Parker R. Cope W2GOM/7
8040 E. Tranquil Blvd.
Prescott Valley AZ 86314

All About Class D Amplifiers

What you do know can't hurt you.

A class D amplifier is a high efficiency amplifier whose efficiencies can be greater than 90%. Some authors restrict class D amplifiers to amplifiers with constant duty cycle, and reserve class S for pulsewidth-modulated (PWM) amplifiers. By whatever name, the amplifiers have very high efficiencies. Just to put us on the same page, I'll call them all class D.

When vacuum tubes were the only active devices, amplifiers were classified as A, B, or C, depending on their conduction angles. For example, a class A amplifier conducted for 360 degrees of the input signal and efficiency could never exceed 50%. A class B amplifier conducted for about 180 degrees and had a maximum efficiency of 78.5%. A class C amplifier conducted for less than 180 degrees and had an efficiency of about 65% to 85%. While other modes of operation were known to the academics, the limitations of tubes restricted their practical application to these three classes.

An in-between class of operation that is appropriate for vacuum tubes is class AB. In class AB the push-pull connection eliminates even harmonics, and makes it possible to extend operation until the instantaneous plate current is reduced to zero for a small portion of each cycle without causing excessive distortion in the output. Class AB amplifiers have efficiencies in the order of 40 or 50%.

Because tubes can have short term dissipation exceeding the rated plate dissipation, class AB was used to squeeze a little more power out of a tube. As an aside, a further distinction of amplifiers was denoted by the subscript 1 or 2 which indicated whether grid current flowed. The subscript 1 meant no grid current and is assumed unless otherwise noted. The subscript 2 meant there was *some* grid current. Class A_2 or AB_2 indicated that grid current flowed (with an attendant increase in distortion). Because the cost of a tube was significant, every effort was made to reduce the tube count and get the most from every stage. Today the cost of the passive components in a circuit outweigh the cost of the semiconductor and no great effort is expended in reducing the transistor count.

Class G amplifiers are a high efficiency class of amplifier that has a maximum efficiency of 84.2%. A complementary class G amplifier requires two power supplies and uses

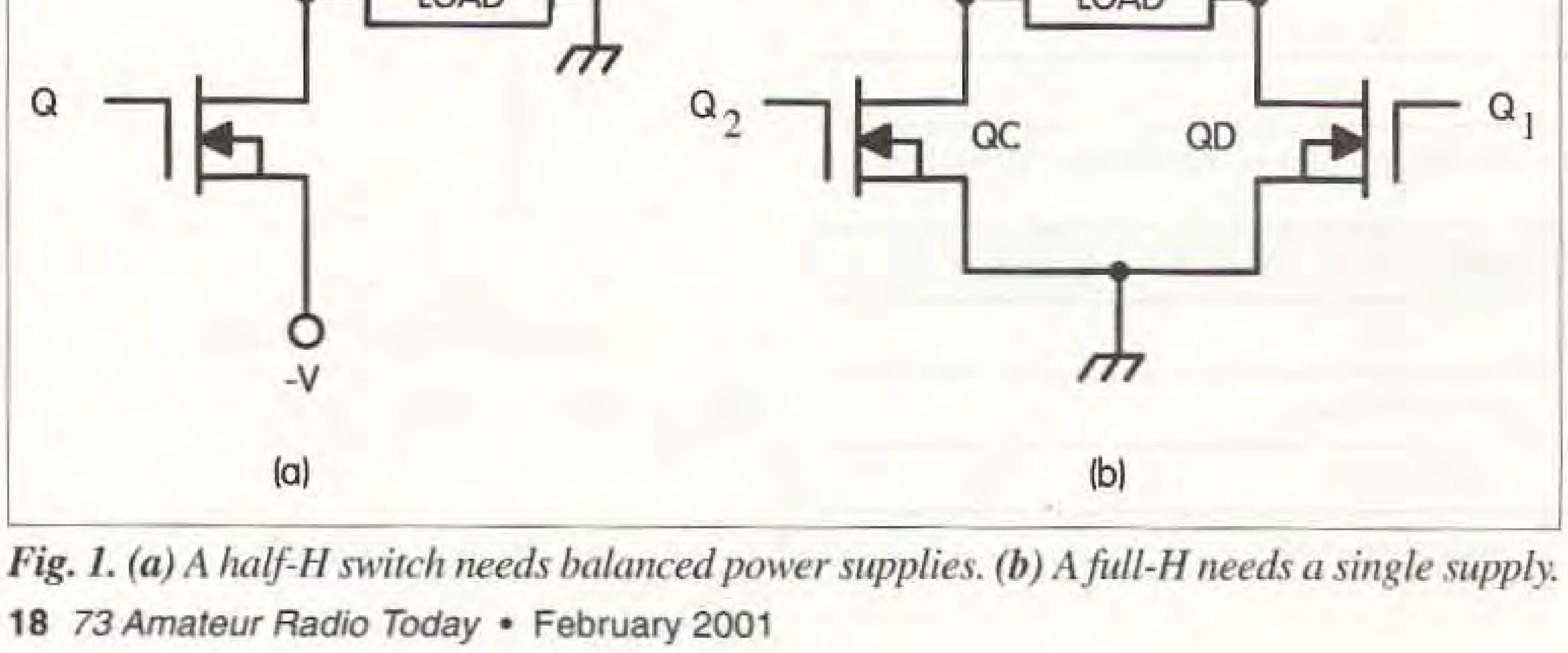

***Fig. 1. (a)** A half-H switch needs balanced power supplies. **(b)** A full-H needs a single supply.*

four transistors, instead of the two used in class B. One pair of transistors operates from a lower supply voltage (e.g. ±1/2 V_{cc}). The first pair of transistors work for low level signals and the second air operates when the input signal exceeds what the first pair can handle. The scheme was described by L. Feldman in "Class G High Efficiency Hi-Fi Amplifiers," in *Radio Electronics,* August 1976. The scheme has enjoyed little success.

The class of operation of an amplifier cannot be divined by just looking at the schematic — it depends on the operating point. For example, the schematic of a class B amplifier looks very much like the schematic of a push-pull class A amplifier, but a class D amplifier is unique.

Power semiconductors have opened many of the formerly esoteric amplifiers to practical uses. Class D amplifiers are an example of a high efficiency class of amplification made practical by semiconductors. Class D amplifiers are basically switches that are either open or closed. When off, the semiconductors conduct no current and dissipate no power. When on, the drop is less than a volt even when conducting amps, and they dissipate little power. On the other hand, vacuum tubes need a plate voltage in the order of a hundred volts to conduct even a moderate current. In short, tubes don't make good switches.

Since the transistors in a class D amplifier are either on or off, they aren't suitable for linear applications as such. But, the average of a PWM signal can be proportional to a varying audio input, and the PWM signal can control efficient power semiconductor switches. Class D amplifiers routinely have efficiencies approaching 95%.

Pulsewidth modulation is widely used in switch mode power supplies (SMPS) to produce a regulated unipolar voltage. The output DC voltage is compared to a reference voltage to produce a PWM signal and the pulsewidth changes to correct the output voltage.

While either MOSFETs or BJTs can act as the switch, the MOSFET is preferred because of its higher switching speed and lower drive requirements. High switching speed is desired because

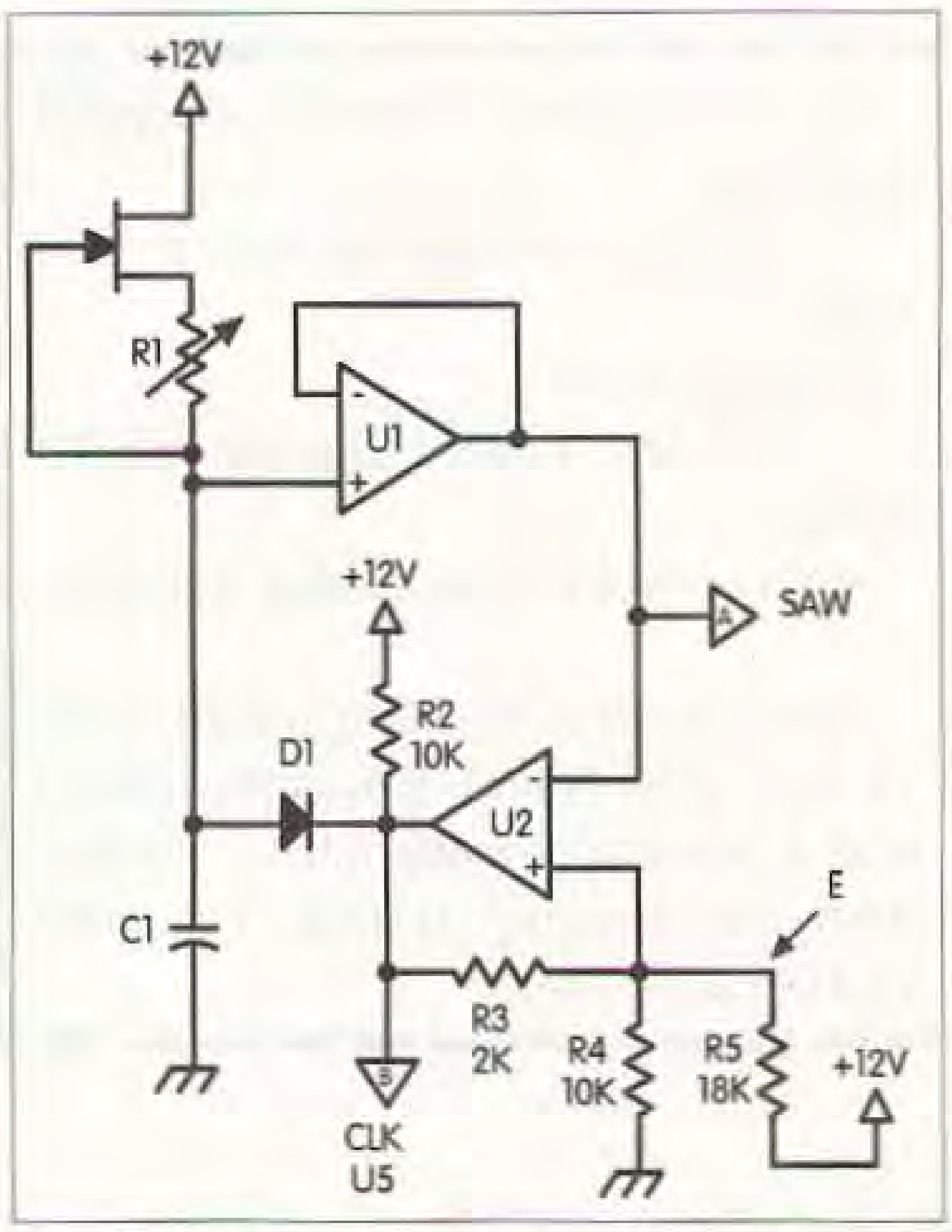

Fig. 2. PWM uses a linear sawtooth.

a high switching frequency reduces the inherent distortion in the recovered signal. A switching frequency ten times the highest audio frequency is sufficient for distortions below 40 dB. Therefore a 20 kHz audio signal requires a switching frequency of 200 kHz or greater. 3 kHz communications audio only requires a switching frequency of 30 kHz.

The efficiency of the switch is ultimately determined by the internal resistance of the switch transistor(s) and the load resistance. When R_S is the switch resistance, the efficiency η is:

$\eta = P_L/(P_L+P_s) = R_L/(R_L+R_S)$

where R_l is the resistance of the load and R_S is the saturation resistance of the transistor(s). For high efficiencies, R_S must be low compared to the load. The internal resistance of a MOSFET is R_{DSon} and the internal resistance of a BJT can be calculated from V_{CESAT} and the collector current I_C, $R_S = V_{CESAT}/I_C$.

An internal resistance of even 0.1 Ω and a 4 Ω load has the potential of achieving an efficiency of about 95%. Recent MOSFET introductions have R_{DSon} of a few milliohms. Since the power dissipated in the transistor is only I^2_{RDSon}, heat sinking of the transistors is seldom needed.

The BJT has a lower collector/emitter saturated voltage and dissipation than the R_{DS} and dissipation of a MOSFET. But this advantage is usually overridden by higher base drive current and delays when turning off the saturated BJT. It's not a simple trade-off.

A price is extracted for using a high switching frequency: The gate drive current for the MOSFET switch becomes a serious consideration for any fast or high frequency application.

Even though the MOSFET is a voltage-controlled device it takes time to charge the input capacitance and affect a switch. The rate of change of voltage with respect to time across the a capacitance may be expressed as:

$dV/dt = I/C$

where I is the charging current in amps and C is the capacity being charged in Farads, t is in seconds. The effective input C of a FET is not obvious.

The input capacity of an FET is more than just the static gate/source capacity, C_{gs}, plus the gate/drain capacitance, C_{gd}. The input capacitance is increased by that old bug-a-boo: "Miller effect." Miller effect is the phenomenon by which the feedback path between input and output provided by the interelectrode capacitance C_{gd} increases the apparent input capacitance.

The input capacitance of an amplifier can be expressed as:

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

where ϕ is the phase angle of the drain load. When the drain load is a resistor, ϕ is 0° and $\cos\phi = 1$. Therefore, when the load is resistive, $C_{in} = C_{gs}+C_{gd}(1+A)$.

Another uncertainty of the input capacitance of a transistor is the variable nature of C_{gd}. C_{gd} of a transistor is not a constant value like a tube's plate-to-grid capacity. The grid/plate capacitance of a tube depends on the physical construction of the tube which is constant, while the gate/drain capacitance of a transistor is dependent on the voltage across the gate/drain junction. As the reverse junction voltage increases, the static junction capacitance decreases. This change in capacitance with voltage is referred to as the varactor effect or the parametric capacitance. The parametric capacitance introduces a complication in determining the capacitance to be charged.

Most power MOSFET manufacturers specify a "total gate charge," Q_T, needed to raise the gate voltage sufficiently to just saturate the drain for the particular drain voltage and rated drain current:

$Q_T = i_g t$

Where i_g is the gate input current in amperes and t is time in seconds needed to charge the gate capacitance. 25 mA into a Q_T of 10 nC (nano-Coulombs)

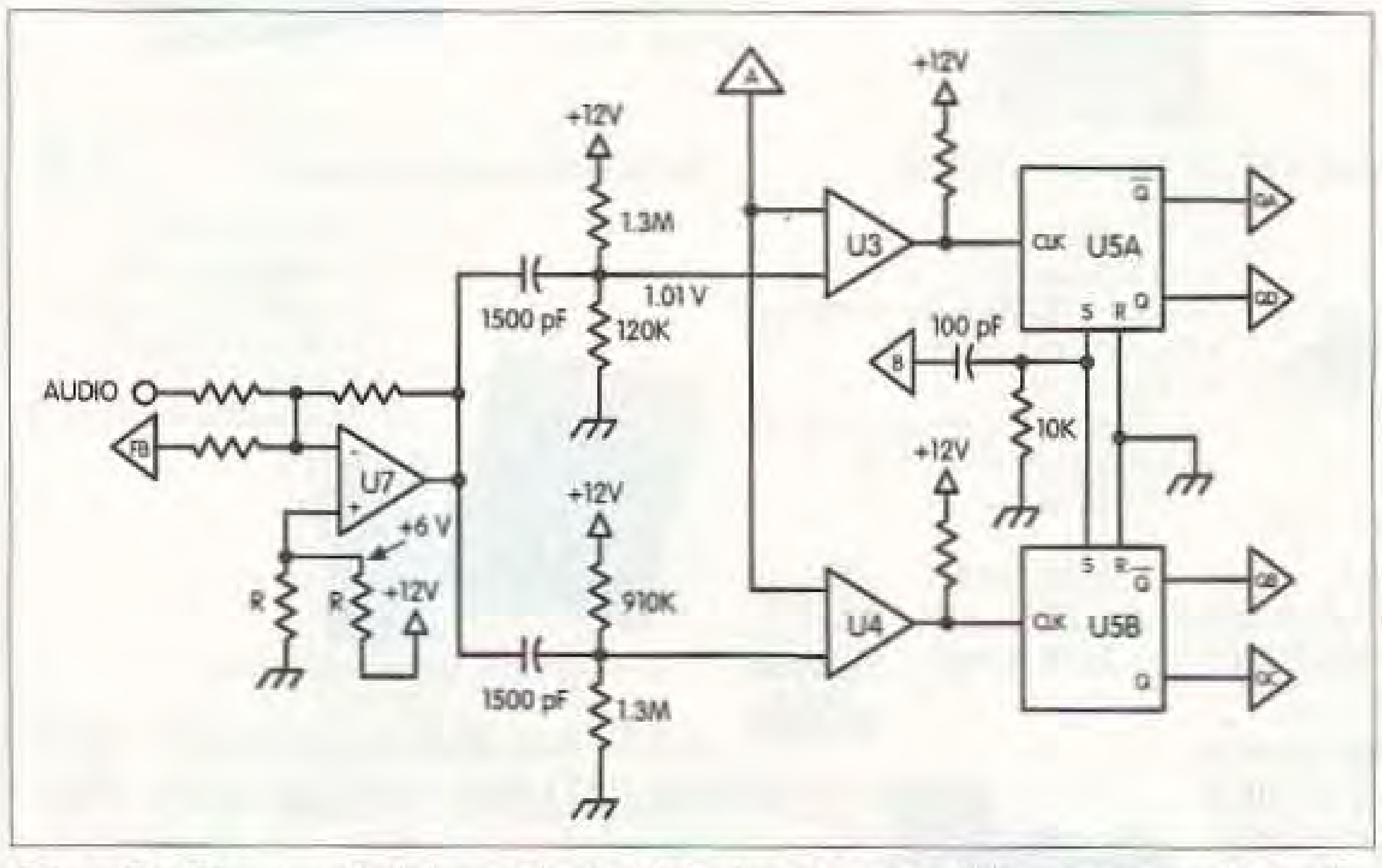

Fig. 3. The variable pulsewidth is obtained by comparing the sawtooth with the audio.

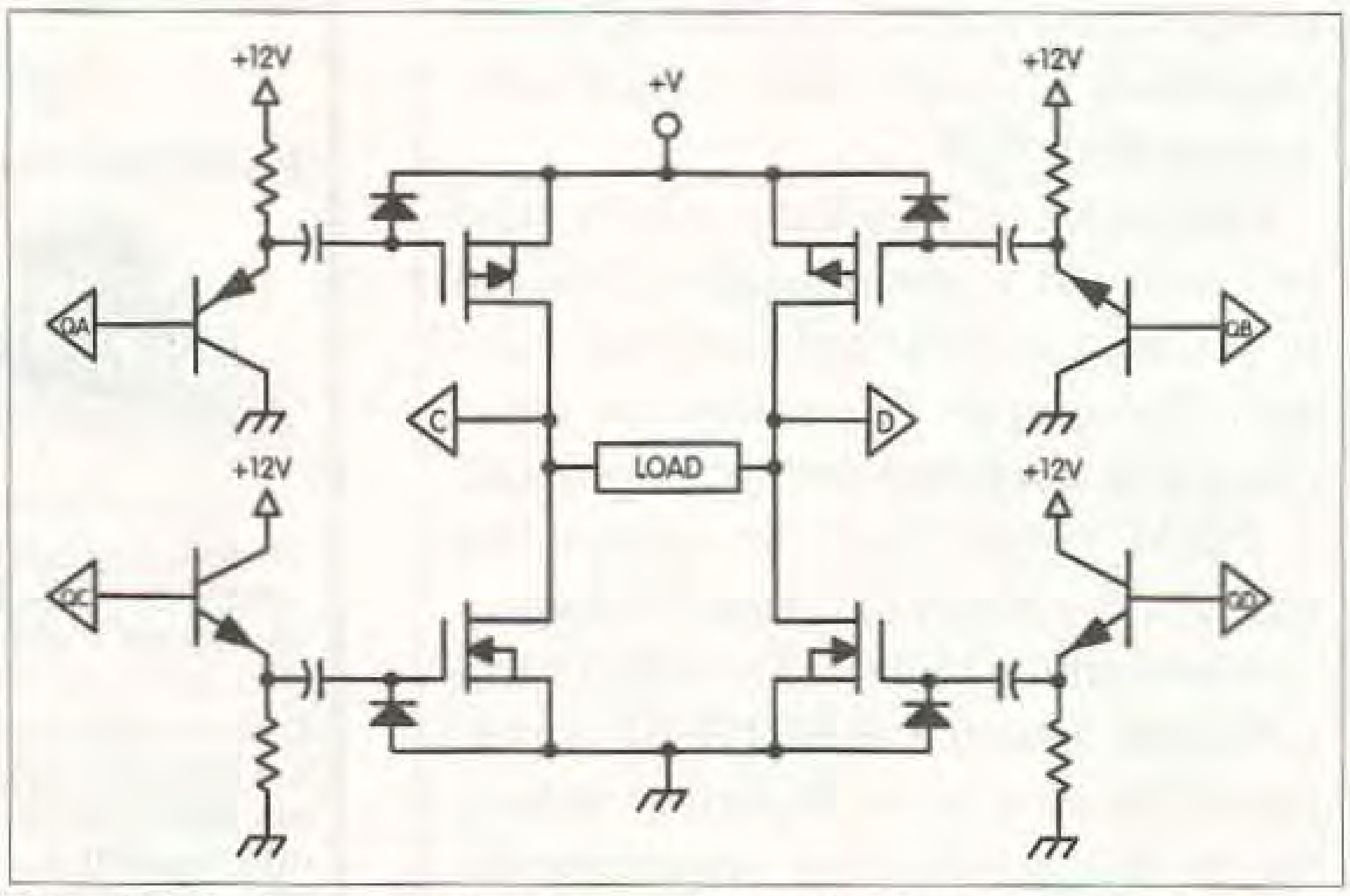

Fig. 4. The drive to high voltage MOSFETs is capacitively coupled.

will change the gate 2.5 V in 1 μsec. The gate drive current is a spike of current that exponentially trails off to zero as the input capacitance becomes charged.

Since the total gate charge is that charge which is required to just switch the device, prudence indicates that overcharge capability be provided. A safety factor in the charging current of two to one is not unreasonable. The compliance of the gate drive current must also be greater than the gate voltage required to saturate the drain current. Most power MOSFETs require a gate voltage in the range of 5 to 10 V to saturate the drain current, but some devices switch with voltages as low as 2.5 V. The charging current then must come from a voltage of 2.5 V to maybe 10 V.

The time needed to charge the gate capacity is important because during the rise and fall of gate voltage, the drain voltage and current are nonzero and power is dissipated in the drain. As a result, the gate drive current influences the efficiency for a particular frequency. A rise/fall time that is 10% of the total period increases the dissipation by a tolerable amount. For example, a 10% rise and fall time decreases an efficiency of 95% to about 94%.

The configuration of the power switch is usually a full-H as shown in **Fig. 1(a)**. When transistors QA and QD are on, current flows in the load in one direction. When QB and QC are on, load current flows in the other direction. For the sake of argument, let's say current flows in QA and QD represent the positive half of the output cycle and QC and QB represent the negative half cycle.

A full-H has the advantage of producing a peak-to-peak output voltage that is nearly twice the single supply voltage. However, a full-H requires complementary MOSFETs and complimentary gate drives. A negative pulse on QA's gate and a positive pulse on QD's gate allows current to flow in one direction in the load. Load current reverses when QB and QC conduct.

A full-H with a 12 V supply has a peak-to-peak load voltage of almost 24 V_{pp}, and the maximum power output to a 4 Ω load will be about $V_{rms}{}^2/4$ or 18 W_{rms}. A half-H shown in **Fig. 1(b)** requires balanced positive and negative supplies and produces an output of

$(V_+ + V_-)2/2R_{load}$.

When the transistors have an R_{DSon} of 0.1 Ω, transistor dissipation ($I_{rms}{}^2 R_{DSon}$) will be about 0.15 W. Obviously, heat sinking is not a major consideration.

If the P-channel transistors QA and QB in **Fig. 1(a)** need a drive of 5 V, they are turned on by a gate voltage of +5 V and off by 0 V. The voltage across the load essentially swings from the positive rail to ground in one direction for a positive input and in the other direction for the negative input.

While the power in the load may be high, transistor dissipation is low. For example, the current in the load and transistor is approximately E_{supply}/R_{load}. If R_{DSon} is specified as 0.1 ohm and the load is 4 ohms, the total resistance in series with the load is 0.2 ohms and 95% of the supply voltage appears across the load. If the load were 16 ohms the efficiency would be over 98% with zero transition time. If the transition time is 10%, the dissipation will increase and efficiency will drop about 1%. Heat sinking is still not much of a factor.

The PWM generator is the critical part of the class D amplifier. PWM controllers intended for power supplies or motor controllers are available from several companies and a few are intended for audio applications. But they all have the same basic internals, though they may be implemented differently.

Fig. 2 shows a generic method of generating linear PWM with discrete components: A constant current charging C1 produces a linear ramp of voltage. U1 is a unity gain buffer that isolates C1 from any loading presented by the following circuits. When the voltage across C1 equals the reference voltage on comparator U2, U2's output goes to near zero and discharges C1 through D1. When U2's input is less than the reference, U2's output goes to +12 V and allows C1 to charge and the cycle repeats. The positive transition of U2's output sets the flip-flops U5A and U5B in **Fig. 3**.

The sawtooth is compared with the audio in comparators U3 and U4,

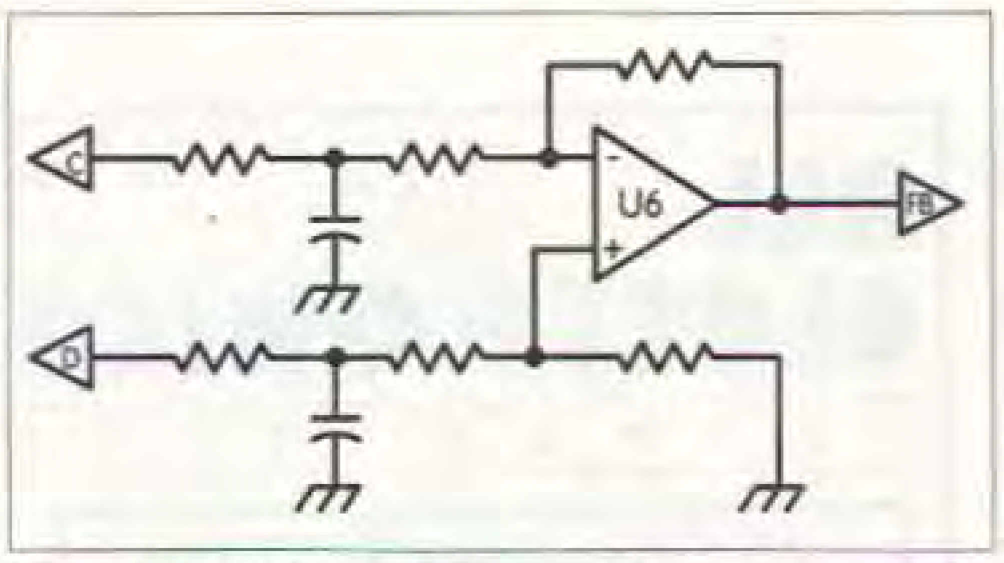

Fig. 5. A differential amplifier converts the floating output to single-ended.

shown in **Fig. 3**, to produce a pulse whose width is proportional to the audio amplitude. U7 provides the gain needed to raise the audio input to about 6 volts peak for use by U3 and U4. U7 also sums the input with the output fed back from U6 for comparison with the input.

The constant current charging C1 is produced by the JFET Q1. R1 in the source of Q1 controls the charging current and consequently the rate of rise of the voltage across C1. The hysteresis of comparator U2 controls the amplitude of the sawtooth. After the amplitude is established by U2, R1 can be adjusted to produce the required frequency. The hysteresis in U2 controls the amplitude of the sawtooth; the current in Q1 controls the rate of rise of the voltage across C1.

For the values shown in **Fig. 3**, the noninverting input of U2 swings from a low, E_{lo}, of about +1 V to a high, E_{hi}, of about +7 V. A swing of about 6 volts. Therefore, the output of U2 switches at inverting inputs of +1 V and +7 V. When the ramp of voltage applied to the inverting input rises to the voltage on the noninverting input, +7 V, the output of U2 goes to zero, and C1 is discharged to near zero through D1. When the voltage on C1 falls to 1 V, the output of U2 switches to +12 V, D1 becomes reverse-biased, and the voltage on C1 rises from about +1 V to about +7 V and the cycle repeats. The positive rise of the output of U2 also resets the flip-flops U5A and U5B.

The hysteresis of U2 and the ramp amplitude can be changed by choosing different values for R2, R3, and R4 and the voltage E at the inverting input can be calculated. When the output of U2 is low, essentially zero, the noninverting input is:

$$E_{lo} = 12x(R3^{-1}+R4^{-1})^{-1}/(R4^{-1}+R3^{-1})^{-1}+R5.$$

When the output of U2 is high, the noninverting input is:

$$E_{hi} = 12x[R5^{-1}+(R2+R3)^{-1})]^{-1}/\{R4+[R5^{-1}+(R2+R3)^{-1}]^{-1}\}.$$

The comparators U3 and U4 compare the voltage ramp to the audio. When the offset audio exceeds the ramp voltage, the comparator switches and produces a pulse whose width is proportional to the audio amplitude. The audio is capacitively coupled to U3 and U4 to remove any DC that may exist on the audio.

The audio input to U3 is offset to +1 V and the positive half of the audio raises the input to U3 above 1 V, and the output switches to +12 V. The audio input to U4 is offset to +7 V and when the negative half of the audio input makes the input less than the sawtooth, U4's output switches.

The positive transition of U3 and U4 triggers U5. The outputs of U5B drive the QB and QC sections of the switch, and the outputs of U5A drive the QA and QD sections of the switch. The flip-flops are reset by the positive transition of U2 when the sawtooth starts its rise. The outputs of U5 drive the switches.

When U5 is a CD4013, the output current is limited so that some current gain must be provided to drive MOSFETs with high Q_T. A 2N3906 PNP emitter follower can be used to drive the P-channel MOSFET and a 2N3905 NPN can be used to drive the N-channel MOSFET. These transistors have h_{fe} of about 60 at 1mA of base current and can supply significant peak currents to drive rather large MOSFETs.

The voltage across the load essentially floats; one side is grounded for positive signals while the other side is grounded for negative signals. This balanced signal is converted to single ended in U6 for feedback to the input of U7 as shown in **Fig. 5**. U6 can also scale the voltage to a range suitable for U7.

The voltage across the load is a series of pulses that must be filtered to remove the switching components before it can be compared to the audio input. The filter is not terribly critical but it can introduce phase shifts in the recovered audio that can lead to instability when fedback. The amplifier will oscillate if the overall phase shift approaches 180° before the gain has fallen to less than one. Therefore, the low-pass filter should cut-off below the switching frequency but well above the highest audio frequency. Ideally, the phase shift should be no more than about 150° at the highest audio frequency.

Kilowatt amplifiers require power supplies in the range of a couple of hundred volts and several amps. The drive required is still only ten volts or so, and the peak gate current will seldom need to be greater than 100 mA. A drive current of 70 mA can switch a MOSFET with a total gate charge of 70 nC in 1 µs.

Shifting the drive to the 200 volt gate voltage of the P-channel MOSFET requires capacitive coupling. **Fig. 4** shows how this can be accomplished. The diodes clamp the gate drive to the high voltage. The coupling capacitors should be large compared to the gate input capacity. But remember, it is the switching frequency being coupled, not the audio frequencies. 0.1 µF capacitors with appropriate voltage ratings should do the job.

Class D amplifier ICs are available that perform the PWM and drive speaker resistances. The Linfinity Microelectronics' LX1720 is a stereo unit that produces 20 W per channel. The data sheet says THD + noise is less than 1% over the 20 Hz to 20 kHz frequency range. Linfinity has a monophonic unit, the LX1710, with power capability of 50 watts that should be available by the summer of 2000. TI also has a monophonic class D IC amplifier, the TPA032D01, that has an output of about 10 W. The IC was designed to drive MOSFETs with Q_T of 10 nC so they have rather limited output current capabilities. For high power, high voltage applications, capacitive coupling and some current gain is in order. The emitter followers are appropriate here as well. While the

Continued on page 59

All About Class D Amplifiers

continued from page 22

THD + noise isn't anything to brag about, it is pretty good.

Big brutes of audio amplifiers, kilowatts, can be built with little or no heat sinks. High efficiency audio amplifiers are no longer just academic curiosities. They are here and now. High power class D amplifiers are certainly in the realm of possibility without being a boat anchor.