Parker R. Cope W2GOM/7
8040 E. Tranquil Blvd.
Prescott Valley AZ 86314
[pamaco@mwaz.com]

Switched Mode Power Supplies

Get out your calculator and have some fun.

Switched mode power supplies (SMPS) are relative newcomers to the electronics world. In days gone by, motor-generators, dynamotors, or vibrator supplies were used to convert a battery voltage to some other voltage. Today SMPSes do that job. They are small, lightweight, and relatively inexpensive, and within the construction capability of the home brewer.

The power density, watts-per-cubic-inch, of newer commercial SMPSes is truly amazing. While some newer SMPSes have densities of 20 watts-per-cubic-inch, most home builders will accept a much lower power density, a larger unit, if they can build it themselves. The home-built supplies may not be the most compact, but they are still a far cry from being boat anchors.

Most conventional power supplies operate from the 60 Hz mains, and have ripple frequencies of 120 Hz. The 60 Hz transformers and chokes are big and heavy, but they are devils we know and have learned to live with. But when we need to operate from a battery, we are pretty much up the creek unless we can work with the battery voltage as it is. Of course, we can still generate 60 Hz with a motor generator and use the 60 Hz supply, but that is like standing up to paddle a canoe — there are better ways.

Instead of generating 60 Hz, why not 60 kHz or 600 kHz? The transformers and inductors will be much smaller and the filtering much easier. Arguing against a high frequency is component availability. 60 Hz magnetics, transformers, and chokes, are commercially available, but high frequency transformers are not. High frequency transformers are small and easy to wind: A few turns on a ferrite core for an SMPS looks pretty good.

As an example, obtaining 28 volts from a 12 volt source can be achieved with a DC/DC converter: A high frequency oscillator followed by an amplifier. Then rectified and filtered. This solution is certainly simple, but it has no regulation and must be manually adjusted. SMPSes are regulated and often have a lot of other bells and whistles not essential to generating a stable voltage such as over or under voltage protection, current limiting, or soft start. Once you know what the necessary parts are and what they do you can add these extra functions.

The output of a bare bones unregulated supply changes as the main battery

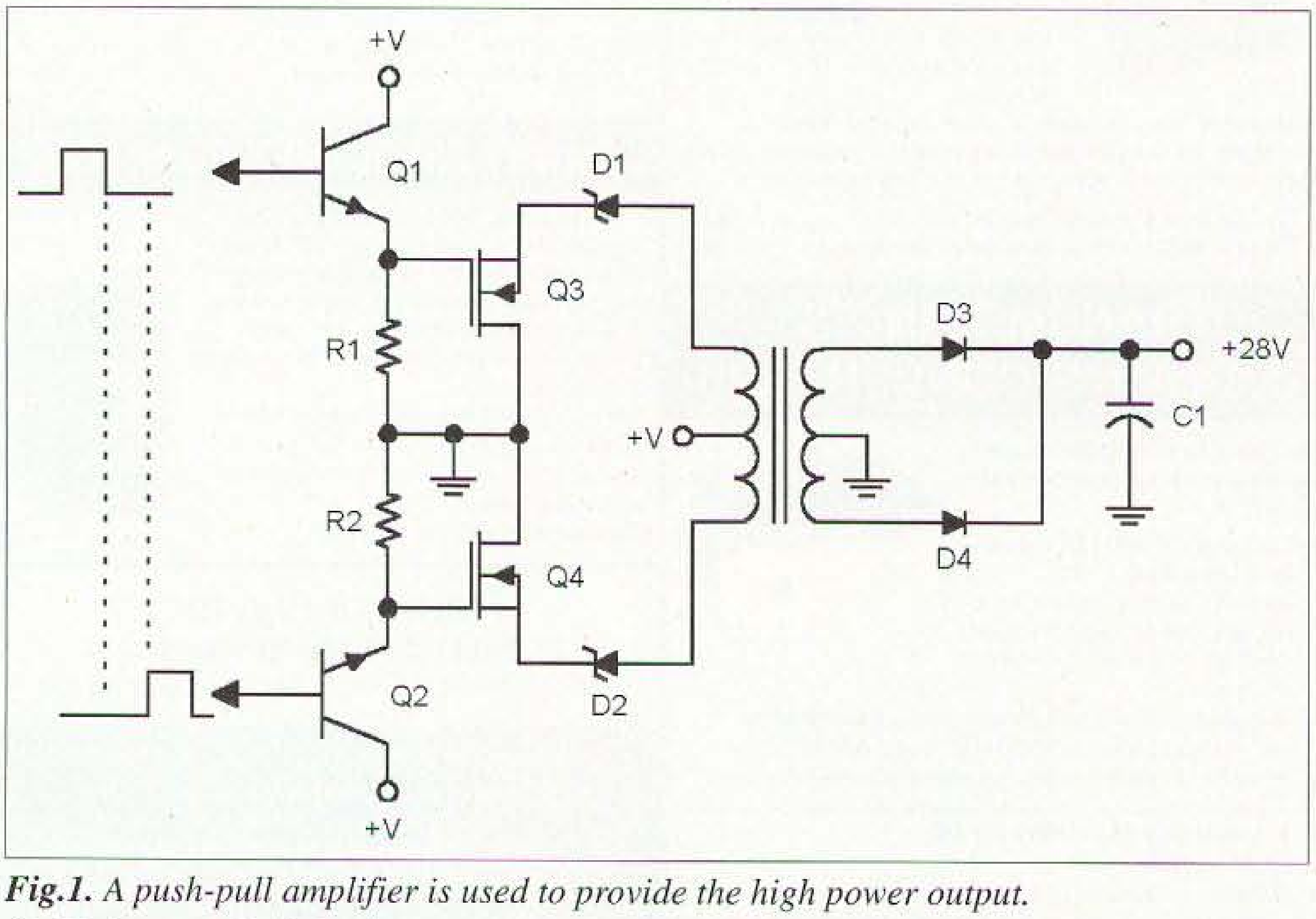

Fig.1. A push-pull amplifier is used to provide the high power output.

voltage changes or as the load changes. An unregulated supply is certainly simple, but adding regulation need not be a deal breaker. It can be as simple as following the unregulated supply with a zenor diode or a three terminal regulator. This is a rather inefficient approach. Another more complex but efficient method is to control the drive to the power oscillator. An even more complex scheme is to sense the DC output and automatically control the drive to the amplifier to maintain the desired DC output voltage. This is an SMPS.

Controlling the output power of an SMPS isn't like controlling the output of a linear amplifier. The output power is controlled by changing the duty cycle of the drive. A push-pull amplifier is shown in **Fig. 1**. When each transistor is on for full alternate half cycles the output will be maximum, but when each transistor is on for only a part of the half cycle, the output will be less. Changing the duty cycle of the drive is the smart way to change the output voltage.

In **Fig. 2**, two sections of U1 a CD4001, U1A and U1B, act as a relaxation oscillator. The leading edges of the square wave are differentiated with C2 and R3, and C3 and R4 and logic NOR'd in U1C, then inverted with U1D to produce the trigger for the monostable multivibrator U2A and U2B, another CD4001. The period of U1A and B is approximately 1/1.39R2C1. R1 just stabilizes the frequency with changes in supply voltage. R1 can have any value but something in the range of three to ten times R2 is typical. Large values of R1 can limit the maximum frequency, so at 200 kHz a value of about three times R2 is safe.

The output of the monostable, U2A and U2B, is approximately 1/0.7C4R5. When R5 is made variable the pulse width can be varied. The negative pulse from U2A is logic OR'd in U2C and D to control the duty cycle of the drive to the amplifier. R5 controls the DC output of the converter.

The outputs of U1A and B and the variable pulse from U22A are logic NOR'd in U2C and U2D to produce a variable width positive pulse on alternate half cycles to drive the amplifier.

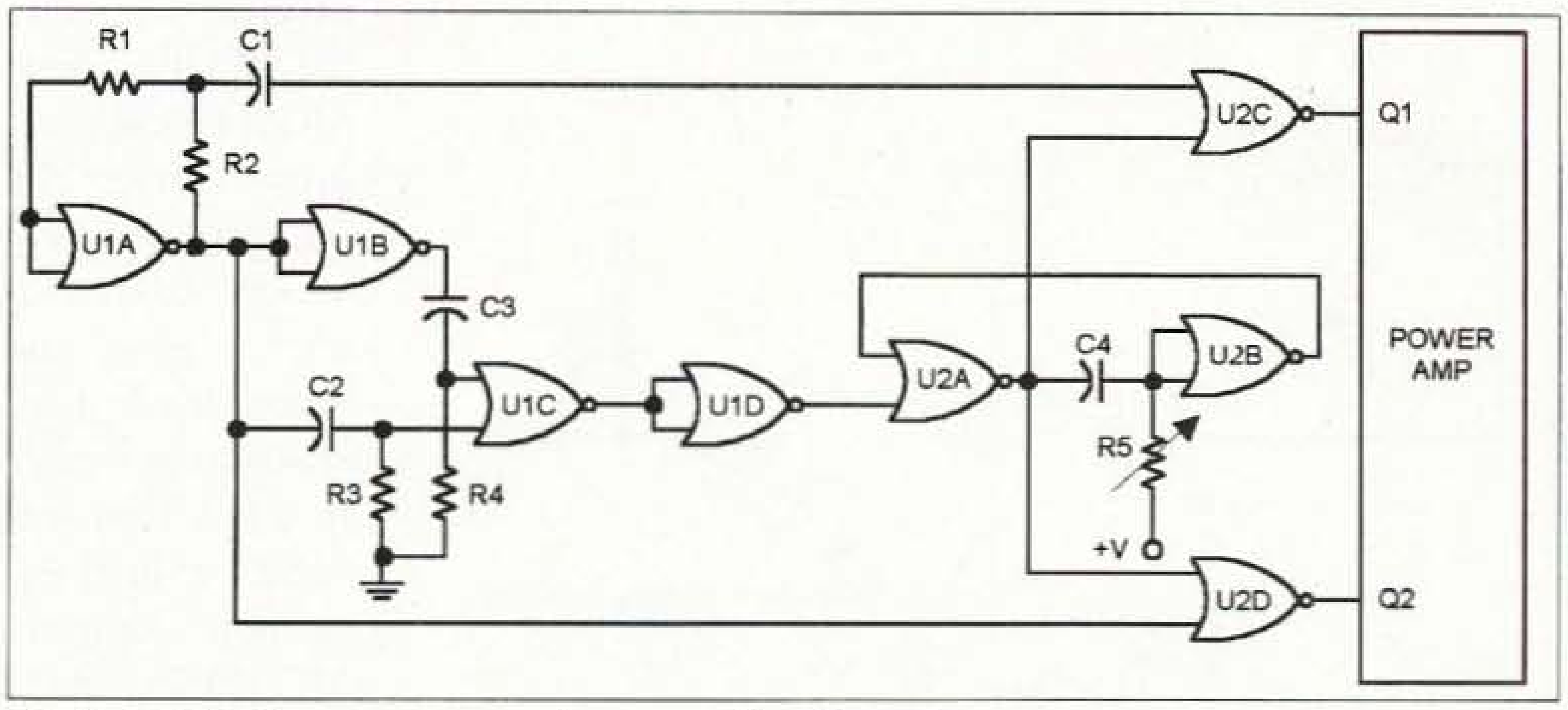

Fig.2. A DC/DC converter can be manually adjusted.

The MOSFETs Q3 and Q4 in **Fig. 1** conduct when their gates are positive: Q3 conducts on one half cycle and Q4 conducts on the next half cycle.

Operating with higher frequencies requires consideration of parasitics that could be ignored at 60 Hz. What were trivial parasitics at 60 Hz become serious problems at 60 kHz and at 600 kHz everything is critical: Leads become inductors, capacitors and inductors are resonant circuits, the equivalent series resistance and inductance limit the effectiveness of capacitors. Reverse recovery time of diodes, and storage time of bipolar transistors add to the design problems.

Bipolar transistor problems can be avoided by using MOSFETs. Since MOSFETs are majority carrier devices, they do not suffer from recovery time, and when driven hard enough can switch in a few nanoseconds. For high frequency supplies MOSFETs are the power devices of choice.

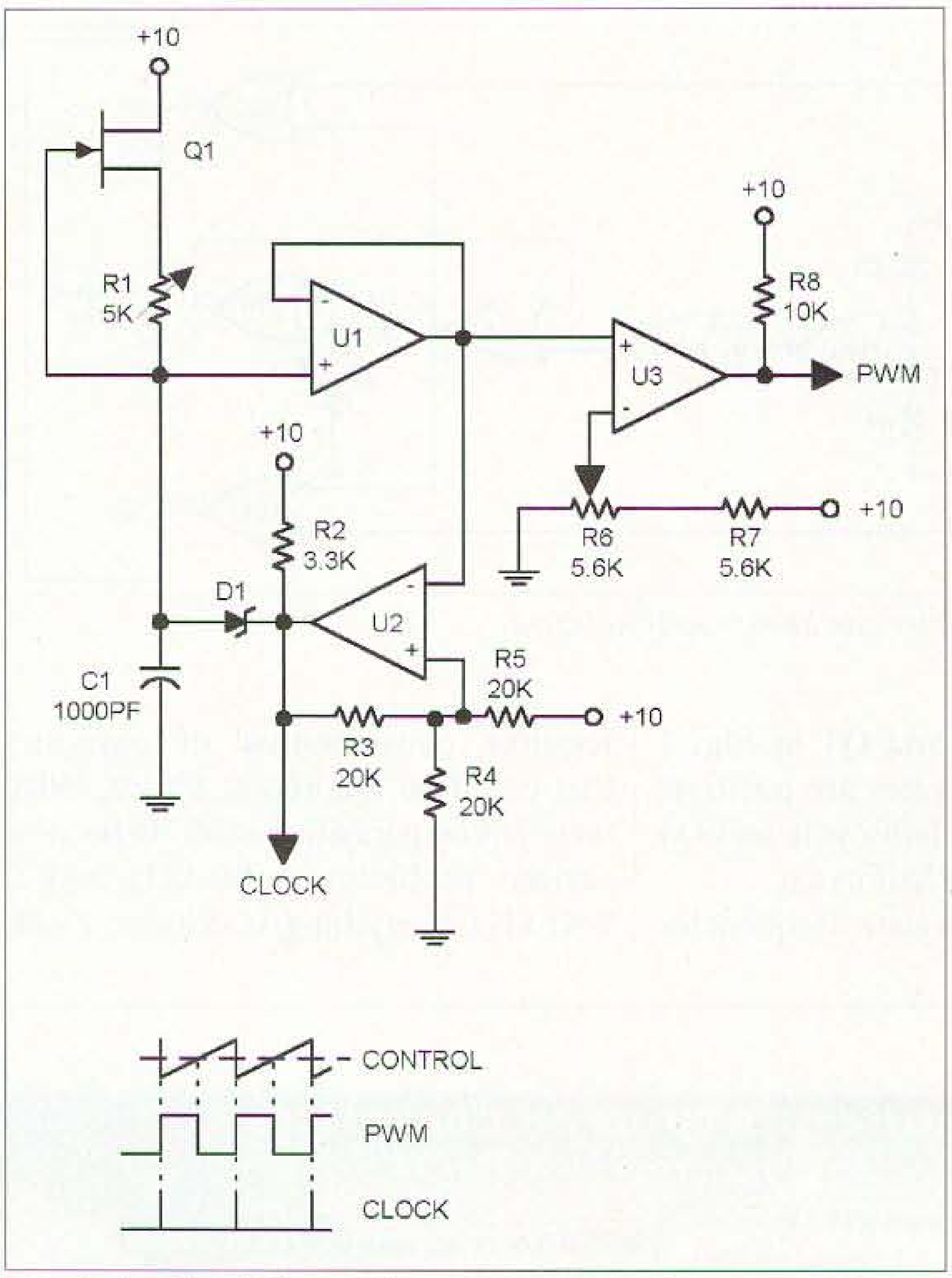

Fig. 3. A pulsewidth modulator is an essential part of an SMPS.

Diode reverse recovery time is a different problem. During the reverse recovery time a diode conducts equally well in the reverse direction, not a good situation.

Reverse recovery times of diodes is short enough to be ignored at 60 Hz, but at higher frequencies reverse recovery time can be a significant fraction of the period of the operating frequency. Ignore reverse recovery time at your peril.

Ordinary silicon rectifiers begin to depart from the ideal at a kilohertz or so. For example, the 1N400X series of silicon rectifiers has recovery times of about 300 nsec. That is, the diode continues to conduct in the reverse direction for 0.3 μsec after the anode becomes negative. When the frequency is 20 kHz the polarity changes every 25 μsec and 0.3 μsec is not negligible. The di- ode essentially conducts in both directions for 10% of the cycle. During the reverse recovery time the filter discharges through the diode(s) and the source. The diode's dissipation is increased and the load on the source is also much higher than expected. The ripple increases as well. All in all a bad situation.

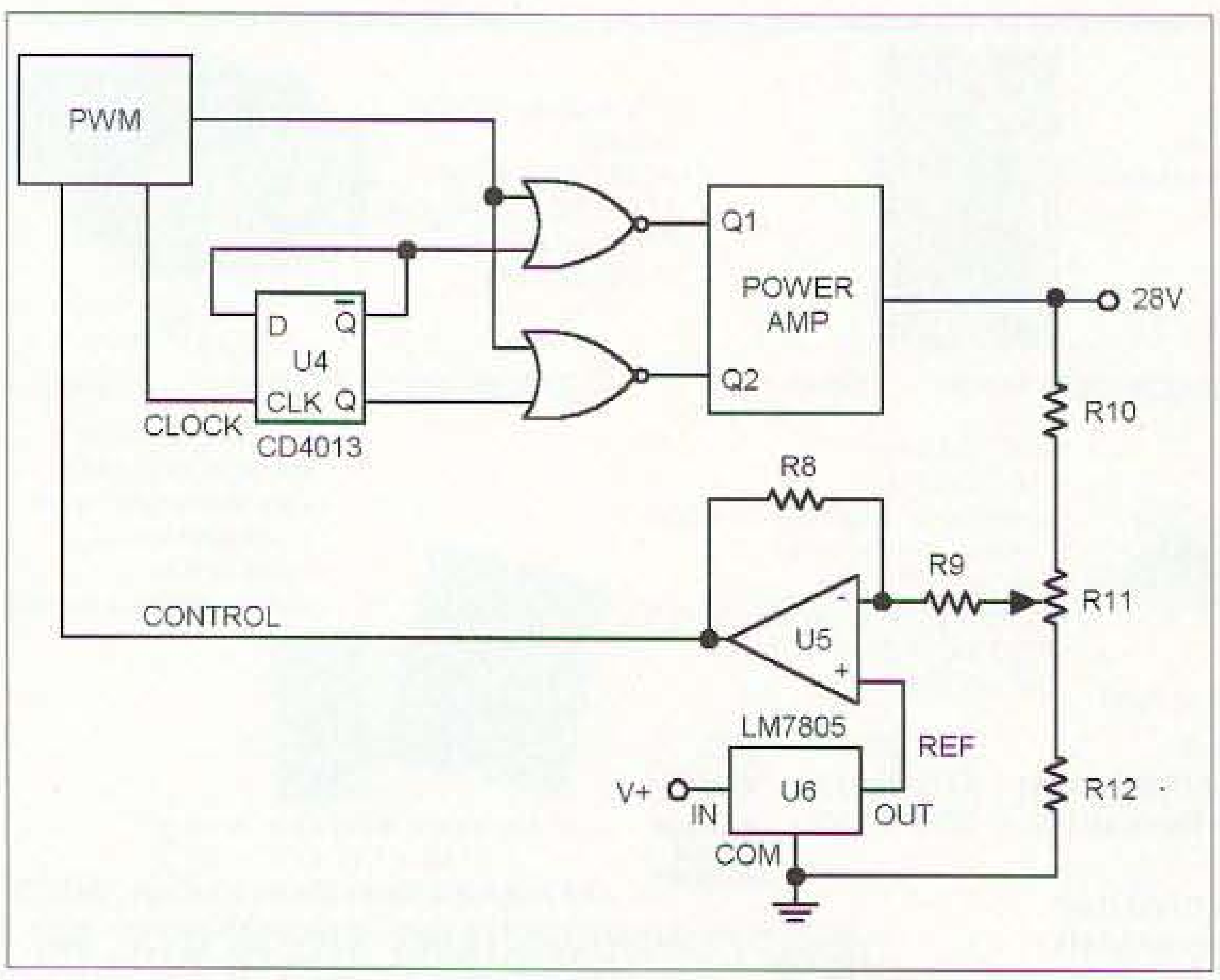

Fig. 4. A sample of the DC output can be fed back to control the output.

Schottky diodes and ultra fast diodes have much shorter reverse recovery times and minimize reverse recovery problems. Schottky diodes are an excellent choice in that they have reverse recovery times of a few picoseconds and forward voltage drops of about 0.3 volts. Unfortunately Schottky diodes have a PIV (peak inverse voltage) of only about 30 or 40 volts. Ultra fast diodes have recovery times of 25 nsec or so and PIVs up to 1 kV. Using ultra fast diodes like the Motorola MUR405 or 410 provide a nice safety factor for operating frequencies of several hundred kilohertz.

The power amplifier and rectifiers are shown in **Fig. 1**. The Schottky diodes D1 and D2 block the negative swing of voltage at the drains of the MOSFETs. (MOSFETs have a parasitic diode that conducts when the drain is negative.)

While MOSFETs are voltage controlled devices, it takes time to charge their input capacitance and raise the gate voltage. The capacity at the input is not just the gate-source capacity plus the gate to drain capacity; the old bug-a-boo Miller effect gets into the act. The input capacity of an amplifier with a resistive load is:

$C_{in} = C_{gs} + C_{gd} \times (1+A)$

C_{gs} is the gate-source capacity, C_{gd} is the gate-drain capacity, and A is the voltage gain. To complicate the matter, C_{gd} is not constant like in a vacuum tube but changes with gate to drain voltage.

MOSFET manufacturers give the input capacity in terms of the total charge needed to be charged to switch the rated drain current when the supply is the rated drain voltage. The total gate capacitance is given as Q_T. The charge of the input capacitance $Q_T = i \times t$, where i is the charging current and t is the time the charging current flows. To switch a MOSFET with a Q_T of 15 nano Coulombs in 150 nsec would require a peak gate current of 100 mA.

The gate current is a spike of current that decays exponentially to zero as the input capacitance is charged. But the peak current is still 100 mA. The output current of the CMOS gates is only a couple of mils so that more current is required to switch quickly. Emitter followers Q1 and Q2 provide the higher current. Since the emitter followers are not saturated their rise and fall times are fast. Of course, MOSFET driver ICs could be used, but 2N3904s can provide 100 mA and they are cheaper and more readily available.

The transformer is a crucial item in the design of the supply. The turns ratio of the transformer dictates the step-up or down of the main battery supply voltage V+. Designing the proper transformer is not a walk in the park, but it's not a deal breaker.

Selection of the wire size depends on the currents involved, and the number of turns depends on the required inductance and the step-up desired. The ampere turns dictate the core size. The turns ratio sets the step up or down of the voltage.

If the main DC source is a 12 V car battery, the voltage can vary from 9 V to 13.8 V; 9 volts is essentially a dead battery and 13.8 volts is a fully charged battery.

If it is desired to build a 28 V 1 A supply powered from a car battery, a 4:1 ratio transformer would step up 9 volts to about 36 volts peak and step up 13.8 volts to about 55 volts peak if the transformer were 100% efficient. With reasonable transformer efficiency and losses in the amplifier and rectifiers, the output should be 30 V minimum and at least 50 V maximum. A toroid will probably be easier to work with than a pot core, but compromises will have to be made. The turns ratios must be whole numbers.

Selecting a core that has the permeability and saturation characteristics is the key consideration. The impedance of the primary with the secondary open should be as large as practical, certainly greater than three times the loaded impedance. The secondary impedance reflected into the primary is the secondary impedance divided by N^2, the turns ratio squared, 1.75 ohms. The magnetizing impedance should be greater than 5.25 ohms, so that at 100 kHz, a primary inductance of at least 8.4 μH is required.

A toroidal transformer wound on a core of Ferronics B material is quite suitable for frequencies up to 500 kHz. As a first cut the 11-282 core is large enough to comfortably accommodate the necessary windings, but as will be shown a single core will saturate. Two cores stacked can increase the effective area and ease that concern.

The magnetic dimensions of two stacked 11-282 cores are: $A_e = 1.044\ \text{cm}^2$ (each core has an $A_e = 0.522\ \text{cm}^2$, $l_e = 5.42$ cm. L computes to be 12 μH for one turn. A usable flux density is less than 3,500 Gauss in each core. For two stacked cores the flux density is 7,000 Gauss.

The inductance of a toroid is given as:

$L = N^2 0.4\pi\mu A_e/l_e \times 10^{-8}$ Henries

L also can be given as the inductive index A_L.

$\mu = B(\text{Gauss})/H(\text{Oersteds})$

$H = 0.4\pi NI/l_e$ Oersteds

where:

A_e is effective area in cm^2.

l_e is effective length in cm.

A_L is the inductance index, the inductance of one turn.

$L = A_L N^2$.

A_L is given as 6,057 nH for the 11-282 core.

The permeability μ and maximum flux density B for B material are given in the data sheets as μ = 5000 and B = 3,500. When two cores are stacked, A_e doubles and doubles the inductive index. Therefore a turn has an inductance $L = A_L N^2 = 12\mu H$ or an impedance of 7.5 ohms at 100 kHz. Remember, each pass of the wire through the hole is one turn.

The transformer has 2 turns centertapped for the primary and 8 turns centertapped for the secondary, or a turns ratio of four to one. The secondary impedance reflected into each half of the primary is $1/N^2$ or 1.75 ohms. The magnetizing impedance of each half of the primary is $2\pi fL_p$ or 7.6 ohms. The total current in the primary is the magnetizing current plus the reflected secondary current. The maximum

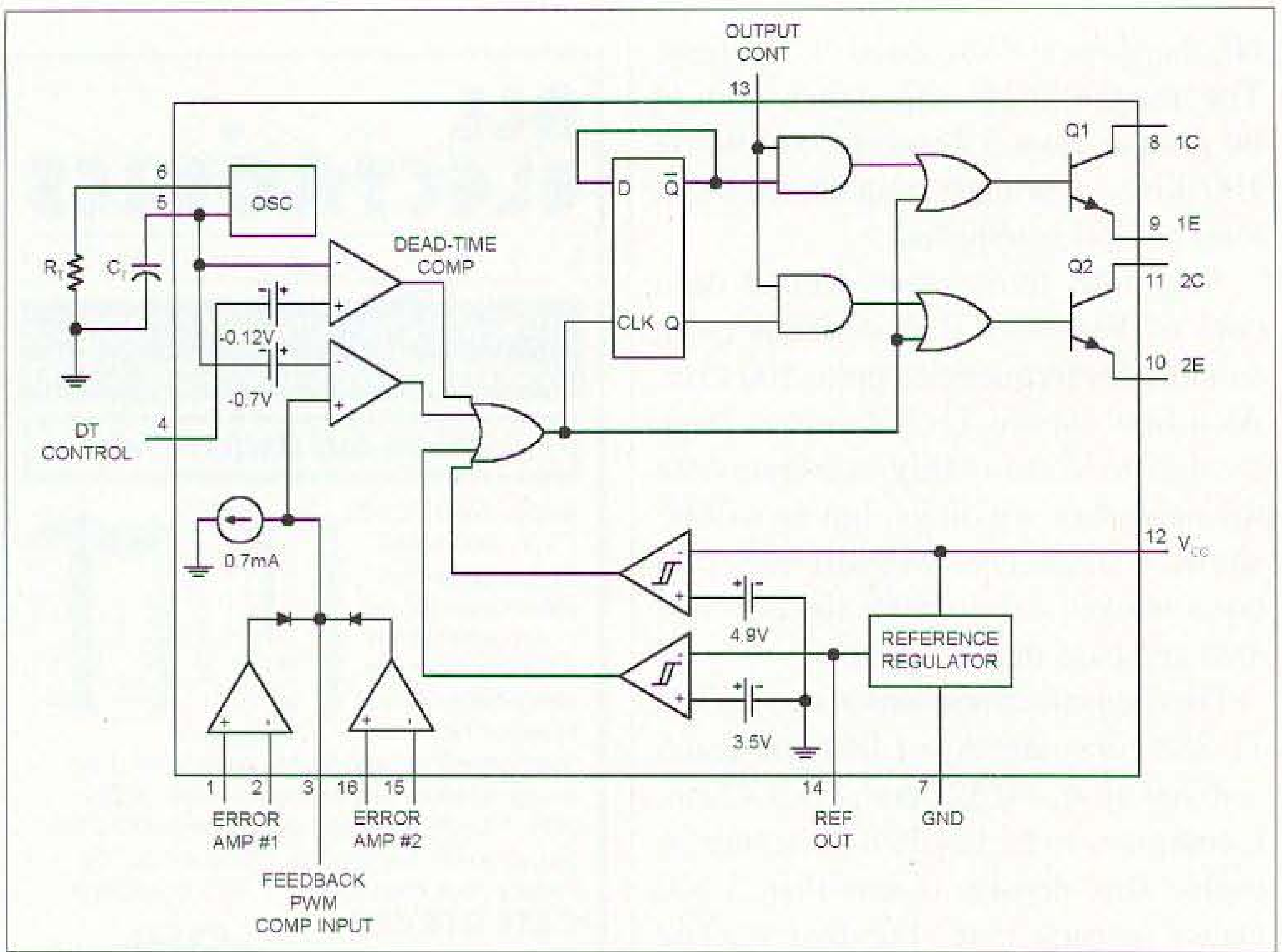

Fig. 5. The TL494 is a switch mode pulsewidth modulation control IC.

magnetizing current exists when the main battery voltage is maximum, $13.8V/7.6\Omega = 1.8A$.

The coercive force in the core is $H = 0.4\pi NI/l_e$ Oersteds. When the total current I is 4A + 1.8A = 5.8A, H = 1.3 Oersteds. The maximum total flux B is μH or B = 1.3 x 5,000 = 6,500 Gauss. With two stacked cores the flux in each core is 3,250 Gauss.

The transformer can be wound on either a ferrite potcore or a toroid core. In most cases a toroid is easier to come by and gaping the core is avoided. In either case, a good rule of thumb is to make the primary's magnetizing current, the current with the secondary open, as low as practical. That is make the inductance as high as practical. A two turn primary would have an inductance of 48 ohms but the field strength would be 2.6, which is too high.

Selecting a core is primarily a matter of finding a core that can support the needed ampere-turns without saturating the core. That is, keeping the magnetic flux (Gauss) below the saturation point of the core. The core material determines the field strength (Oersteds) that can be supported.

The wire size used depends on the RMS current the wire will be carrying. For a single layer winding the wire size should be greater than 200 circular mils per amp. In a multi-layer winding transformer the wire size should be greater than 750 circular mils per amp. As an aside, MIL specs limit the wire size in a harness to at least 500 circular mils per amp. Therefore #18 AWG can be used for the primary and #22 AWG for the secondary. The 11-282 core has enough ID to easily allow #18 AWG to be used for both the primary and secondary. The temperature rise of #18 should be well within the temperature limits of any enameled wire, but Formvar® insulation is recommended and is ideal for any homebrewer's applications. Formvar may be a bit difficult to strip, but its toughness makes it a good all-purpose insulation.

A pulsewidth modulator (PWM) shown in **Fig. 3** can also generate a variable pulse width to control the drive to the power amplifier. The PWM is a basic functional block of all switched mode power supplies.

The variable pulsewidth in the output of U3 of **Fig. 3** is obtained by comparing a control voltage from R6 to the sawtooth output of op amp U1 in the comparator U3. Decreasing the voltage narrows the output pulse.

The sawtooth is generated by a constant current linearly charging the capacitor C1. The voltage on C1 is isolated from the following circuits by the buffer amplifier U1. U2 is a comparator that controls the amplitude of the sawtooth. U3 compares the sawtooth voltage with a pulsewidth control voltage.

The sawtooth amplitude is determined by the hysteresis around U2. When the voltage on C1 (the output of U1) exceeds the voltage on the noninverting input of the comparator U2, the output of U2 goes to near zero and C1 is discharged through D1 toward 0.3 V, the forward drop of D1. When the voltage on C1 falls below the voltage on the inverting input, about 0.5 V, the output goes high allowing C1 to recharge and repeat the cycle. With a 10 volt supply and the values shown in **Fig. 3**, the sawtooth swings from about 0.5 V to 4.9 V.

The positive going output of U2 also provides the clock for U4 of **Fig. 4**.

The frequency of the sawtooth is determined by how rapidly the voltage on C1 rises from 0.5 V to 4.9 V. With a constant charging current, the rate of change of voltage on C1 is dV/dt = I/C. A charging current of 0.88 mA will change the voltage on 1,000 pF from 0.5 V to 4.9 V, a change of 4.4, in 5 microseconds, and produce a 200 kHz sawtooth.

The constant current charging C1 is obtained with Q1, a 2N5457 N-channel depletion mode JFET (a MPF102 is a fair substitute for the 2N5457). R1 controls the gate source voltage, and consequently the drain current. For the 2N5457, drain current is independent of drain voltage when drain-source voltage is a couple of volts and drain current is less than 1 mA. The gate source voltage is about 1.4 volts when drain current is 0.88 mA. Under these conditions R1 would be 1.6k.

Unfortunately, the drain current vs. gate-source voltage is not a tightly controlled parameter so a trimmer R1 is needed to set the current and frequency to a particular value. A variable R1 of 5k will probably suffice. While the value of R1 can be calculated, a more practical solution is to just adjust the source resistor to produce 0.88 mA, and forget about the analytical stuff.

If you want to do the math, an accurate prediction of the value of R1 can be made when the parameters of the particular JFET are known. The values given in the data sheets for the 2N5457 are pretty loose: I_{DSS}, the drain current with gate-source voltage zero, is between 5 mA and 1 mA, and cutoff voltage V_{off}, the gate-source voltage that reduces the drain current to less than 1 μA, is between 0.5 V and 6 V. The MPF102 limits are looser. Not very close, but all is not lost.

The relationship between drain current and gate voltage of a JFET can be expressed as:

$$I_D = I_{DSS}(1 - Vgs/V_{off})^2 \quad \text{(Eq 1)}$$

where:

I_D = drain current for the particular gate voltage.

I_{DSS} = drain current with the gate-source voltage zero.

V_{off} = gate-source voltage that reduces I_D to essentially zero (less than 1 μA).

V_{gs} = the gate-source voltage that produces I_D.

Equation 1 can be rewritten to solve for V_{gs}/V_{off} and V_{off}:

$$V_{gs}/V_{off} = 1 - \sqrt{\frac{I_D}{I_{DSS}}} \quad \text{(Eq 2)}$$

$$V_{off} = V_{gs}/[1 - \sqrt{\frac{I_D}{I_{DSS}}}] \quad \text{(Eq 3)}$$

The values of I_{DSS} and V_{off} can be determined with a couple of simple tests. A supply of 6 to 20 volts or so, a resistor in the order of 10k and a voltmeter and milliammeter will be needed. Connect the positive voltage to the drain and the negative side to the source. Short the gate to source of the JFET and measure the drain current I_{DSS}. Connect a 10k or so resistor in the source. Measure the voltage V_{gs} across the resistor and the resulting drain current I_D. With V_{gs} and I_D known, the value of V_{off} can be computed with equation 3. With these values of I_{DSS} and V_{off} established, the gate-source voltage needed to produce a particular drain current can be calculated with equation 1.

The drain current of a JFET is essentially independent of the drain-source voltage when the drain-source voltage is above pinch-off. For a typical 2N5457, pinch-off is a couple of volts for drain current of less than 1 mA.

In **Fig. 3**, PWM is accomplished by comparing the sawtooth voltage to a variable control voltage obtained from potentiometer R6. When the sawtooth voltage on the inverting input of U3 is less positive than the voltage on the noninverting input, the output is high and stays high until the sawtooth exceeds the voltage on the inverting input. Reducing the positive voltage to the inverting input from R6 decreases the width of the positive pulse on the output of U3, and increases the width of the low.

In **Fig. 4**, U4, a type D flip-flop like the CD4013, is connected to divide the sawtooth period by two. In a D flip-flop, the D input is transferred to the Q output on application of a positive clock pulse. The outputs are 100 kHz square waves (200 kHz divided by 2)

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that control the outputs of the NOR gates that drive the power amplifier.

The rise of the output of U2 of the PWM provides the clock for the CD4013. Therefore, on alternate cycles of the clock (sawtooth) the low output of U4 switches from the Q output to the not-Q output.

The outputs of the NOR gates are high when both inputs are low. That is, when the output of U3 of **Fig. 3** is low and one output of U4 of **Fig. 4** is low, the outputs of a NOR gate is high and one of the MOSFETs is on.

The DC output can be regulated when a sample of the output is used to control the U3 instead of the voltage from the manually controlled pot R6, as shown in **Fig. 4**.

Feeding a sample of the output back to control the output is simple in concept, but the devil is in the details. The feedback considerations are exactly the same as those for any feedback amplifier: the feedback must be 180° out of phase with the input until the gain of the amplifier has fallen to less than one. When the voltage fedback is in phase with the input, or nearly so, the system will oscillate. Therefore care must be taken to insure that the phase of the fedback signal is always at least 150° out of phase with the input.

The operational amplifier, U5 in **Fig. 4**, compares a sample of the output DC with a fixed reference DC voltage to provide the control voltage for the PWM. An increase in the output of the DC voltage causes the output of the error amplifier U5 to be less positive and narrows the PWM's positive pulse.

The feedback to the PWM forces the inverting input of U5 to be 5 volts. Therefore, adjusting R11 in **Fig. 4** changes the voltage division and varies the DC output voltage. R9 minimizes the change in input resistance of U5 and consequently the gain of the amplifier as R11 is changed.

The stability of the output voltage is essentially the stability of the reference voltage and the closed loop gain of the amplifier. The gain of the error amplifier U5 is the chief determinant of regulation with changes in load. A small change in the output voltage produces a significant change in the control voltage fed back to the PWM.

In the example given, 9 to 13 volts was assumed across each half of the transformer. In reality, the voltage across the transformer depends on the kinds of transistors and blocking diodes used in the amplifier. When the transistors are MOSFETs, the losses in the drain-source resistance can be very low. In newer devices the drain-source resistance R_{DS} is as low as 30 mΩ and drops less than 0.3 V even when drain current is 10 A. When the blocking diodes are Schottkies, the total drop will be less than 0.6 V.

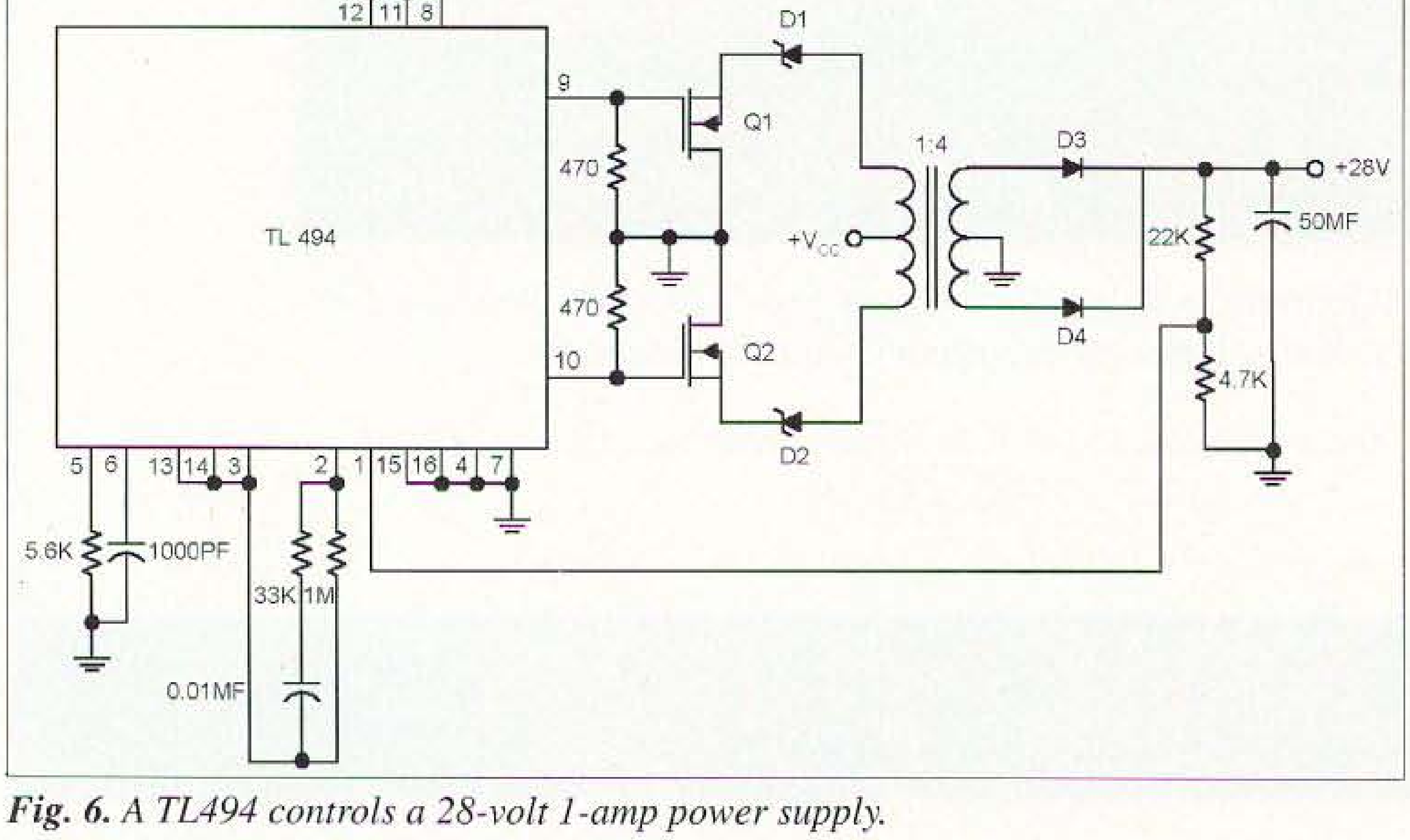

Fig. 6. A TL494 controls a 28-volt 1-amp power supply.

While common emitter bipolar transistors could be considered, their storage time tends to be long when saturated and limits the frequency that can be used. Emitter followers are fast enough because they do not saturate. Unfortunately unsaturated means their collector-emitter voltage is a volt or so. In many SMPSes emitter followers are used and the losses accepted.

Switch mode power supply controllers are available as ICs that are reasonably priced. The ICs are very similar even though the methods of implementation may vary and some extra bells and whistles may be included. **Fig. 5** shows the Motorola TL494, a basic PWM controller IC.

The TL494 has two error amplifiers, a dead band amplifier, a regulated 5 volt reference voltage, an on chip oscillator, a PWM, and two uncommitted bipolar output transistors that can supply 200 mA. Since both emitter and collector are uncommitted, the transistors can be used as either common emitter or emitter followers. All this in a 16-pin DIP. Unless you just like to build things, the IC is the way to go. Still, understanding what each of the functional blocks in the IC does will give a much better appreciation of how to apply the IC to your particular needs.

The extra error amplifier can be used to shut down the supply when there is an over current or the voltage exceeds some preset values. In the figure, pin 13 OC is the output control, a high enables the supply. The dead time controls the maximum duty cycle the power amplifier can have and is controlled by the voltage on pin 4, the minimum dead band occurs when pin 4 is grounded. The compensation network between pins 3 and 4, 33k and 0.01 µF, reduces the error amplifier gain above 3 kHz.

The internal oscillator is controlled by the R_T at pin 6 and C_T at pin 5. The oscillator frequency (twice the output switching frequency) is approximately $1.1/R_TC_T$. For an oscillator frequency of 200 kHz C_T can be 1000 pF and R_T is 5.6 kΩ.

The diodes D1 and D2 are

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