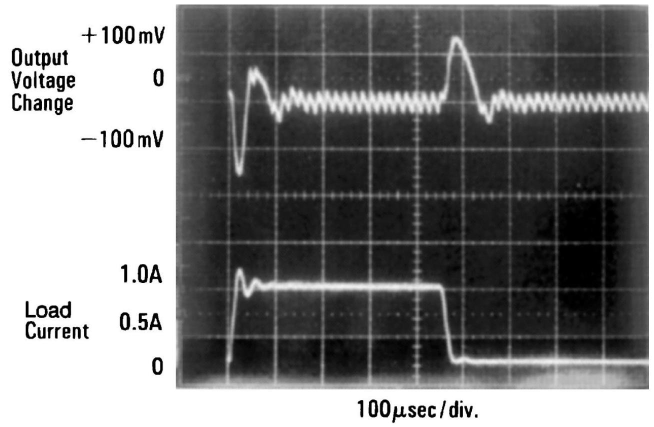
C: Inductor Current, 0.5A/div

D: Output Ripple Voltage, 20 mV/div,

AC-Coupled

Horizontal Time Base: 5 µs/div

Load Transient Response

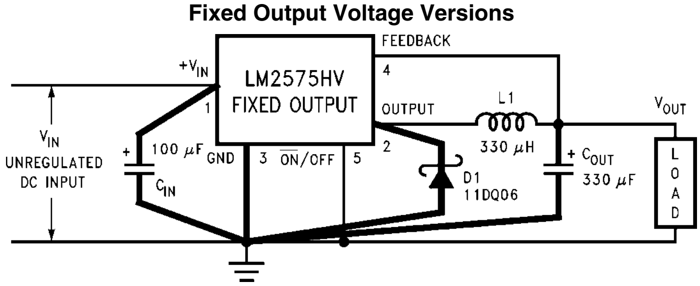


1147507

Test Circuit and Layout Guidelines

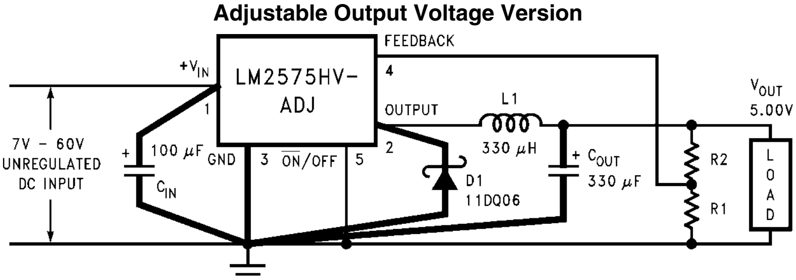
As in any switching regulator, layout is very important. Rapidly switching currents associated with wiring inductance generate voltage transients which can cause problems. For minimal inductance and ground loops, the length of the leads indicated by heavy lines should be kept as short as possible. Single-point grounding (as indicated) or ground plane construction should be used for best results. When using the Adjustable version, physically locate the programming resistors near the regulator, to keep the sensitive feedback wiring short.

Fixed Output Voltage Versions



C_{IN} — 100 μF, 75V, Aluminum Electrolytic
C_{OUT} — 330 μF, 25V, Aluminum Electrolytic
D1 — Schottky, 11DQ06
L1 — 330 μH, PE-52627 (for 5V in, 3.3V out, use 100 μH, PE-92108)

Adjustable Output Voltage Version



$$V_{OUT} = V_{REF}\left(1 + \frac{R2}{R1}\right)$$

$$R2 = R1\left(\frac{V_{OUT}}{V_{REF}} - 1\right)$$

where V_{REF} = 1.23V, R1 between 1k and 5k.
R1 — 2k, 0.1%
R2 — 6.12k, 0.1%
Note: Pin numbers are for the TO-220 package.

FIGURE 2.

LM2575 Series Buck Regulator Design Procedure

PROCEDURE (Fixed Output Voltage Versions)	EXAMPLE (Fixed Output Voltage Versions)
Given: V_{OUT} = Regulated Output Voltage (3.3V, 5V, 12V, or 15V) V_{IN}(Max) = Maximum Input Voltage I_{LOAD}(Max) = Maximum Load Current	**Given:** V_{OUT} = 5V V_{IN}(Max) = 20V I_{LOAD}(Max) = 0.8A
1. Inductor Selection (L1) **A.** Select the correct Inductor value selection guide from *Figures 3*, *4*, *5*, *6* (Output voltages of 3.3V, 5V, 12V or 15V respectively). For other output voltages, see the design procedure for the adjustable version. **B.** From the inductor value selection guide, identify the inductance region intersected by V_{IN}(Max) and I_{LOAD}(Max), and note the inductor code for that region. **C.** Identify the inductor value from the inductor code, and select an appropriate inductor from the table shown in *Figure 9*. Part numbers are listed for three inductor manufacturers. The inductor chosen must be rated for operation at the LM2575 switching frequency (52 kHz) and for a current rating of 1.15 × I_{LOAD}. For additional inductor information, see the inductor section in the Application Hints section of this data sheet.	**1. Inductor Selection (L1)** **A.** Use the selection guide shown in *Figure 4*. **B.** From the selection guide, the inductance area intersected by the 20V line and 0.8A line is L330. **C.** Inductor value required is 330 µH. From the table in *Figure 9*, choose AIE 415-0926, Pulse Engineering PE-52627, or RL1952.
2. Output Capacitor Selection (C_{OUT}) **A.** The value of the output capacitor together with the inductor defines the dominate pole-pair of the switching regulator loop. For stable operation and an acceptable output ripple voltage, (approximately 1% of the output voltage) a value between 100 µF and 470 µF is recommended. **B.** The capacitor's voltage rating should be at least 1.5 times greater than the output voltage. For a 5V regulator, a rating of at least 8V is appropriate, and a 10V or 15V rating is recommended. Higher voltage electrolytic capacitors generally have lower ESR numbers, and for this reason it may be necessary to select a capacitor rated for a higher voltage than would normally be needed.	**2. Output Capacitor Selection (C_{OUT})** **A.** C_{OUT} = 100 µF to 470 µF standard aluminum electrolytic. **B.** Capacitor voltage rating = 20V.
3. Catch Diode Selection (D1) **A.** The catch-diode current rating must be at least 1.2 times greater than the maximum load current. Also, if the power supply design must withstand a continuous output short, the diode should have a current rating equal to the maximum current limit of the LM2575. The most stressful condition for this diode is an overload or shorted output condition. **B.** The reverse voltage rating of the diode should be at least 1.25 times the maximum input voltage.	**3. Catch Diode Selection (D1)** **A.** For this example, a 1A current rating is adequate. **B.** Use a 30V 1N5818 or SR103 Schottky diode, or any of the suggested fast-recovery diodes shown in *Figure 8*.
4. Input Capacitor (C_{IN}) An aluminum or tantalum electrolytic bypass capacitor located close to the regulator is needed for stable operation.	**4. Input Capacitor (C_{IN})** A 47 µF, 25V aluminum electrolytic capacitor located near the input and ground pins provides sufficient bypassing.

Inductor Value Selection Guides

(For Continuous Mode Operation)

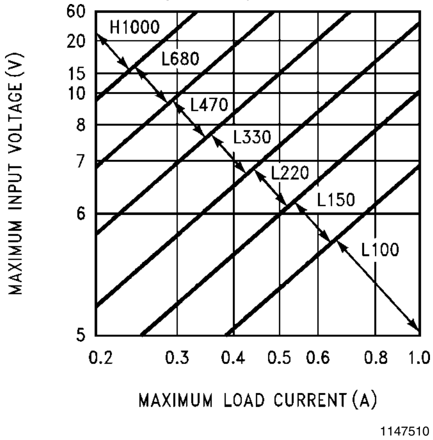


FIGURE 3. LM2575(HV)-3.3

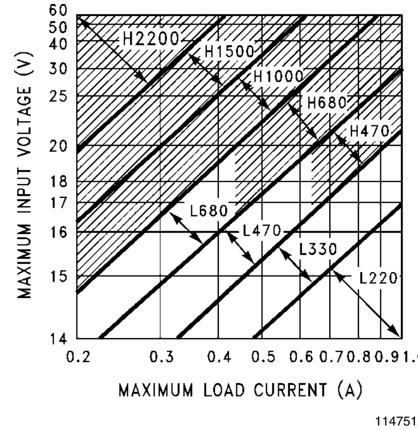


FIGURE 5. LM2575(HV)-12

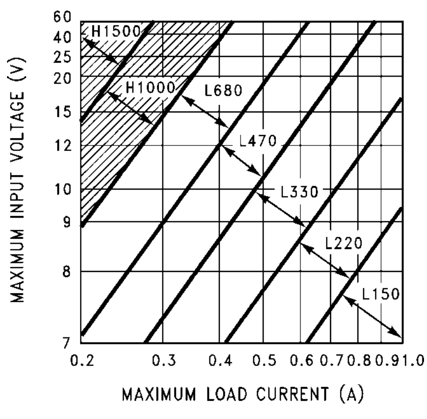


FIGURE 4. LM2575(HV)-5.0

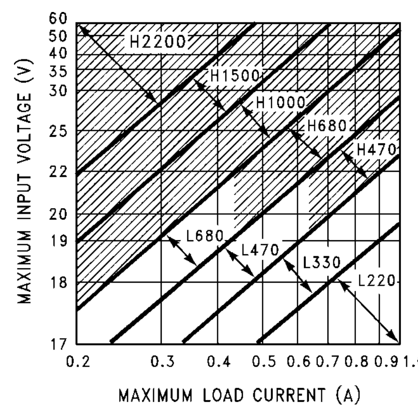


FIGURE 6. LM2575(HV)-15

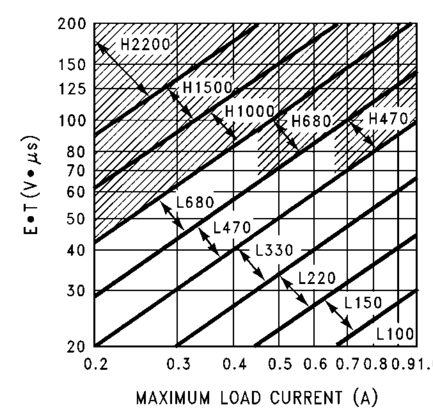


FIGURE 7. LM2575(HV)-ADJ

PROCEDURE (Adjustable Output Voltage Versions)	EXAMPLE (Adjustable Output Voltage Versions)
Given: V_{OUT} = Regulated Output Voltage V_{IN}(Max) = Maximum Input Voltage I_{LOAD}(Max) = Maximum Load Current F = Switching Frequency *(Fixed at 52 kHz)*	**Given:** V_{OUT} = 10V V_{IN}(Max) = 25V I_{LOAD}(Max) = 1A F = 52 kHz
1. Programming Output Voltage *(Selecting R1 and R2, as shown in Figure 2)* Use the following formula to select the appropriate resistor values. $V_{OUT} = V_{REF}\left(1 + \frac{R2}{R1}\right)$ where $V_{REF} = 1.23V$ R_1 can be between 1k and 5k. *(For best temperature coefficient and stability with time, use 1% metal film resistors)* $R2 = R1\left(\frac{V_{OUT}}{V_{REF}} - 1\right)$	**1.Programming Output Voltage** ***(Selecting R1 and R2)*** $V_{OUT} = 1.23\left(1 + \frac{R2}{R1}\right)$ Select R1 = 1k $R2 = R1\left(\frac{V_{OUT}}{V_{REF}} - 1\right) = 1k\left(\frac{10V}{1.23V} - 1\right)$ R2 = 1k (8.13 – 1) = 7.13k, closest 1% value is 7.15k
2. Inductor Selection (L1) **A.** Calculate the inductor Volt • microsecond constant, E • T (V • µs), from the following formula: $E \bullet T = (V_{IN} - V_{OUT})\frac{V_{OUT}}{V_{IN}} \bullet \frac{1000}{F\ (in\ kHz)}\ (V \bullet \mu s)$ **B.** Use the E • T value from the previous formula and match it with the E • T number on the vertical axis of the **Inductor Value Selection Guide** shown in *Figure 7*. **C.** On the horizontal axis, select the maximum load current. **D.** Identify the inductance region intersected by the E • T value and the maximum load current value, and note the inductor code for that region. **E.** Identify the inductor value from the inductor code, and select an appropriate inductor from the table shown in *Figure 9*. Part numbers are listed for three inductor manufacturers. The inductor chosen must be rated for operation at the LM2575 switching frequency (52 kHz) and for a current rating of 1.15 × I_{LOAD}. For additional inductor information, see the inductor section in the application hints section of this data sheet.	**2. Inductor Selection (L1)** **A.** Calculate E • T (V • µs) $E \bullet T = (25 - 10) \bullet \frac{10}{25} \bullet \frac{1000}{52} = 115\ V \bullet \mu s$ **B.** E • T = 115 V • µs **C.** I_{LOAD}(Max) = 1A **D.** Inductance Region = H470 **E.** Inductor Value = 470 µH *Choose from* ***AIE*** *part #430-0634,* ***Pulse Engineering*** *part #PE-53118, or* ***Renco*** *part #RL-1961.*
3. Output Capacitor Selection (C_{OUT}) **A.** The value of the output capacitor together with the inductor defines the dominate pole-pair of the switching regulator loop. For stable operation, the capacitor must satisfy the following requirement: $C_{OUT} \geq 7{,}785\frac{V_{IN}(Max)}{V_{OUT} \bullet L(\mu H)}\ (\mu F)$ The above formula yields capacitor values between 10 µF and 2000 µF that will satisfy the loop requirements for stable operation. But to achieve an acceptable output ripple voltage, (approximately 1% of the output voltage) and transient response, the output capacitor may need to be several times larger than the above formula yields. **B.** The capacitor's voltage rating should be at last 1.5 times greater than the output voltage. For a 10V regulator, a rating of at least 15V or more is recommended. Higher voltage electrolytic capacitors generally have lower ESR numbers, and for this reason it may be necessary to select a capacitor rate for a higher voltage than would normally be needed. *(Continued)*	**3. Output Capacitor Selection (C_{OUT})** **A.** $C_{OUT} > 7{,}785\frac{25}{10 \bullet 150} = 130\ \mu F$ However, for acceptable output ripple voltage select $C_{OUT} \geq 220\ \mu F$ C_{OUT} = 220 µF electrolytic capacitor *(Continued)*

PROCEDURE (Adjustable Output Voltage Versions)	EXAMPLE (Adjustable Output Voltage Versions)
4. Catch Diode Selection (D1)	**4. Catch Diode Selection (D1)**
A. The catch-diode current rating must be at least 1.2 times greater than the maximum load current. Also, if the power supply design must withstand a continuous output short, the diode should have a current rating equal to the maximum current limit of the LM2575. The most stressful condition for this diode is an overload or shorted output. See diode selection guide in *Figure 8*. **B.** The reverse voltage rating of the diode should be at least 1.25 times the maximum input voltage.	**A.** For this example, a 3A current rating is adequate. **B.** Use a 40V MBR340 or 31DQ04 Schottky diode, or any of the suggested fast-recovery diodes in *Figure 8*.
5. Input Capacitor (C_{IN})	**5. Input Capacitor (C_{IN})**
An aluminum or tantalum electrolytic bypass capacitor located close to the regulator is needed for stable operation.	A 100 µF aluminum electrolytic capacitor located near the input and ground pins provides sufficient bypassing.

To further simplify the buck regulator design procedure, National Semiconductor is making available computer design software to be used with the Simple Switcher line of switching regulators. ***Switchers Made Simple*** *(version 3.3) is available on a (3½") diskette for IBM compatible computers from a National Semiconductor sales office in your area.*

V_R	Schottky		Fast Recovery	
	1A	3A	1A	3A
20V	1N5817 MBR120P SR102	1N5820 MBR320 SR302	The following diodes are all rated to 100V 11DF1 MUR110 HER102	The following diodes are all rated to 100V 31DF1 MURD310 HER302
30V	1N5818 MBR130P 11DQ03 SR103	1N5821 MBR330 31DQ03 SR303		
40V	1N5819 MBR140P 11DQ04 SR104	IN5822 MBR340 31DQ04 SR304		
50V	MBR150 11DQ05 SR105	MBR350 31DQ05 SR305		
60V	MBR160 11DQ06 SR106	MBR360 31DQ06 SR306		

FIGURE 8. Diode Selection Guide

Inductor Code	**Inductor Value**	**Schott** (Note 15)	**Pulse Eng.** (Note 16)	**Renco** (Note 17)
L100	100 μH	67127000	PE-92108	RL2444
L150	150 μH	67127010	PE-53113	RL1954
L220	220 μH	67127020	PE-52626	RL1953
L330	330 μH	67127030	PE-52627	RL1952
L470	470 μH	67127040	PE-53114	RL1951
L680	680 μH	67127050	PE-52629	RL1950
H150	150 μH	67127060	PE-53115	RL2445
H220	220 μH	67127070	PE-53116	RL2446
H330	330 μH	67127080	PE-53117	RL2447
H470	470 μH	67127090	PE-53118	RL1961
H680	680 μH	67127100	PE-53119	RL1960
H1000	1000 μH	67127110	PE-53120	RL1959
H1500	1500 μH	67127120	PE-53121	RL1958
H2200	2200 μH	67127130	PE-53122	RL2448

Note 15: Schott Corp., (612) 475-1173, 1000 Parkers Lake Rd., Wayzata, MN 55391.

Note 16: Pulse Engineering, (619) 674-8100, P.O. Box 12236, San Diego, CA 92112.

Note 17: Renco Electronics Inc., (516) 586-5566, 60 Jeffryn Blvd. East, Deer Park, NY 11729.

FIGURE 9. Inductor Selection by Manufacturer's Part Number

Application Hints

INPUT CAPACITOR (C_{IN})

To maintain stability, the regulator input pin must be bypassed with at least a 47 µF electrolytic capacitor. The capacitor's leads must be kept short, and located near the regulator.

If the operating temperature range includes temperatures below −25°C, the input capacitor value may need to be larger. With most electrolytic capacitors, the capacitance value decreases and the ESR increases with lower temperatures and age. Paralleling a ceramic or solid tantalum capacitor will increase the regulator stability at cold temperatures. For maximum capacitor operating lifetime, the capacitor's RMS ripple current rating should be greater than

$$1.2 \times \left(\frac{t_{ON}}{T}\right) \times I_{LOAD}$$

where $\frac{t_{ON}}{T} = \frac{V_{OUT}}{V_{IN}}$ for a buck regulator

and $\frac{t_{ON}}{T} = \frac{|V_{OUT}|}{|V_{OUT}| + V_{IN}}$ for a buck-boost regulator.

INDUCTOR SELECTION

All switching regulators have two basic modes of operation: continuous and discontinuous. The difference between the two types relates to the inductor current, whether it is flowing continuously, or if it drops to zero for a period of time in the normal switching cycle. Each mode has distinctively different operating characteristics, which can affect the regulator performance and requirements.

The LM2575 (or any of the Simple Switcher family) can be used for both continuous and discontinuous modes of operation.

The inductor value selection guides in *Figure 3* through *Figure 7* were designed for buck regulator designs of the continuous inductor current type. When using inductor values shown in the inductor selection guide, the peak-to-peak inductor ripple current will be approximately 20% to 30% of the maximum DC current. With relatively heavy load currents, the circuit operates in the continuous mode (inductor current always flowing), but under light load conditions, the circuit will be forced to the discontinuous mode (inductor current falls to zero for a period of time). This discontinuous mode of operation is perfectly acceptable. For light loads (less than approximately 200 mA) it may be desirable to operate the regulator in the discontinuous mode, primarily because of the lower inductor values required for the discontinuous mode.

The selection guide chooses inductor values suitable for continuous mode operation, but if the inductor value chosen is prohibitively high, the designer should investigate the possibility of discontinuous operation. The computer design software ***Switchers Made Simple*** will provide all component values for discontinuous (as well as continuous) mode of operation.

Inductors are available in different styles such as pot core, toriod, E-frame, bobbin core, etc., as well as different core materials, such as ferrites and powdered iron. The least expensive, the bobbin core type, consists of wire wrapped on a ferrite rod core. This type of construction makes for an inexpensive inductor, but since the magnetic flux is not completely contained within the core, it generates more electromagnetic interference (EMI). This EMI can cause problems in sensitive circuits, or can give incorrect scope readings because of induced voltages in the scope probe.

The inductors listed in the selection chart include ferrite pot core construction for AIE, powdered iron toroid for Pulse Engineering, and ferrite bobbin core for Renco.

An inductor should not be operated beyond its maximum rated current because it may saturate. When an inductor begins to saturate, the inductance decreases rapidly and the inductor begins to look mainly resistive (the DC resistance of the winding). This will cause the switch current to rise very rapidly. Different inductor types have different saturation characteristics, and this should be kept in mind when selecting an inductor.

The inductor manufacturer's data sheets include current and energy limits to avoid inductor saturation.

INDUCTOR RIPPLE CURRENT

When the switcher is operating in the continuous mode, the inductor current waveform ranges from a triangular to a sawtooth type of waveform (depending on the input voltage). For a given input voltage and output voltage, the peak-to-peak amplitude of this inductor current waveform remains constant. As the load current rises or falls, the entire sawtooth current waveform also rises or falls. The average DC value of this waveform is equal to the DC load current (in the buck regulator configuration).

If the load current drops to a low enough level, the bottom of the sawtooth current waveform will reach zero, and the switcher will change to a discontinuous mode of operation. This is a perfectly acceptable mode of operation. Any buck switching regulator (no matter how large the inductor value is) will be forced to run discontinuous if the load current is light enough.

OUTPUT CAPACITOR

An output capacitor is required to filter the output voltage and is needed for loop stability. The capacitor should be located near the LM2575 using short pc board traces. Standard aluminum electrolytics are usually adequate, but low ESR types are recommended for low output ripple voltage and good stability. The ESR of a capacitor depends on many factors, some which are: the value, the voltage rating, physical size and the type of construction. In general, low value or low voltage (less than 12V) electrolytic capacitors usually have higher ESR numbers.

The amount of output ripple voltage is primarily a function of the ESR (Equivalent Series Resistance) of the output capacitor and the amplitude of the inductor ripple current (ΔI_{IND}). See the section on inductor ripple current in Application Hints.

The lower capacitor values (220 µF–680 µF) will allow typically 50 mV to 150 mV of output ripple voltage, while larger-value capacitors will reduce the ripple to approximately 20 mV to 50 mV.

$$\text{Output Ripple Voltage} = (\Delta I_{IND})\ (\text{ESR of } C_{OUT})$$

To further reduce the output ripple voltage, several standard electrolytic capacitors may be paralleled, or a higher-grade capacitor may be used. Such capacitors are often called “high-frequency,” “low-inductance,” or “low-ESR.” These will reduce the output ripple to 10 mV or 20 mV. However, when operating in the continuous mode, reducing the ESR below 0.05Ω can cause instability in the regulator.

Tantalum capacitors can have a very low ESR, and should be carefully evaluated if it is the only output capacitor. Because of their good low temperature characteristics, a tantalum can be used in parallel with aluminum electrolytics, with the tantalum making up 10% or 20% of the total capacitance.

The capacitor's ripple current rating at 52 kHz should be at least 50% higher than the peak-to-peak inductor ripple current.

CATCH DIODE

Buck regulators require a diode to provide a return path for the inductor current when the switch is off. This diode should be located close to the LM2575 using short leads and short printed circuit traces.

Because of their fast switching speed and low forward voltage drop, Schottky diodes provide the best efficiency, especially in low output voltage switching regulators (less than 5V). Fast-Recovery, High-Efficiency, or Ultra-Fast Recovery diodes are also suitable, but some types with an abrupt turn-off characteristic may cause instability and EMI problems. A fast-recovery diode with soft recovery characteristics is a better choice. Standard 60 Hz diodes (e.g., 1N4001 or 1N5400, etc.) are also **not suitable.** See *Figure 8* for Schottky and “soft” fast-recovery diode selection guide.

OUTPUT VOLTAGE RIPPLE AND TRANSIENTS

The output voltage of a switching power supply will contain a sawtooth ripple voltage at the switcher frequency, typically about 1% of the output voltage, and may also contain short voltage spikes at the peaks of the sawtooth waveform.

The output ripple voltage is due mainly to the inductor sawtooth ripple current multiplied by the ESR of the output capacitor. (See the inductor selection in the application hints.)

The voltage spikes are present because of the fast switching action of the output switch, and the parasitic inductance of the output filter capacitor. To minimize these voltage spikes, special low inductance capacitors can be used, and their lead lengths must be kept short. Wiring inductance, stray capacitance, as well as the scope probe used to evaluate these transients, all contribute to the amplitude of these spikes.

An additional small LC filter (20 µH & 100 µF) can be added to the output (as shown in *Figure 15*) to further reduce the amount of output ripple and transients. A 10 × reduction in output ripple voltage and transients is possible with this filter.

FEEDBACK CONNECTION

The LM2575 (fixed voltage versions) feedback pin must be wired to the output voltage point of the switching power supply. When using the adjustable version, physically locate both output voltage programming resistors near the LM2575 to avoid picking up unwanted noise. Avoid using resistors greater than 100 kΩ because of the increased chance of noise pickup.

$\overline{\text{ON}}$ /OFF INPUT

For normal operation, the $\overline{\text{ON}}$ /OFF pin should be grounded or driven with a low-level TTL voltage (typically below 1.6V). To put the regulator into standby mode, drive this pin with a high-level TTL or CMOS signal. The $\overline{\text{ON}}$ /OFF pin can be safely pulled up to $+V_{IN}$ without a resistor in series with it. The $\overline{\text{ON}}$ /OFF pin should not be left open.

GROUNDING

To maintain output voltage stability, the power ground connections must be low-impedance (see *Figure 2*). For the TO-3 style package, the case is ground. For the 5-lead TO-220 style package, both the tab and pin 3 are ground and either connection may be used, as they are both part of the same copper lead frame.

With the N or M packages, all the pins labeled ground, power ground, or signal ground should be soldered directly to wide printed circuit board copper traces. This assures both low inductance connections and good thermal properties.

HEAT SINK/THERMAL CONSIDERATIONS

In many cases, no heat sink is required to keep the LM2575 junction temperature within the allowed operating range. For each application, to determine whether or not a heat sink will be required, the following must be identified:

1. Maximum ambient temperature (in the application).
2. Maximum regulator power dissipation (in application).
3. Maximum allowed junction temperature (150°C for the LM1575 or 125°C for the LM2575). For a safe, conservative design, a temperature approximately 15°C cooler than the maximum temperature should be selected.
4. LM2575 package thermal resistances θ_{JA} and θ_{JC}.

Total power dissipated by the LM2575 can be estimated as follows:

$$P_D = (V_{IN})\,(I_Q) + (V_O/V_{IN})\,(I_{LOAD})\,(V_{SAT})$$

where I_Q (quiescent current) and V_{SAT} can be found in the Characteristic Curves shown previously, V_{IN} is the applied minimum input voltage, V_O is the regulated output voltage, and I_{LOAD} is the load current. The dynamic losses during turn-on and turn-off are negligible if a Schottky type catch diode is used.

When no heat sink is used, the junction temperature rise can be determined by the following:

$$\Delta T_J = (P_D)\,(\theta_{JA})$$

To arrive at the actual operating junction temperature, add the junction temperature rise to the maximum ambient temperature.

$$T_J = \Delta T_J + T_A$$

If the actual operating junction temperature is greater than the selected safe operating junction temperature determined in step 3, then a heat sink is required.

When using a heat sink, the junction temperature rise can be determined by the following:

$$\Delta T_J = (P_D)\,(\theta_{JC} + \theta_{\text{interface}} + \theta_{\text{Heat sink}})$$

The operating junction temperature will be:

$$T_J = T_A + \Delta T_J$$

As above, if the actual operating junction temperature is greater than the selected safe operating junction temperature, then a larger heat sink is required (one that has a lower thermal resistance).

When using the LM2575 in the plastic DIP (N) or surface mount (M) packages, several items about the thermal properties of the packages should be understood. The majority of the heat is conducted out of the package through the leads, with a minor portion through the plastic parts of the package. Since the lead frame is solid copper, heat from the die is readily conducted through the leads to the printed circuit board copper, which is acting as a heat sink.

For best thermal performance, the ground pins and all the unconnected pins should be soldered to generous amounts of printed circuit board copper, such as a ground plane. Large areas of copper provide the best transfer of heat to the surrounding air. Copper on both sides of the board is also helpful in getting the heat away from the package, even if there is no direct copper contact between the two sides. Thermal resistance numbers as low as 40°C/W for the SO package, and 30°C/W for the N package can be realized with a carefully engineered pc board.

Included on the ***Switchers Made Simple*** design software is a more precise (non-linear) thermal model that can be used to determine junction temperature with different input-output parameters or different component values. It can also calculate the heat sink thermal resistance required to maintain the regulators junction temperature below the maximum operating temperature.

Additional Applications

INVERTING REGULATOR

Figure 10 shows a LM2575-12 in a buck-boost configuration to generate a negative 12V output from a positive input voltage. This circuit bootstraps the regulator's ground pin to the negative output voltage, then by grounding the feedback pin, the regulator senses the inverted output voltage and regulates it to −12V.

For an input voltage of 12V or more, the maximum available output current in this configuration is approximately 0.35A. At lighter loads, the minimum input voltage required drops to approximately 4.7V.

The switch currents in this buck-boost configuration are higher than in the standard buck-mode design, thus lowering the available output current. Also, the start-up input current of the buck-boost converter is higher than the standard buck-mode regulator, and this may overload an input power source with a current limit less than 1.5A. Using a delayed turn-on or an undervoltage lockout circuit (described in the next section) would allow the input voltage to rise to a high enough level before the switcher would be allowed to turn on.

Because of the structural differences between the buck and the buck-boost regulator topologies, the buck regulator design procedure section can not be used to select the inductor or the output capacitor. The recommended range of inductor values for the buck-boost design is between 68 µH and 220 µH, and the output capacitor values must be larger than what is normally required for buck designs. Low input voltages or high output currents require a large value output capacitor (in the thousands of micro Farads).

The peak inductor current, which is the same as the peak switch current, can be calculated from the following formula:

$$I_p \approx \frac{I_{LOAD}\,(V_{IN} + |V_O|)}{V_{IN}} + \frac{V_{IN}\,|V_O|}{V_{IN} + |V_O|} \times \frac{1}{2\,L_1\,f_{osc}}$$

Where f_{osc} = 52 kHz. Under normal continuous inductor current operating conditions, the minimum V_{IN} represents the worst case. Select an inductor that is rated for the peak current anticipated.

Also, the maximum voltage appearing across the regulator is the absolute sum of the input and output voltage. For a −12V output, the maximum input voltage for the LM2575 is +28V, or +48V for the LM2575HV.

The ***Switchers Made Simple*** (version 3.3) design software can be used to determine the feasibility of regulator designs using different topologies, different input-output parameters, different components, etc.

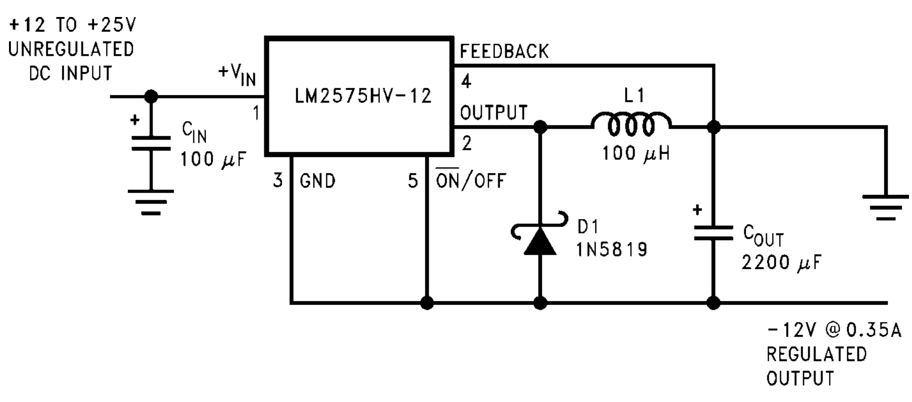
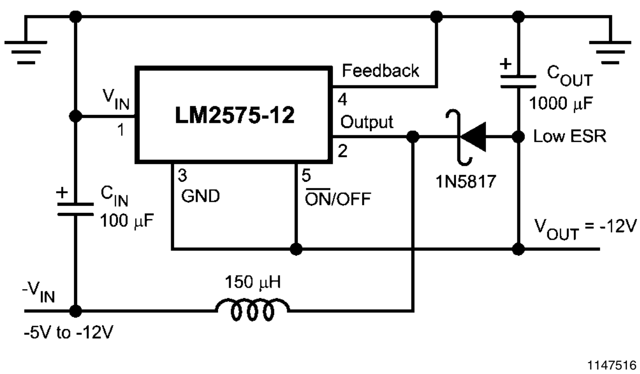


FIGURE 10. Inverting Buck-Boost Develops −12V

NEGATIVE BOOST REGULATOR

Another variation on the buck-boost topology is the negative boost configuration. The circuit in *Figure 11* accepts an input voltage ranging from −5V to −12V and provides a regulated −12V output. Input voltages greater than −12V will cause the output to rise above −12V, but will not damage the regulator.

Because of the boosting function of this type of regulator, the switch current is relatively high, especially at low input voltages. Output load current limitations are a result of the maximum current rating of the switch. Also, boost regulators can not provide current limiting load protection in the event of a shorted load, so some other means (such as a fuse) may be necessary.



Typical Load Current

200 mA for V_{IN} = −5.2V

500 mA for V_{IN} = −7V

Note: Pin numbers are for TO-220 package.

FIGURE 11. Negative Boost

UNDERVOLTAGE LOCKOUT

In some applications it is desirable to keep the regulator off until the input voltage reaches a certain threshold. An undervoltage lockout circuit which accomplishes this task is shown in *Figure 12*, while *Figure 13* shows the same circuit applied to a buck-boost configuration. These circuits keep the regulator off until the input voltage reaches a predetermined level.

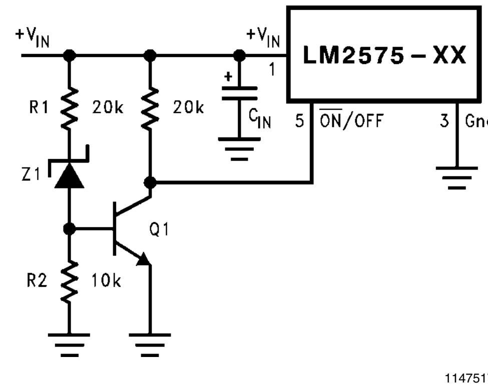
$$V_{TH} \approx V_{Z1} + 2V_{BE}\ (Q1)$$

DELAYED STARTUP

The $\overline{ON}$ /OFF pin can be used to provide a delayed startup feature as shown in *Figure 14*. With an input voltage of 20V and for the part values shown, the circuit provides approximately 10 ms of delay time before the circuit begins switching. Increasing the RC time constant can provide longer delay times. But excessively large RC time constants can cause problems with input voltages that are high in 60 Hz or 120 Hz ripple, by coupling the ripple into the $\overline{ON}$ /OFF pin.

ADJUSTABLE OUTPUT, LOW-RIPPLE POWER SUPPLY

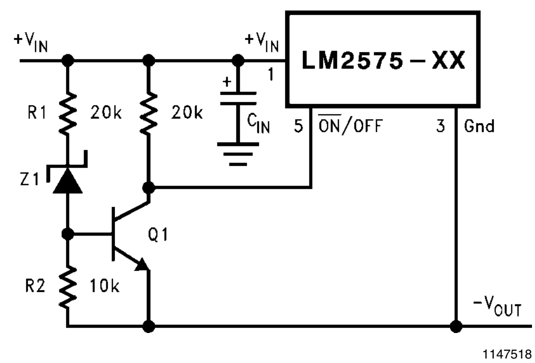
A 1A power supply that features an adjustable output voltage is shown in *Figure 15*. An additional L-C filter that reduces the output ripple by a factor of 10 or more is included in this circuit.



Note: Complete circuit not shown.

Note: Pin numbers are for the TO-220 package.

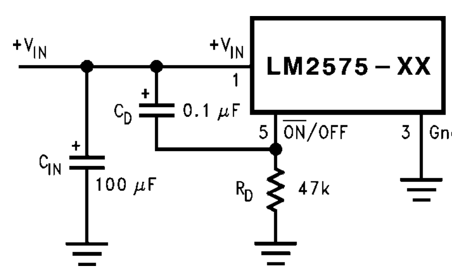
FIGURE 12. Undervoltage Lockout for Buck Circuit



Note: Complete circuit not shown (see *Figure 10*).

Note: Pin numbers are for the TO-220 package.

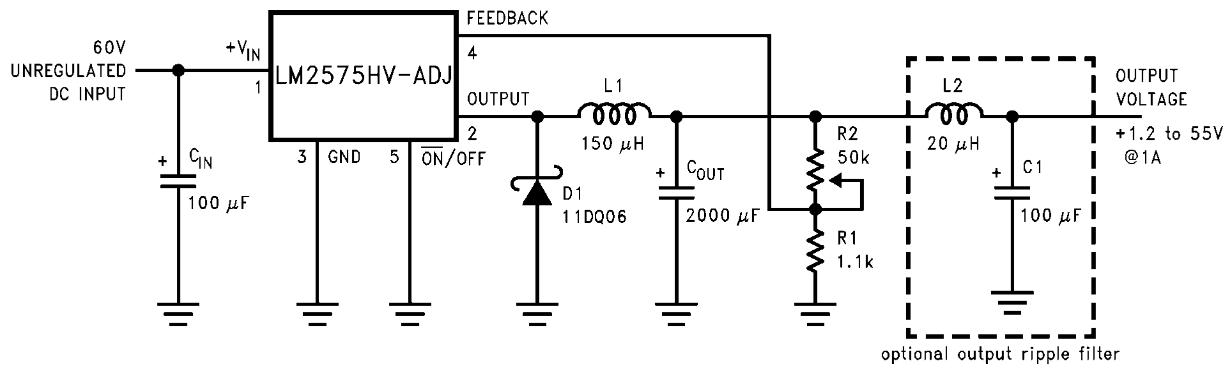
FIGURE 13. Undervoltage Lockout for Buck-Boost Circuit



Note: Complete circuit not shown.

Note: Pin numbers are for the TO-220 package.

FIGURE 14. Delayed Startup



Note: Pin numbers are for the TO-220 package.

FIGURE 15. 1.2V to 55V Adjustable 1A Power Supply with Low Output Ripple

Definition of Terms

BUCK REGULATOR

A switching regulator topology in which a higher voltage is converted to a lower voltage. Also known as a step-down switching regulator.

BUCK-BOOST REGULATOR

A switching regulator topology in which a positive voltage is converted to a negative voltage without a transformer.

DUTY CYCLE (D)

Ratio of the output switch's on-time to the oscillator period.

for buck regulator $$D = \frac{t_{ON}}{T} = \frac{V_{OUT}}{V_{IN}}$$

for buck-boost regulator $$D = \frac{t_{ON}}{T} = \frac{|V_O|}{|V_O| + V_{IN}}$$

CATCH DIODE OR CURRENT STEERING DIODE

The diode which provides a return path for the load current when the LM2575 switch is OFF.

EFFICIENCY (η)

The proportion of input power actually delivered to the load.

$$\eta = \frac{P_{OUT}}{P_{IN}} = \frac{P_{OUT}}{P_{OUT} + P_{LOSS}}$$

CAPACITOR EQUIVALENT SERIES RESISTANCE (ESR)

The purely resistive component of a real capacitor's impedance (see *Figure 16*). It causes power loss resulting in capacitor heating, which directly affects the capacitor's operating lifetime. When used as a switching regulator output filter, higher ESR values result in higher output ripple voltages.

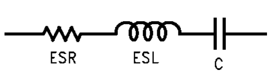


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FIGURE 16. Simple Model of a Real Capacitor

Most standard aluminum electrolytic capacitors in the 100 µF–1000 µF range have 0.5Ω to 0.1Ω ESR. Higher-grade capacitors ("low-ESR", "high-frequency", or "low-inductance"') in the 100 µF–1000 µF range generally have ESR of less than 0.15Ω.

EQUIVALENT SERIES INDUCTANCE (ESL)

The pure inductance component of a capacitor (see *Figure 16*). The amount of inductance is determined to a large extent on the capacitor's construction. In a buck regulator, this unwanted inductance causes voltage spikes to appear on the output.

OUTPUT RIPPLE VOLTAGE

The AC component of the switching regulator's output voltage. It is usually dominated by the output capacitor's ESR multiplied by the inductor's ripple current (ΔI_{IND}). The peak-to-peak value of this sawtooth ripple current can be determined by reading the Inductor Ripple Current section of the Application hints.

CAPACITOR RIPPLE CURRENT

RMS value of the maximum allowable alternating current at which a capacitor can be operated continuously at a specified temperature.

STANDBY QUIESCENT CURRENT (I_{STBY})

Supply current required by the LM2575 when in the standby mode ($\overline{ON}$ /OFF pin is driven to TTL-high voltage, thus turning the output switch OFF).

INDUCTOR RIPPLE CURRENT (ΔI_{IND})

The peak-to-peak value of the inductor current waveform, typically a sawtooth waveform when the regulator is operating in the continuous mode (vs. discontinuous mode).

CONTINUOUS/DISCONTINUOUS MODE OPERATION

Relates to the inductor current. In the continuous mode, the inductor current is always flowing and never drops to zero, vs. the discontinuous mode, where the inductor current drops to zero for a period of time in the normal switching cycle.

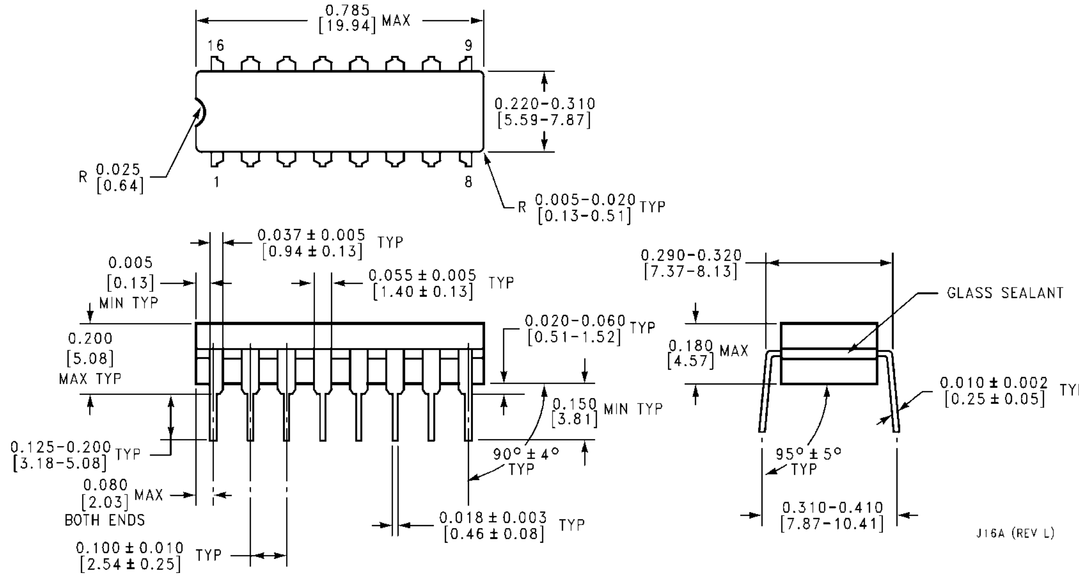
INDUCTOR SATURATION

The condition which exists when an inductor cannot hold any more magnetic flux. When an inductor saturates, the inductor appears less inductive and the resistive component dominates. Inductor current is then limited only by the DC resistance of the wire and the available source current.

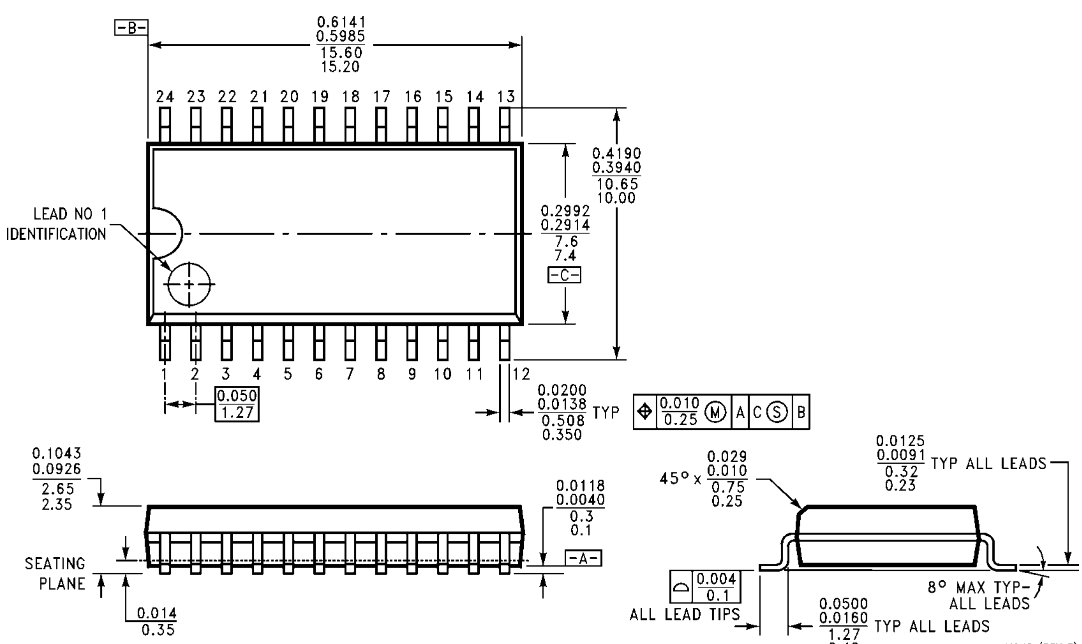
OPERATING VOLT MICROSECOND CONSTANT ($E \bullet T_{op}$)

The product (in Volt•µs) of the voltage applied to the inductor and the time the voltage is applied. This $E \bullet T_{op}$ constant is a measure of the energy handling capability of an inductor and is dependent upon the type of core, the core area, the number of turns, and the duty cycle.

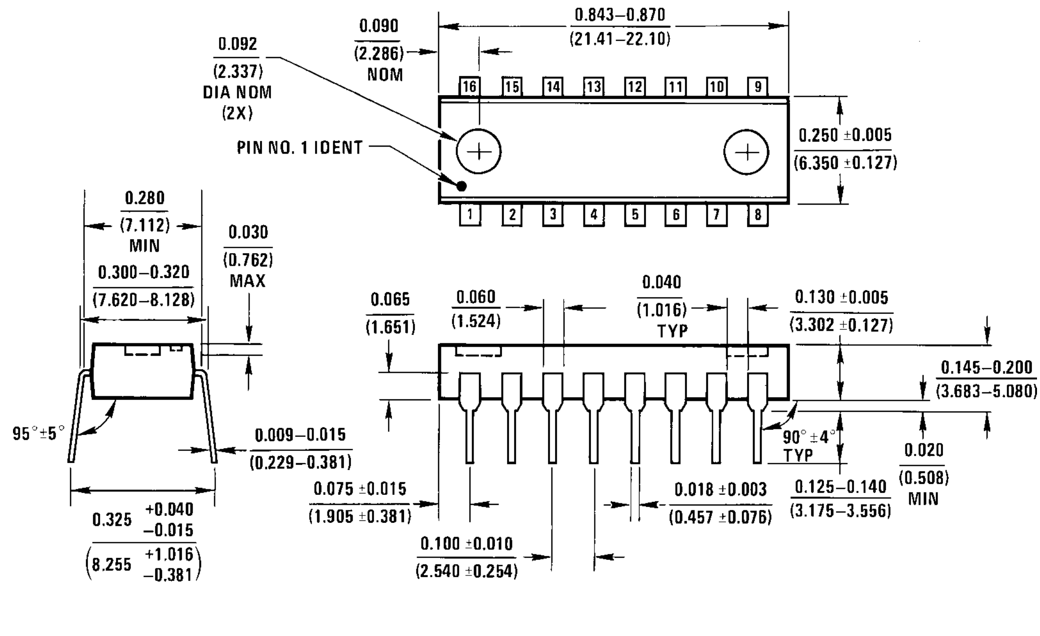
Physical Dimensions inches (millimeters) unless otherwise noted



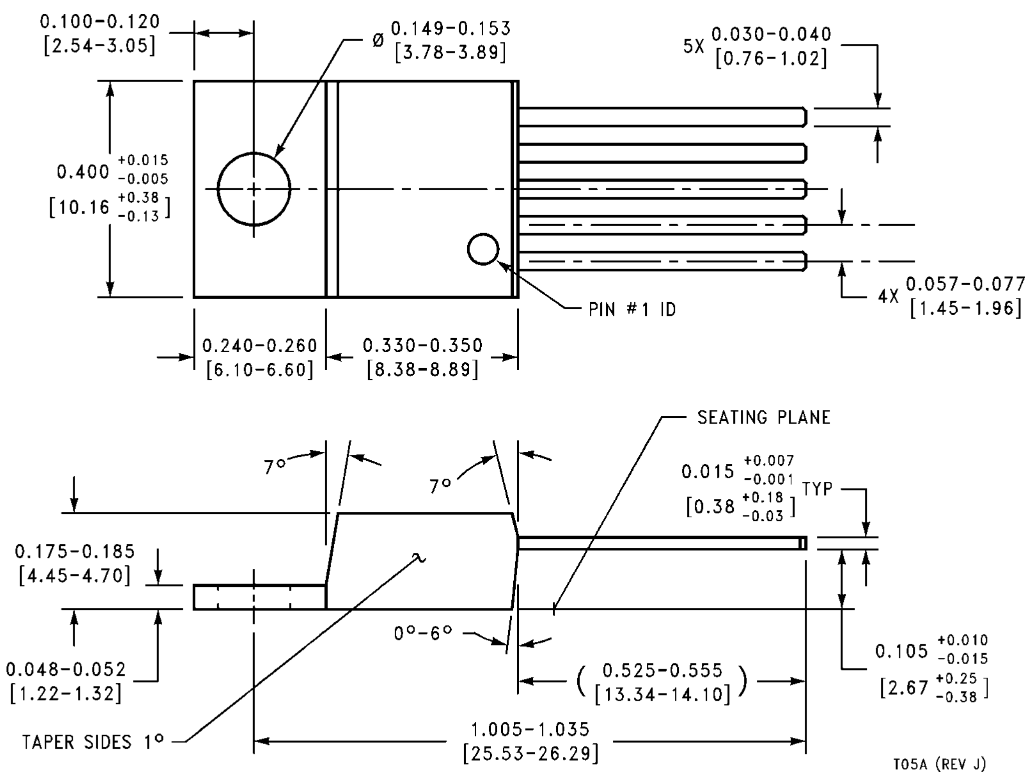
16-Lead Ceramic Dual-in-Line (J)
Order Number LM1575J-3.3/883, LM1575J-5.0/883,
LM1575J-12/883, LM1575J-15/883, or LM1575J-ADJ/883
NS Package Number J16A



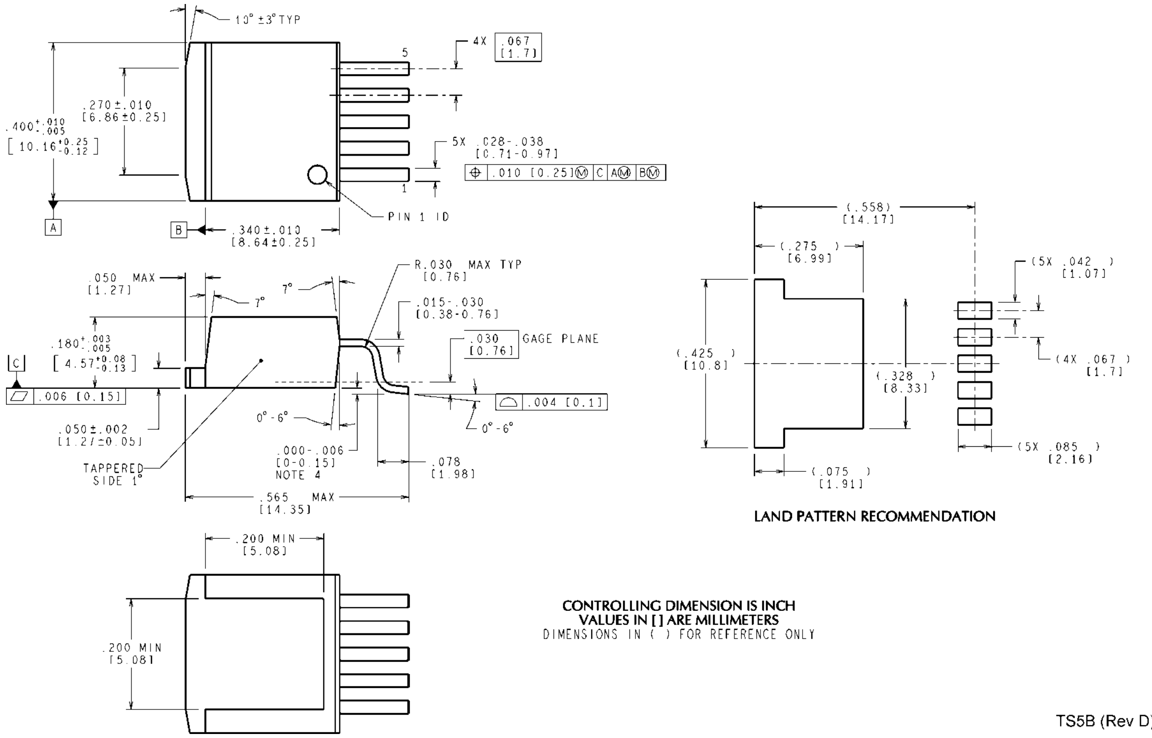
24-Lead Wide Surface Mount (WM)
Order Number LM2575M-5.0, LM2575HVM-5.0, LM2575M-12,
LM2575HVM-12, LM2575M-15, LM2575HVM-15,
LM2575M-ADJ or LM2575HVM-ADJ
NS Package Number M24B



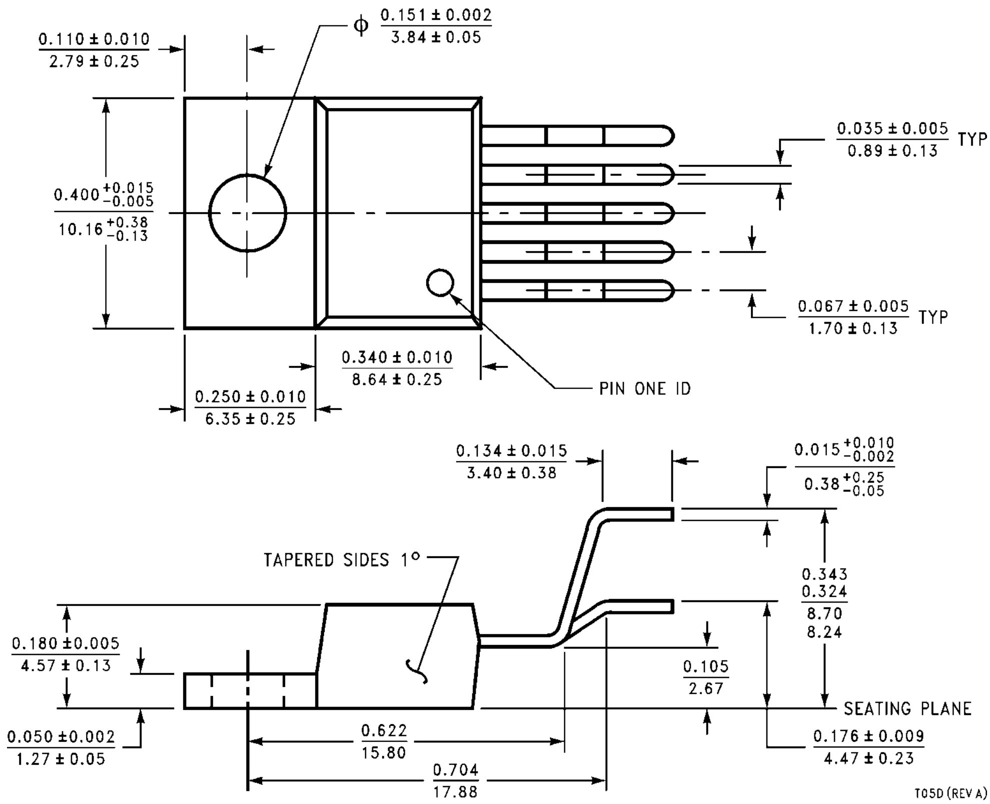
16-Lead Molded DIP (N)
Order Number LM2575N-5.0, LM2575HVN-5.0, LM2575N-12, LM2575HVN-12, LM2575N-15, LM2575HVN-15, LM2575N-ADJ or LM2575HVN-ADJ
NS Package Number N16A



5-Lead TO-220 (T)
Order Number LM2575T-3.3, LM2575HVT-3.3, LM2575T-5.0, LM2575HVT-5.0, LM2575T-12, LM2575HVT-12, LM2575T-15, LM2575HVT-15, LM2575T-ADJ or LM2575HVT-ADJ
NS Package Number T05A



TO-263, Molded, 5-Lead Surface Mount
Order Number LM2575S-3.3, LM2575HVS-3.3, LM2575S-5.0, LM2575HVS-5.0, LM2575S-12, LM2575HVS-12, LM2575S-15, LM2575HVS-15, LM2575S-ADJ or LM2575HVS-ADJ
NS Package Number TS5B



Bent, Staggered 5-Lead TO-220 (T)
Order Number LM2575T-3.3 Flow LB03, LM2575HVT-3.3 Flow LB03,
LM2575T-5.0 Flow LB03, LM2575HVT-5.0 Flow LB03,
LM2575T-12 Flow LB03, LM2575HVT-12 Flow LB03,
LM2575T-15 Flow LB03, LM2575HVT-15 Flow LB03,
LM2575T-ADJ Flow LB03 or LM2575HVT-ADJ Flow LB03
NS Package Number T05D

Notes

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REVISITING THE 12-V POWER SUPPLY

◊ Some readers have caught the "power dissipated in the filter capacitor" goof that was in the November 1992 *QST* power-supply article.[1] The readers are absolutely correct: An *ideal* capacitor does not dissipate power. This error occurred in an attempt to address an additional requirement when choosing a filter capacitor: the capacitor's ability to handle ripple current.

In a high-current power-supply design, the filter capacitor has to supply a large amount of current between the peaks of the rectified dc output of the full-wave bridge. This drives up the physical capacitor size, both to achieve the required capacitance and to reduce the capacitor's internal resistances. *Power is dissipated in the capacitor only because of the current and the capacitor's internal resistances.* The power dissipated should be small as long as a physically large capacitor is used.

It's best to select a capacitor that provides a low ripple voltage on the unregulated dc prior to the regulator. This allows the use of a lower transformer secondary voltage which, in turn, reduces the pass-transistor power dissipation. Use the following formula (from the *Handbook*) to determine the capacitance required for a given output current and ripple voltage. This allows selection of capacitors for various load currents. Remember: Capacitors can be connected in parallel, if needed, to get the desired capacitance value.

In this equation, C is in microfarads (μF), f is in hertz (120 Hz for a full-wave rectifier), and E_{ripple} is expressed in peak-to-peak volts.

$$C = \frac{(I_l)\left(\frac{1}{f}\right)}{E_{ripple}} \times 10^6 \qquad \text{(Eq 1)}$$

Using this equation yields a value of 45,000 μF for the filter capacitor in a 15-A supply with 1 V RMS (approximately 2.8 V P-P) ripple on the unregulated dc. Additional capacitance proportionally reduces the ripple voltage; a higher load current proportionally increases the required capacitance value.

In my prototype supply, I used a filter capacitor I had on hand. The preceding formula shows a 2.8 V P-P ripple with a 45,000 μF capacitor. The ripple voltage on the prototype (with a 19,000 μF capacitor) measured about 3.3 V P-P at a load current of 15 A. The transformer secondary voltage under load is high enough so that the regulator provides the required output without losing regulation because of the ripple voltage. However, the ripple voltage causes additional power to be dissipated in the pass transistors.

Heat-Sink Size

One reader commented that the internal heat sink was rather small for the amount of power dissipated in the transistors, and that the fins should be oriented vertically rather than horizontally. The heat-sink arrangement in my prototype supply was determined by parts availability and is adequate for my application because the supply is used for intermittent operation—such as with an SSB transmitter.

Those contemplating constructing a *continuous-duty* supply should definitely increase the heat-sink area and orient the fins vertically. The heat sink should also be mounted *outside* the cabinet where it is offered unrestricted airflow. For additional information on thermal design, see the *Handbook's* power-supply section.

Fusing

One reader suggested placing a fuse between the filter capacitor and the regulator. If this was done in the design presented, the fuse *would never open!* Even with a short on the output, the supply *limits the output current* to a set maximum value (assuming the fuse rating was higher than the current-limit value). However, adding a circuit breaker or fuse between the filter capacitor and the regulator, and moving the *hot* side of the overvoltage SCR to the regulator side of the circuit breaker can provide additional protection.

With the SCR gate still sensing the output voltage, an overvoltage condition will turn on the SCR. When that happens, the SCR looks like a short circuit across the filter capacitor. Then, the circuit breaker will open, disconnecting power from the regulator. This provides better protection against overvoltage because the power is disconnected from the output. Also, power is no longer dissipated in the pass transistors when the SCR is conducting. The fuse or circuit breaker also acts as a backup in the event that the regulator's current-limiting circuit malfunctions. For best results, a low-value resistor should be added in series with the SCR to create a "soft-short" condition. Size the resistor to draw enough current to open the fuse or circuit breaker used.

Pass Transistors

Another reader had some important information about the pass transistors. The 2N3055 transistors should have a resistor connected from base to emitter to ensure that they can be completely cut off under a no-load condition. The value of the base-to-emitter resistors should be about 33 ohms. Connect the resistors directly between the base and emitter of each transistor (Q1-Q3).

The base-emitter resistors are required because the collector-to-base leakage current can be as high as 5 mA when the transistors are hot and operated at maximum V_{CE}. Because the V_{CE} of the transistors is well under 20 V, we can use a lower value for the leakage current.

Assuming that we use the three transistors, and that since the V_{CE} is under 20 V, a value of about 3 mA for the maximum leakage of a new transistor would be reasonable. If no base-to-emitter resistors are used, the resulting emitter current can be calculated by multiplying the transistor gain, plus one, by the leakage current. Over time, the transistor leakage will increase (it can be up to 10 times higher for parts operated near their limits). The parts in my design are not operated near their limits, therefore the resistors are sized to provide a base-to-emitter voltage of about 0.3 with a 9-mA leakage current. This value is a compromise based on the use of three transistors, a low V_{CE}, and a collector current of only 5 A per transistor. Because the base leads are tied together—and the emitter leads are also tied together through the emitter resistors—the net resistance of the three 33-ohm resistors in parallel is about 11 ohms.

The measured V_{BE} of the transistors with a 15-A load is about 1 V, for a power in each 33-ohm resistor of 30 mW. A standard ½-W resistor can be used and fits easily across the base and emitter leads of the TO-3 package transistors. The TIP-112 has similar resistors built into its case, so we don't have to add them externally.

To accommodate the leakage current of the transistors (now reduced by the base-emitter resistors), it's best to have an internal load on the supply output. The regulator's voltage divider is a 1.8-mA load at an output of 12 V dc. That may be enough for new transistors, but is not enough to accommodate a possible 30- to 40-mA leakage current from aged transistors. A resistor of about 300 ohms or less connected from the junction of R1 through R4 to ground provides a path for the current. The maximum value of the resistor can be calculated by R = E ÷ I, where E is the regulated output voltage and I is the leakage current of the transistors. The next-smallest standard value should be used. Moving the 75-ohm bleeder resistor to this location would also work.

[1]E. Oscarson, "A 12-V, 15-A Power Supply," *QST*, Nov 1992, pp 36-41.

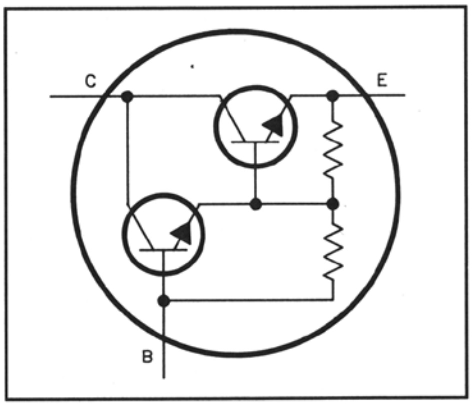

Fig 1—Correct representation of Q5, the TIP112 transistor.

Cosmetics

Lastly, a cosmetic error that got into the design is the schematic representation of Q5, the TIP-112 transistor. The correct diagram is shown in Fig 1.

In the Fig 1 caption, R10 should be identified as 330 Ω, ½ W. Delete R11.

I would like to thank all the readers for their comments and for the additional information.—*Ed Oscarson, WA1TWX, 70 Behrens Rd, New Hartford, CT 06057*

MORE ON ELEVATED RADIALS

◊ I've been doing some NEC modeling of AM broadcast antennas, and I believe that the results will be of interest to others. This latest research indicates that elevated radials can be used in conjunction with a *ground-mounted* vertical monopole to achieve results that are as good as can be obtained from a conventional ground-mounted tower with 120 buried radials.

The general layout is given in Fig 2. The ground rods can be omitted without hurting the electrical performance of the antenna, and the masts, which are used to support the elevated radials, can be conductive (metal) or nonconductive (the difference in radiated field strength is 3% or less).

The vertical monopole (ground-mounted on a base insulator) has a physical length of 0.25 λ. The four radials slant upward from the feed point at a 45° angle until reaching the desired height (H), then extend outward horizontally from that point. The computer analysis appears to indicate that the height of the radials (H) can be as little as *4 to 5 feet* on 80 meters, but a height of 10 to 15 feet produces a slight increase in signal strength, and would be better from a safety standpoint.

How long should the radials be for best performance? As a *general* rule, if the radials are suspended at a height of H above ground, then their length should be equal to 0.25 λ *plus* H. In other words, if H = 10 feet, then at 3.8 MHz, the length of the radials should be 0.25 λ (which is 64.7 feet) + 10 feet = 74.7 feet. The feed-point impedance depends on many variables, but NEC-GSD predicts values in the range of 22 to 31 ohms. Complete information can be found in A. Christman and R. Radcliff, "Using Elevated Radials with Ground-Mounted Towers," *IEEE Transactions on Broadcasting*, Vol 37, No 3, Sep 1991, pp 77-82. —*Al Christman, KB8I, Grove City College, 100 Campus Dr, Grove City, PA 16127-2104*

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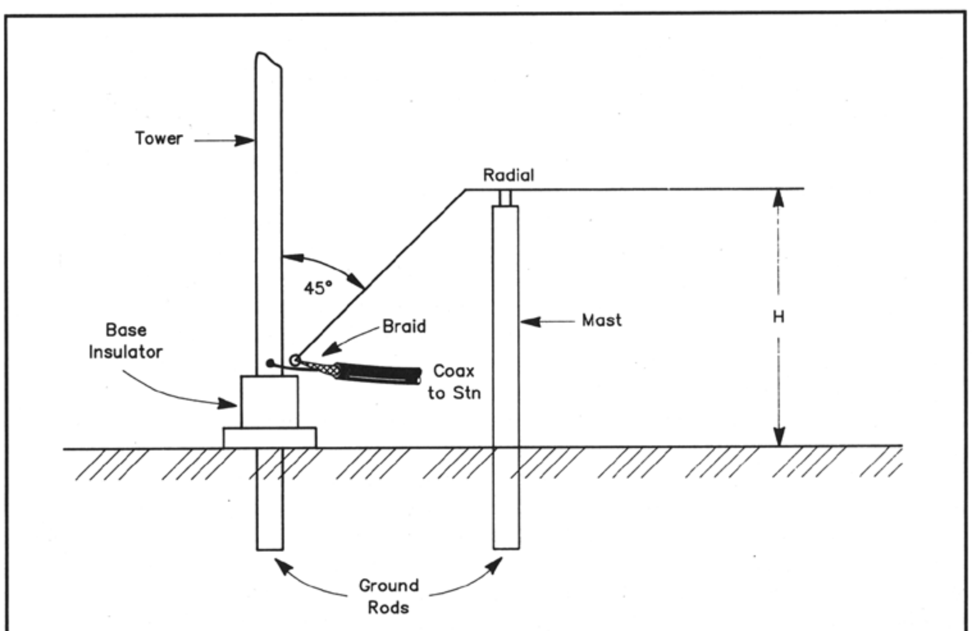


Fig 2—Elevated radials can be used in conjunction with a ground-mounted vertical monopole to achieve results equal to those of a conventional ground-mounted tower with 120 buried radials. The ground rods can be omitted without hurting the electrical performance of the antenna. The masts supporting the elevated radials can be conductive (metal) or nonconductive (the difference in radiated field strength is 3% or less).

ARRL Handbook CD

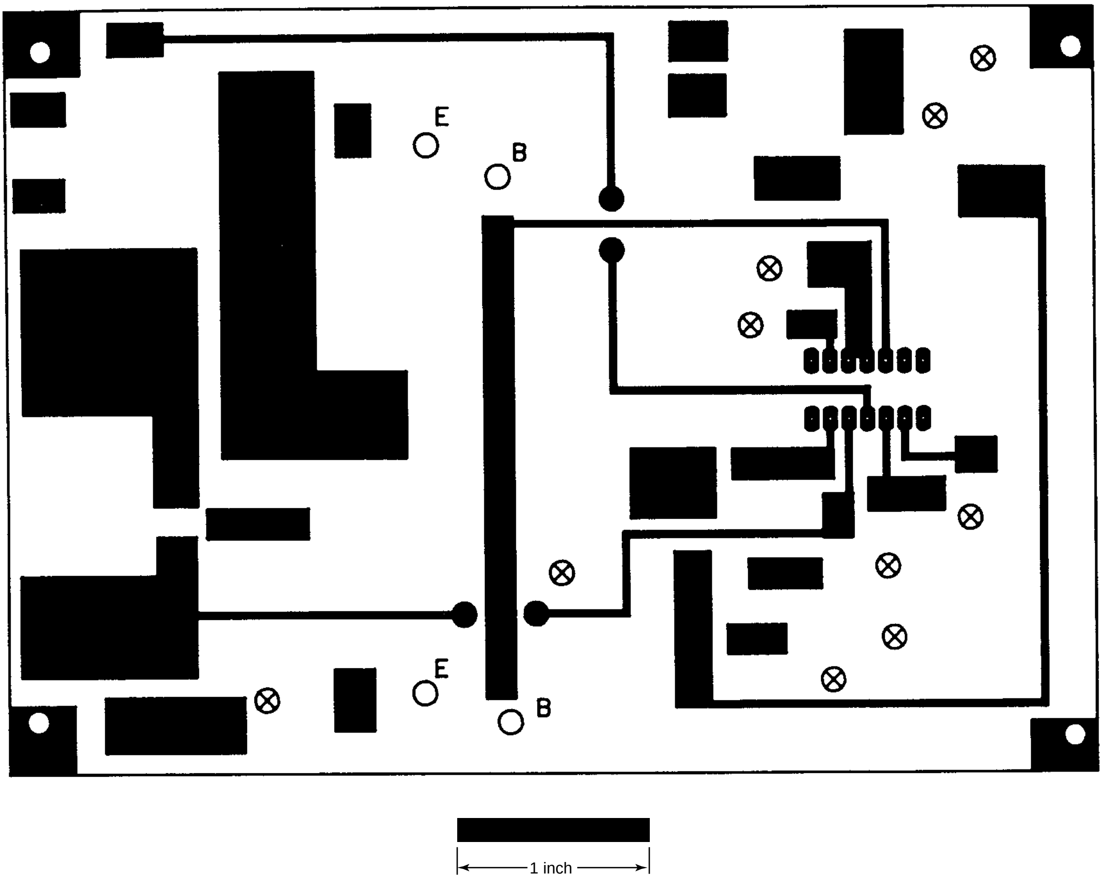
Template File

Title:	Sabin Power Supply
Chapter:	7
Topic:	A Series-Regulated 4.5- to 25-V, 2.5-A Power Supply

Template contains:

PC board etching pattern.

Sabin Power Supply



A Series Regulated 4.5- to 25-V, 2.5-A Power Supply

The Series Regulator Power Supply: A Closer Look

By William E. Sabin, WØIYH
1400 Harold Dr, SE, Cedar Rapids, IA 52403

For many applications, the rapidly advancing technology of the switching regulator power supply has made it the preferable approach, especially where light weight, small size and high efficiency are very important. But for my basement laboratory requirements I finally decided to build a series regulator supply. During the lab-bench development of sensitive low-level circuitry it is necessary to be sure that the power supply is beyond reproach and not contributing, in confusing ways, to various problems. The "switchers" can be a later addition to the equipment design.

In the course of the initial design work it occurred to me that my understanding of the series regulator was inadequate. The excellent material in references 1 and 2 helped, but several other questions came up. I would like to share with you my additional investigations, and describe the design and construction of the supply.

Requirements

The requirements which I believed essential are listed below. Compromises in cost and complexity are also apparent in these specs.

1) Continuously variable output voltage from 4.5 to 25.0 V. The extra circuitry required to go to zero was not justified.

2) Load currents from 0.0 to 2.5 A, continuous duty.

3) Tight load regulation, better than 0.03%, no load to 2.0 A, 0.1% to 2.5 A.

4) Line regulation 0.01% at 2.0 A dc for 117 to 122 V ac.

5) Very low ac ripple, less than 2 *microvolts* RMS at 2.0 A load.

6) Very low random noise, less than 2 *microvolts* RMS in the 0.1 Hz to 500 kHz band.

7) Use off-the-shelf transformer and other easily obtainable parts.

8) Excellent response to load fluctuations and transients; low output impedance.

After reviewing the switching regulator literature, in particular references 2 and 3, I felt that I could meet these difficult specs much more easily with the series regulator approach, especially since size, efficiency and heat dissipation were not important constraints in this case.

Implementation

Fig 1 is a system diagram of the supply, showing the various elements involved in the up front design. The analyses, simulations, various tests performed and the wiring interconnect approach can all be discussed with respect to this diagram. Fig 1 also includes three test circuits.

A type 723 regulator chip was used because of its simplicity and because its reference voltage is brought out to a separate pin so that I could filter the reference noise, typical of Zener diodes, to a very low level with C5, as suggested in the data sheet for the 723 and later verified to be true. The current-limiting circuitry is also accessible at pins 2 and 3 and is activated by the voltage drop across R2 + R3.

The most important regulator considerations can be described as follows:

A) When the output is 25.0 V at 2.5 A at a line voltage of 117 V ac, the Vcb of Q1 and Q2, and also the difference between pins 11, 12 and pin 10 of the regulator chip, when the ripple waveform on C1 is at its minimum (trough) value, must be sufficient to avoid a dropout of regulation and an increase in ripple output. A large value of C1 is used to reduce ripple voltage. Also, the R1, C2 combination reduces the ac on the regulator chip by a factor of 25 and this helped to avoid the need for an extremely large value for C1. Recall also, that the current flow in Q1 and Q2 is not strongly influenced by collector voltage variations if the base-to-emitter voltage is constant. The alternative to these steps would have been a special higher voltage transformer, with which I did not want to become involved.

B) When the output voltage is 4.5 V at 2.5 A at a line voltage of 122 V ac, the power dissipation in Q1 and Q2 is about 65 W. The heat sink requirements are established at this condition. A room temperature of 20° C is assumed. To minimize heating, it is desirable to have a power transformer with as low a voltage as possible, and the steps taken in A) help to assure this. Minimizing other voltage drops that occur between the emitters of Q1 and Q2 and the output terminal helped to assure that a standard 25.2-V ac transformer would do the job. A bend-back circuit prevents overheating when the output is short circuited.

C) The series regulator is a good example of a feedback control system. The open-loop gain and bandwidth, the phase and gain margins and the transient response are important factors. The goal was to maximize the closed-loop performance of the regulator. The approach was to use a high value of open-loop gain and to establish the open-loop frequency response mainly by means of (a) the RC lowpass filter consisting of C6, the resistances which separate Q1 and Q2 and the output resistance of Q1 and Q2 and (b) a single small capacitor C4 at the regulator chip.

D) The mechanical construction should emphasize heat removal, but a cooling fan would not be used. The maximum load current would be scaled to a level which the components could tolerate. A bend-back circuit would be used to limit the maximum heat dissipation with a short circuited output.

Current Limiting

In Fig 1, when the voltage drop from pin 2 to pin 3 of the 723 regulator reaches about 0.62 V, the output current becomes limited to the value of 0.62 / (R2 + R3) A. R3 is a 0.11-Ω resistor which acts as a shunt for the digital meter which measures load current. Resistor R2 is switch selectable (shown in Fig 4) to produce three values of maximum current, approximately 0.1 A, 0.5 A and 2.5 A. This very simple approach will protect delicate circuits and PC boards from burnout destruction.

Test Circuits

Fig 1 shows three test circuits. One is an adjustable load test circuit which can be modulated linearly (almost) by a sine or triangle wave using a function generator which has a dc offset adjustment (so that the waveform always has positive polarity), or by a bidirectional square wave. This circuit is used to test the response to various kinds of load fluctuations and has proved to be very informative, as discussed later.

The second test circuit (loop gain tester) is inserted into the regulator loop so that a test signal can be inserted in series with the loop in order to measure the *open-loop* gain and frequency response. But you will notice that, at dc and very low ac frequencies, the loop is closed through Rb and Ra, and the dc output voltage is being pretty well regulated, something which is essential to the loop testing. By observing the magnitude (and *rate* of change) of the frequency response it is possible to deduce information about the phase shift[4]. With this information available, the gain and phase margins and therefore the regulation, stability, transient response and output impedance of the closed-loop regulator can be estimated.

The third test circuit is a two-stage opamp preamplifier and oscilloscope, to measure very small signals in the 0.1 Hz to 400 kHz range.

Open-Loop Testing

The test signal which is applied to points A, A' is reduced 60 dB by Rd and Rc (for ease of adjustment). Capacitor C couples V_a, the voltage across Rc, to the 723 chip through Ra. Ra is roughly the resistance which the 723 sees in normal operation. The test signal is amplified by, at most, 74 dB on its way clockwise around the loop to the right-hand end of Rb. It is then attenuated by the factor 20 log (Rb / Rc) = 100 dB. This means that the "leak through" back to the 723 input is much smaller than the V_a that we started with, if the frequency is 2 Hz or greater. At dc and very low frequencies the regulator functions *somewhat* normally. Above 2 Hz, then, the magnitude of the open-loop gain at the test frequency is very nearly the ratio of $|V_o| / |V_a|$.

The first benefit of this tester was that it isolated an instability in the 723 chip. An oscillation at several hundred kHz was cured by C4 (33 pF) and C3 (100 µF / 50 V with very short leads). Normally, one would suspect the oscillation to involve the overall loop, but this was not the case. This kind of instability is common in feedback control systems, where everything appears to be functioning

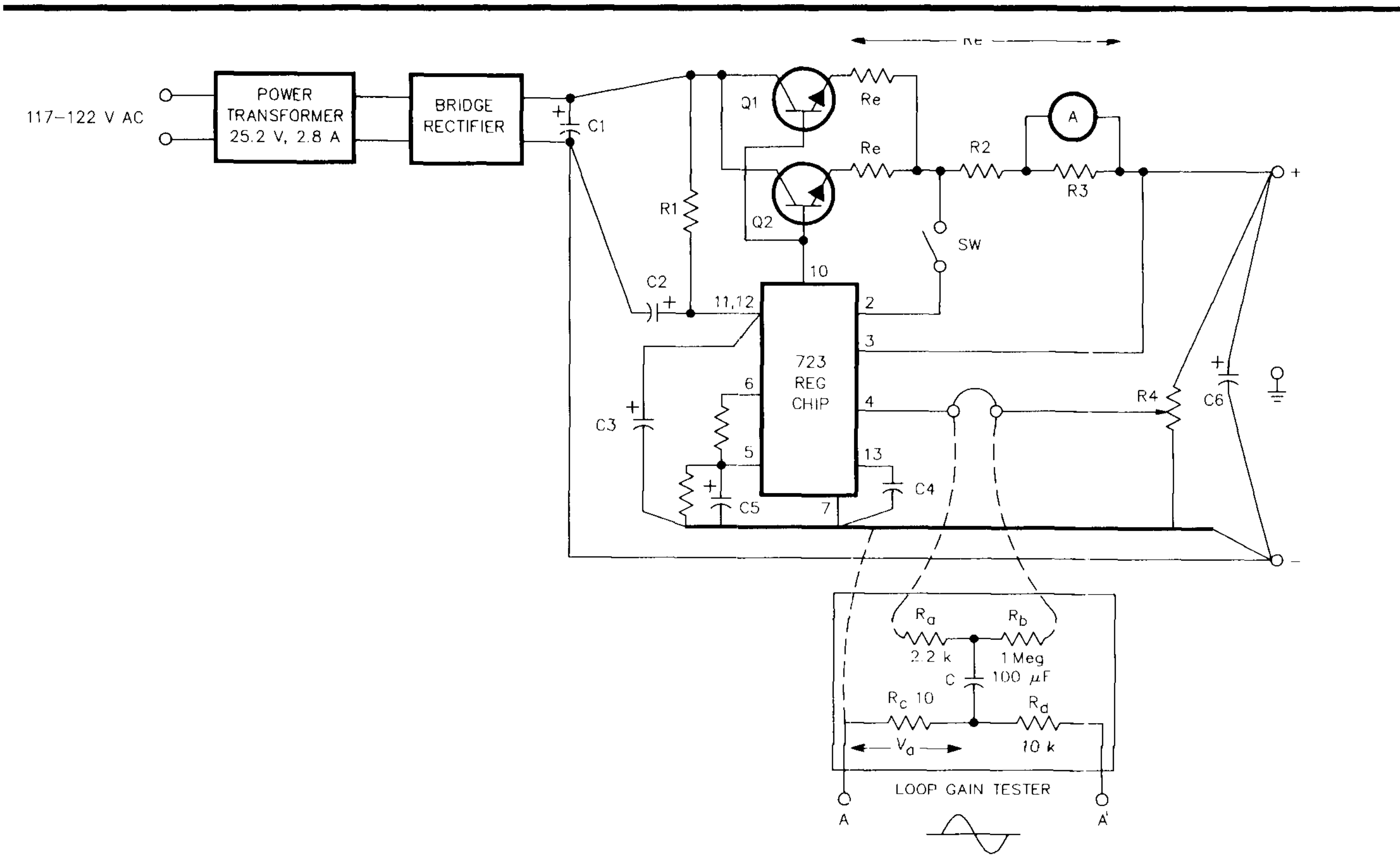


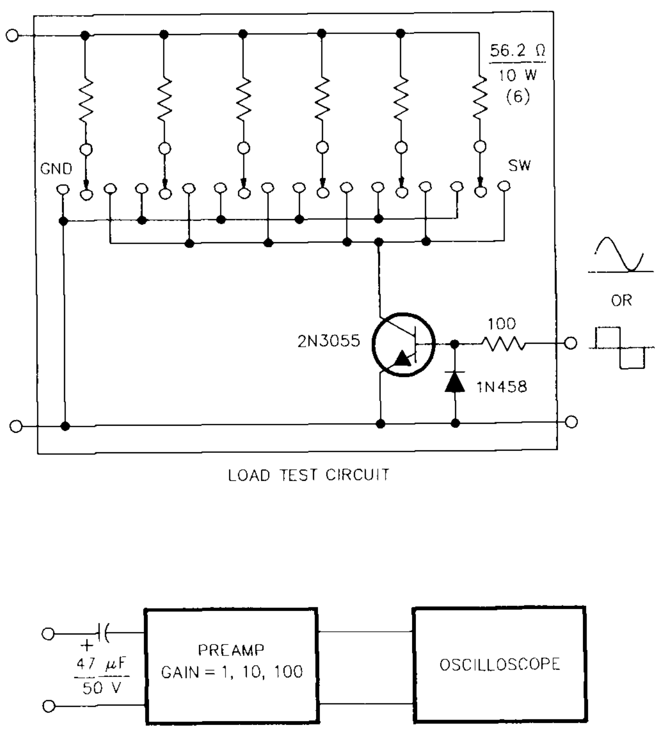
Fig 1—Simplified diagram used to discuss design principles.

(usually not to full specification) but an embedded element is not stable.

Open-Loop Frequency Response

Looking at Fig 1, the test signal is amplified by about 74 dB (on a voltage basis), on its way clockwise to the emitters of Q1 and Q2. It is then lowpass filtered by C6 and RE (the combination of Re + R2 + R3). It is then divided down by potentiometer R4 (when the output is 25 V, this division is greatest because the pot position is nearest to ground). The open loop gain is the product of these three factors and its greatest value is 59 dB at 25-V output and 74 dB at 4.5 V.

Fig 2 is the open-loop frequency response to the top of R4 when RE is set for the 2.5-A current range. At very low frequencies, the drop-off is due to the gradual closure of the feedback loop, as mentioned above. At higher frequencies the roll off is due to the combined effects of C4 and C6 and occurs at a 6 dB per octave rate (within the errors of my instrumentation). The corner frequency is about 280 Hz, which is $1 / (2 \times \pi \times RE \times C6)$ where RE = 0.57 Ω and C6 = 1000 μF. For comparison, a reference curve (6 dB per octave at the high and low ends) is superimposed. At about 1.2 kHz or so the reactance of C6 is roughly equal to the ESR (equivalent series resistance) of C6 (about 0.13 Ω for a small 1000 μF aluminum electrolytic, verified by direct measurement). Beyond this frequency, the impedance of C6 does not diminish and C4 takes over, thereby maintaining the 6 dB per octave roll off rate. Careful measurements and computer simulations of the regulator loop verified that C4 and C6 do in fact collaborate in this manner pretty well.



This was the effect I wanted. At 120 Hz (the major ripple frequency) the loop gain is maximum so that the regulator loop is working hard to suppress ripple output. At higher frequencies, the roll off rate of 6 dB per octave implies a loop phase shift in the neighborhood of 90°, which assures closed-loop stability and good transient response. Closed-loop transient response tests using a square-wave signal into the load test circuit verify the absence of ringing and large overshoots.

When RE is set to the 0.5 A or 0.1 A positions, the corner frequencies are 58 Hz or 12 Hz and the roll off rate remains 6 dB per octave as before. In these positions, the loop gain at 120 Hz is reduced, however the ripple voltage across C1 is also greatly reduced at these lighter load currents, and the final result is that the output ripple remains very low.

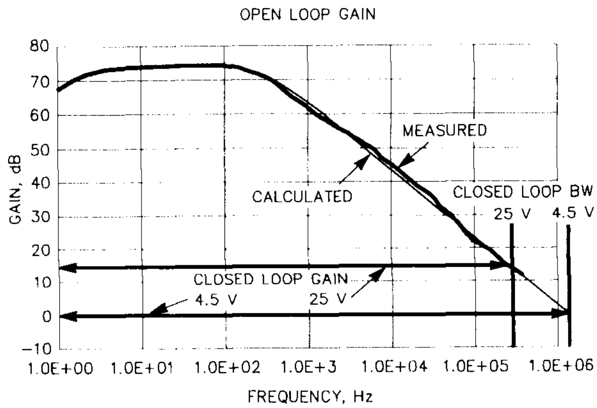


Fig 2—Regulator loop frequency response.

Closed-Loop Response

The closed-loop gain of the regulator is 20 times the log of the ratio of the output voltage, 4.5 V min to 25.0 V max, to the reference voltage, 4.5 V. Fig 2 shows the locations of the min and max gain values and also the corresponding closed-loop bandwidths. By locating the 280-Hz corner frequency fairly close to the 120-Hz ripple frequency, we have made the closed-loop bandwidth no wider than is necessary, which is commendable in a voltage regulator.

Another important parameter in a regulator is its closed-loop output impedance. Fig 3 shows a computer simulation of this. Mathematical analysis and actual measurements using the load test circuit with a sine-wave test signal corroborate the simulations quite well. Two results are shown. In (a) the C6 component (Fig 1) is removed and C4 is increased so that the 280-Hz corner frequency is maintained, as we discussed before. At low frequency, the output impedance should be RE (0.57 Ω) divided by the open-loop voltage gain (5000 max), which equals about 0.11 milliΩ). Above 280 Hz, though, the output impedance increases rapidly because the open-loop gain is decreasing. It will eventually reach the value of RE.

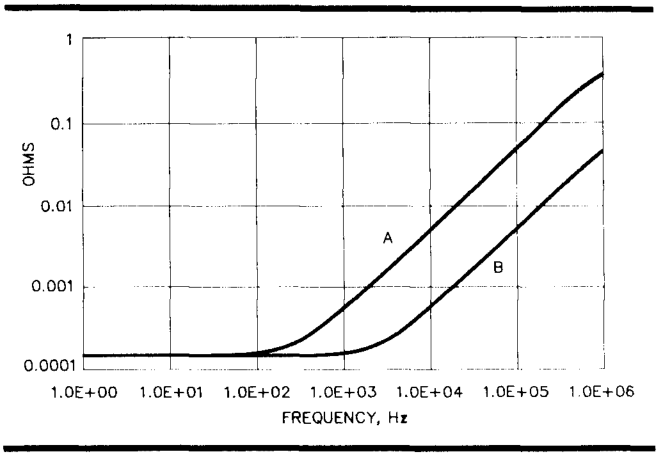


Fig 3—Output impedance magnitude.

In curve (b) the original values of C6 and C4 are used and the output impedance remains low out to about 1200 Hz and then increases but still remains much lower than in curve (a). This is because C4 is now *smaller* than in curve (a) and *it* mainly determines the frequency characteristic. In other words, the impedance of C6 (its reactance plus its ESR) is in parallel with the much lower output impedance of a high-gain feedback amplifier and therefore it is much less influential in determining the power supply output impedance. This situation gradually changes as frequency gets higher, as some study of the following equation will show.

$$\frac{1}{Z_{out}} = \frac{1}{RE}\left[\frac{K\beta}{1 + jf/f4} + 1\right] + \frac{1}{ESR}\left[\frac{jf/f6}{1 + jf/f6}\right]$$

K = amplifier gain
β = R4 divider ratio
f4 = corner frequency for C4
f6 = corner frequency for C6 and its ESR
RE see Fig 1

The result of this discussion is that the output impedance characteristic of the power supply is reduced at frequencies which may be significant in certain applications. Furthermore, it can be reduced to levels (by virtue of the feedback) which a practical capacitor may not be capable of. Of course, to take advantage of this lower impedance the regulator must be located extremely close to the application (remote sensing is also a possibility). Recall, also, that some small value of C4 was needed to stabilize the 723 chip.

Regulation and Wiring

When the line voltage was changed from 117 to 122-V ac, at about 25-V dc, 2.0 A, the output voltage varied less than 0.01%. When the dc load was changed from 0 to 2.0 A the output changed less than 0.03%. Heavy-duty binding posts reduced a small but significant voltage drop from the rear to the front of the front panel. When extremely tight regulation and low output impedance are important, the power leads to the load must be very short, heavy straps. Multiple loads should "fan out," both plus and minus, from the binding posts, not connected in tandem.

Fig 1 details the method of wiring the critical circuits and the following items are enumerated. Each was verified to be important to achieve the clean performance described above.

A) C1 is wired with short heavy leads to present minimum impedance for ac ripple. C2 returns to the negative side of C1.

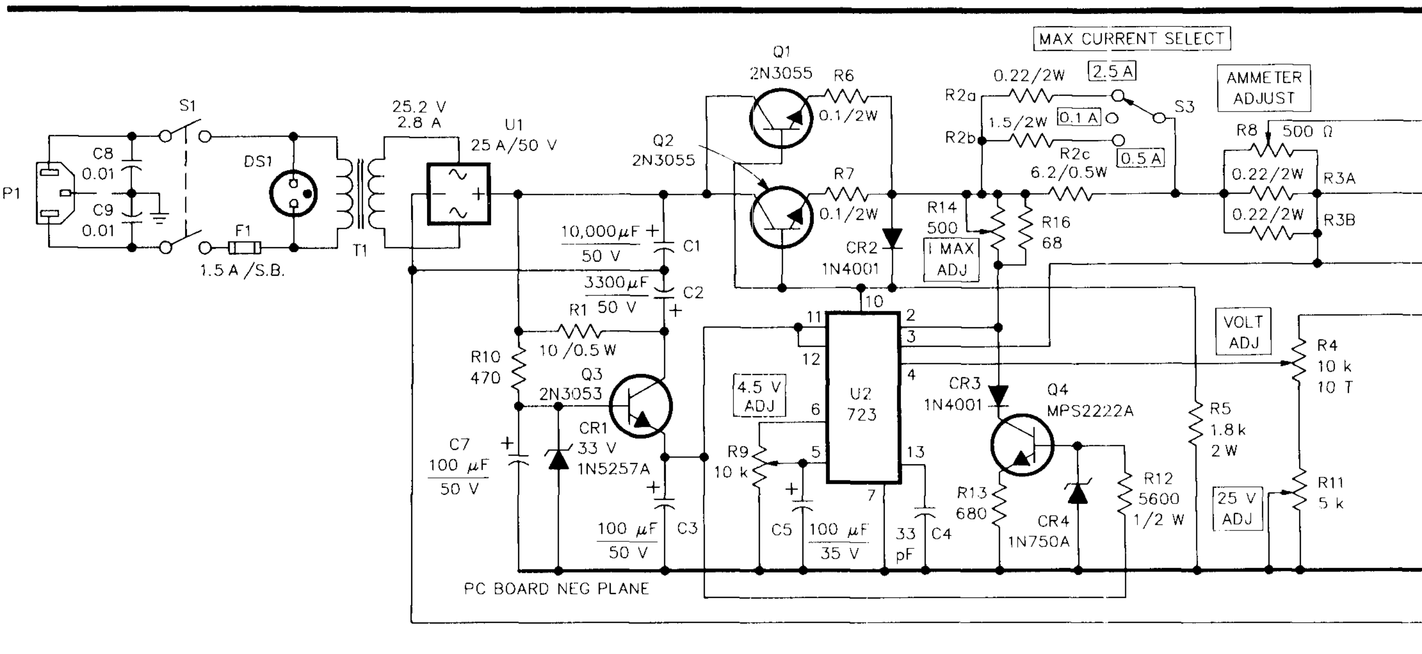


Fig 4—Complete schematic of regulated power supply.

B) C6 is connected directly to the binding posts. R4 returns to the regulator negative plane.

C) The regulator PC board negative plane is connected to the negative output terminal at a single point.

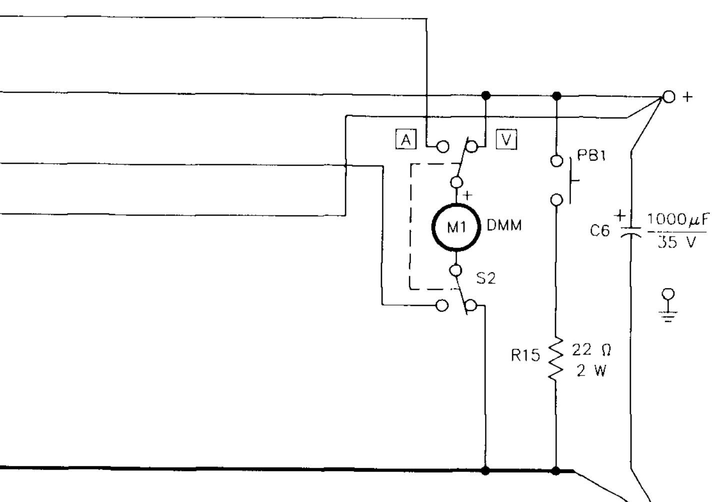
D) The bottom of C1 connects directly to the negative binding post with a heavy lead so that load current fluctuations prefer not to flow on the regulator PC board ground plane and thus influence the operation of the regulator in unpredictable ways.

Complete Schematic

The complete schematic in Fig 4 contains a few features not previously mentioned. The circuitry of Q3, R10, C7 and CR1 prevents the voltage on pins 11 and 12 of U2 from exceeding the 40-V max rating, especially at light loading and high line voltage. C7 eliminates a very small ac ripple at the dc output. As load current increases, the voltage at C1 decreases and Q3 then goes into saturation.

When R4 is turned in a direction to reduce output voltage, U2, Q1 and Q2 are turned off until C6 can discharge to the lower voltage. It was noticed that the emitter-to-base junctions of Q1 and Q2 were going into reverse breakdown (about 2.0 V) so that C6 could discharge through R5. The purpose of CR2 is to prevent this breakdown, which may or may not be dangerous (no actual problem was noticed). The purpose of R5 is to provide a minimum output loading on U2.

The bend-back circuit is interesting. Q4, CR4 and R13 provide a constant current through, and therefore a constant voltage drop across, R16. This is needed to make the current limiting, pins 2 and 3 of U2, work properly over the entire 4.5 to 25.0-V range. But as the current limiting action pulls the voltage at the top of R16 below about 4.0 V, diode CR3 quickly drops out of conduction, the drop across R16 goes toward zero, and the load current falls to and remains at about 1.9 A, thereby limiting the dissipation in Q1, Q2 and T1 and allowing short circuit protection for an indefinite time. This is a regenerative, positive feedback process. R14 is adjusted so that the bend-back starts at about 2.6 A. This circuit was modeled and perfected using a simulation program prior to breadboarding. See reference 1 for further discussion of bend-back circuitry.



The metering is done with a Heath Model SM-2300-A auto-ranging DMM which sells for $20. It is mounted on the front panel and is dedicated to the power supply. The voltage across R3 is 0.11 times the load current. R8 provides the required calibration of the ammeter (0.10 times load current). R3 is an ordinary wire-wound and not a high-grade ammeter shunt, but it is adequate since it does not heat up significantly at 2.5 A. The purpose of R9 is to set the reference voltage on pin 5 of U2 at 4.5 V so that the output can have that minimum value. R11 sets the 25.0-V upper limit. R4 is a ten-turn helipot, for easier adjustment. C8 and C9 are 2.0-kV ceramic, suitable for the ac line bypass function. ***A three-wire line cord is used so that the supply chassis is always tied to building ground, for safety reasons.*** The dc output floats with respect to chassis ground and performance is independent of the grounding connection. Push button PB1 gives a fast discharge of the capacitors (through R15) after turnoff if the load current is very small.

Construction

Figs 5(a) and (b) show the general construction method that I used. The cabinet and chassis surface are 0.062-inch aluminum plates connected by aluminum angle stock, drilled and tapped for 6-32 screws. Ventilation screens at the top and rear provide an excellent chimney effect. The chassis plate is tightly joined to the side plates for better heat transfer. The heat sink selection, one each for Q1 and Q2 (Wakefield 403A), was done according to the excellent discussion in reference 1 and need not be repeated here. The worst case 2N3055 junction temperature was calculated at 145° C, based on a measured (using a Radio Shack 271-110 thermistor that I calibrated myself) case temperature of 95° C and dissipation of 32 W each for Q1 and Q2. This temperature is a little higher than reference 1 recommends, but I consider it acceptable for intermittent lab usage.

The DMM is epoxied to a narrow aluminum strip which is screw mounted to the front panel. The battery compartment at the rear of the DMM is accessed by removing these screws. I really like the dedicated DMM arrangement because of its ability to resolve small changes in volts and amps. I also like the three-position maximum current-selector switch better than a continuous-adjustment potentiometer.

Fig 6 shows the underneath. The PC board is mounted on standoff insulators and the positive and negative output leads are close to the binding posts. All components and wiring are on one side of the board (with the help of a few jumper wires) and the other side is entirely ground plane, except for small circular areas where component through-pads are located. Silastic and pieces of Kraft paper are used to cover up the exposed 120-V ac.

My cabinet construction style is somewhat labor intensive, involving a lot of metal work, and the reader is encouraged to think of simpler approaches, for example,

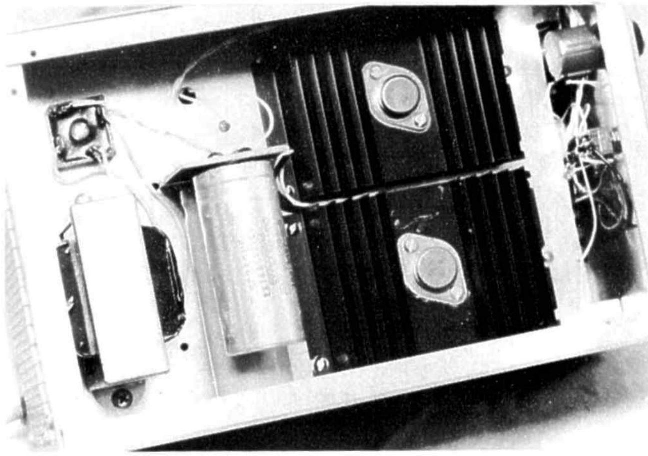


Fig 5—(a) Cabinet (b) top view showing two Wakefield 403A heat sinks.

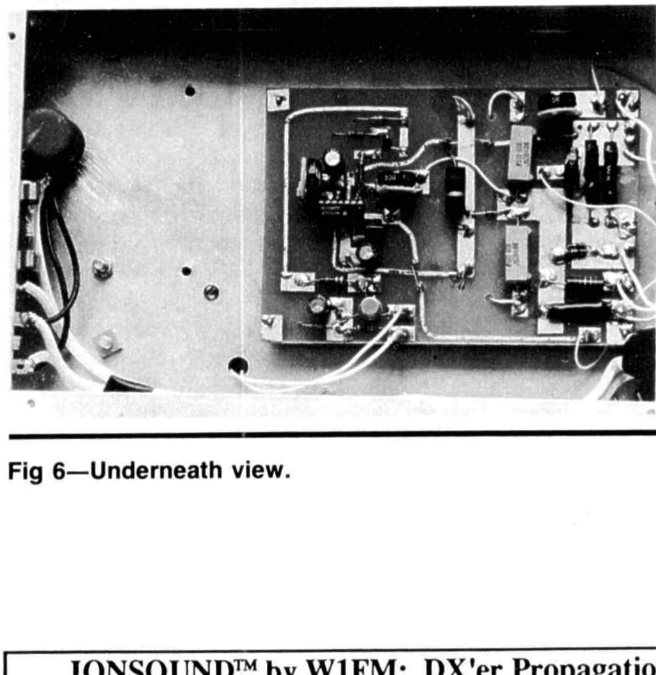


Fig 6—Underneath view.

a 7 × 11 × 2-inch chassis with bottom cover and rubber feet, a 7.5 × 6-inch front panel and some kind of perforated metal cover.

References

[1] *The ARRL Handbook*, 1991, chapters 6 and 27.

[2] DeMaw, Doug, "A 1.25 to 25 V 2.5 A Regulated Power Supply," *QST*, September 1989, and DeMaw, Doug, "Some Power Supply Design Hints," *QST*, November 1989.

[3] Brown, Marty, "Practical Switching Power Supply Design," Academic Press Inc, San Diego, 1990 (Motorola Series in Solid State Electronics).

[4] Van Valkenburg, *Modern Network Synthesis*, chapter 8, Wiley Book Co, 1967.

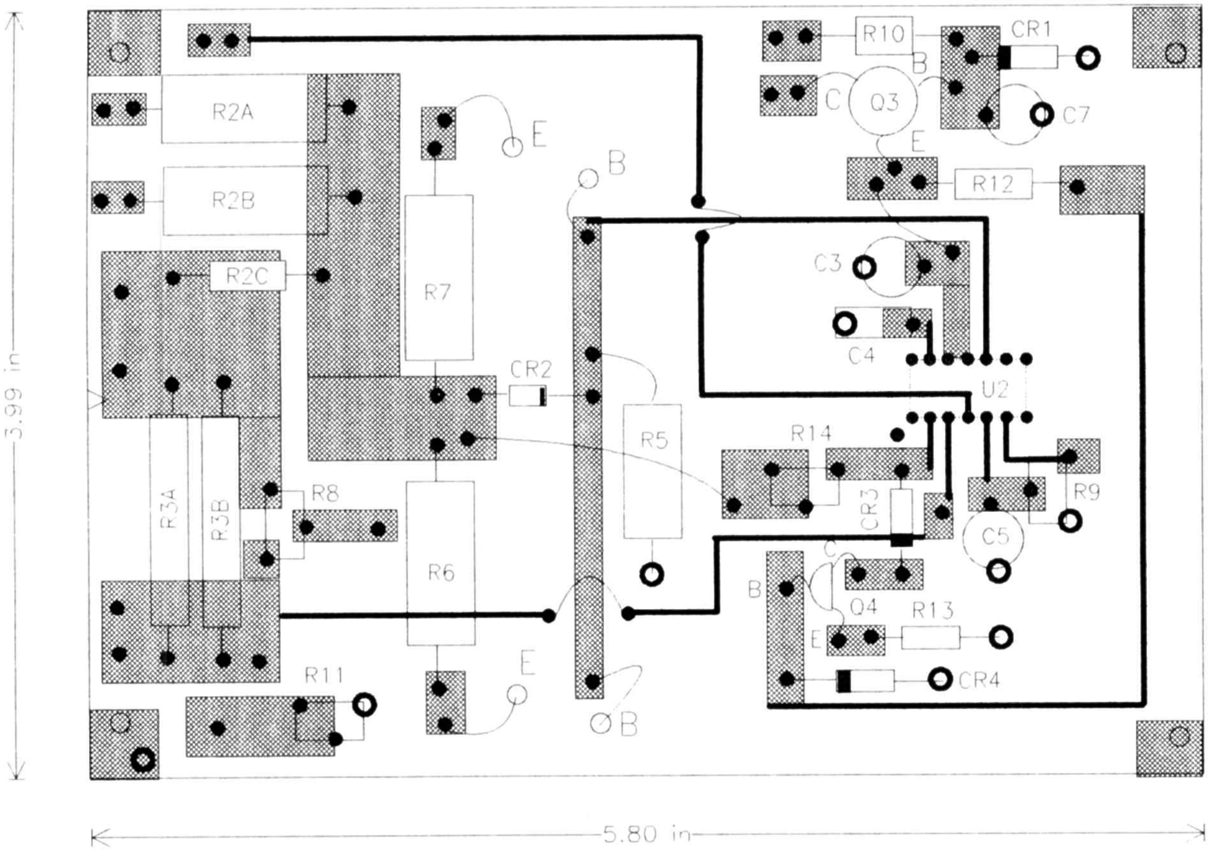


Fig 7—Parts placement.

Electrical Parts List for Power Supply

(RS signifies Radio Shack part number)

C1—10,000 μF, 50 V	CDE 10000-50-AC	$ 8.45
C2—3300 μF, 50 V	CDE 3300-50-AA	6.20
C3,C7—100 $n = muF, 50 V	RS-272-1044	0.89
C4—33 pF, 50 V		0.20
C5—100 μF, 50 V	RS-272-1028	0.79
C6—1000 μF, 35 V	RS-272-1032	1.59
C8,C9—0.01 μF, 2 kV	RS-272-160	0.99
CR1—1N5257A	Motorola	0.40
CR2,CR3—1N4001	RS-276-1101	0.50
CR4—1N750A	Motorola	0.20
DS1—Neon, 120 V ac	RS-272-704	0.90
F1—1.5 A SLO-BLO	RS-270-1284	0.65
M1—DMM	Heath SM-2300-A	20.00
PB1—Push button	RS-275-1547	0.70
Q1,Q2—2N3055	RS-276-2041	3.98
Q3—2N3053	RS-276-2030	0.79
Q4—MPS2222A	RS-276-2009	0.59
R1—10 Ω, 0.5 W		0.25
R2a—0.22 Ω, 2 W	Tresco PW3	0.50
R2b—1.5 Ω, 2 W		0.50
R2c—6.2 Ω, 5%, 0.5 W		0.65
R3A,R3B—0.15 Ω, 2W		0.81
R4—10-kΩ, 10-turn pot	Bourns 3540S	10.49
R5—1.8 kΩ, 2 W		0.25
R6,R7—0.1 Ω, 2 W		1.00
R8,R14—500-Ω pot	RS-271-226	1.40
R9—10-kΩ pot	RS-271-218	0.69
R10—470 Ω, 0.25 W		0.10
R11—5-kΩ pot	RS-271-217	0.69
R12—5.6 kΩ, 0.5 W		0.25
R13—680 Ω, 0.25 W		0.10
R15—22 Ω, 2 W		0.20
R16—68 Ω, 0.25 W		0.10
S1,S2—DPDT	RS-275-652	4.98
S3—SPDT (center off)	RS-275-654	2.39
T1—25.2 V ac, 2.8 A	Stancor P-8388	18.31
U1—Rectifier Bridge, 25 A, 50 V	RS-276-1185	2.69
U2—LM723 regulator	RS-27-1740	0.99
Heatsink (2)	Wakefield 403A	15.00
Heatsink grease	RS-276-1372	1.59
TO3 HDWE for Q1, Q2	RS-276-1371	1.98
Fuse holder	RS-270-739	0.50
Line cord, 6 foot	RS-278-1258	3.00
	Total Electrical Parts	117.23

SWITCHING REGULATORS

Introduction

The switching regulator is increasing in popularity because it offers the advantages of higher power conversion efficiency and increased design flexibility (multiple output voltages of different polarities can be generated from a single input voltage).

This paper will detail the operating principles of the four most commonly used switching converter types:

Buck: used the reduce a DC voltage to a lower DC voltage.

Boost: provides an output voltage that is higher than the input.

Buck-Boost (invert): an output voltage is generated opposite in polarity to the input.

Flyback: an output voltage that is less than or greater than the input can be generated, as well as multiple outputs.

Also, some multiple-transistor converter topologies will be presented:

Push-Pull: A two-transistor converter that is especially efficient at low input voltages.

Half-Bridge: A two-transistor converter used in many off-line applications.

Full-Bridge: A four-transistor converter (usually used in off-line designs) that can generate the highest output power of all the types listed.

Application information will be provided along with circuit examples that illustrate some applications of Buck, Boost, and Flyback regulators.

Switching Fundamentals

Before beginning explanations of converter theory, some basic elements of power conversion will be presented:

THE LAW OF INDUCTANCE

If a voltage is forced across an inductor, a current will flow through that inductor (and this current will vary with time). Note that the **current flowing in an inductor will be time-varying even if the forcing voltage is constant**.

It is equally correct to say that if a time-varying current is forced to flow in an inductor, a voltage across the inductor will result.

The fundamental law that defines the relationship between the voltage and current in an inductor is given by the equation:

$$v = L \; (di/dt)$$

Two important characteristics of an inductor that follow directly from the law of inductance are:

1) A voltage across an inductor results **only** from **a current that changes with time**. A **steady (DC) current flowing in an inductor causes no voltage across it** (except for the tiny voltage drop across the copper used in the windings).

2) A **current flowing in an inductor can not change value instantly** (in zero time), as this would require infinite voltage to force it to happen. However, **the faster the current is changed in an inductor, the larger the resulting voltage will be**.

Note: Unlike the current flowing in the inductor, the **voltage across it <u>can</u> change instantly** (in zero time).

The principles of inductance are illustrated by the information contained in Figure 25.

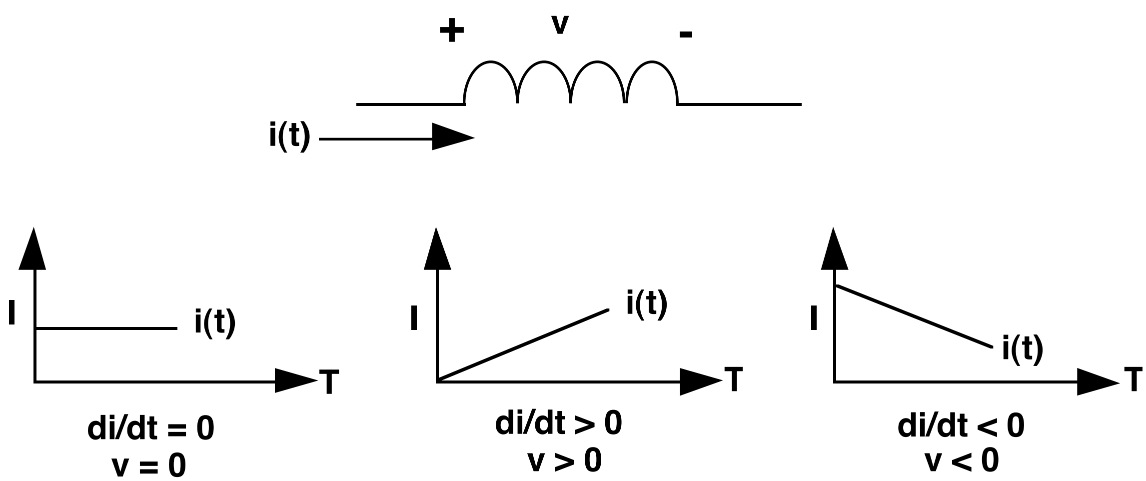


FIGURE 25. INDUCTOR VOLTAGE/CURRENT RELATIONSHIP

The important parameter is the di/dt term, which is simply a measure of how the current changes with time. When the current is plotted versus time, the **value of di/dt is defined as the slope of the current plot at any given point**.

The graph on the left shows that current which is constant with time has a di/dt value of zero, and results in no voltage across the inductor.

The center graph shows that a current which is increasing with time has a positive di/dt value, resulting in a positive inductor voltage.

Current that decreases with time (shown in the right-hand graph) gives a negative value for di/dt and inductor voltage.

It is important to note that a **linear current ramp in an inductor** (either up or down) **occurs only when it has a constant voltage across it**.

TRANSFORMER OPERATION

A transformer is a device that has two or more magnetically-coupled windings. The basic operation is shown in Figure 26.

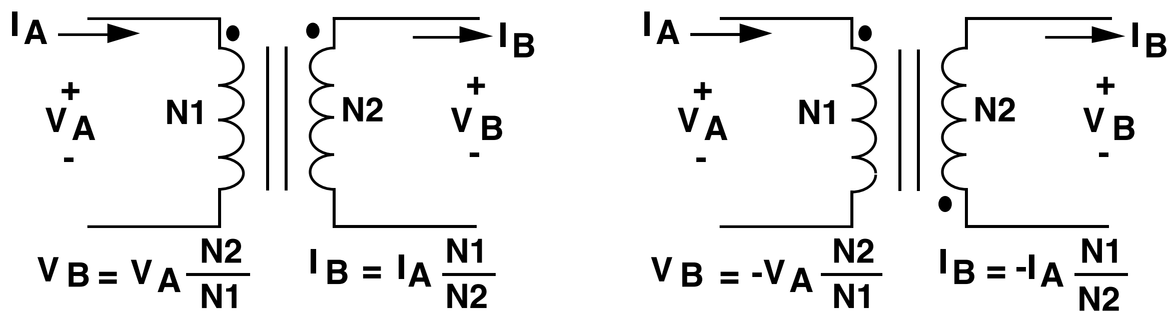


FIGURE 26. TRANSFORMER THEORY

The action of a transformer is such that a time-varying (AC) voltage or current is transformed to a higher or lower value, as set by the transformer turns ratio. The transformer does not add power, so it follows that the power (V X I) on either side must be constant. That is the reason that the **winding with more turns has higher voltage but lower current**, while the **winding with less turns has lower voltage but higher current**.

The **dot on a transformer winding identifies its polarity** with respect to another winding, and **reversing the dot results in inverting the polarity**.

Example of Transformer Operation:

An excellent example of how a transformer works can be found under the hood of your car, where a transformer is used to generate the 40 kV that fires your cars spark plugs (see Figure 27).

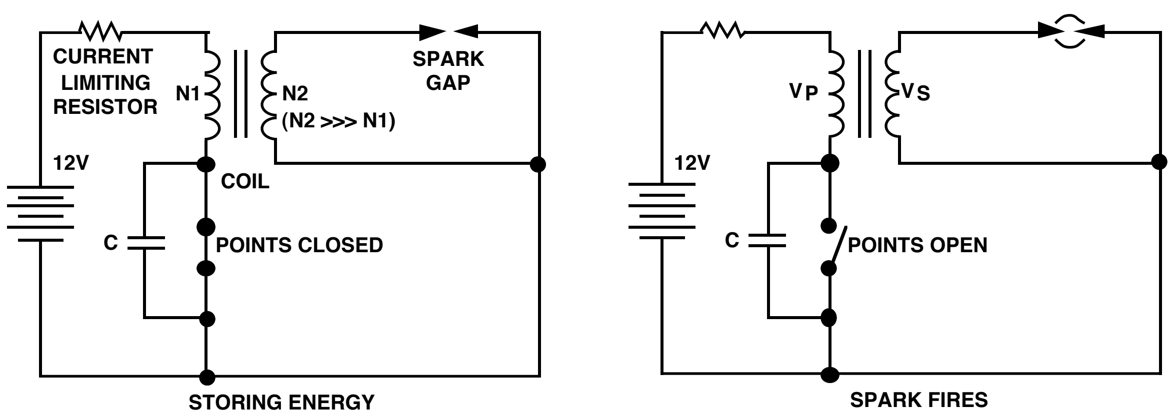


FIGURE 27. SPARK FIRING CIRCUIT

The "coil" used to generate the spark voltage is actually a transformer, with a very high secondary-to-primary turns ratio.

When the points first close, current starts to flow in the primary winding and eventually reaches the final value set by the 12V battery and the current limiting resistor. At this time, the current flow is a fixed DC value, which means no voltage is generated across either winding of the transformer.

When the points open, the current in the primary winding collapses very quickly, causing a large voltage to appear across this winding. This voltage on the primary is magnetically coupled to (and stepped up by) the secondary winding, generating a voltage of 30 kV - 40 kV on the secondary side.

As explained previously, the law of inductance says that it is not possible to instantly break the current flowing in an inductor (because an infinite voltage would be required to make it happen).

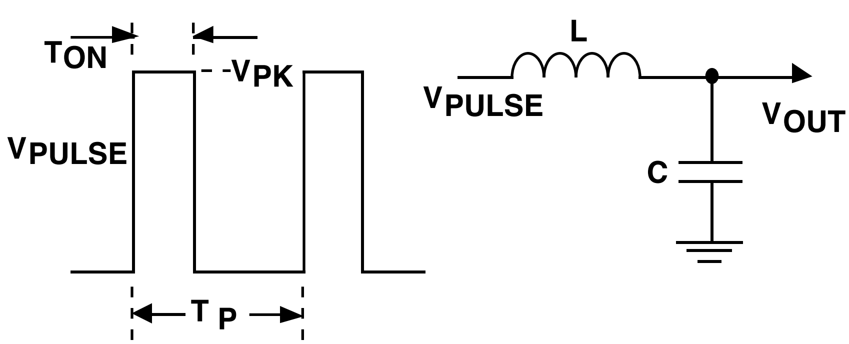
This principle is what causes the arcing across the contacts used in switches that are in circuits with highly inductive loads. When the switch just begins to open, the high voltage generated allows electrons to jump the air gap so that the current flow does not actually stop instantly. Placing a capacitor across the contacts helps to reduce this arcing effect.

In the automobile ignition, a capacitor is placed across the points to minimize damage due to arcing when the points "break" the current flowing in the low-voltage coil winding (in car manuals, this capacitor is referred to as a "condenser").

PULSE WIDTH MODULATION (PWM)

All of the switching converters that will be covered in this paper use a form of output voltage regulation known as **Pulse Width Modulation (PWM)**. Put simply, the feedback loop adjusts (corrects) the output voltage by changing the ON time of the switching element in the converter.

As an example of how PWM works, we will examine the result of applying a series of square wave pulses to an L-C filter (see Figure 28).



$$V_{OUT} = \frac{T_{ON}}{T_P} \times V_{PK}$$

FIGURE 28. BASIC PRINCIPLES OF PWM

The series of square wave pulses is filtered and provides a **DC output voltage that is equal to the peak pulse amplitude multiplied times the duty cycle** (duty cycle is defined as the switch ON time divided by the total period).

This relationship explains how the output voltage can be directly controlled by changing the ON time of the switch.

Switching Converter Topologies

The most commonly used DC-DC converter circuits will now be presented along with the basic principles of operation.

BUCK REGULATOR

The most commonly used switching converter is the Buck, which is used to down-convert a DC voltage to a lower DC voltage of the same polarity. This is essential in systems that use distributed power rails (like 24V to 48V), which must be locally converted to 15V, 12V or 5V with very little power loss.

The Buck converter uses a transistor as a switch that alternately connects and disconnects the input voltage to an inductor (see Figure 29).

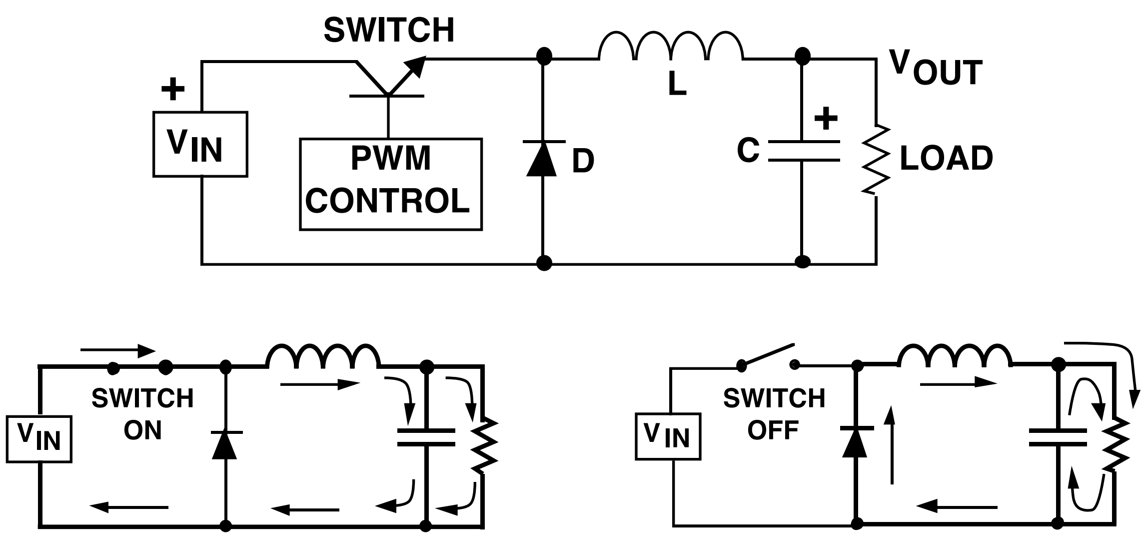


FIGURE 29. BUCK REGULATOR

The lower diagrams show the current flow paths (shown as the heavy lines) when the switch is on and off.

When the switch turns on, the input voltage is connected to the inductor. The difference between the input and output voltages is then forced across the inductor, causing current through the inductor to increase.

During the **on time**, the inductor current flows into both the load and the output capacitor (the capacitor **charges** during this time).

When the switch is turned off, the input voltage applied to the inductor is removed. However, since the current in an inductor can not change instantly, the voltage across the inductor will adjust to hold the current constant.

The input end of the inductor is forced negative in voltage by the decreasing current, eventually reaching the point where the diode is turned on. The inductor current then flows through the load and back through the diode.

The capacitor discharges into the load during the off time, contributing to the total current being supplied to the load (the total load current during the switch off time is the sum of the inductor and capacitor current).

The shape of the current flowing in the inductor is similar to Figure 30.

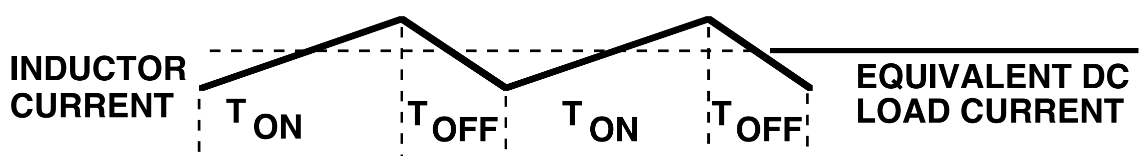


FIGURE 30. BUCK REGULATOR INDUCTOR CURRENT

As explained, the current through the inductor ramps up when the switch is on, and ramps down when the switch is off. The DC load current from the regulated output is the average value of the inductor current.

The peak-to-peak difference in the inductor current waveform is referred to as the **inductor ripple current**, and the inductor is typically selected large enough to keep this ripple current less than 20% to 30% of the rated DC current.

CONTINUOUS vs. DISCONTINUOUS OPERATION

In most Buck regulator applications, the inductor current never drops to zero during full-load operation (this is defined as **continuous mode operation**). Overall performance is usually better using continuous mode, and it allows maximum output power to be obtained from a given input voltage and switch current rating.

In applications where the maximum load current is fairly low, it can be advantageous to design for discontinuous mode operation. In these cases, operating in discontinuous mode can result in a smaller overall converter size (because a smaller inductor can be used).

Discontinuous mode operation at lower load current values is generally harmless, and **even converters designed for continuous mode operation at full load will become discontinuous as the load current is decreased** (usually causing no problems).

BOOST REGULATOR

The Boost regulator takes a DC input voltage and produces a DC output voltage that is higher in value than the input (but of the same polarity). The Boost regulator is shown in Figure 31, along with details showing the path of current flow during the switch on and off time.

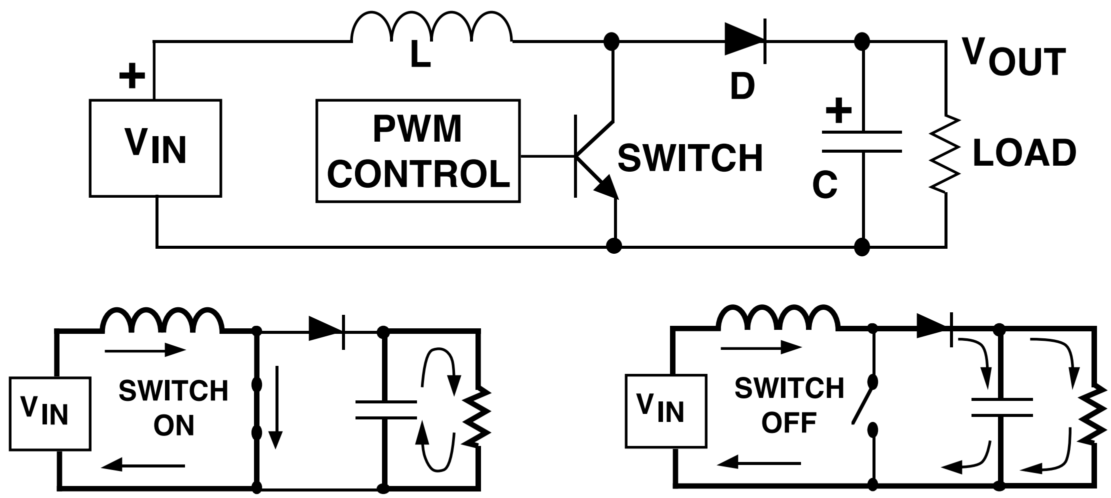


FIGURE 31. BOOST REGULATOR

Whenever the switch is on, the input voltage is forced across the inductor which causes the current through it to increase (ramp up).

When the switch is off, the decreasing inductor current forces the "switch" end of the inductor to swing positive. This forward biases the diode, allowing the capacitor to charge up to a voltage that is higher than the input voltage.

During steady-state operation, the inductor current flows into both the output capacitor and the load during the switch off time. When the switch is on, the load current is supplied only by the capacitor.

OUTPUT CURRENT AND LOAD POWER

An important design consideration in the Boost regulator is that the **output load current and the switch current are not equal**, and the **maximum available load current is always less than the current rating of the switch transistor**.

It should be noted that the **maximum total power available for conversion in <u>any</u> regulator** is equal to the **input voltage multiplied times the maximum average input current** (which is **less than** the current rating of the switch transistor).

Since the output voltage of the Boost is **higher than the input voltage**, it follows that the **output current must be lower than the input current**.

BUCK-BOOST (INVERTING) REGULATOR

The Buck-Boost or Inverting regulator takes a DC input voltage and produces a DC output voltage that is opposite in polarity to the input. The negative output voltage can be either larger or smaller in magnitude than the input voltage.

The Inverting regulator is shown in Figure 32.

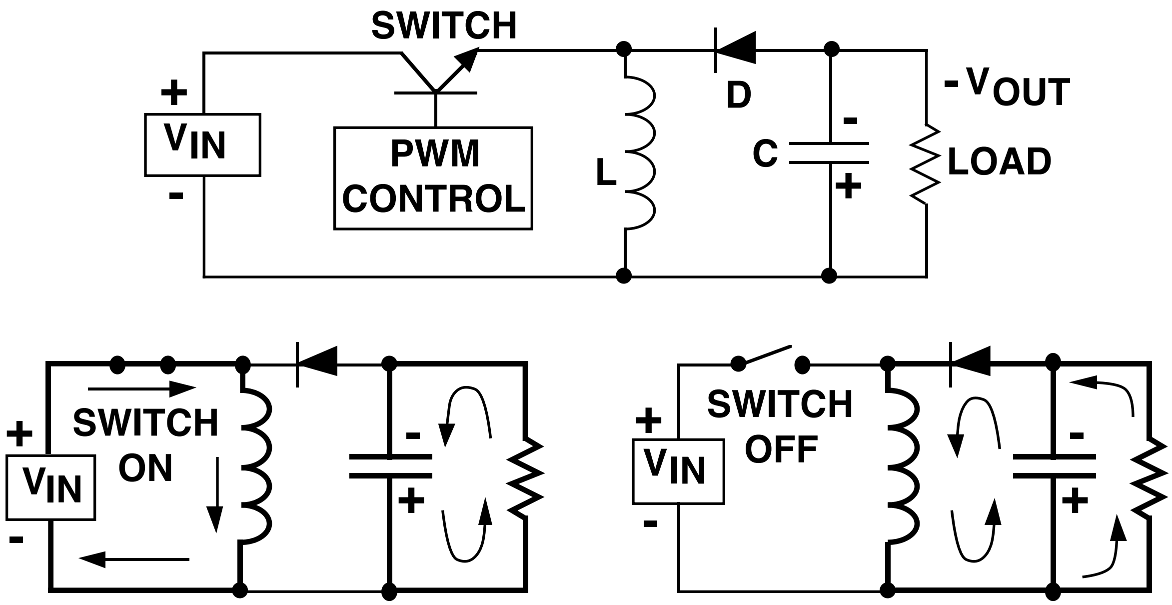


FIGURE 32. BUCK-BOOST (INVERTING) REGULATOR

When the switch is on, the input voltage is forced across the inductor, causing an increasing current flow through it. During the on time, the discharge of the output capacitor is the only source of load current.

This requires that the charge lost from the output capacitor during the on time be replenished during the off time.

When the switch turns off, the decreasing current flow in the inductor causes the voltage at the diode end to swing negative. This action turns on the diode, allowing the current in the inductor to supply **both the output capacitor and the load**.

As shown, **the load current is supplied by inductor when the switch is off, and by the output capacitor when the switch is on**.

FLYBACK REGULATOR

The Flyback is the most versatile of all the topologies, allowing the designer to create one or more output voltages, some of which may be opposite in polarity.

Flyback converters have gained popularity in battery-powered systems, where a single voltage must be converted into the required system voltages (for example, +5V, +12V and -12V) with very high power conversion efficiency.

The basic single-output flyback converter is shown in Figure 33.

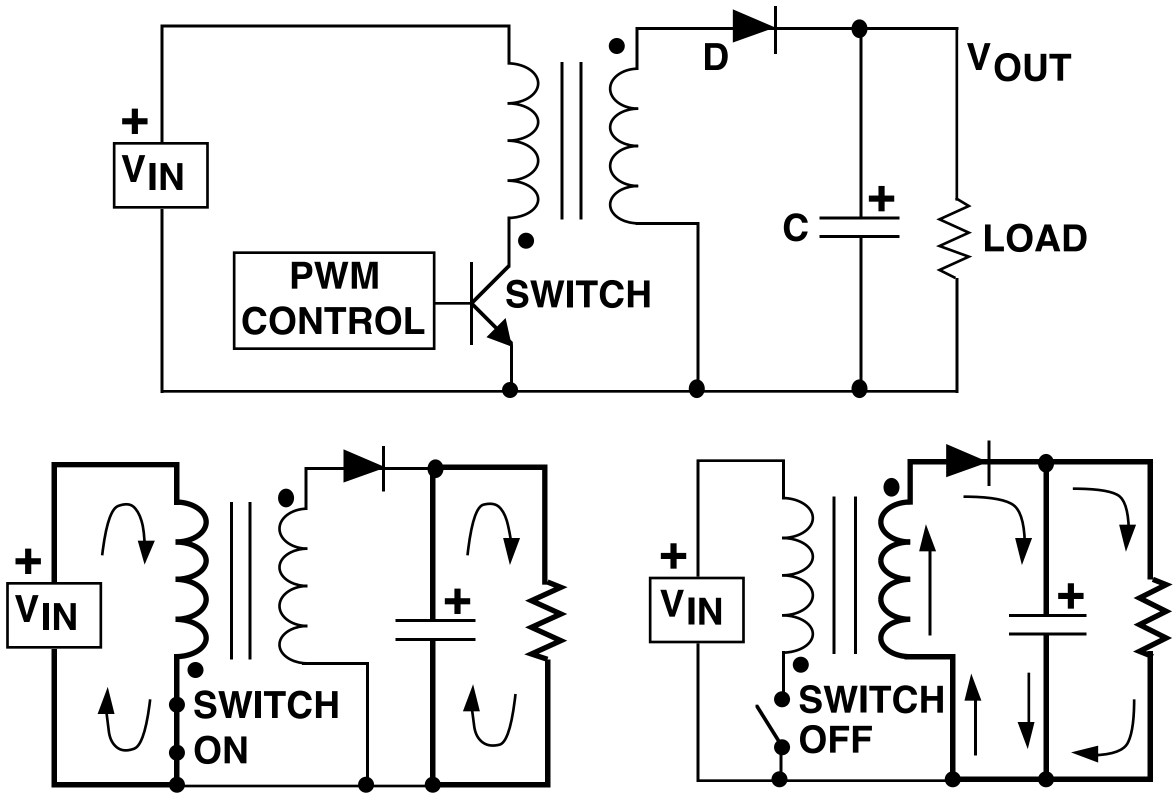


FIGURE 33. SINGLE-OUTPUT FLYBACK REGULATOR

The most important feature of the Flyback regulator is the transformer phasing, as shown by the dots on the primary and secondary windings.

When the switch is on, the input voltage is forced across the transformer primary which causes an increasing flow of current through it.

Note that the **polarity of the voltage on the primary is dot-negative** (more negative at the dotted end), causing a voltage with the same polarity to appear at the transformer secondary (the magnitude of the secondary voltage is set by the transformer seconday-to-primary turns ratio).

The dot-negative voltage appearing across the secondary winding turns off the diode, preventing current flow in the secondary winding during the switch on time. During this time, the load current must be supplied by the output capacitor alone.

When the switch turns off, the decreasing current flow in the primary causes the voltage at the dot end to swing positive. At the same time, the primary voltage is reflected to the secondary with the same polarity. The dot-positive voltage occurring across the secondary winding turns on the diode, allowing current to flow into both the load and the output capacitor. The output capacitor charge lost to the load during the switch on time is replenished during the switch off time.

Flyback converters operate in either **continuous mode** (where the secondary current is always >0) or **discontinuous mode** (where the secondary current falls to zero on each cycle).

GENERATING MULTIPLE OUTPUTS

Another big advantage of a Flyback is the capability of providing multiple outputs (see Figure 34). In such applications, one of the outputs (usually the highest current) is selected to provide PWM feedback to the control loop, **which means this output is directly regulated.**

The other secondary winding(s) are **indirectly regulated**, as their pulse widths will follow the regulated winding. The load regulation on the unregulated secondaries is not great (typically 5 - 10%), but is adequate for many applications.

If tighter regulation is needed on the lower current secondaries, an LDO post-regulator is an excellent solution. The secondary voltage is set about 1V above the desired output voltage, and the LDO provides excellent output regulation with very little loss of efficiency.

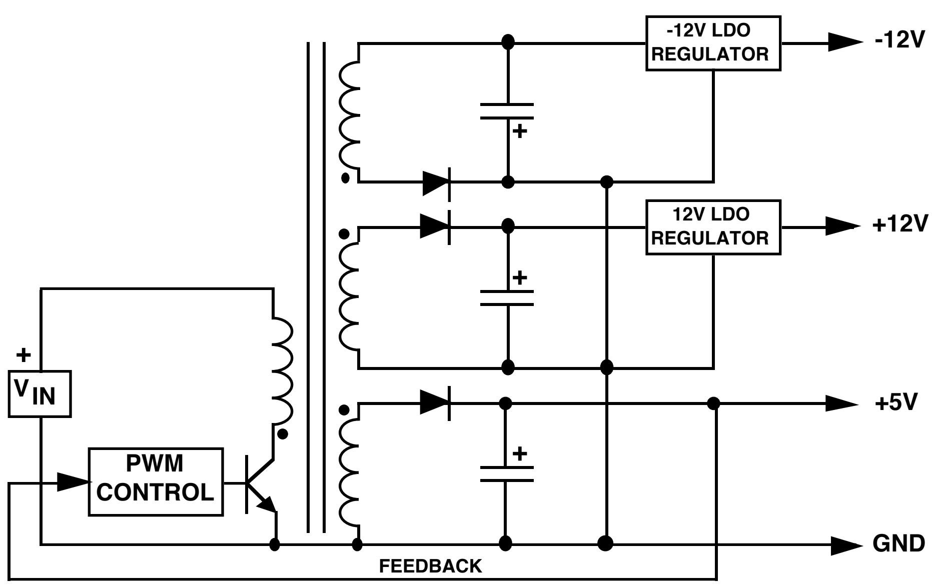


FIGURE 34. TYPICAL MULTIPLE-OUTPUT FLYBACK

PUSH-PULL CONVERTER

The Push-Pull converter uses two to transistors perform DC-DC conversion (see Figure 35).

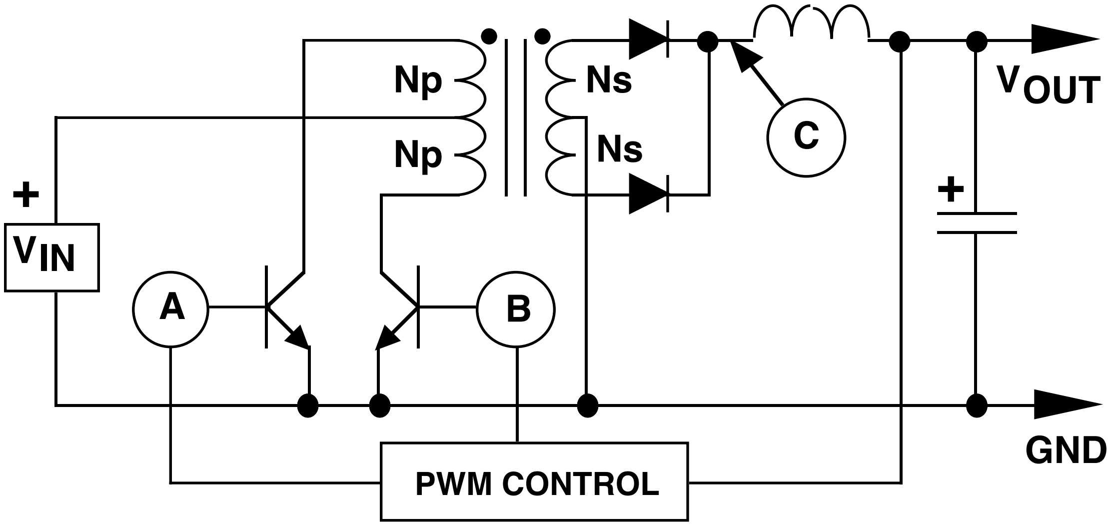


FIGURE 35. PUSH-PULL CONVERTER

The converter operates by turning on each transistor on alternate cycles (the two transistors are **never** on at the same time). **Transformer secondary current flows at the same time as primary current** (when either of the switches is on).

For example, when transistor "A" is turned on, the input voltage is forced across the upper primary winding with dot-negative polarity. On the secondary side, a dot-negative voltage will appear across the winding which turns on the bottom diode. This allows current to flow into the inductor to supply both the output capacitor and the load.

When transistor "B" is on, the input voltage is forced across the lower primary winding with dot-positive polarity. The same voltage polarity on the secondary turns on the top diode, and current flows into the output capacitor and the load.

An important characteristic of a Push-Pull converter is that the switch transistors have to be able the stand off more than twice the input voltage: when one transistor is on (and the input voltage is forced across one primary winding) the same magnitude voltage is induced across the other primary winding, but it is "floating" on top of the input voltage. This puts the collector of the turned-off transistor at **twice** the input voltage with respect to ground.

The "double input voltage" rating requirement of the switch transistors means the Push-Pull converter is best suited for lower input voltage applications. It has been widely used in converters operating in 12V and 24V battery-powered systems.

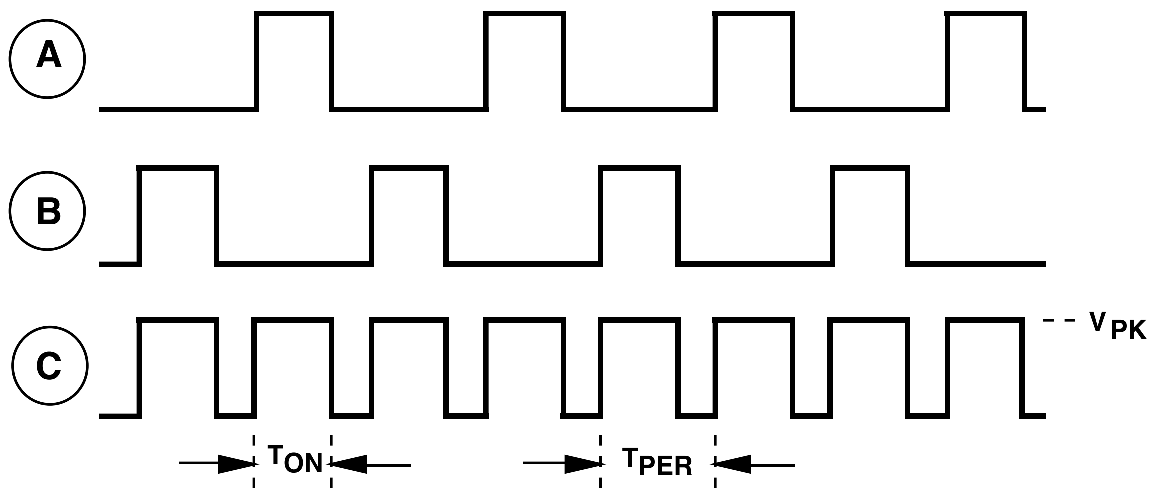


FIGURE 36. TIMING DIAGRAM FOR PUSH-PULL CONVERTER

Figure 36 shows a timing diagram which details the relationship of the input and output pulses.

It is important to note that frequency of the secondary side voltage pulses is twice the frequency of operation of the PWM controller driving the two transistors. For example, if the PWM control chip was set up to operate at 50 kHz on the primary side, the frequency of the secondary pulses would be 100 kHz.

The DC output voltage is given by the equation:

$$V_{OUT} = V_{PK} \times (T_{ON} / T_{PER})$$

The peak amplitude of the secondary pulses (V_{PK}) is given by:

$$V_{PK} = (V_{IN} - V_{SWITCH}) \times (N_S / N_P) - V_{RECT}$$

This highlights why the Push-Pull converter is well-suited for low voltage converters. The voltage forced across each primary winding (which provides the power for conversion) is the full input voltage minus only the saturation voltage of the switch.

If MOS-FET power switches are used, the voltage drop across the switches can be made extremely small, resulting in very high utilization of the available input voltage.

Another advantage of the Push-Pull converter is that it can also generate multiple output voltages (by adding more secondary windings), some of which may be negative in polarity. This allows a power supply operated from a single battery to provide all of the voltages necessary for system operation.

A disadvantage of Push-Pull converters is that they require very good matching of the switch transistors to prevent unequal on times, since this will result in saturation of the transformer core (and failure of the converter).

HALF-BRIDGE CONVERTER

The Half-Bridge is a two-transistor converter frequently used in high-power designs. It is well-suited for applications requiring load power in the range of 500W to 1500W, and is almost always operated directly from the AC line.

Off-line operation means that no large 60 Hz power transformer is used, eliminating the heaviest and costliest component of a typical transformer-powered supply. All of the transformers in the Half-Bridge used for power conversion operate at the switching frequency (typically 50 kHz or higher) which means they can be very small and efficient.

A very important advantage of the Half-Bridge is **input-to-output isolation** (the regulated DC output is electrically isolated from the AC line). But, this means that all of the PWM control circuitry must be referenced to the DC output ground.

The voltage to run the control circuits is usually generated from a DC rail that is powered by a small 60 Hz transformer feeding a three-terminal regulator. In some designs requiring extremely high efficiency, the switcher output takes over and provides internal power after the start-up period.

The switch transistor drive circuitry must be isolated from the transistors, requiring the use of base drive transformers. The added complexity of the base drive circuitry is a disadvantage of using the Half-Bridge design.

If a 230 VAC line voltage is rectified by a full-wave bridge and filtered by a capacitor, an unregulated DC voltage of about 300V will be available for DC-DC conversion. If 115 VAC is used, a voltage doubler circuit is typically used to generate the 300V rail.

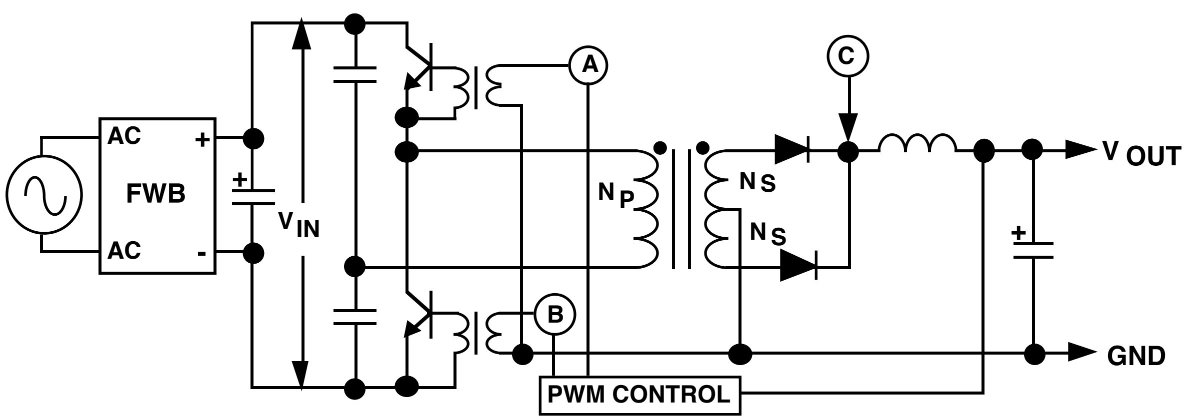


FIGURE 37. HALF-BRIDGE CONVERTER

The basic Half-bridge converter is shown in Figure 37. A capacitive divider is tied directly across the unregulated DC input voltage, providing a reference voltage of $1/2V_{IN}$ for one end of the transformer primary winding. The other end of the primary is actively driven up and down as the transistors alternately turn on and off.

The switch transistors force one-half of the input voltage across the primary winding during the switch on time, reversing polarity as the transistors alternate. The switching transistors are **never** on at the same time, or they would be destroyed (since they are tied directly across V_{IN}). The timing diagram for the Half-Bridge converter is shown in Figure 38 (it is the same as the Push-Pull).

When the "A" transistor is on, a dot-positive voltage is forced across the primary winding and reflected on the secondary side (with the magnitude being set by the transformer turns ratio). The dot-positive secondary voltage turns on the upper rectifier diode, supplying current to both the output capacitor and the load.

When the "A" transistor turns off and the "B" transistor turns on, the polarity of the primary voltage is reversed. The secondary voltage polarity is also reversed, turning on the lower diode (which supplies current to the output capacitor and the load).

In a Half-Bridge converter, **primary and secondary current flow in the transformer at the same time** (when either transistor is on), supplying the load current and charging the output capacitor. The output capacitor discharges into the load **only during the time when both transistors are off**.

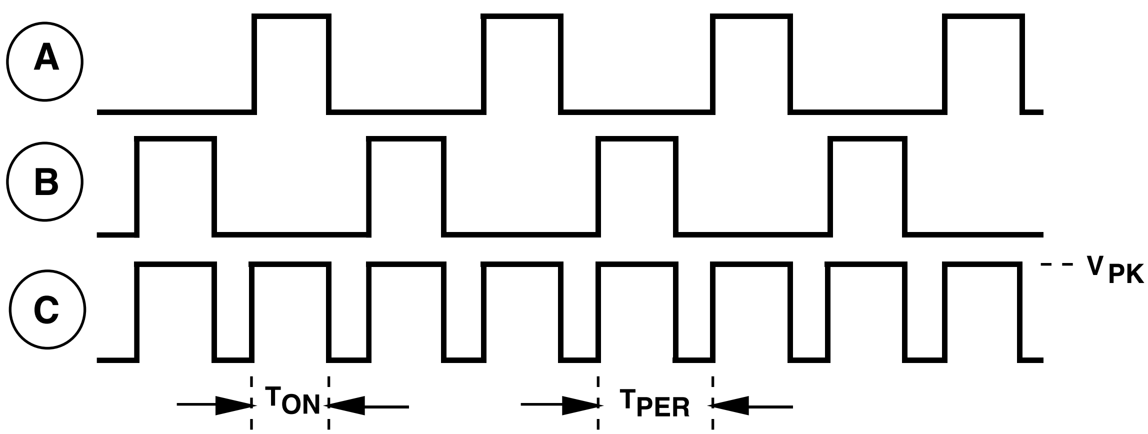


FIGURE 38. TIMING DIAGRAM FOR HALF-BRIDGE CONVERTER

It can be seen that the voltage pulses on the transformer secondary side (applied to the L-C filter) are occurring at twice the frequency of the PWM converter which supplies the drive pulses for the switching transistors.

The output voltage is again given by:

$$V_{OUT} = V_{PK} \times (T_{ON} / T_{PER})$$

The peak amplitude of the secondary pulses (V_{PK}) is given by:

$$V_{PK} = (1/2\ V_{IN} - V_{SWITCH}) \times (N_S / N_P) - V_{RECT}$$

FULL-BRIDGE CONVERTER

The Full-Bridge converter requires a total of four switching transistors to perform DC-DC conversion. The full bridge is most often seen in applications that are powered directly from the AC line, providing load power of 1 kW to 3 kW.

Operating off-line, the Full Bridge converter typically uses about 300V of unregulated DC voltage for power conversion (the voltage that is obtained when a standard 230 VAC line is rectified and filtered).

An important feature of this design is the isolation from the AC line provided by the switching transformer. The PWM control circuitry is referenced to the the output ground, requiring a dedicated voltage rail (usually powered from a small 60 Hz transformer) to run the control circuits.

The base drive voltages for the switch transistors (which are provided by the PWM chip) have to be transformer-coupled because of the required isolation.

Figure 39 shows a simplified schematic diagram of a Full-Bridge converter.

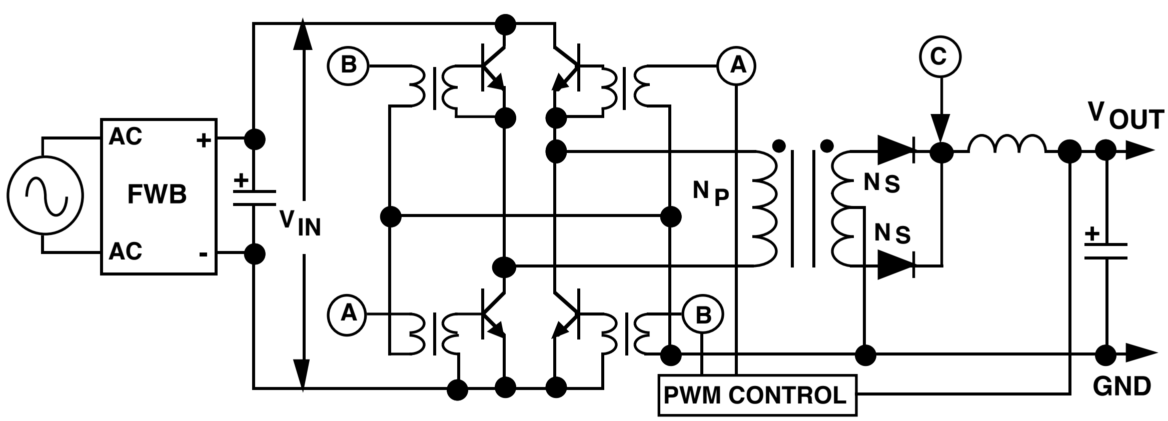


FIGURE 39. FULL BRIDGE CONVERTER

The transformer primary is driven by the **full voltage V_{IN}** when either of the transistor sets ("A" set or "B" set) turns on. The full input voltage utilization means the **Full-Bridge can produce the most load power of all the converter types**. The timing diagram is identical to the Half-Bridge, as shown in Figure 38.

Primary and secondary current flows in the transformer during the switch on times, while the output capacitor discharges into the load when both transistors are off.

The equation for the output voltage is (see Figure 38):

$$V_{OUT} = V_{PK} \times (T_{ON} / T_{PER})$$

The peak voltage of the transformer secondary pulses (V_{PK}) is given by:

$$V_{PK} = (V_{IN} - 2V_{SWITCH}) \times (N_S / N_P) - V_{RECT}$$

Application Hints For Switching Regulators

Application information will be provided on topics which will enhance the designers ability to maximize switching regulator performance.

Capacitor Parasitics Affecting Switching Regulator Performance

All capacitors contain parasitic elements which make their performance less than ideal (see Figure 40).

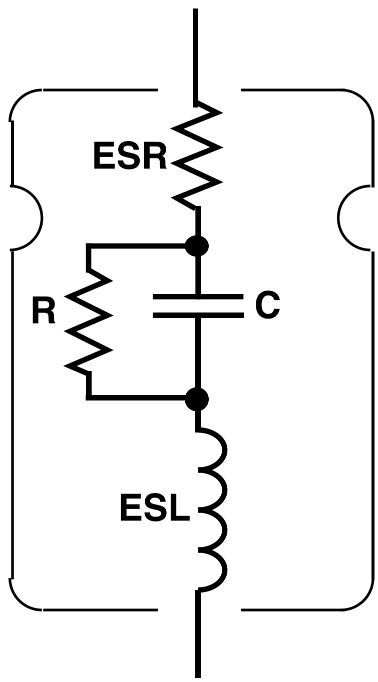


FIGURE 40.
CAPACITOR PARASITICS

Summary of Effects of Parasitics:

ESR: The ESR (**Equivalent Series Resistance**) causes internal heating due to power dissipation as the ripple current flows into and out of the capacitor. The capacitor can fail if ripple current exceeds maximum ratings.

Excessive output voltage ripple will result from high ESR**, and regulator loop instability is also possible**. ESR is highly dependent on temperature, increasing very quickly at temperatures below about 10 °C.

ESL: The ESL (**Effective Series Inductance**) limits the high frequency effectiveness of the capacitor. High ESL is the reason electrolytic capacitors need to be bypassed by film or ceramic capacitors to provide good high-frequency performance.

The ESR, ESL and C within the capacitor form a resonant circuit, whose frequency of resonance should be as high as possible. Switching regulators generate ripple voltages on their outputs with very high frequency (>10 MHz) components, which can cause ringing on the output voltage if the capacitor resonant frequency is low enough to be near these frequencies.

INPUT CAPACITORS

All of the switching converters in this paper (and the vast majority in use) operate as DC-DC converters that "chop" a DC input voltage at a very high frequency. As the converter switches, it has to draw current pulses from the input source. The **source impedance is extremely important**, as even a small amount of inductance can cause significant ringing and spiking on the voltage at the input of the converter.

The best practice is to **always provide adequate capacitive bypass as near as possible to the switching converter input**. For best results, an electrolytic is used with a film capacitor (and possibly a ceramic capacitor) in parallel for optimum high frequency bypassing.

OUTPUT CAPACITOR ESR EFFECTS

The primary function of the output capacitor in a switching regulator is filtering. As the converter operates, current must flow into and out of the output filter capacitor.

The ESR of the output capacitor directly affects the performance of the switching regulator. ESR is specified by the manufacturer on good quality capacitors, **but be certain that it is specified at the frequency of intended operation**.

General-purpose electrolytics usually only specify ESR at 120 Hz, but capacitors intended for high-frequency switching applications will have the ESR guaranteed at high frequency (like 20 kHz to 100 kHz).

Some ESR dependent parameters are:

Ripple Voltage: In most cases, **the majority of the output ripple voltage results from the ESR of the output capacitor**. If the ESR increases (as it will at low operating temperatures) the output ripple voltage will increase accordingly.

Efficiency: As the switching current flows into and out of the capacitor (through the ESR), power is dissipated internally. This "wasted" power reduces overall regulator efficiency, and **can also cause the capacitor to fail if the ripple current exceeds the maximum allowable specification for the capacitor**.

Loop Stability: The ESR of the output capacitor can affect regulator loop stability. Products such as the LM2575 and LM2577 are compensated for stability assuming the ESR of the output capacitor will stay within a specified range.

Keeping the ESR within the "stable" range is not always simple in designs that must operate over a wide temperature range. **The ESR of a typical aluminum electrolytic may increase by 40X as the temperature drops from 25°C to -40°C**.

In these cases, an aluminum electrolytic must be paralleled by another type of capacitor with a flatter ESR curve (like Tantalum or Film) so that the effective ESR (which is the parallel value of the two ESR's) stays within the allowable range.

Note: if operation **below -40°C** is necessary, aluminum electrolytics are probably not feasible for use.

BYPASS CAPACITORS

High-frequency bypass capacitors are always recommended on the supply pins of IC devices, but if the devices are used in assemblies near switching converters **bypass capacitors are absolutely required.**

The components which perform the high-speed switching (transistors and rectifiers) generate significant EMI that easily radiates into PC board traces and wire leads.

To assure proper circuit operation, all IC supply pins must be bypassed to a clean, low-inductance ground (for details on grounding, see next section).

Proper Grounding

The "ground" in a circuit is supposed to be at one potential, but in real life it is not. When ground currents flow through traces which have non-zero resistance, voltage differences will result at different points along the ground path.

In DC or low-frequency circuits, "ground management" is comparatively simple: the only parameter of critical importance is the **DC resistance of a conductor**, since that defines the voltage drop across it for a given current. In high-frequency circuits, it is the **inductance** of a trace or conductor that is much more important.

In switching converters, peak currents flow in high-frequency (> 50 kHz) pulses, which can cause severe problems if trace inductance is high. Much of the "ringing" and "spiking" seen on voltage waveforms in switching converters is the result of high current being switched through parasitic trace (or wire) inductance.

Current switching at high frequencies tends to flow near the surface of a conductor (this is called "skin effect"), which means that ground traces must be very wide on a PC board to avoid problems. It is usually best (when possible) to use one side of the PC board as a **ground plane**.

Figure 41 illustrates an example of a terrible layout:

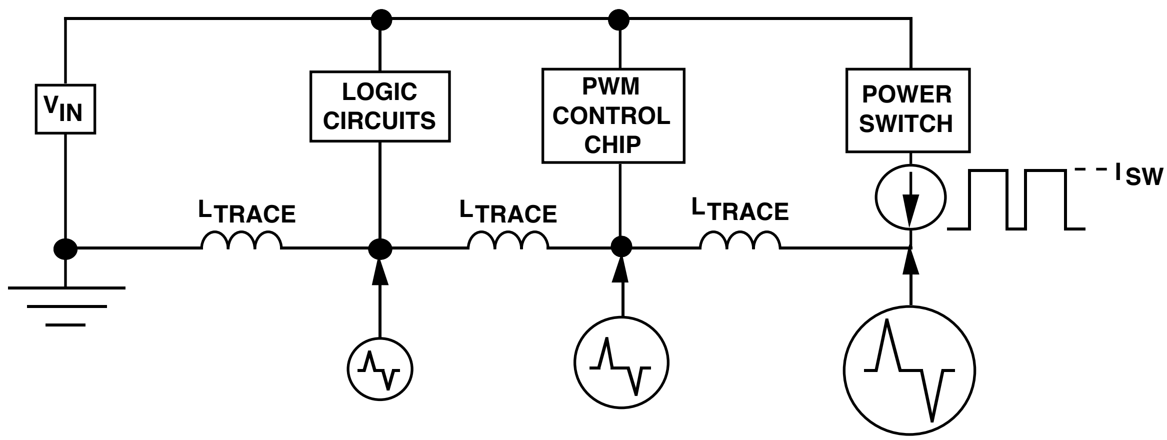


FIGURE 41. EXAMPLE OF POOR GROUNDING

The layout shown has the high-power switch return current passing through a trace that also provides the return for the PWM chip and the logic circuits. The switching current pulses flowing through the trace will cause a voltage spike (positive and negative) to occur as a result of the rising and falling edge of the switch current. This voltage spike follows directly from the **v = L (di/dt)** law of inductance.

It is important to note that the **magnitude of the spike will be different at all points along the trace**, being largest near the power switch. Taking the ground symbol as a point of reference, this shows how all three circuits would be bouncing up and down with respect to ground. More important, **they would also be moving with respect to each other**.

Mis-operation often occurs when sensitive parts of the circuit "rattle" up and down due to ground switching currents. This can induce noise into the reference used to set the output voltage, resulting in excessive output ripple.

Very often, regulators that suffer from ground noise problems appear to be unstable, and break into oscillations as the load current is increased (which increases ground currents).

A much better layout is shown in Figure 42.

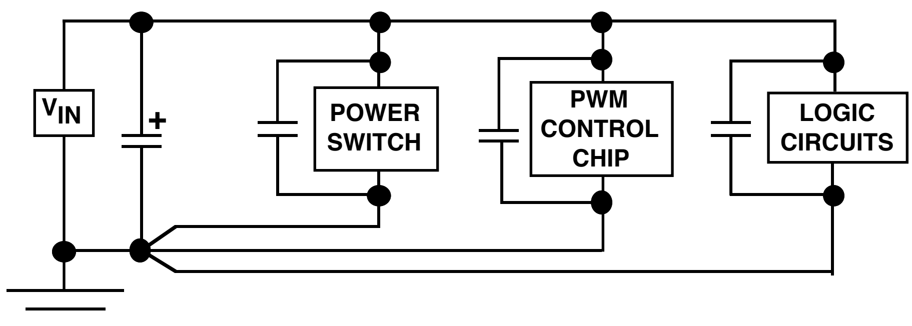


FIGURE 42. EXAMPLE OF GOOD GROUNDING

A big improvement is made by using **single-point grounding**. A good high-frequency electrolytic capacitor (like solid Tantalum) is used near the input voltage source to provide a good ground point.

All of the individual circuit elements are returned to this point **using separate ground traces**. This prevents high current ground pulses from bouncing the logic circuits up and down.

Another important improvement is that **the power switch** (which has the highest ground pin current) **is located as close as possible to the input capacitor**. This minimizes the trace inductance along its ground path.

It should also be pointed out that all of the individual circuit blocks have "local" bypass capacitors tied directly across them. The purpose of this capacitor is RF bypass, so it must be a ceramic or film capacitor (or both).

A good value for bypassing logic devices would be 0.01 μF ceramic capacitor(s), distributed as required.

If the circuit to be bypassed generates large current pulses (like the power switch), **more capacitance is required**. A good choice would be an aluminum electrolytic bypassed with a film and ceramic capacitor. Exact size depends on peak current, but the more capacitance used, the better the result.

Transformer/Inductor Cores and Radiated Noise

The type of core used in an inductor or transformer directly affects its cost, size, and radiated noise characteristics. Electrical noise radiated by a transformer is extremely important, as it may require shielding to prevent erratic operation of sensitive circuits located near the switching regulator.

The most commonly used core types will be presented, listing the advantages and disadvantages of each.

The important consideration in evaluating the electrical noise that an inductor or transformer is likely to generate is the **magnetic flux path**. In Figure 43, the **slug core** and **toroidal core** types are compared.

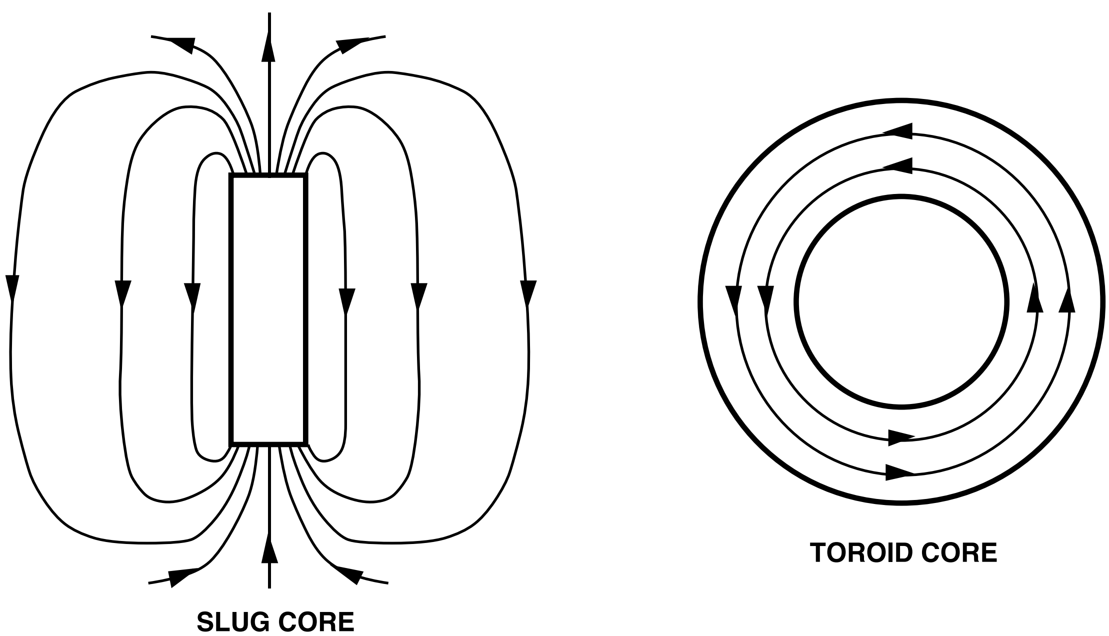


FIGURE 43. FLUX PATHS IN SLUG AND TOROID CORES

The flux in the slug core leaves one end, travels through the air, and returns to the other end. The **slug core is the highest (worst) for radiated flux noise**. In most cases, the slug core device will give the smallest, cheapest component for a given inductor size (it is very cheap to manufacture).

The magnetic flux path in the toroid is **completely contained within the core**. For this reason it has the **lowest (best) radiated flux noise**. Toroid core components are typically larger and more expensive compared to other core types. Winding a toroid is fairly difficult (and requires special equipment), driving up the finished cost of the manufactured transformer.

There is another class of cores commonly used in magnetic design which have radiated flux properties that are much better than the slug core, but not as good as the toroid. These cores are two-piece assemblies, and are assembled by gluing the core pieces together around the bobbin that holds the winding(s).

The cores shown are frequently "gapped" to prevent saturation of the Ferrite core material. The air gap is installed by grinding away a small amount of the core (the gap may be only a few thousandths of an inch).

Figure 44 shows the popular E-I, E-E and Pot cores often used in switching regulator transformers. The cores show the locations where an air gap is placed (if required), but the bobbins/windings are omitted for clarity.

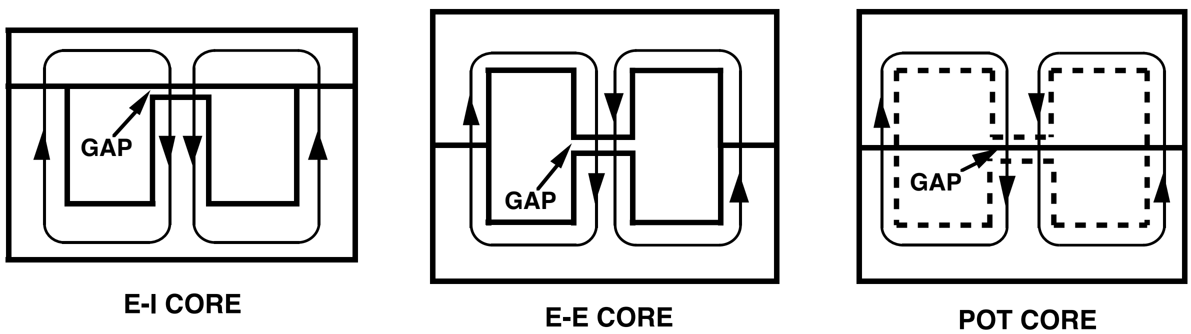


FIGURE 44. FLUX PATHS IN E-I, E-E AND POT CORES

The air gap can emit flux noise because there is a high flux density in the vicinity of the gap, as the flux passing through the core has to jump the air gap to reach the other core piece.

The E-E and E-I cores are fairly cheap and easy to manufacture, and are very common in switching applications up to about 1 kW. There is a wide variety of sizes and shapes available, made from different Ferrite "blends" optimized for excellent switching performance. The radiated flux from this type of core is still reasonably low, and can usually be managed by good board layout.

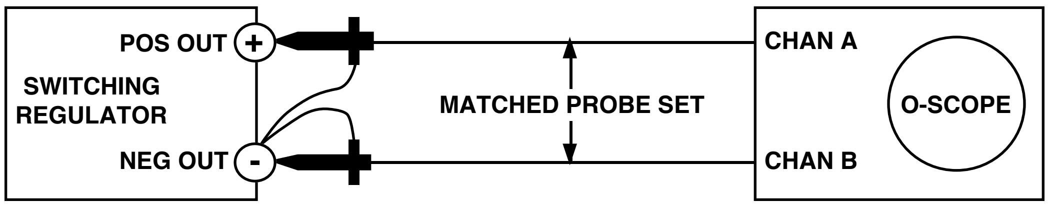
The Pot core (which is difficult to accurately show in a single view drawing), benefits from the shielding effect of the core sides (which are not gapped). This tends to keep the radiated flux contained better than an E-E or E-I core, making the Pot core second best only to the toroid core in minimizing flux noise.

Pot cores are typically more expensive than E-E or E-I cores of comparable power rating, but they have the advantage of being less noisy. Pot core transformers are much easier to manufacture than toroid transformers because the windings are placed on a standard bobbin and then the core is assembled around it.

Measuring Output Ripple Voltage

The ripple appearing on the output of the switching regulator can be important to the circuits under power. Getting an accurate measurement of the output ripple voltage is not always simple.

If the output voltage waveform is measured using an oscilloscope, an accurate result can only be obtained using a **differential** measurement method (see Figure 45).



NOTE: INVERT CHANNEL B AND ADD TO CHANNEL A TO REMOVE COMMON-MODE SIGNAL

FIGURE 45. DIFFERENTIAL OUTPUT RIPPLE MEASUREMENT

The differential measurement shown uses the second channel of the oscilloscope to "cancel out" the signal that is common to both channels (by inverting the B channel signal and adding it to the A channel).

The reason this method must be used is because the fast-switching components in a switching regulator generate voltage spikes that have significant energy at very high frequencies. These signals can be picked up very easily by "antennas" as small as the 3" ground lead on the scope probe.

Assuming the probes are reasonably well matched, the B channel probe will pick up the same radiated signal as the A channel probe, which allows this "common-mode" signal to be eliminated by adding the inverted channel B signal to channel A.

It is often necessary to measure the RMS output ripple voltage, and this is usually done with some type of digital voltmeter. If the reading obtained is to be meaningful, the following must be considered:

1) **The meter must be true-RMS reading**, since the waveforms to be measured are very non-sinusoidal.

2) **The 3dB bandwidth of the meter should be at least 3X the bandwidth of the measured signal** (the output voltage ripple frequency will typically be > 100 kHz).

3) **Subtract the "noise floor" from the measurement**. Connect both meter leads to the negative regulator output and record this value. Move the positive meter lead to positive regulator output and record this value. **The actual RMS ripple voltage is the difference between these two readings**.

Measuring Regulator Efficiency of DC-DC Converters

The **efficiency (η)** of a switching regulator is defined as:

$$\eta = P_{LOAD} / P_{TOTAL}$$

In determining converter efficiency, the first thing that must be measured is the total consumed power (P_{TOTAL}). Assuming a DC input voltage, P_{TOTAL} is defined as the **total power drawn from the source,** which is **equal to:**

$$P_{TOTAL} = V_{IN} \times I_{IN}\,(AVE)$$

It must be noted that the input current value used in the calculation must be the **average value** of the waveform (**the input current will not be DC or sinusoidal**).

Because the total power dissipated must be constant from input to output, P_{TOTAL} is also **equal to the load power plus the internal regulator power losses**:

$$P_{TOTAL} = P_{LOAD} + P_{LOSSES}$$

Measuring (or calculating) the power to the load is very simple, since the output voltage and current are both DC. The load power is found by:

$$P_{LOAD} = V_{OUT} \times I_{LOAD}$$

Measuring the input power drawn from the source is not simple. Although the **input voltage to the regulator is DC, the current drawn at the input of a switching regulator is not**. If a typical "clip-on" current meter is used to measure the input current, the taken data will be essentially meaningless.

The average input current to the regulator can be measured with reasonable accuracy by using a wide-bandwidth current probe connected to an oscilloscope.

The average value of input current can be closely estimated by drawing a horizontal line that divides the waveform in such a way that the **area of the figure above the line will equal the "missing" area below the line** (see Figure 46). In this way, the "average" current shown is equivalent to the value of DC current that would produce the same input power.

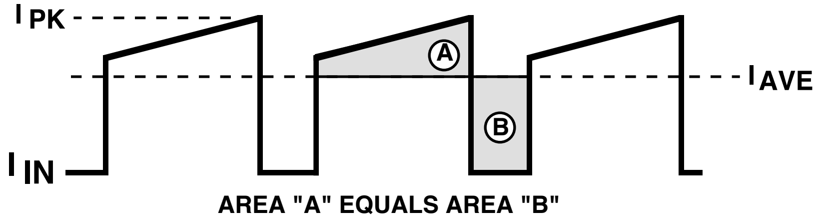


FIGURE 46. AVERAGE VALUE OF TYPICAL INPUT CURRENT WAVEFORM

If more exact measurements are needed, it is possible to **force** the current in the line going to the input of the DC-DC converter to be DC by using an L-C filter between the power source and the input of the converter (see Figure 47).

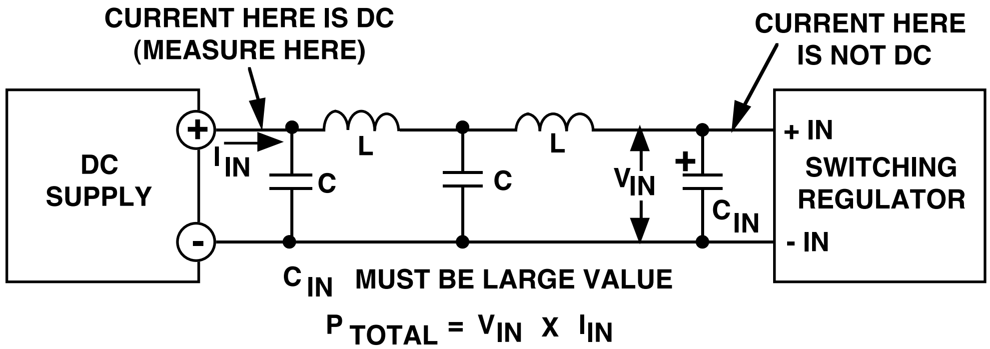


FIGURE 47. L-C FILTER USED IN DC INPUT CURRENT MEASUREMENT

If the L-C filter components are adequate, the current coming from the output of the DC power supply will be DC current (**with no high-frequency switching component**) which means it can be accurately measured with a cheap clip-on ammeter and digital volt meter.

It is essential that a large, low-ESR capacitor be placed at C_{IN} to support the input of the switching converter. The L-C filter that the converter sees looking back into the source presents a high impedance for switching current, which means C_{IN} is necessary to provide the switching current required at the input of the converter.

Measuring Regulator Efficiency of Off-Line Converters

Off-Line converters are powered directly from the AC line, by using a bridge rectifier and capacitive filter to generate an unregulated DC voltage for conversion (see Figures 37 and 38).

Measuring the total power drawn from the AC source is fairly difficult because of the **power factor.** If **both the voltage and current are sinusoidal, power factor is defined as the cosine of the phase angle between the voltage and current waveforms.**

The capacitive-input filter in an off-line converter causes the input current to be **very non-sinusoidal.** The current flows in narrow, high-amplitude pulses (called Haversine pulses) which requires that the power factor be re-defined in such cases.

For capacitive-input filter converters, power factor is defined as:

$$P.F. = P_{REAL} / P_{APPARENT}$$

The **real power drawn from the source** (P_{REAL}) is the power (in **Watts**) which equals the **sum of the load power and regulator internal losses**.

The **apparent power** ($P_{APPARENT}$) is equal to the **RMS input current times the RMS input voltage**. Re-written, the importance of power factor is shown

$$I_{IN}\,(RMS) = P_{REAL} \;/\; \left(V_{IN}\,(RMS) \times PF \right)$$

The RMS input current that the AC line must supply (for a given **real power** in Watts) increases directly as the power factor reduces from unity. Power factor for single-phase AC-powered converters is typically about 0.6. If three-phase power is used, the power factor is about 0.9.

If the efficiency of an off-line converter is to be measured, power analyzers are available which will measure and display input voltage, input power, and power factor. These are fairly expensive, so they may not be available to the designer.

Another method which will give good results is to measure the power **after the rectifier bridge and input capacitor** (where the voltage and current are DC). This method is shown in Figure 48.

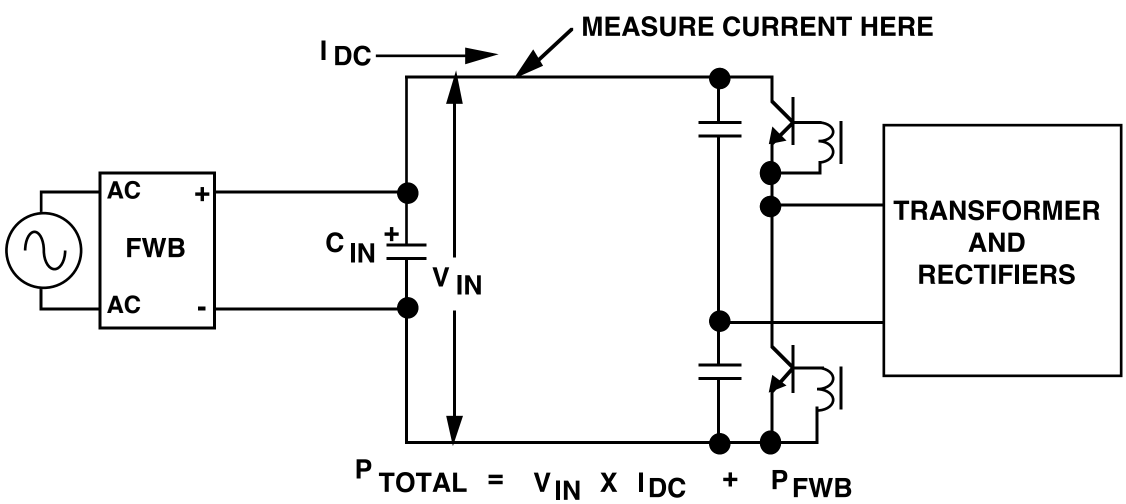


FIGURE 48. MEASURING INPUT POWER IN OFF-LINE CONVERTER

The current flowing from C_{IN} to the converter should be very nearly DC, and the average value can be readily measured or approximated (see previous section).

The total power drawn from the AC source is the sum of the power supplied by C_{IN} (which is V_{IN} X I_{DC}) and the power dissipated in the input bridge rectifier. The power in the bridge rectifier is easily estimated, and is actually negligible in most off-line designs.

Application Circuits

Introduction

Application circuits will be detailed which will demonstrate some examples of switching regulator designs.

LM2577: A Complete Flyback/Boost Regulator IC

The LM2577 is an IC developed as part of the **SIMPLE SWITCHER**™ product family. It is a **current-mode control** switching regulator, with a built-in NPN switch rated for 3A switch current and 65V breakdown voltage.

The most commonly used applications are for Flyback or Boost regulators (see Figure 49). In the Boost regulator, the output is always greater than the input. In the Flyback, the output may be greater than, less than, or equal to the input voltage.

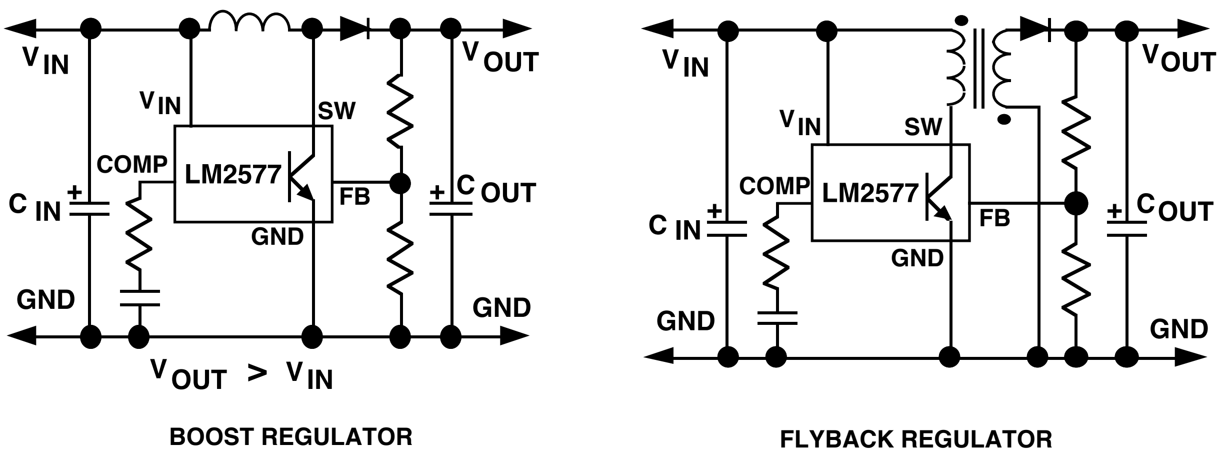


FIGURE 49. BASIC APPLICATION CIRCUITS FOR THE LM2577

The theory of operation of the Flyback and Boost Converters has been previously covered, and will not be repeated here.

The LM2577 is targeted for applications with load power requirements up to a maximum of about 25 Watts, and can be used to implement Boost or Flyback regulators (with multiple output voltages available if Flyback is selected).

The **SIMPLE SWITCHER** product family is supported by design-aid software titled "Switchers Made Simple", which allows finished designs to be generated directly from the computer.

The next sections will show the LM2577 being used in circuits which are more advanced than the typical applications (these circuits were generated as solutions for specific customer requirements).

Increasing Available Load Power in an LM2577 Boost Regulator

One of the most frequently requested circuits is a method to squeeze more power out of a boost converter. The maximum load power available at the output is directly related to the input power available to the DC-DC converter.

When the input voltage is a low value (like 5V), this greatly reduces the amount of power that can be drawn from the source (because the maximum input current is limited by what the switch can handle). In the case of the LM2577, the maximum switch current is 3A (peak).

Increased load power can be obtained with the LM2577 by paralleling two devices (see Figure 50). Because current-mode control is used in the LM2577, the two converters will automatically share the load current demand.

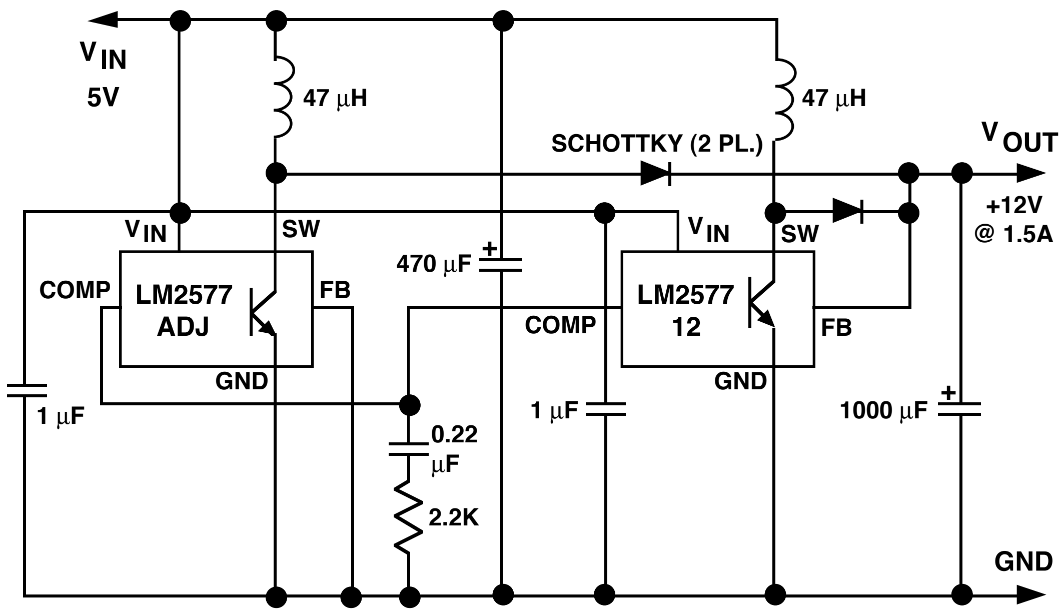


FIGURE 50. DUAL LM2577 BOOST CIRCUIT

The right-hand regulator (which is a fixed 12V version) is the master that sets the duty cycle of both regulators (tying the **Compensation** pins together forces the duty cycles to track).

The master regulator has direct feedback from the output, while the other regulator has its Feedback pin grounded. Grounding the Feedback pin makes the regulator attempt to run "wide open" (at maximum duty cycle), but the master regulator controls the voltage at **both** Compensation pins, which adjusts the pulse widths as required to hold the output voltage at 12V.

LM2577 Negative Buck Regulator

The LM2577 can be used in a Buck regulator configuration that takes a **negative input voltage and produces a regulated negative output voltage** (see Figure 51).

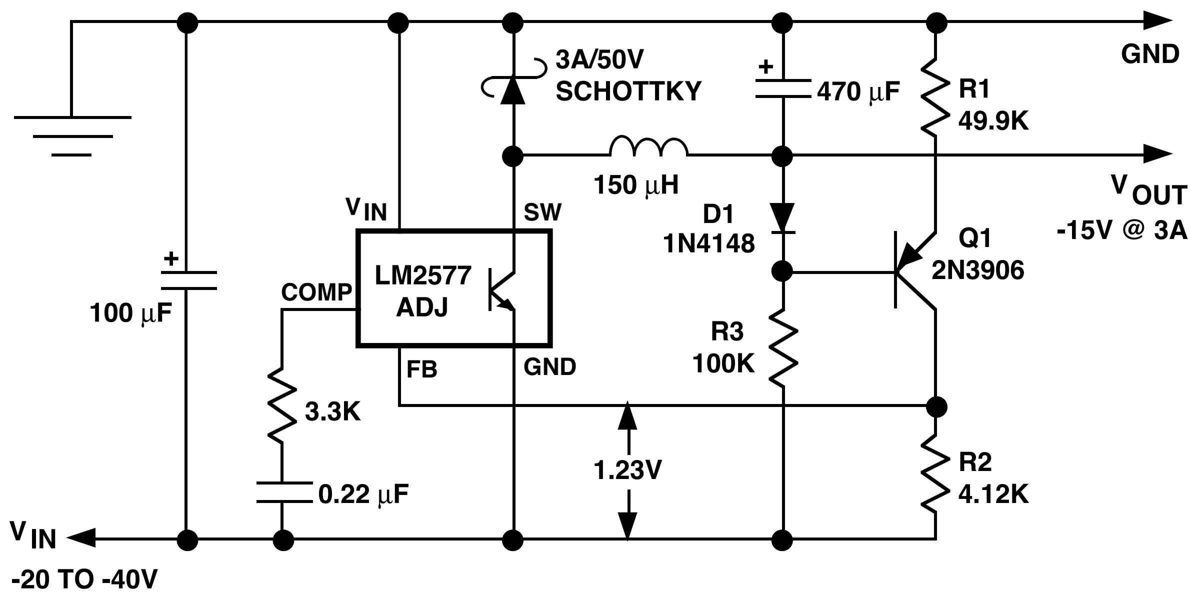


FIGURE 51. NEGATIVE BUCK REGULATOR

The LM2577 is referenced to the negative input, which means the feedback signal coming from the regulated output must be DC level shifted. R1, D1, and Q1 form a current source that sets a current through R2 that is directly proportional to the output voltage (D1 is included to cancel out the V_{BE} of Q1).

Neglecting the base current error of Q1, the current through R2 is equal to:

$$I_{R2} = V_{OUT} / R1$$ (which is **300 µA** for this example.)

The voltage across R2 provides the 1.23V feedback signal which the LM2577 requires for its feedback loop.

The operation of the power converter is similar to what was previously described for the Buck regulator:

When the switch is ON, current flows from ground through the load, into the regulator output, through the inductor, and down through the switch to return to the negative input. The output capacitor also charges during the switch ON time.

When the switch turns OFF, the voltage at the diode end of the inductor flies positive until the Schottky diode turns on (this allows the inductor current to continue to flow through the load during the OFF time). The output capacitor also discharges through the load during the OFF time, providing part of the load current.

LM2577 Three-Output, Isolated Flyback Regulator

Many applications require electrical isolation between the input and output terminals of the power supply (for example, **medical monitoring instruments** require isolation to assure patient safety).

Figure 52 shows an example of a three-output Flyback regulator[2] built using the LM2577 that has electrical isolation between the input and output voltages.

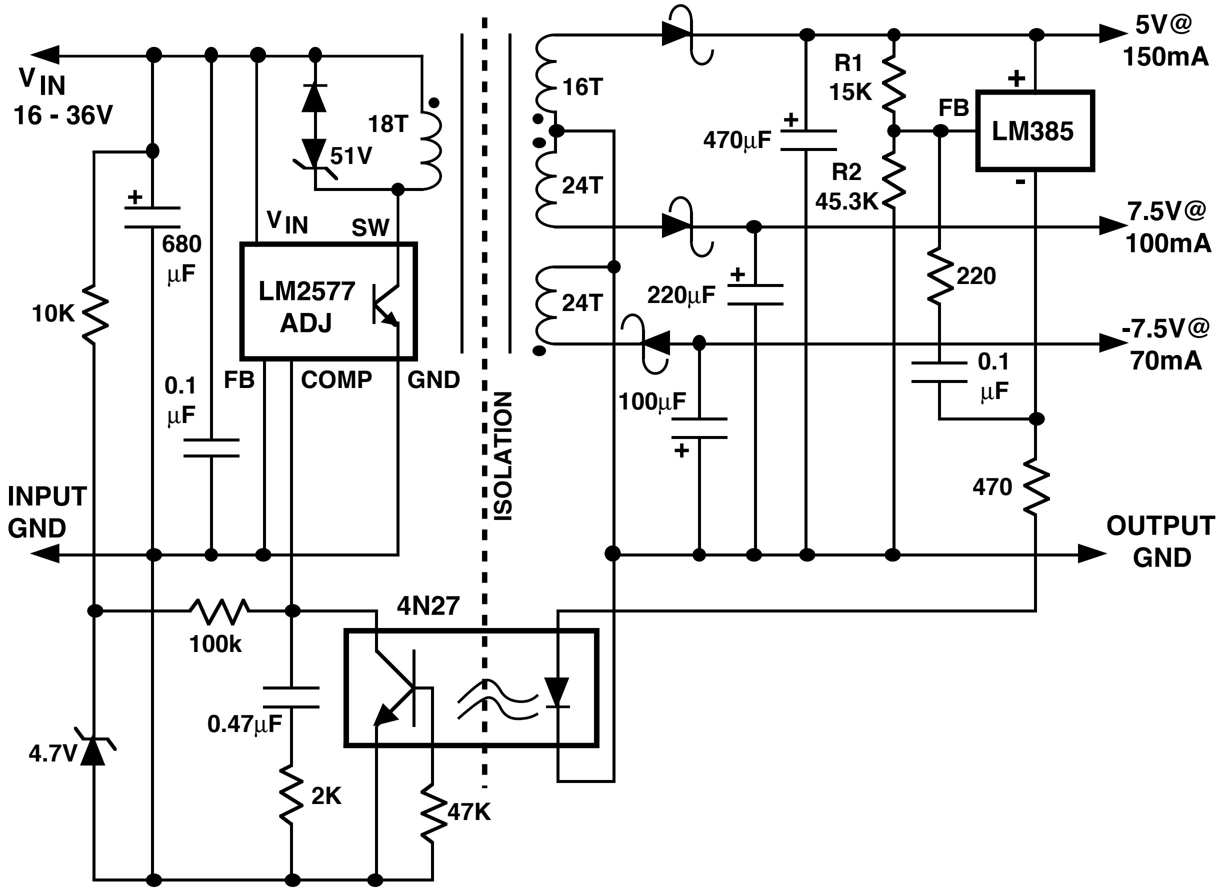


FIGURE 52. THREE-OUTPUT ISOLATED FLYBACK REGULATOR

Three output voltages are obtained from three separate transformer secondary windings, with voltage feedback being taken from the 5V output.

To maintain electrical isolation, the feedback path uses a 4N27 opto-coupler to transfer the feedback signal across the isolation barrier.

The 5V output is regulated using an LM385 adjustable reference, whose voltage is set by R1 and R2. The LM385 operates by forcing a 1.24V reference voltage between the positive terminal and the feedback pin, so the set voltage across the LM385 is given by:

$$V_{REF} = 1.24 \times (R2/R1 + 1)$$

For the values shown in this example, the voltage will be 5V.

The function of the LM385 in the circuit can be described as an "ideal" Zener diode, because the current flowing through the LM385 is very small until the voltage at its positive terminal reaches 5V with respect to ground. At that point, it tries to regulate its positive terminal to 5V by conducting current (which flows out of the negative terminal of the LM385 and through the 470Ω resistor into the diode side of the opto-coupler).

When the LM385 starts conducting current through the opto-coupler diode, the collector of the transistor in the opto-coupler pulls down on the compensation pin of the LM2577, which reduces the duty cycle (pulse widths) of the switching converter. In this way, a negative feedback loop is established which holds the output at 5V.

The feedback signal from the collector of the opto-coupler is fed into the compensation pin (not the feedback pin) of the LM2577 in order to bypass the internal error amplifier of the LM2577. The gain of the LM385 is so high that using the error amplifier inside the loop would make it difficult to stabilize (and is not necessary for good performance).

Test data taken with the input voltage set to 26V and all outputs fully loaded showed the frequency response of the control loop had a 0dB crossover point of 1kHz with a phase margin of 90°.

The 7.5V and -7.5V outputs are not directly regulated, which means their voltages are set by the pulse width of the regulated (5V) winding. As a result, the load regulation of these two outputs is not quite as good as the 5V output.

Summary of test performance data:

Output Voltages	Line Regulation (@ Full Load)	Load Regulation (VIN=26V)	Output Ripple Voltage (25°C)
5V	0.2%	0.04% (30mA - 150mA)	50mV
7.5V	0.3%	3% (20mA - 100mA)	50mV
-7.5V	0.3%	2% (12mA - 70mA)	50mV

LM2575 and LM2576 Buck Regulators

The LM2575 and LM2576 products are Buck regulators developed as part of the **SIMPLE SWITCHER**™ product family.

The LM2575 is rated for 1A of continuous load current, while the LM2576 can supply 3A. The maximum input voltage for the parts is 40V (60V for the "HV" versions), with both adjustable and fixed output voltages available. Both parts are included in the "Switchers Made Simple" design software.

The basic Buck regulator application circuit is shown in Figure 53.

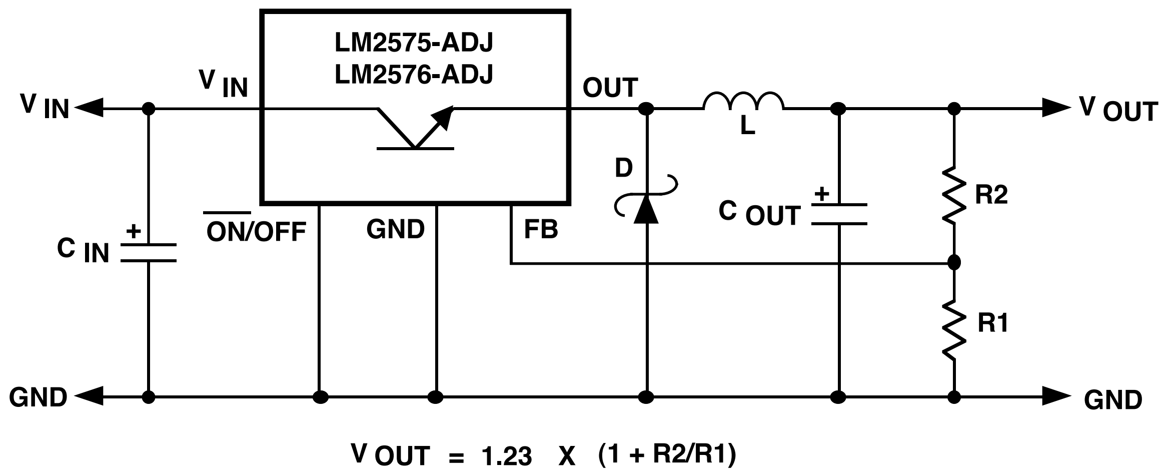


FIGURE 53. LM2575 AND LM2576 BUCK REGULATOR APPLICATION

The LM2575 and LM2576 can also be used in an inverting (Buck-Boost) configuration which allows a positive input voltage to be converted to a negative regulated output voltage (see Figure 54).

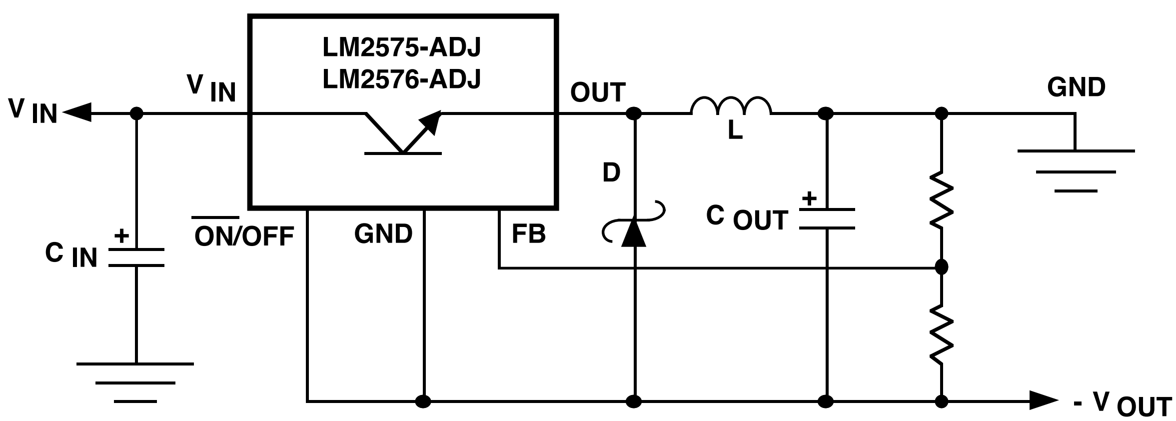


FIGURE 54. LM2575 AND LM2576 INVERTING APPLICATION

Low Dropout, High Efficiency 5V/3A Buck Regulator

A circuit was developed which provides a 5V/3A regulated output voltage with very high efficiency and very low dropout voltage (see Figure 55). The customer required that the circuit be able to operate with an input voltage range of 6V to 12V, allowing **only 1V of dropout** at the lowest input voltage.

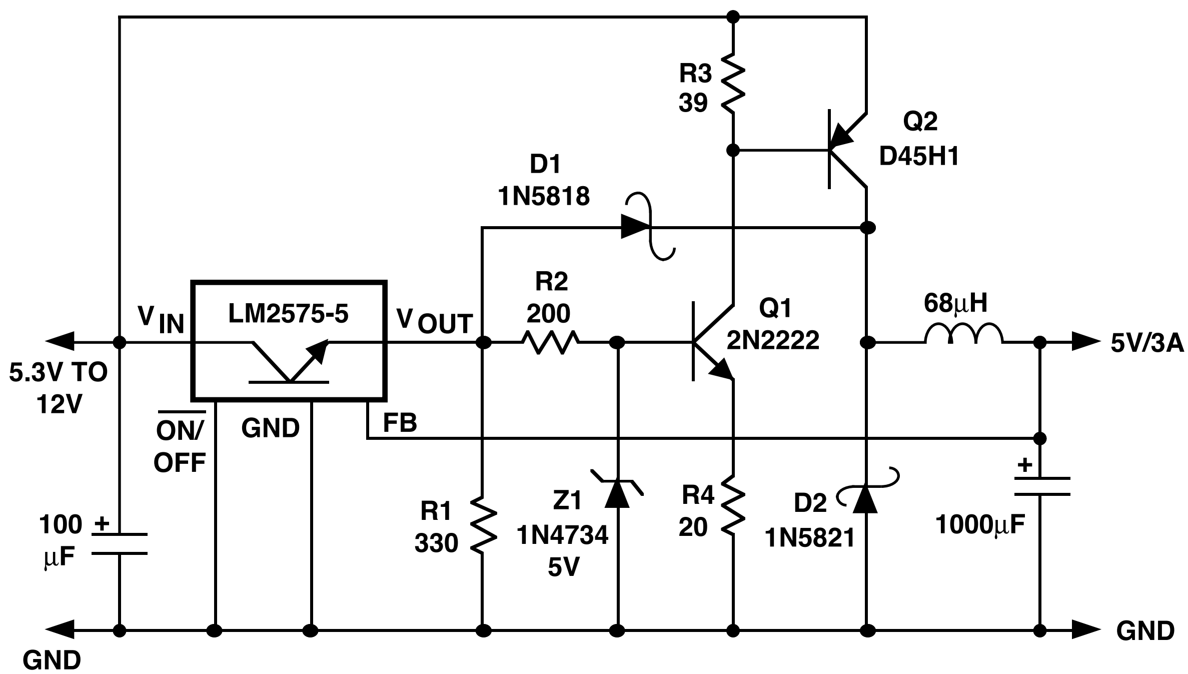


FIGURE 55. LOW-DROPOUT 5V/3A REGULATOR

An unusual feature of this circuit is that it can stay in regulation with **only 300 mV across the regulator**. Also, the **efficiency is highest (89%) at the lowest input voltage** (buck converters are typically more efficient at higher input voltages).

The low (**<300mV**) dropout voltage is achieved by using an external PNP power transistor (Q2) as the main switching transistor (the other transistors in the circuit are drivers for Q2). With components values shown, **Q2 has a saturation voltage of 200mV @ 3A**, which allows the 300 mV input-output differential requirement for the regulator to be met.

The switch inside the LM2575 drives the base of Q1 through R2. Note that the maximum collector current of Q1 (and the maximum base drive available for Q2) is limited by Z1 and R4. When Z1 clamps at 5V, the maximum current through Q1 is:

$$I_{Q1}\ (MAX) = (5 - V_{BE}) / R4 = 215mA$$

The maximum Q1 current (215mA) limits the amount of base drive available to Q2, forcing the collector current of Q2 to "beta limit" as the output is overloaded (this means the maximum collector current of Q2 will be limited by the gain of the transistor and the base drive provided). Although this is not a precise current limiter, it is adequate to protect Q2 from damage during an overload placed on the output.

If the regulator output is **shorted to ground**, the output short-circuit current flows from the output of the LM2575 (through D1 and the inductor), **which means the regulator short-circuit current is limited to the value set internally to the LM2575 (which is about 2A)**.

Note also that when the regulator output is shorted to ground, the cathode of D1 will also be near ground. This allows D1 to clamp off the base drive to Q1 off, preventing current flow in the switch transistor Q2.

If the input voltage does not exceed 8V, R2 and Z1 are not required in the circuit.

This circuit was tested with 6V input and was **able to deliver more than 4A of load current with 5V out**. Other test data taken are:

Measured Performance Data:	
Line Regulation:	
5.3V to 12V @ 1A	32mV
5.3V to 12V @ 3A	45mV
Load Regulation:	
0.3A to 3A @ 5.3V Input	10mV
0.3A to 3A @ 12V Input	17mV
Efficiency @ 3A Load:	
VIN = 5.3V	89%
VIN = 12V	80%
Output Ripple Voltage:	
VIN = 7.2V, IL = 3A	35 mV(p-p)

Batteries in a Portable World

A Handbook on Rechargeable Batteries for Non-Engineers

Isidor Buchmann

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Third Edition

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Chapter 9 | **Testing and Monitoring**

Difficulties with Battery Testing

A German manufacturer of luxury cars points out that one out of two starter batteries returned under warranty is working and has no problem. It is possible that battery testers used in service garages did not detect the batteries correctly before they were returned under warranty. ADAC* reported in 2008 that 40 percent of all roadside automotive failures are battery-related. In Japan, battery failure is the largest single complaint among new car owners. The average car is driven 13km (8 miles) per day and mostly in congested cities. The most common reason for battery failure is undercharge. Battery performance is important; problems during the warranty period tarnish customer satisfaction.

Battery malfunction during the warranty period is seldom a factory defect; driving habits are the culprits. A manufacturer of German-made starter batteries stated that factory defects account for only 5 to 7 percent of warranty claims. The battery remains a weak link, and is evident when reviewing the *ADAC 2008 report* for the year 2007. The study examines the breakdowns of 1.95 million vehicles six years old or less, and Table 9-1 provides the reasons.

Percentage of Failure	Cause of Failure
52%	Battery
15%	Flat tire
8%	Engine
7%	Wheels
7%	Fuel injection
6%	Heating, cooling
5%	Fuel systems

Table 9-1: Most common car failures

Batteries cause the most common failures requiring road assistance.

Source: ADAC 2008

The cellular phone industry experiences an even more astonishing battery return pattern. Nine out of 10 batteries returned under warranty have no problem or can easily be serviced. This is no fault of the manufacturers but they pay a price that is ultimately charged to the user.

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* The ADAC (Allgemeiner Deutscher Automobil-Club e.V.) originated in Germany in 1903 and is Europe's largest automobile club, with over 16 million members.

Part of the problem lies in the difficulty of testing batteries at the consumer level, and this applies to storefronts and service garages alike. Battery rapid-test methods seem to dwell in medieval times, and this is especially evident when comparing advancements made on other fronts. We don't even have a reliable method to estimate state-of-charge — most of such measurements using voltage and coulomb counting are guesswork. Assessing capacity, the most reliable health indicator of a battery, dwells far behind.

The battery user may ask why the industry is lagging so far behind. The answer is simple: battery testing and monitoring is far more complex than outsiders perceive it. As there is no single diagnostic device that can assess the health of a person, so are there no instruments that can quickly check the state-of-health of a battery. Like the human body, batteries can have many hidden deficiencies that no single tester is able to identify with certainly. Yes, we can apply a discharge, but this takes the battery out of service and induces stress, especially on large systems. In some cases, even a discharge does not provide conclusive results either, as we will learn later (see "Discharge Methods" on page 221).

As doctors will examine a patient with different devices, so also does a battery need several approaches to find anomalies. A dead battery is easy to measure and all testers can do this. The challenge comes in evaluating a battery in the 80 to 100 percent performance range. This chapter examines current and futuristic methods and how they stand up. One thing to remember is this: batteries cannot be measured; the appropriate instruments can only make predictions or estimations. This is synonymous with a doctor examining a patient, or the weatherman predicting the weather. All findings are estimations with various degrees of accuracies.

How to Measure Internal Resistance

The resistance of a battery provides useful information about its performance and detects hidden trouble spots. High resistance values are often the triggering point to replace an aging battery, and determining resistance is especially useful in checking stationary batteries. However, resistance comparison alone is not effective, because the value between batches of lead acid batteries can vary by eight percent. Because of this relatively wide tolerance, the resistance method only works effectively when comparing the values for a given battery from birth to retirement. Service crews are asked to take a snapshot of each cell at time of installation and then measure the subtle changes as the cells age. A 25 percent increase in resistance over the original reading hints to an overall performance drop of 20 percent.

Manufacturers of stationary batteries typically honor the warranty if the internal resistance increases by 50 percent. Their preference is to get true capacity readings by applying a full discharge. It is their belief that only a discharge can provide reliable readings and they ask users to perform the service once a year. While this advice has merit, a full discharge requires a temporary disconnection of the battery from the system, and on a large battery such a test takes an entire day to complete. In the real world, very few battery installations receive this type of service and most measurements are based on battery resistance readings.

Measuring the internal resistance is done by reading the voltage drop on a load current or by AC impedance. The results are in ohmic values. There is a notion that internal resistance is related to capacity, and this is false; the resistance of many batteries stays flat through most of the service life. Figure 9-2 shows the capacity fade and internal resistance of lithium-ion cells.

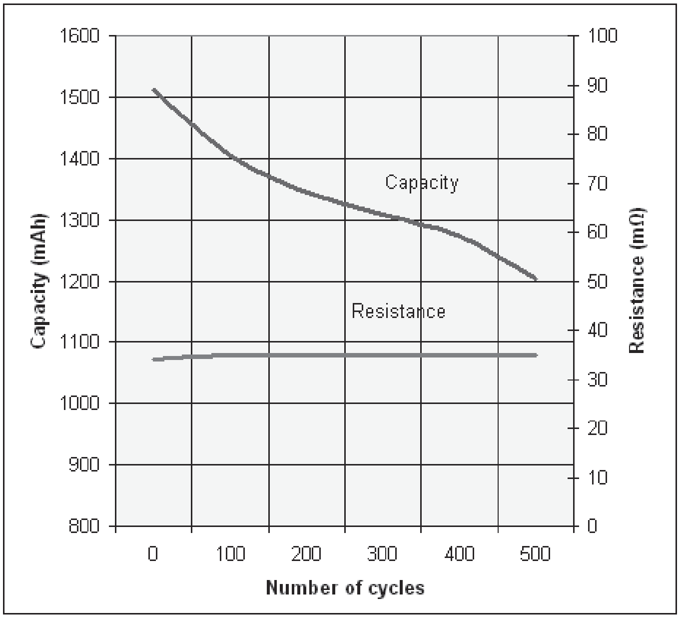

Figure 9-2: Relationship between capacity and resistance as part of cycling

Resistance does not reveal the state-of-health of a battery. The internal resistance often stays flat with use and aging.

Cycle test on Li-ion batteries at 1C:
Charge: 1,500mA to 4.2V, 25°C
Discharge: 1,500 to 2.75V, 25°C

Courtesy of Cadex

What Is Impedance?

Before exploring the different methods of measuring the internal resistance of a battery, let's examine what electrical resistance means, and let's differentiate between a pure *resistance* (R) and *impedance* (Z) that includes reactive elements such as coils and capacitors. Both values are given in Ohms (Ω), a measure formulated by the German physicist Georg Simon Ohm, who lived from 1798 to 1854. (One Ohm produces a voltage drop of 1V with a current flow of 1A.) The difference between resistance and impedance lies in the *reactance.* Let me explain.

The electrical resistance of a pure load, such as a heating element, has no reactance. Voltage and current flow in unison and there is no advancing or trailing phase shift that would occur with a reactive load, such as an electric motor or a florescent light fixture. The ohmic resistance on a pure resistive load is the same with direct current (DC) as is with alternating current (AC). The Power Factor (pf) is 1, which provides the most accurate metering of the power consumed.

Most electrical loads, as well as a battery as power source, have reactance. They consist of *capacitive* reactance (capacitor) and *inductive* reactance (coil). The resistor of a reactance varies with the frequency of the electrical power. The capacitive resistance decreases with higher frequency while the inductive resistance increases. (To explain resistance change with frequency, we compare an oil damper that has a stiffer resistance when moved fast. See also Chapter 1, "Watts and Volt-amps (VA)," on page 32.) A battery has resistive, capacitive and inductive resistance, and the term *impedance* includes all three in one.

Impedance can best be illustrated with the Randles model. Figure 9-3 illustrates the basic model of a lead acid battery, which reflects resistors and a capacitor (R1, R2 and C). The inductive reactance is commonly omitted because it plays a negligible role in a battery, especially at a low frequency.

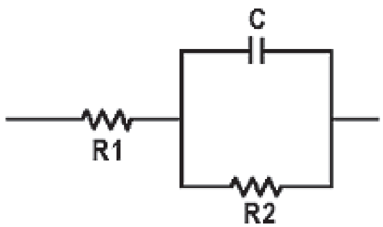

Figure 9-3: Randles model of a lead acid battery

The overall battery resistance consists of ohmic resistance, as well as inductive and capacitive reactance. The schematic and electrical values differ for every battery.

Now that we have learned the basics of internal battery resistance and how they can be applied to rapid-test batteries at different frequencies, this section examines current and future battery test methods. It also discusses advantages and shortfalls.

DC Load Method

Ohmic measurement is one of the oldest and most reliable test methods. The battery receives a brief discharge lasting a few seconds. A small pack gets an ampere or less and a starter battery is loaded with 50A and more. A voltmeter measures the voltage drop and Ohm's law calculates the resistance value (voltage divided by current equals resistance).

DC load measurements work well to check large stationary batteries, and the ohmic readings are very accurate and repeatable. Manufacturers of test instruments claim resistance readings in the 10 micro-ohm range. Many garages use the carbon pile to measure starter batteries, and with experience mechanics familiar with this loading device get a reasonably good assessment of the battery. The invasive test is in many ways more reliable than non-invasive methods.

The DC load method has a limitation in that it blends R1 and R2 of the Randles model into one combined resistor and ignores the capacitor (see Figure 9-4). "C" is an important component of a battery that represents 1.5 farads per 100Ah capacity. In essence, the DC method sees the battery as a resistor and can only provide ohmic references.

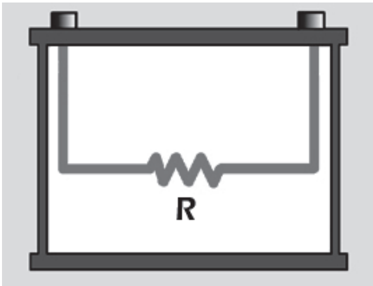

Figure 9-4: DC load method

The true integrity of the Randles model cannot be seen. R1 and R2 appear as one ohmic value.

Courtesy of Cadex

The *two-tier DC load* method offers an alternative method by applying two sequential discharge loads of different currents and time durations. The battery first discharges at a low current for 10 seconds, followed by a higher current for three seconds (see Figure 9-5), and Ohm's law calculates the resistance values. Evaluating the voltage signature under the two load conditions offers additional information about the battery, but the values are strictly resistive and do not reveal SoC and capacity estimations.

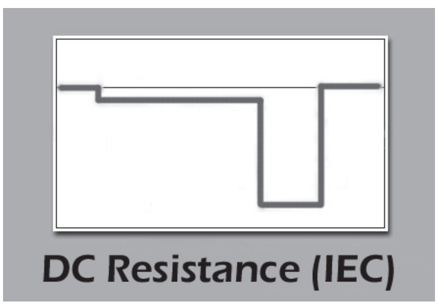

Figure 9-5: Two-tier DC load

The two-tier DC load follows the IEC 60285 and IEC 61436 standards and provides lifelike test conditions for many battery applications. The load test is the preferred method for batteries powering DC loads.

Courtesy of Cadex

AC Conductance

The AC conductance method replaces the DC load and injects an alternating current into the battery. At a set frequency of between 80 and 90 hertz, the capacitive and inductive reactance converge, resulting in a negligible voltage lag that minimizes the reactance. Manufacturers of AC conductance equipment claim battery resistance readings in the 50 micro-ohm range, and these instruments are commonly used in North American car garages. The single-frequency technology as illustrated in Figure 9-6 sees the components of the Randles model as one complex impedance called the *modulus of Z.*

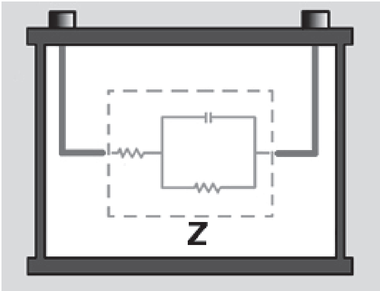

Figure 9-6: AC conductance method

The individual components of the Randles model are molten together and cannot be distinguished.

Courtesy of Cadex

Smaller batteries often use the popular 1000-hertz (Hz) ohm test method. A 1000Hz signal excites the battery, and the Ohm's law calculates the resistance. It is important to note that the AC method shows different values to the DC load, and both are correct. For example, Li-ion in an 18650 cell produces about 36mOhm with a 1000Hz AC signal and roughly 110mOhm with a DC load. Since both readings are correct, and yet are so far apart, the user needs to consider the application. The pulse DC load method provides the best indication for a DC application such as driving a motor or powering a light, while the 1000Hz method better reflects the performance of a digital load, such as a cellular phone that relies to a large extent on the capacitor characteristics of a battery. Figure 9-7 illustrates the 100Hz method.

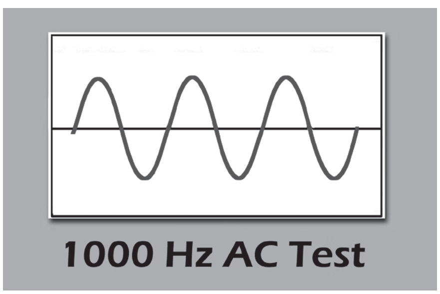

Figure 9-7: 1000-hertz method

The IEC 1000-hertz is the preferred method to take impedance snapshots of batteries powering digital devices.

Courtesy of Cadex

Electrochemical Impedance Spectroscopy

Electrochemical impedance spectroscopy (EIS) enables more than resistance readings; it can estimate state-of-charge and capacity. Research laboratories have been using EIS for many years to evaluate battery characteristics, but high equipment cost, slow test times and the need for trained professionals to decipher large volumes of data have limited this technology to laboratory environments. EIS is able to read each component of the Randles model individually; however, analyzing the value at different frequencies and correlating the data is an enormous task. Fuzzy logic and advanced digital signal processor (DSP) technology have simplified this task. Figure 9-8 illustrates the battery component, which EIS technology is capable of reading.

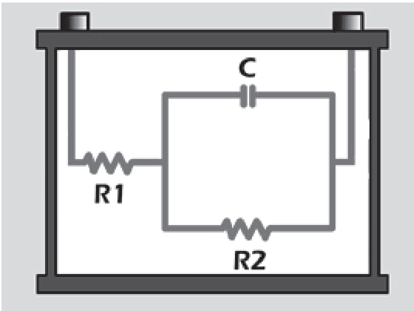

Figure 9-8: Spectro™ method

R1, R2 and C are measured separately, which enables state-of-charge and capacity measurements.

Courtesy of Cadex

How to Measure State-of-charge

Voltage Method

Measuring state-of-charge by voltage is the simplest method, but it can be inaccurate. Cell types have dissimilar chemical compositions that deliver varied voltage profiles. Temperature also plays a role. Higher temperature lowers the open-circuit voltage, a lower temperature raises it, and this phenomenon applies to all chemistries in varying degrees.

The most blatant error of voltage-based SoC occurs when disturbing the battery with a charge or discharge. This agitation distorts the voltage and no longer represents the true state-of-charge. To get accurate measurements, the battery needs to rest for at least four hours to attain equilibrium; battery manufacturers recommend 24 hours. Adding the element of time to neutralize voltage polarization does not sit well with batteries in active duty. One can see that this method is ill suited for fuel gauging.

Each battery chemistry delivers a unique discharge signature that requires a tailored model. While voltage-based SoC works reasonably well for a lead acid battery that has rested, the flat discharge curve of nickel- and lithium-based batteries renders the voltage method impracticable. And yet, voltage is commonly used on consumer products. A “rested” Li-cobalt of 3.80V/cell in open circuit indicates a SoC of roughly 50 percent.

The discharge voltage curves of Li-manganese, Li-phosphate and NMC are very flat, as 80 percent of the charge is stored in a tight voltage window. This profile assists applications requiring a steady voltage but presents a challenge for designers relying on voltage to read SoC. The only practical use of the voltage method on these batteries is to indicate “low charge.”

Lead acid has diverse plate compositions that must be considered when measuring SoC by voltage. Calcium, an additive that makes the battery maintenance-free, raises the voltage by 5–8 percent. Temperature also affects the open-circuit voltage; heat lowers it while cold causes it to rise. Surface charge further fools SoC estimations by showing an elevated voltage immediately after charge; a brief discharge before measurement counteracts the error. Finally, AGM batteries produce a slightly higher voltage than the flooded equivalent.

When measuring SoC by voltage, the battery voltage must be truly "floating," with no load attached. If the battery is installed in a car, any parasitic load can quickly falsify the readings. In spite of the notorious inaccuracies, most SoC measurements rely on the open circuit voltage (OCV) because it's simple, whereas alternative methods are too expensive and need calibration. Voltage-based state-of-charge is popular for wheelchairs, scooters and golf cars.

Hydrometer

The hydrometer offers an alternative to measuring SoC, but this only applies to flooded lead acid and flooded nickel-cadmium. Here is how it works: As the battery accepts charge, the sulfuric acid gets heavier, causing the specific gravity (SG) to increase. As the SoC decreases through discharge, the sulfuric acid removes itself from the electrolyte and binds to the plate, forming lead sulfate. The density of the electrolyte becomes lighter and more water-like, and the specific gravity gets lower. Table 9-9 provides the BCI readings of starter batteries.

Approximate state-of-charge	Average specific gravity	Open circuit voltage			
		2V	6V	8V	12V
100%	1.265	2.10	6.32	8.43	12.65
75%	1.225	2.08	6.22	8. 30	12.45
50%	1.190	2.04	6.12	8.16	12.24
25%	1.155	2.01	6.03	8.04	12.06
0%	1.120	1.98	5.95	7.72	11.89

Table 9-9: BCI standard for SoC estimation of a maintenance-free starter battery with antimony. The readings are taken at room temperature of 26°C (78°F); the battery had rested for 24 hours after charge or discharge.

While BCI specifies the specific gravity of a fully charged starter battery at 1.265, battery manufacturers may go for 1.280 and higher. When increasing the specific gravity, the SoC readings on the look-up table will adjust upwards accordingly. Besides charge level and acid density, the SG can also vary due to low fluid levels, which raises the SG reading because of higher concentration. Alternatively, the battery can be overfilled, which lowers the number. When adding water, allow time for mixing before taking the SG measurement.

The specific gravity also varies according to battery type. Deep-cycle batteries use a dense electrolyte with an SG of up to 1.330 to get maximum runtime; aviation batteries have a SG of 1.285; traction batteries for forklifts are at 1.280; starter batteries come in at 1.265 and stationary batteries are at a low 1.225. Low specific gravity reduces corrosion. The resulting lower specific energy of stationary batteries is not as critical as longevity.

Nothing in the battery world is absolute. The specific gravity of fully charged deep-cycle batteries of the same model can range from 1.270 to 1.305; fully discharged, these batteries may vary between 1.097 and 1.201. Temperature is another variable that alters the specific gravity reading. The colder the temperature is, the higher (more dense) the SG value becomes. Table 9-10 illustrates the SG gravity of a deep-cycle battery at various temperatures.

Temperature of the Electrolyte		Gravity at full charge
40°C	104°F	1.266
30°C	86°F	1.273
20°C	68°F	1.280
10°C	50°F	1.287
0°C	32°F	1.294

Table 9-10: Relation of specific gravity and temperature of deep-cycle battery

Colder temperatures provide higher specific gravity readings.

Errors can also occur if the acid has stratified, meaning the concentration is light on top and heavy on the bottom (Figure 8-15 on page 187). High acid concentration artificially raises the open circuit voltage, which can fool SoC estimations through false SG and voltage indication. The electrolyte needs to stabilize after charge and discharge before taking the SG reading.

Coulomb Counting

Laptops, medical equipment and other professional portable devices use coulomb counting as a SoC indication. This method works on the principle of measuring the current that flows in and out of the battery. If, for example, a battery was charged for one hour at one ampere, the same energy should be available on discharge. This is not the case. Inefficiencies in charge acceptance, especially towards the end of charge, as well as losses during discharge and storage reduce the total energy delivered and skew the readings. The available energy is always less than what had been fed to the battery, and compensation corrects the shortage.

Disregarding these irregularities, coulomb counting works reasonably well, especially for Li-ion. However, the one percent accuracy some device manufacturers advertise is only possible in an ideal world and with a new battery. Independent tests show errors of up to 10 percent when in typical use. Aging causes a gradual deviation from the working model on which the coulomb counter is based. The result is a laptop promising 30 minutes of remaining runtime and all of a sudden the screen goes dark. Periodic calibration by applying a full discharge and charge to reset the flags reduces the error. (See Chapter 6, “Calibration,” on page 148.)

There is a move towards electrochemical impedance spectroscopy (page 214), and even magnetism (page 219) to measure state-of-charge. These new technologies get more accurate estimation than with voltage and can be used when the battery is under load. Furthermore, temperature, surface charge and acid stratification do not affect the readings noticeably.

Impedance Spectroscopy

Impedance spectroscopy evaluates the battery on the impedance values of the Randles model and works on flooded and sealed lead acid. The battery does not need to rest before taking the reading and parasitic loads do not affect the outcome. Figure 9-11 illustrates an incorrect SoC reading because of voltage drop when a load is applied; Figure 9-12 shows the correct result under the same conditions with impedance spectroscopy.

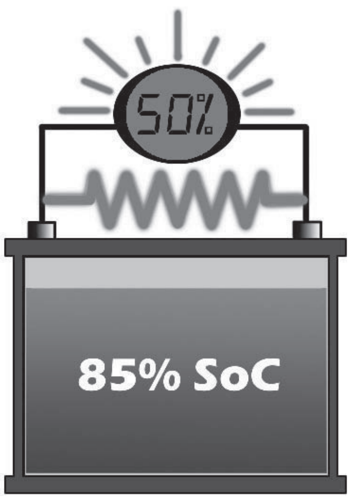

Figure 9-11: BCI*-based SoC reading. A parasitic load distorts voltage-based SoC readings. Voltage recovery takes 4–8 hours.

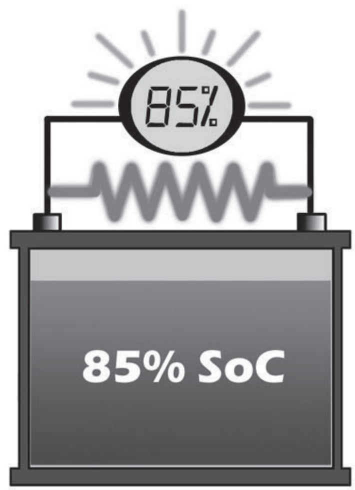

Figure 9-12: SoC based on impedance spectroscopy. A parasitic load does not affect the SoC reading.

* BCI (Battery Council International) measures state-of-charge by open circuit voltage. The voltage methods works well if the battery has no load and has rested after charge or discharge.

Courtesy of Cadex

Quantum Magnetism

In pursuit of a better way to measure battery state-of-charge, researchers are exploring radically new methods, one of which is *quantum magnetism* (Q-Mag™). Q-Mag by Cadex reads magnetism through spin-dependent tunneling. Here is how it works.

When discharging a lead acid battery, the negative plate changes from lead to lead sulfate, which has a different magnetic susceptibility to lead. Measuring the resulting change of the magnetic field with a sensor responding to magnetism provides linear SoC information. The magnetic change also works with lithium-ion, and the feedback is more pronounced than with lead acid. Figure 9-13 shows the concept on a starter battery.

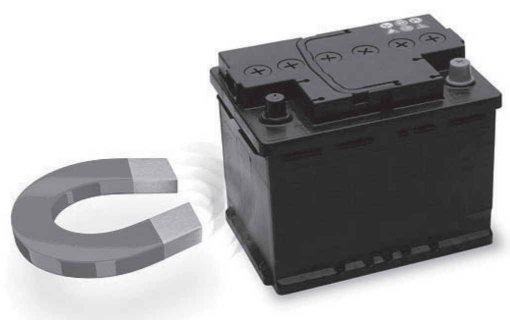

Figure 9-13: State-of-charge measurement by quantum magnetism

Lead fights the applied magnetism less than lead sulfite, allowing SoC measurement by magnetism. Li-ion also responds well to magnetic SoC measurement.

Courtesy of Cadex

The sensor consists of two metal alloys separated by a thin insulator in the nanometer range (thickness of few atoms). The electrons in a magnetic field tunnel through the insulator more easily than in a neutral state, leading to a resistive change. Q-Mag™ interprets state-of-charge using mathematical models. The error is +/–7 percent over the entire SoC range, an accuracy that is unthinkable with voltage measurement, hydrometer and coulomb counters.

All batteries behave in a similar way in that the composition of the electrodes changes, which affects the magnetic characteristics. Q-Mag works on new as well as aged batteries and the technology is immune to voltage distortion caused by loading, charging or surface charge on lead acid. Figure 9-14 shows how magnetic measurements can track discharge/ charge activities of a lead acid battery independent of voltage. The circles represent the voltage under charge and the triangles reveal the state-of-charge.

Measuring the intrinsic state of a battery rather than relying on voltage enables more precise full-charge detection. This feature can be used to improve charge methods and diagnose battery deficiencies, including predicting end-of-life by measuring battery capacity. Q-Mag works also with lithium-ion in non-ferric enclosures. Many of these technologies are proprietary and are in various experimental stages at Cadex.

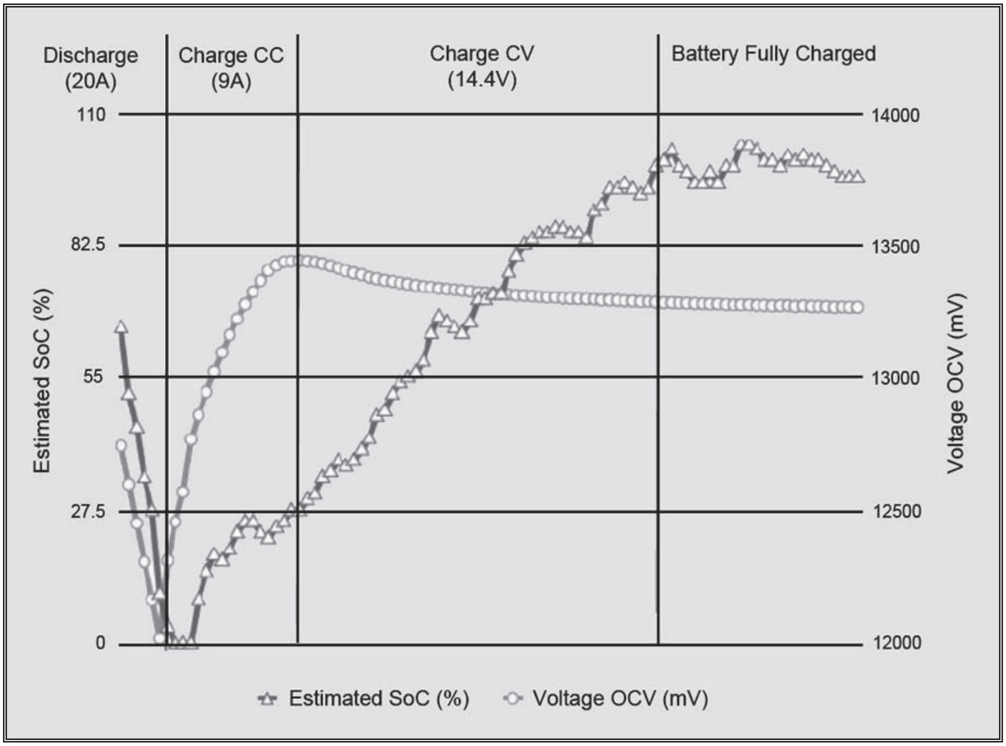

Figure 9-14: Discharge/charge profile of a starter battery
Magnetism traces SoC from 0 to 100% against voltage.

Test method: The battery was first discharged at 20A, followed by a constant charge of 9A to 14.4V and subsequent float charge. (October 2009)

Laboratories of Cadex

How to Measure Capacity

The traditional charge/discharge/charge cycle still offers a dependable way to measure battery capacity. Alternate methods have been tried but none deliver reliable readings. Inaccuracies have led users to adhere to the proven discharge methods even if the process is time-consuming and removes the battery from service for the duration of the test.

While portable batteries can be discharged and recharged relatively quickly, a full discharge and recharge on large lead acid batteries gets quite involved, and service personnel continue to seek faster methods even if the readings are less accurate. This section explains what's available in new technologies, but first we look at the discharge method more closely.

Discharge Method

One would assume that capacity measurement with discharge is accurate but this is not always the case, especially with lead acid batteries. In fact, there are large variations between identical tests, even when using highly accurate equipment and following established charge and discharge standards, with temperature control and mandated rest periods. This behavior is not fully understood except to consider that batteries exhibit human-like qualities. Our IQ levels also vary depending on the time of day and other conditions. Nickel- and lithium-based chemistries provide more consistent results than lead acid on discharge/charge tests.

To verify the capacity on repeat tests, Cadex checked 91 starter batteries with diverse performance levels and plotted the results in Figure 9-15. The horizontal x-axis shows the batteries from weak to strong, and the vertical y-axis reflects capacity. The batteries were prepared in the Cadex laboratories according to SAE J537 standards by giving them a full charge and a 24-hour rest. The capacity was then measured by applying a regulated 25A discharge to 10.50V (1.75V/cell) and the results plotted in diamonds (Test 1). The test was repeated under identical conditions and the resulting capacities added in squares (Test 2). The second reading exhibits differences in capacity of +/–15 percent across the battery population. Other laboratories that test lead acid batteries experience similar discrepancies.

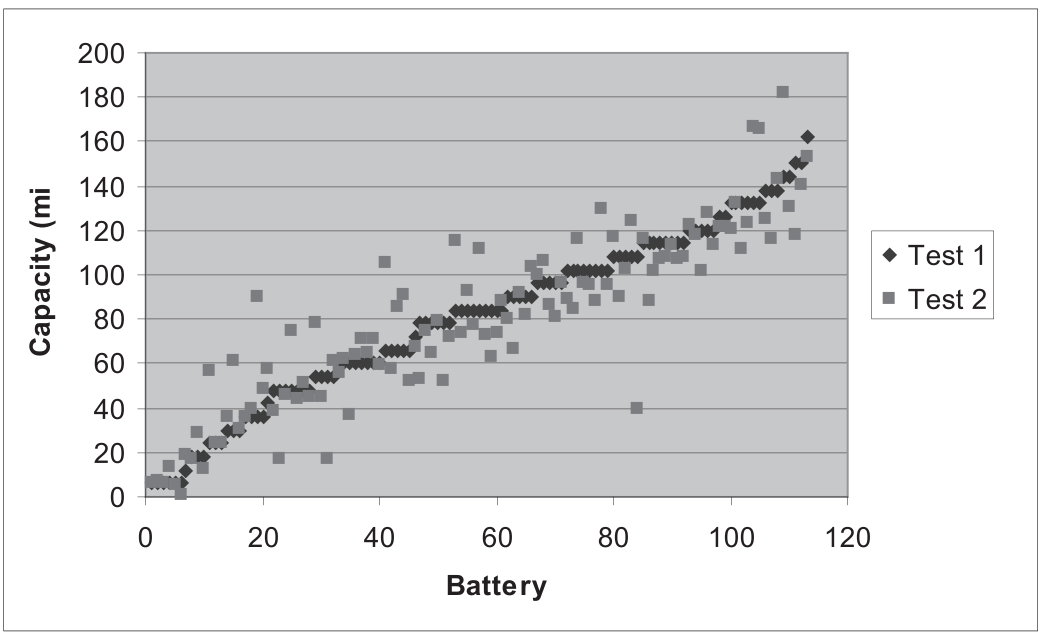

Figure 9-15: Capacity fluctuations on two identical charge/discharge tests of 91 starter batteries. The capacities differ +/–15% between Test 1 and Test 2.

Courtesy of Cadex (2005)

Capacity vs. CCA

Starter batteries have two distinct values, *CCA* and *capacity.* These two readings are close to each other like lips and teeth, but the characteristics are uniquely different; one cannot predict the other. (See Chapter 8, "How Age Affects Capacity and Resistance," on page 188.)

Measuring the internal battery resistance, which relates to CCA on a starter battery, is relatively simple but the reading provides only a snapshot of the battery at time of measurement. Resistance alone cannot predict the end of life of a battery. For example, at a CCA of 560A and a capacity of 25 percent, for example, a starter battery will still crank well but it can surprise the motorist with a sudden failure of not turning the engine (as I have experienced).

The leading health indicator of a battery is *capacity,* but this estimation is difficult to read. A capacity test by discharge is not practical with starter batteries; this would cause undue stress and take a day to complete. Most battery testers do not measure capacity but look at the internal resistance, which is an approximation of CCA. The term *approximation* is correct — laboratory tests at Cadex and at a German luxury car manufacturer reveal that the readings are only about 70 percent accurate. A full CCA test is seldom done; one battery can take a week to measure.

The SAE J537 CCA test by BCI mandates to cool a fully charged battery to -18°C (0°F) for 24 hours, and while at subfreezing temperature apply a high-current discharge that simulates the cranking of an engine. A 500 CCA battery would need to supply 500A for 30 seconds and stay above 7.2V (1.2V/cell) to pass. If it fails the test, the battery has a CCA rating of less than 500A. To find the CCA rating, the test must be repeated several times with different current settings to find the triggering point when the battery passes through 7.2V line. Between each test, the battery must be brought to ambient temperature for recharging and cooled again for testing. (For CCA with DIN and IEC norms, refer to "Test Method" on page 223)

To examine the relationship between CCA and capacity, Cadex measured CCA and capacity of 175 starter batteries at various performance levels. Figure 9-16 shows the CCA on the vertical y-axis and reserve capacity* readings on the horizontal x-axis. The batteries are arranged from low to high, and the values are given as a percentage of the original ratings.

The table shows noticeable discrepancies between CCA and capacity, and there is little correlation between these readings. Rather than converging along the diagonal reference line, CCA and RC wander off in both directions and resemble the stars in a clear sky. A closer look reveals that CCA gravitates above the reference line, leaving the lower right vacant. High CCA with low capacity is common, however, low CCA with high capacity is rare. In our table, one battery has 90 percent CCA and produces a low 38 percent capacity; another delivers 71 percent CCA and delivers a whopping 112 percent capacity (these are indicated by the dotted lines).

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* North America marks the reserve capacity (RC) of starter batteries in minutes; RC applies a 25A discharge to 1.75V/cell and measures the elapsed time in minutes. Europe and other parts of the world use ampere-hours (Ah). The RC-to-Ah conversion formula is as follows: RC divided by 2 plus 16.

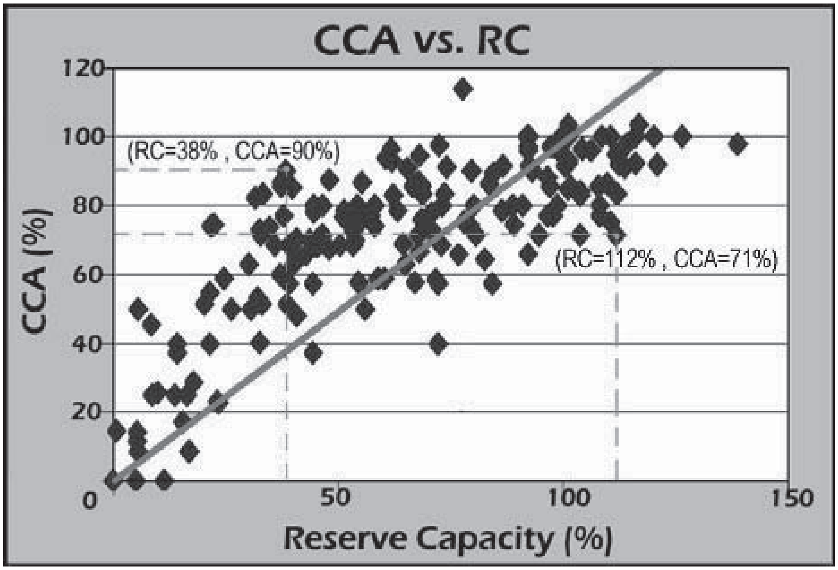

Figure 9-16: CCA and reserve capacity (RC) of 175 aging starter batteries

The CCA of aging starter batteries gravitates above the diagonal reference line. (Few batteries have low CCA and high capacity.)

Courtesy of Cadex

Test method: The CCA and RC readings were obtained according to SAE J537 standards (BCI). CCA (BCI) loads a fully charged battery at –18°C (0°F) for 30s at the CCA-rated current of the battery. The voltage must stay above 7.2V to pass. CCA DIN and IEC norms are similar with these differences: DIN discharges for 30s to 9V, and 150s to 6V; IEC discharges for 60s to 8.4V. RC applies a 25A discharge to 1.75V/cell and measures the elapsed time in minutes.

As discussed earlier, a battery check must include several test points. An analogy can be made with a medical doctor who examines a patient with several instruments to find the diagnosis. A serious illness could escape the doctor's watchful eyes if only blood pressure or temperature was taken. While medical staff are well trained to evaluate multiple data points, most battery personnel do not have the knowledge to read a Nyquist plot and other data on a battery scan. Nor are test devices available that give reliable diagnosis of all battery ills.

Testing Lead Acid

Many manufacturers of battery testers claim to measure battery health on the fly. These instruments work well in finding battery defects that involve voltage anomalies and elevated internal resistance, but other performance criteria remain unknown. Stating that a battery tester based on internal resistance can also measure capacity is misleading. Advertising features that are outside the equipment's capabilities confuses the industry into believing that multifaceted results are attainable with basic methods. Manufacturers of these instruments are aware of the complexity involved, but some like to add a flair of mystery in their marketing scheme, similar to a maker of a shampoo product promising to grow lush hair on a man's bald head. Here is a brief history of battery testers for lead acid and what they can do.

The *carbon pile,* introduced in the 1980s, applies a DC load of short duration to a starter battery, simulating cranking. The voltage drop and recovery time provide a rough indication of battery health. The test works reasonably well and offers evidence that power is present. A major advantage is the ability to detect batteries that have failed due to a shorted cell (low specific gravity in one cell due to high self-discharge). Capacity estimation, however, is not possible, and a battery that simply has a low state-of-charge appears as *weak.* In addition, the tester must rely on voltage to estimate state-of-charge. A skilled mechanic can, however, detect a faulty battery based on the voltage signature and loading behavior.

The *AC conductance* meters appeared in 1992 and were hailed as a breakthrough. The non-invasive method injects an AC signal into the battery to measure the internal resistance. Today, these testers are commonly used to check the CCA of starter batteries and verify resistance change in stationary batteries. While small and easier to use, AC conductance cannot read capacity, and the resistive value gives only an approximation of the real CCA of a starter battery. A shorted cell could pass as good because in such a battery the overall conductivity and terminal voltage are close to normal, even though the battery cannot crank the motor. AC conductance testers are common in North America; Europe prefers the DC load method.

Critical progress has been made towards *electrochemical impedance spectroscopy* (EIS). Cadex took the EIS technology a step further and developed battery specific models that are able to estimate the health of a lead acid battery. *Multi-model electrochemical impedance spectroscopy,* or Spectro™ for short, reads battery capacity, CCA and state-of-charge in a single, non-invasive test. Figure 9-17 illustrates the Spectro CA-12 handheld battery tester.

Figure 9-17: Spectro CA-12 battery tester

Compact battery rapid tester displays capacity, CCA and state-of-charge in 15 seconds.

Courtesy Cadex

The Spectro CA-12 handheld device, in which the Spectro™ technology is embedded, excites the battery with signals from 20–2000Hz. A DSP deciphers the 40 million transactions churned out during the 15-second test into readable results. To check a battery, the user simply selects the battery voltage, Ah and designated matrix. Tests can be done under a steady load of up to 30A and a partial charge, however, if the state-of-charge is less than 40 percent, the instrument advises the user to charge and retest.

The Spectro method is a further development of EIS, a technology that had been around for several decades. What's new is the use of multi models and faster process times. Cost and size have also shrunk. Earlier models cost tens of thousands of dollars and traveled on wheels. The heart of Spectro is not so much the mechanics but the algorithm. No longer do modern EIS devices accompany a team of scientist to decipher tons of data. Experts predict that the battery industry is moving towards the multi-model EIS technology to estimate batter performance

Nowhere is the ability to read capacity more meaningful than with deep-cycle batteries in golf cars, aerial work platforms and wheelchairs, as well as military and naval carriers. Getting a readout in seconds without putting the vehicles out of commission allows for a quick performance check on a suspect battery before deployment in the field. Figures 9-18, 9-19 and 9-20 show typical battery problems and how modern test technologies can detect them.

Figure 9-18: Low charge

Drive is sluggish; Spectro™ reads low SoC. Capacity estimation is correct in spite of low charge.

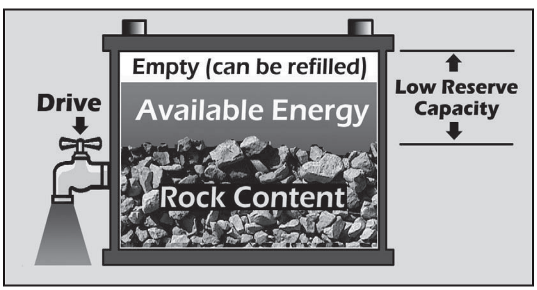

Figure 9-19: Low capacity

Battery has good drive but short runtimes. Spectro™ reads good impedance but low capacity.

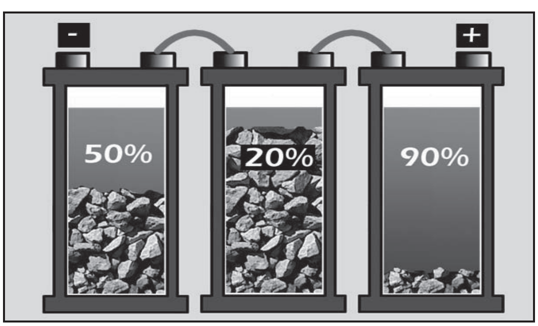

Figure 9-20: Faulty set

Spectro™ finds low performing and shorted blocks in a string. Good batteries can be regrouped and reused.

All figures Courtesy of Cadex

Matrices
Measurement devices, such as the Spectro CA-12, are not universal instruments capable of estimating the capacity of any battery that may come along; they require battery specific matrices, also known as pattern recognition algorithm. A matrix is a multi dimensional lookup table against which the measured readings are compared. Text recognition, fingerprint identification and visual imaging operate on a similar principle in that a model exists, with which to equate the derived readings.

This book identifies three commonly used measuring methods. The principle in all is to take one or several sets of readings and compare them against known reference settings or images to disclose the characteristics of a battery. The three methods are as follows.

Scalar: The *single value scalar test* takes a reading and compares the result with a stored reference value. In battery testing this could be measuring a voltage, interrogating the battery by applying discharge pulses or injecting a frequency and then comparing the derived result against a single reference point. This is the simplest test, and most DC load and single-frequency AC conductance testers use this method.

Vector: The *vector method* applies pulses of different currents, or excites the battery with several frequencies, and evaluates the results against preset vector points to study the battery under various stress conditions. Typical applications for this one-dimensional scalar model are battery testers that apply multi-tier DC loads or inject several test frequencies. Because of added complexity in evaluating the different data points and limited benefits, the vector method is seldom used.

Matrix: The *matrix method*, which was introduced on page 224, scans a battery with a frequency spectrum as if to capture the image of a landscape and compare the imprint with a stored model of known characteristics. This multi-dimensional set of scalars, which form the foundation of Spectro™, provides the most in-depth information but is complex in terms of evaluating the data generated. With a proprietary algorithm, the Spectro™ technology is able to estimate battery capacity, CCA and SoC.

Matrices are primarily used to estimate battery capacity, however, CCA and state-of-charge also require matrices. These are easier to assemble and serve a broad range of starter batteries. While the Spectro™ method offers an accuracy of 80 to 90 percent on capacity, CCA is 95 percent exact. This compares to 60 to 70 percent with battery testers using the scalar method. Service personnel are often unaware of the low accuracy; verifications are seldom done, as this would involve several days of laboratory testing.

A further drawback of scalar battery testers is obtaining a reading that is neither resistance nor CCA. While there are similarities between the two, no standard exists and each instrument gives different values. In terms of assessing a dying battery, however, this method is adequate as it reflects conductivity. The larger disadvantage is not being able to read capacity. Table 9-21 illustrates test accuracies using scalar, vector and matrix methods.

Measuring units	Scalar Single value	Vector One-dimensional set of scalars	Matrix Multi-dimensional set of scalars
CCA	60–70% accurate		90–95% accurate
Capacity	N/A		80–90% accurate
SoC	Voltage-based; requires rest after charge and discharge		90–95% accurate (with new battery)

Table 9-21: Accuracy in battery readings with different measuring methods
Scalar and vector provide resistance with references to CCA on starter batteries. The matrix method is more accurate and provides capacity estimations but needs reference matrices.

To generate a matrix, batteries with different state-of-health are scanned. The more batteries of the same model but diverse capacity mix are included in the mix, the stronger the matrix will become. If, for example, the matrix consists only of two batteries, one showing a capacity of 60 percent and the other 100 percent, then the accuracy would be low for the batteries in between. Adding a third battery with an 80 percent capacity will solidify the matrix, similar to placing a pillar at the center of a bridge. To cover the full spectrum, a well-developed matrix should include battery samples capturing capacities of 50, 60, 70, 80, 90 and 100 percent. Batteries much below 50 percent are less important because they constitute a fail.

It is difficult to obtain aged batteries, especially with newer models. Forced aging by cycling in an environmental chamber is of some help; however, age-related stresses from the field are not represented accurately. It also helps to include batteries from different regions to represent unique environmental user patterns. A starter battery in a Las Vegas taxi has different strains than that of a car driven by a grandmother in northern Germany.

Different state-of-charge levels increase the complexity to estimate battery health. The SoC on a new battery can be determined relatively easily with impedance spectroscopy, however, the formula changes as the battery ages. A battery tester should therefore be capable of examining new and old batteries with a charge level of 40 to 100 percent. With ample data, this is possible because natural aging of a battery is predictable and the scanned information can be massaged to calculate age. This is similar to face recognition that correctly identifies a person even if he or she has developed a few wrinkles and has grown gray hair.

Simplifications in matrix development are possible by grouping batteries that share the same chemistry, voltage and a similar capacity range into a generic matrix. This simplifies logistics; however, the readout is classified into categories rather than numbers. Figure 9-22 illustrates the classification scheme of Good, Low and Poor. Good passes as a good battery; Low is suspect and predicts the end of life; and Poor is a fail that mandates replacement.

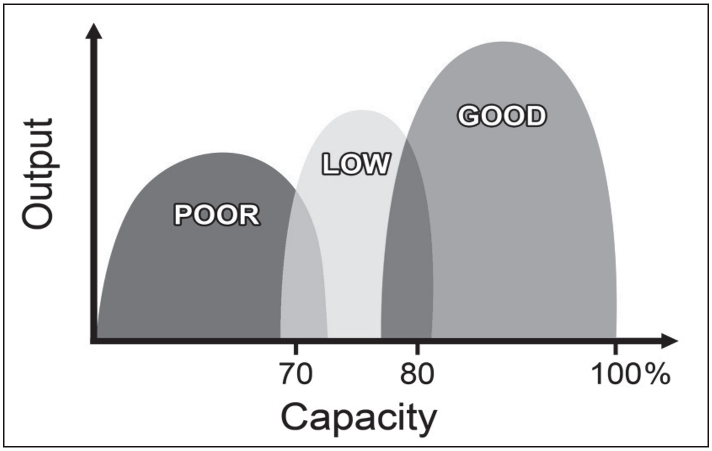

Figure 9-22: Classifying batteries into categories

The classification method provides an intelligent assessment of what constitutes a usable battery for a given application. Some classifications have pass/fail; others provide GOOD, LOW and POOR.

Courtesy of Cadex

Service personnel appreciate the classification method because it gives them an intelligent assessment of what constitutes a usable battery for a given application and eliminates customer interference. If numeric capacity readings are mandatory for a given battery type, a designated matrix can be developed and downloaded into the tester from the Internet.

Testing Nickel-based Batteries

Nickel-based batteries have unique properties, and Cadex developed a rapid-test method for these battery systems called *QuickTest*™. The process takes three minutes and uses an inference algorithm. Figure 9-23 illustrates the general structure of the algorithm applied.

QuickTest™ fuses data from six variables, which are capacity, internal resistance, self-discharge, charge acceptance, discharge capabilities and mobility of electrolyte. A trend-learning algorithm combines the data to provide a dependable state-of-health (SoH) reading in percentage. The system uses battery-specific matrices stored in battery adapters of a designated battery analyzer (Cadex). The user can create a matrix in the field by scanning two or more batteries on the analyzer's *Learn* program. The battery must be at least 20 percent charged.

Among other parameters, QuickTest™ relies on the internal resistance of a battery pack, and the welding joints between the cells might cause a problem, especially on packs with 10 cells or more. Although seemingly insignificant in terms of added resistance, mechanical linkages behave differently to a chemical cell and this causes an unwanted error. The linkage error is not seen on a conventional discharge test or when doing a resistance check but interferes with rapid-test methods on voltages above 20V. It is also possible that each cell of a multi-cell pack behaves on its own when excited with a common signal and the result gets muddled.

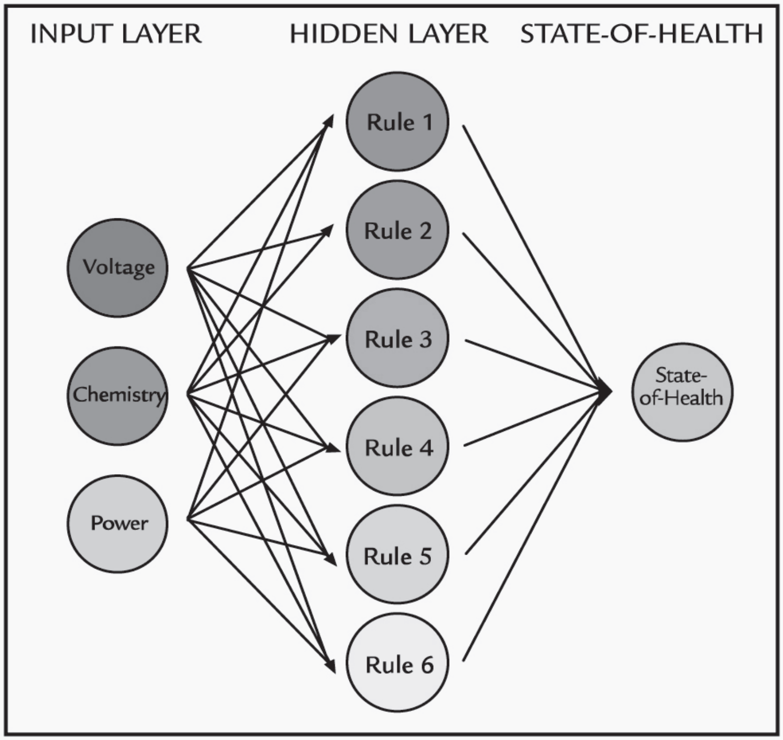

Figure 9-23: QuickTest™ structure

Multiple variables are fed to the micro controller, "'fuzzified" and processed by parallel logic. The data is averaged and weighted according to battery application.

Courtesy of Cadex

Testing Lithium-based Batteries

With the large number of lithium-ion batteries in use and the population growing rapidly, developing an effective testing method has become an urgent task. *QuickSort*™ (Cadex) is a further development of QuickTest™ using a generic matrix. The simplification was made possible by limiting the battery population to single-cell Li-ion from 500 to 1,500mAh. (Larger cells and higher voltages will need a different generic matrix.) Rather than capacity readout in percentage, QuickSort™ classifies the battery health as Good, Low or Poor.

Electrochemical dynamic response, the method used for QuickSort™, measures the mobility of ion flow between the electrodes on a digital load. The response can be compared with a mechanical arm under load. A strong arm resembling a good battery remains firm, and a weak arm synonymous to a faded battery bends and becomes sluggish under load.

The test takes 30 seconds, is 90 percent accurate regardless of battery cathode material and can be performed with a state-of-charge range of between 40 and 100 percent. QuickSort™ requires the correct mAh, and setting a wrong value does not shift the reading on a linear scale from good to poor, as one would expect, but makes the sorting less accurate. The system does not rely on internal resistance per se. This would produce unreliable readings because modern lithium-ion maintains low resistance with use and time (see Figure 9-2 on page 211). An overall resistance check is only done at the conclusion of the test. Figure 9-24 shows the concept.

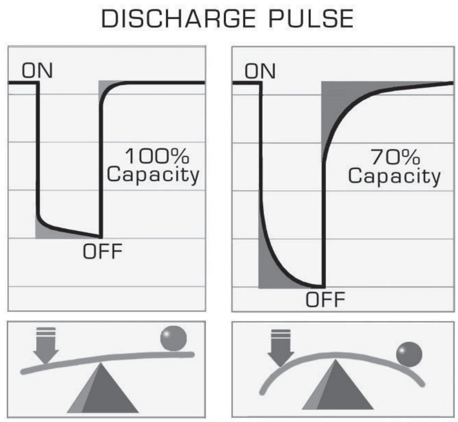

Figure 9-24:
Electrochemical dynamic response

The electrochemical dynamic response measures the ion flow between the positive and negative plates. This process can be compared to a mechanical arm under load.

Courtesy of Cadex

Lithium-ion batteries have different diffusion rates, and in terms of electrochemical dynamic response, Li-ion polymer with gelled electrolyte appears to be faster than Li-ion containing liquefied electrolyte. Li-polymer may need a different matrix to produce accurate readings.

Scientists explore new ways to evaluate the health of a battery with scanning frequencies ranging from several kilohertz to milihertz. High frequencies reveal the resistive qualities of a battery, which presents a bird-eye's view in landscape form. By lowering the frequency, diffusion begins to provide insight into unique battery characteristics that allow capacity estimation, sulfation detection and revealing of dry-out condition.

Evaluating batteries at sub one-hertz frequency needs long test times. At one milihertz, for example, a cycle takes 1,000 seconds and several data points are required to assess a battery with certainty. Low-frequency tests can take several minutes for one measurement, however, with clever software simulation, the duration can be shortened to just a few seconds.

Research engineers at Cadex are working on a technique called *Low Frequency Pulse Train* (LFPT), also known as *diffusion technology*. Diffusion works with most chemistries and the information retrieved provides vital information relating to battery capacity and underlying deficiencies. This knowledge enables the all-important *state-of-life estimation*, the ultimate goal for advanced battery management systems (BMS).

There is a critical need for practical battery testers that can examine the state-of-health of batteries in medical equipment, military instruments, computing devices, power tools and UPS systems. There are currently no instruments that can reliably predict battery state-of-life on the fly, although many device manufacturers may claim their instruments will do so.

How to Monitor a Battery

One of the most urgent requirements for battery-powered devices is the development of a reliable and economical way to monitor battery state-of-function (SoF). This is a demanding task when considering that there is still no dependable method to read state-of-charge, the most basic characteristic of a battery. Even if SoC were displayed accurately, charge information alone has limited benefits without knowing the capacity. The objective is to identify *battery readiness,* which describes what the battery can deliver at a given moment. SoF includes capacity (the amount of energy the battery can hold), internal resistance (the delivery of power), and state-of-charge (the amount of energy the battery holds at that moment).

Stationary batteries were among the first to include monitoring systems, and the most common form of supervision is voltage measurement of individual cells. Some systems also include cell temperature and current measurement. Knowing the voltage drop of each cell at a given load reveals cell resistance. Cell failure caused by rising resistance through plate separation, corrosion and other malfunctions can thus be identified. Battery monitoring also serves in medical, defense and communication devices, as well as wheeled mobility and electric vehicle applications.

In many ways, present battery monitoring falls short of meeting the basic requirements. Besides assuring *readiness*, battery monitoring should also keep track of aging and offer end-of-life predictions so that the user knows when to replace a fading battery. This is currently not being done in a satisfactory manner. Most monitoring systems are tailored for new batteries and adjust poorly to aging ones. As a result, battery management systems (BMS) tend to lose accuracy gradually until the information obtained gets so far off that it becomes a nuisance. This is not an oversight by the manufacturers; engineers know about this shortcoming. The problem lies in technology, or lack thereof.

Another limitation of current monitoring systems is the bandwidth in which battery conditions can be read. Most systems only reveal anomalies once the battery performance has dropped below 70 percent and the performance is being affected. Assessment in the | all-important 80–100 percent operating range is currently impossible, and systems give the batteries a good bill of health. This complicates end-of-life predictions, and the user needs to wait until the battery has sufficiently deteriorated to make an assessment. Measuring a battery once the performance has dropped or the battery has died is ineffective, and this complicates battery exchange systems proposed for the electric vehicle market. One maker of a battery tester proudly states in a brochure that their instrument "Detects any faulty battery." So, eventually, does the user.

Some medical devices use date stamp or cycle count to determine the end of service life of a battery. This does not work well either, because batteries that are used little are not exposed to the same stresses as those in daily operation. To reduce the risk of failure, authorities may mandate an earlier replacement of all batteries. This causes the replacement of many packs that are still in good working condition. Old habits are hard to break, and it is often easier to leave the procedure as written rather than to revolt. This satisfies the battery vendor but increases operating costs and creates environmental burdens.

Portable devices such as laptops use coulomb counting that keeps track of the in- and out flowing currents. Such a monitoring device should be flawless, but as mentioned earlier, the method is not ideal either. Internal losses and inaccuracies in capturing current flow add to an unwanted error that must be corrected with periodic calibrations.

Over-expectation with monitoring methods is common, and the user is stunned when suddenly stranded without battery power. Let's look at how current systems work and examine up-and-coming technologies that may change the way batteries are monitored.

Voltage-Current-Temperature Method

The Volkswagen Beetle in simpler days had minimal battery problems. The only management system was ensuring that the battery was being charged while driving. Onboard electronics for safety, convenience, comfort and pleasure have greatly added to the demands on the battery in modern cars since then. For the accessories to function reliably, the state-of-charge of the battery must be known at all times. This is especially critical with start-stop technologies, a mandated requirement on new European cars to improve fuel economy.

When the engine stops at a red light, the battery draws 25–50 amperes of current to feed the lights, ventilators, windshield wipers and other accessories. When the light changes, the battery must have enough charge to crank the engine, which requires an additional 350A. With the engine started again and accelerating to the posted speed limit, the battery begins charging after a 10-second delay.

Realizing the importance of battery monitoring, car manufacturers have added battery sensors that measure voltage, current and temperature. Packaged in a small housing that forms part of the positive clamp, the *electronic battery monitor* (EBM) provides useful information about the battery and provides an accuracy of about +/–15 percent when the battery is new. As the battery ages, the EBM begins drifting and the accuracy drops to 20-30 percent. The model used for monitoring the battery is simply not able to adjust. To solve this problem, EBM would need to know the state-of-health of the battery, and that includes the all-important capacity. No method exists today that is fully satisfactory, and some mechanics disconnect the battery management system to stop the false warning messages.

A typical start-stop vehicle goes through about 2,000 micro cycles per year. Test data obtained from automakers and the Cadex laboratories indicate that with normal usage in a start-stop configuration, the battery capacity drops to approximately 60 percent in two years. (See Figure 8-19 on page 191.) Field use reveals that the standard flooded lead acid lacks robustness, and carmakers are reverting to a modified version lead acid battery.

Automakers want to ensure that no driver gets stuck in traffic with a dead battery. To conserve energy, modern cars automatically turn off unnecessary accessories when the battery is low and the motor stays running at a stoplight. Even with this measure, state-of-charge can remain low if commuting in gridlock conditions because motor idling does not provide much charge to the battery, and with essential accessories like lights and windshield wipers on, the net effect could be a small discharge.

Battery monitoring is also important on hybrid vehicles to optimize charge levels. Intelligent charge management prevents stressful overcharge and avoids deep discharges. When the charge level is low, the internal combustion (IC) engine engages earlier than normal and is left running longer for additional charge. On a fully charged battery, the IC engine turns off and the car moves on the electrical motor in slow traffic.

Improved battery management is of special interest to the manufacturers of the electric vehicle. In terms of state-of-charge, a discerning driver expects similar accuracies in energy reserve as are possible with a fuel-powered vehicle, and current technologies do not yet allow this. Furthermore, the driver of an EV anticipates a fully charged battery will power the vehicle for the same distance as the car ages. This is not the case and the drivable distance will get shorter with each passing year. Distances will also be shorter when driving in cold temperatures because of reduced battery performance.

Magnetic Method

Under "How to Measure State-of-charge" in this chapter on page 219 we explored an improved way to measure state-of-charge by using magnetism. We now take this technology further and apply it to battery monitoring. Figure 9-25 illustrates the installation of the Q-Mag™ sensor on the side of a starter battery in close proximity to the negative plate. The technology works for lead- and lithium-based batteries.

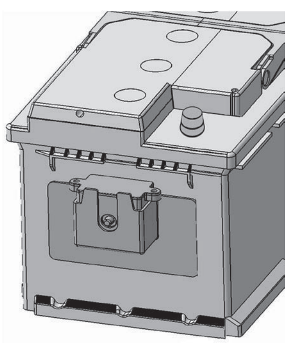

Figure 9-25:
Q-Mag™ sensor installed on the side of a starter battery

The sensor measures the SoC of a battery by magnetism. When discharging a lead acid battery, the negative plate changes from lead to lead sulfate. Lead sulfate has a different magnetic susceptibility to lead, which a magnetic sensor can measure.

Courtesy of Cadex (2009)

The potential of the Q-Mag™ technology is multifold, and this book addresses only the most basic functions. A key advantage is measuring SoC while the battery is being charged or is under load. In a charger, this allows optimal service under all conditions, including hot and cold temperature charging. Knowing the true SoC and tailoring the charge to best charge acceptance is of special interest to automotive and uninterruptible power supply (UPS) markets.

A Q-Mag-controlled charger can prolong the life of chronically undercharged lead acid batteries by applying maximum current when the opportunity arises without causing undue damage to the battery. Being relieved of voltage feedback, an intelligent charger based on Q-Mag™ can balance the state-of-charge of a fully charged battery by only replenishing the current that is lost through loading and self-discharge. Maintaining a "neutral" charge state saves energy and prolongs battery life by eliminating sulfation or overcharge.

As battery supervisor, Q-Mag™ can recognize sulfation and acid stratification on lead acid batteries. Coupled with an intelligent charger, the system can apply a corrective charge to fix the battery before the condition becomes irreversible. Furthermore, an imbalance between the terminal voltage and the Q-Mag-estimated SoC points to a battery with high self-discharge (partially shorted cell). Observing the SoC level during rest periods allows the assessment of self-discharge and the estimation of battery end of life.

The ability to measure SoC while a battery is on charge or on a load enables the estimation of battery capacity. Several proprietary techniques are possible, all of which offer a critical improvement to present systems. The voltage and impedance methods used today reveal only an anomaly when the battery is failing, and coulomb counters lose accuracy as the battery ages. One of the most critical measuring requirements of a battery test system is to know the usable capacity between 70 and 100 percent capacity.

Battery monitoring without touching the poles of the individual cells makes Q-Mag™ attractive for stationary batteries. The installation involves placing the sensors between the batteries and collecting SoC data, among other battery information, with the help of a controller on low voltage. It is conceivable that battery manufacturers in the future will include the sensors in the housing as part of production. Economical pricing at high volume and small size could make this feasible.

Q-Mag™ works across several battery chemistries, and the magnetic measuring technique may one day solve the critical need for improved battery monitoring in hybrid and electric vehicles. Research engineers at Cadex will also examine nickel-based batteries; however, the ferrous enclosure of the cylindrical cells may pose limitations. A solid aluminum enclosure on Li-phosphate does not inhibit the magnetic measurement, as the tests at Cadex are showing.

Q-Mag™ may one day also assist in the consumer market to test batteries by magnetism. Placing the battery on a test mat, similar to charging a battery, may one day be possible.

Battery Test Equipment

Conventional battery test methods measure the stored energy through a full discharge. This procedure is time-consuming and stresses the battery. There is a move towards methods that take only seconds instead of hours; however, rapid testing provides only estimated state-of-health values, and the accuracies vary according to the method used. Public safety, medical and defense organizations still depend on tests involving periodic full discharge/charge cycles.

Battery Analyzer

Battery analyzers became popular in the 1980s and 1990s to restore nickel-cadmium batteries affected by "memory," as well as to prolong battery life as part of maintenance. The Cadex C7000 Series serves the industry well and set new standards for what a battery analyzer could do. These workhorses accommodate lead-, nickel- and lithium-based batteries, and operate stand-alone or with a PC. Figure 9-26 illustrates a C7400 battery analyzer servicing a variety of batteries in configured adapters that set the analyzer to the correct setting. Each of the four independent stations allows unique service programs.

Figure 9:26:
Cadex C7400 battery analyzer

Two- and four-station analyzers service batteries from 1.2 to 15V, programmable up to 4A per station. The extended version goes to 36V and 6A charge and discharge currents. The service programs include QuickSort™ for rapid-test of Li-ion batteries.

Courtesy of Cadex

Connecting various shapes and forms of batteries has always been a challenge, and technicians have invented unique contraptions with springs and levers so complicated that only the inventor dares to touch. There is no simple way to connect batteries, especially when dealing with small packs that have tiny surface contacts.

Cadex solved the battery interface challenge with *custom adapters* for common batteries and *universal adapters* for specialty packs. The custom adapters are easiest to use; they are specially designed and the batteries go in only one way. The adapters are smart and are able to hold configuration codes for up to 10 different battery types. This allows the servicing of batteries with identical footprints but different electrical values. The user can edit the parameters with the menu function on the analyzer or with the PC.

The universal adapters consist of user-programmable *Smart Cables* that accommodate virtually any battery type. With the proliferation of cellular batteries and the need for a quick and simple battery interchange, Cadex developed the *RigidArm*™ (Figure 9-27). This adapter features spring-loaded arms that meet the battery contacts from the top down and apply correct pressure to the contacts. Lockable mechanisms allow quick and repetitive testing of same-type batteries. The retractable floor holds the battery in a vertical position, and magnetic guides keep the battery in place if laid horizontally. For added safety, a temperature sensor monitors the battery during the test.

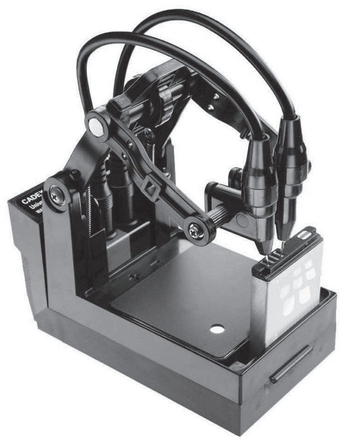

Figure 9-27:
RigidArm™ for cellular batteries

The universal adapter simplifies the interface with small batteries. The adapter holds 10 of the most commonly used mAh ratings and is compatible with Cadex battery analyzers.

Courtesy of Cadex

Servicing Cellular Batteries

Advancements made in battery test equipment make it feasible to service the over four billion cellular batteries in global use at storefronts while the customer waits. Hooking up the battery still needs some skill, and once the contacts are established the service technician may need to enter the capacity in mAh and other battery specifications.

Most cellular batteries have three or four contacts. The positive [+] terminal is normally at the outer edge and the negative [–] one is positioned towards the inside. The third contact is the thermistor measuring the battery temperature, and unless the battery adapter is specially made for the battery type, the thermistor is normally not hooked up for the test; a universal adapter often has its own temperature protection. The fourth contact, if available, may offer code identification for configuration. Figure 9-28 illustrates a typical contact positioning.

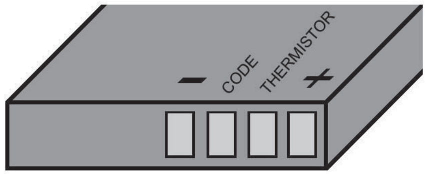

Figure 9-28:
Typical contacts on a cellular battery

The positive [+] is normally at the outer right and the negative [–] is on the inside. Most batteries have a thermistor; some also offer a code.

Returned batteries are either discarded or shipped to service centers where they are tested and redistributed as Class B packs. Looking closer at the tonnage of these returned batteries reveals that nine out of 10 packs have no problem and can be serviced. Seeing an opportunity for business, large refurbishing centers have sprung up in the USA that test 400,000 batteries per month, with volumes anticipated to increase to one million per month.

Storefront testing reduces waste, and the motto goes: "To the storefront and no further." Battery analyzers featuring rapid-test programs are offered that give a clear assessment of a battery in a few seconds while the customer waits. Figure 9-29 illustrates a service concept for storefront testing while the customer waits. If the battery needs charging or has a genuine fault, an alternate pack is given from the pool of previously tested batteries.

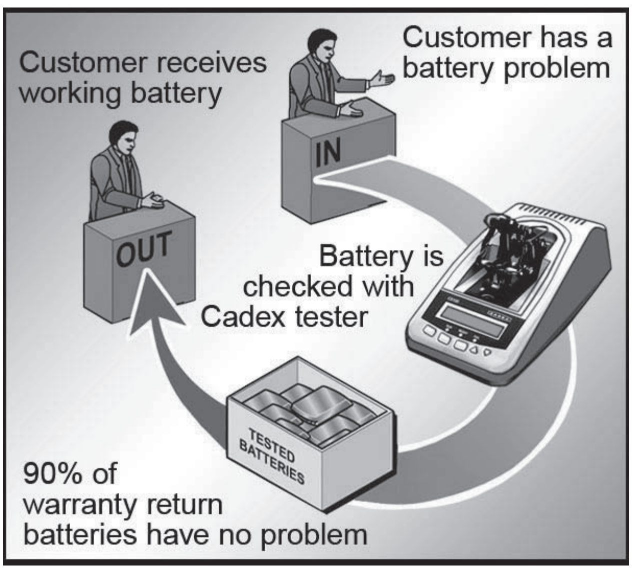

Figure 9-29:
Storefront service

Batteries are serviced while the customer waits. A faulty pack is replaced from the pool of previously serviced batteries. Storefront testing reduces handling, lessens disposal and improves customer satisfaction.

Courtesy of Cadex

One of the difficulties of storefront testing has been the availability of suitable battery diagnostic equipment. The older units lacked accuracy in rapid testing, and had a predictive capacity that resembled a ticket in a Las Vegas lottery; many potential users hesitated to buy such equipment. QuickSort™ provides 90 percent accuracy across the population of cellular batteries. (See "Testing Lithium-based Batteries," on page 229.) With a PC, some analyzers allow service reports to be printed, and the Internet enables a central manager to monitor the activity of each store. Figure 9-30 illustrates a battery analyzer designed for storefront use.

Figure 9-30: Cadex C5100 analyzer for lithium-ion batteries

This analyzer rapid tests, charges and cycles batteries. The RigidArm™ adapter allows easy interface to cellular batteries; also accepts preprogrammed adapters. QuickSort™ tests batteries in 30 seconds.

Courtesy of Cadex

Maintaining Fleet Batteries

A battery analyzer assures that fleet batteries meet the minimum performance standards. The device also helps to restore low performers, if such a service is possible with the battery types in question. In addition, a battery analyzer supervises the all-important function of a timely replacement at the end of a productive life. Manufacturers of portable equipment support battery maintenance because well-performing batteries reflect positively on the equipment, a win-win situation for both manufacturer and user.

Many battery analyzers come with PC application software. With BatteryShop™ (by Cadex), for example, the PC becomes the command center and all functions are processed through the keyboard, as well as other input devices. Clicking the mouse on any of the 2,000 batteries listed in the database configures the analyzer to the correct setting, eliminating the need for further programming. The user has the liberty to add, remove and edit the batteries listed should the specification change.

Labeling each fleet battery with a permanent ID number simplifies logistics and traceability. A printer connected to PC BatteryShop™ generates these labels in bar code format. The user simply scans the label, which in turn configures the analyzer and retrieves the performance history for review. Besides capacity readings and service dates, purchasing date, vendor information and pricing can also be entered. Figure 9-31 illustrates the battery scan, service and data examination.

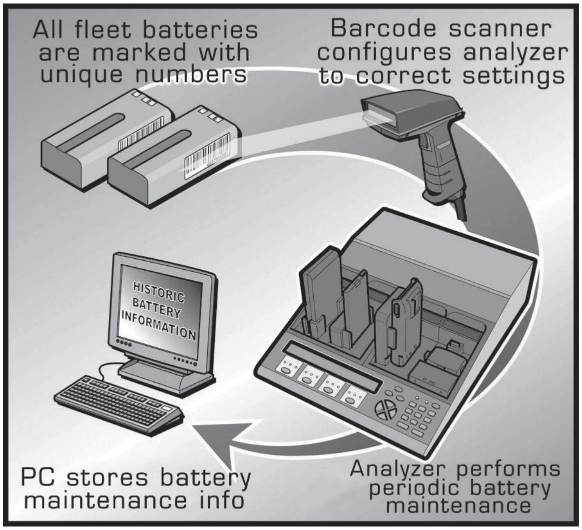

Figure 9-31: Fleet battery management

Labeling each battery with a unique number simplifies battery service. Swiping the barcode label reveals the history of the battery.

Courtesy of Cadex

Another tracking method for fleet batteries is attaching a removable label that shows the battery information at a glance between services, as illustrated in Figure 9-32. The system is self-governing in that all batteries must regularly be serviced as part of quality control. This is made possible by providing a time period between the last service and the new date due. With this information on hand, the prudent battery user only picks a battery that has been serviced and meets this quality assurance (QA) test protocol. Setting up the maintenance system is simple and managing it requires only about 30 minutes per day.

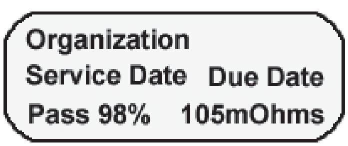

Figure 9-32: Sample of removable battery label

The label shows battery information at a glance and includes name of organization for traceability, capacity in percent, as well as past and future service dates.

Setting up a battery maintenance system requires a battery analyzer that is capable of printing battery stick-on labels. The analyzer should also offer a program that automatically restores nickel-based batteries if the set capacity threshold cannot be met. Cadex analyzers meet these requirements and go one step further by offering adjustable capacity target settings to select the minimum performance criteria for the given operation.

Most fleet operations use 80 percent as their battery pass/fail criterion. Increasing the threshold to 85 percent tightens the performance tolerance but passes fewer batteries; lowering the settings extends service life but offers less stringent performance standards. When choosing the setting, the organization must ensure that the lowest-level battery in the fleet is able to fulfill its assigned duty. Figures 9-33, 9-34 and 9-35 illustrate the battery label system.

Rechargeable batteries do not die suddenly but gradually get weaker with time. A service every one to three months offers plenty of confidence that all batteries will meet the minimum required capacity and last through the shift with some energy to spare.

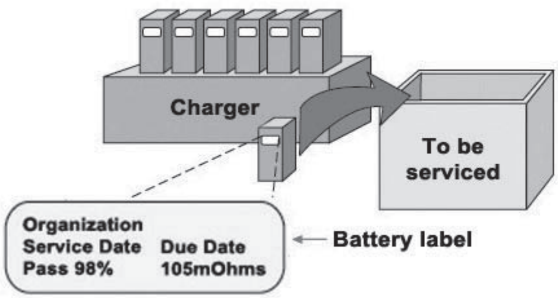

Figure 9-33:
Sorting batteries for servicing

When taking a battery from the charger, the user checks the service date, and if expired the battery is placed in the "To be serviced" box.

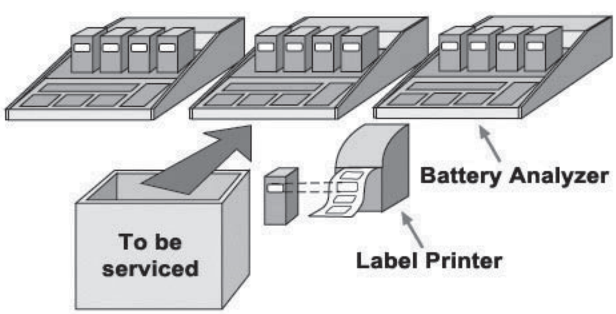

Figure 9-34:
Servicing expired batteries

The analyzers service the batteries and recondition them if low in capacity (only nickel-based batteries receive recondition). Passing batteries are relabeled showing capacity and the next service date.

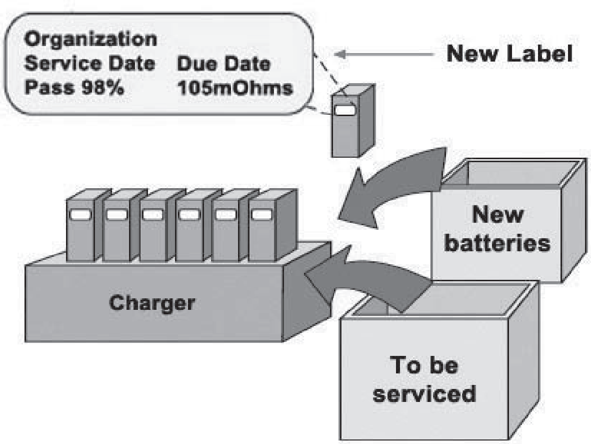

Figure 9-35:
Reinstating batteries

The failed batteries are removed from service and replaced with new packs. The new and serviced batteries go back into service by being charged.

All figures Courtesy of Cadex

Battery Test Systems

While battery analyzers are tools to service batteries; battery test systems provide multi-purpose test functions for research laboratories. Typical applications are life cycle testing and verifying cell balance in field imitation. Such tests can often be automated with a custom program. *Load capture* allows storing load signatures for playback simulations. Many battery test systems also control external load units and environmental chambers. Other uses of such systems are quality inspections and verifying warranty claims. Figure 9-36 illustrates a typical battery test system.

Figure 9-36:
Cadex C8000 Battery Test System

Four independent channels provide up to 10A each and 36V. Maximum charge power is 400W, discharge is 320W. The discharge power can be enhanced with external load banks.

Courtesy of Cadex

The alternate to a battery test system is a programmable power supply controlled by a computer. Such a platform offers flexibility but requires careful programming to prevent stress to the battery and possible damage or fire if an anomaly were to occur. A battery test system, such as the Cadex C8000, offers protected charge and discharge programs that identify a faulty battery and terminate a service safely. The system can be overridden to do destructive tests.

Simple Guidelines to Choosing a Battery Test System

- Similar to a medical test or the weather forecast, battery testers provide only estimations. No single instrument can do it all; several methods are needed to attain a full assessment.
- Most batteries keep a low internal resistance while the capacity drops gradually with age.
- Battery resistance provides only a snapshot and cannot provide the end-of-life prediction.
- Capacity is the leading health indicator but this measurement is difficult to estimate.
- A charge or discharge agitates the voltage and the battery needs several hours of normalize.
- Coulomb counting requires periodic calibration to keep accuracy.
- Battery management prevents surprise failure and allows for a scheduled retirement.
- Storefront battery testing provides on-site troubleshooting to verify performance.