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

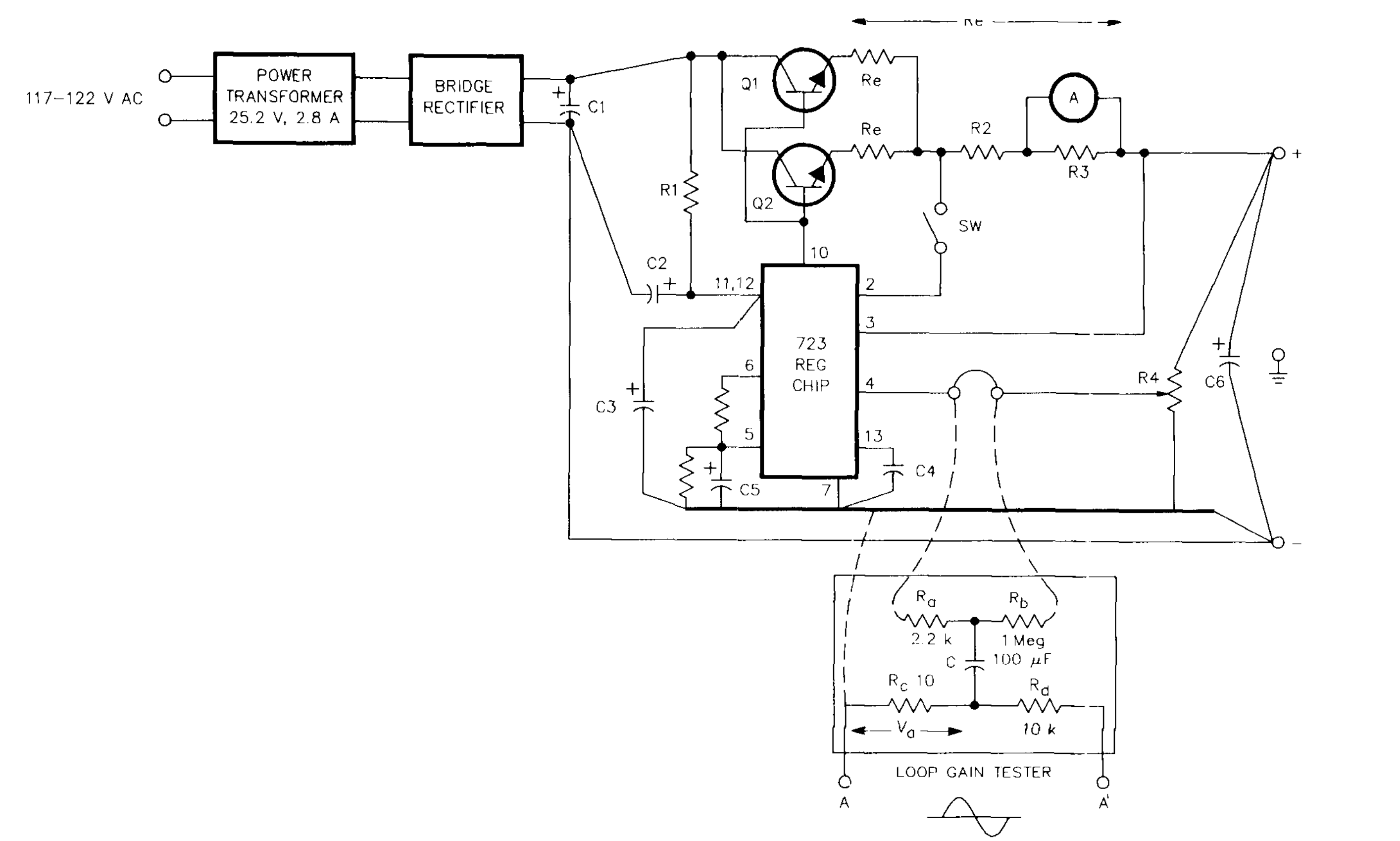


Fig 1—Simplified diagram used to discuss design principles.

(usually not to full specification) but an embedded element is not stable.

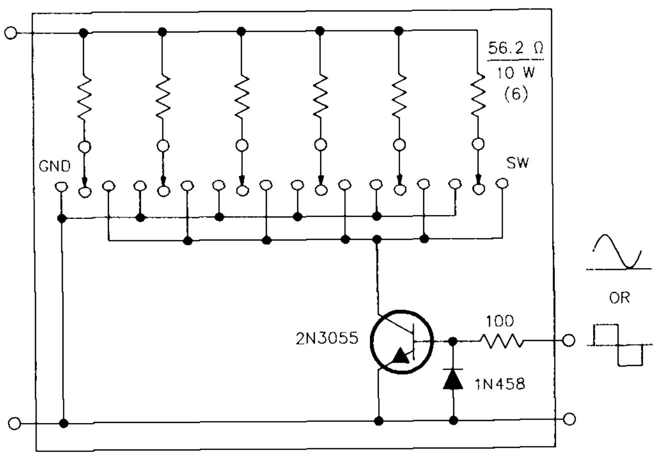
Open-Loop Frequency Response

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.

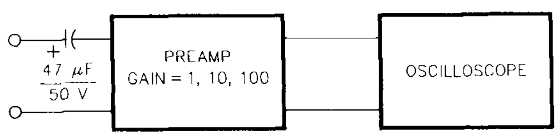
Fig 2 is the open-loop frequency response to the top of R4 when RE is set for the 2.5-A current range. At very low frequencies, the drop-off is due to the gradual closure of the feedback loop, as mentioned above. At higher frequencies the roll off is due to the combined effects of C4 and C6 and occurs 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.



LOAD TEST CIRCUIT



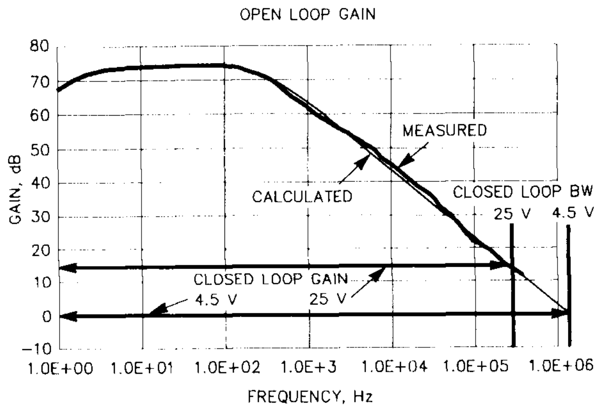


Fig 2—Regulator loop frequency response.

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.

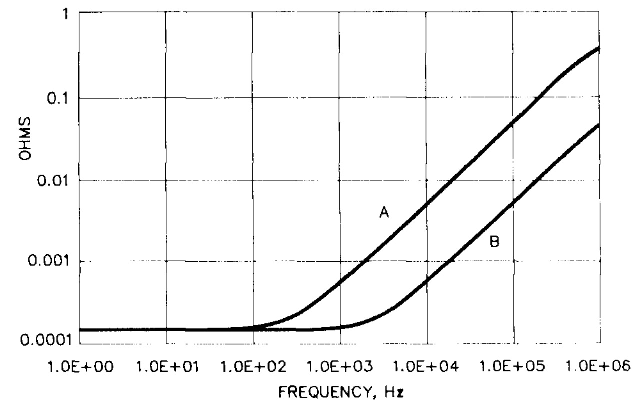


Fig 3—Output impedance magnitude.

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.

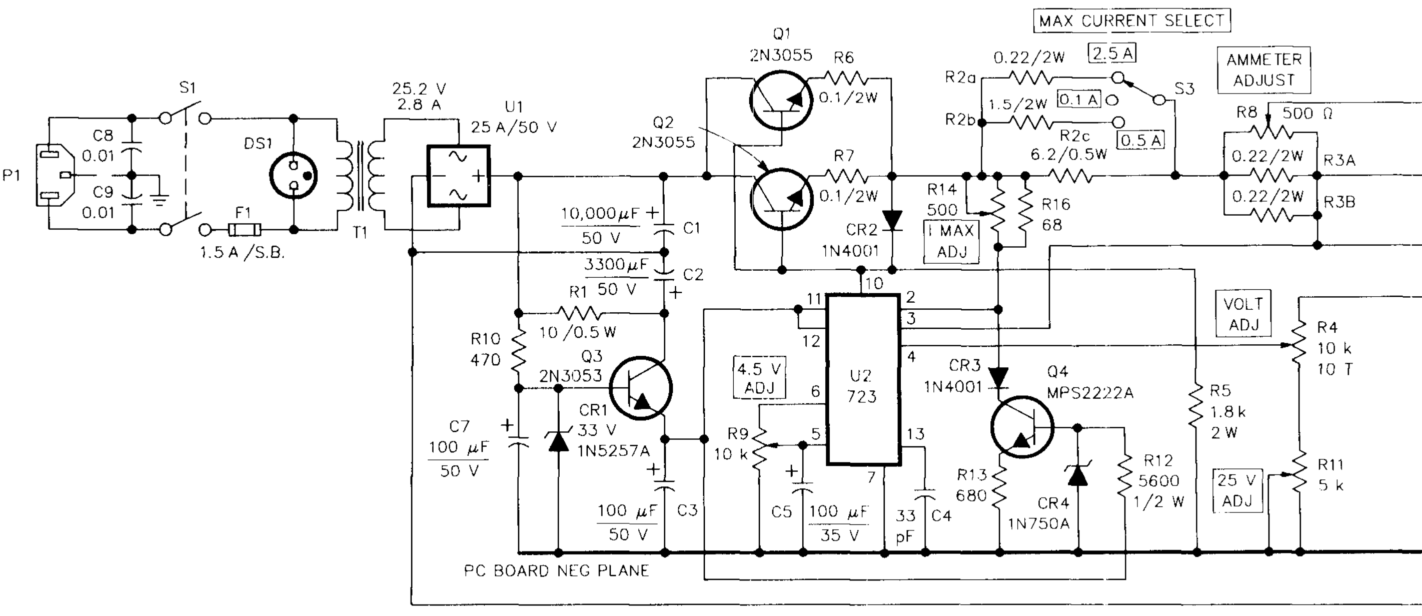


Fig 4—Complete schematic of regulated power supply.

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.

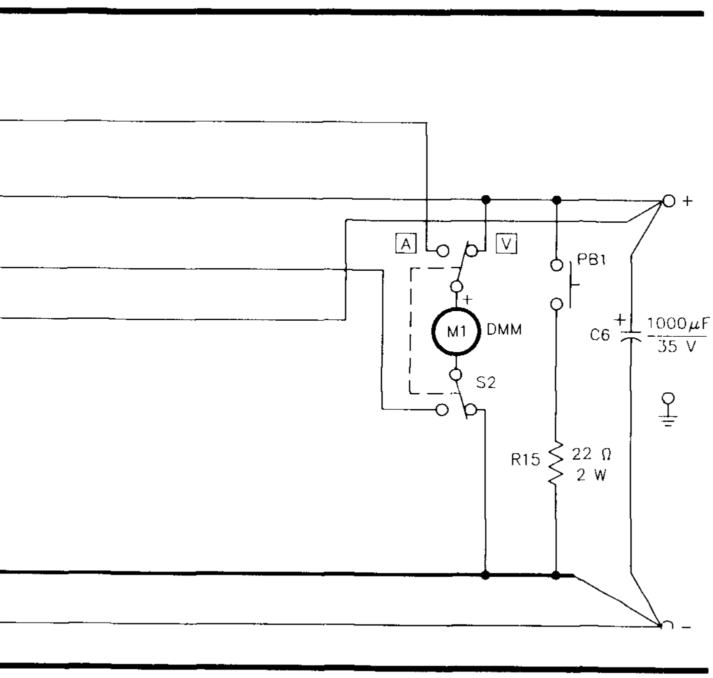
D) The bottom of C1 connects directly to the negative binding post with a heavy lead so that load current fluctuations prefer not to flow on the regulator PC board ground plane and thus influence the operation of the regulator in unpredictable ways.

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,

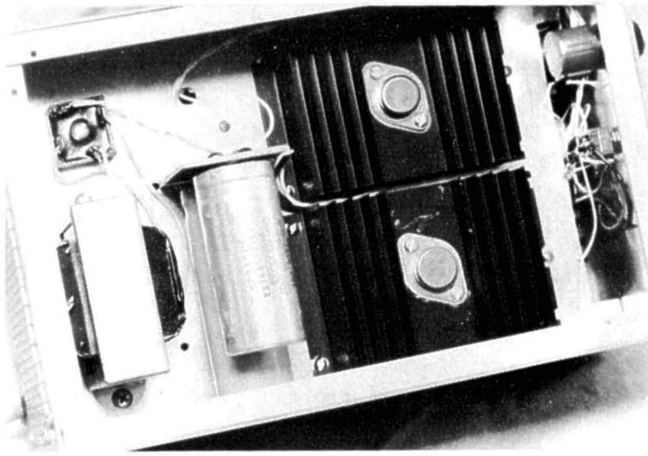


Fig 5—(a) Cabinet (b) top view showing two Wakefield 403A heat sinks.

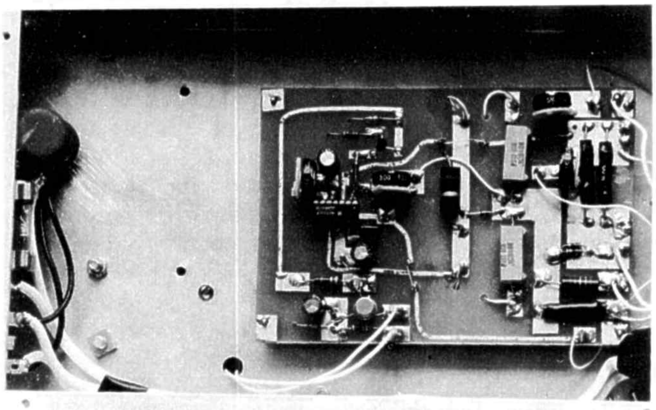


Fig 6—Underneath view.

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.

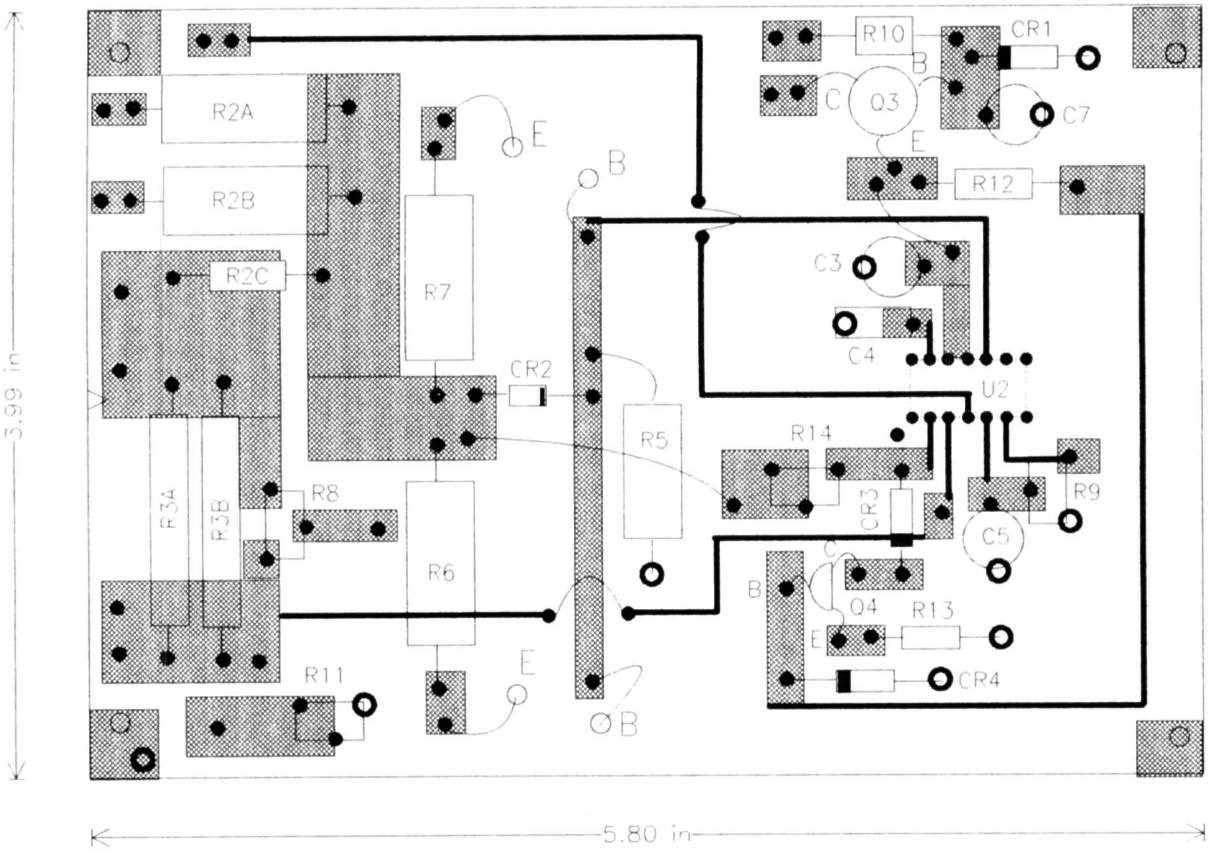


Fig 7—Parts placement.

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