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Class D amplifiers provide high efficiency for audio systems

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Advances in MOSFET technology and integrated half- and full-bridge predrivers now make class D amplifiers a practical alternative to linear amplifiers in many applications. The biggest benefit compared to more traditional amplifier topologies is high efficiency.

Many engineers are familiar with motor-control circuits that pulse-width-modulate a voltage to control both the direction and speed of a dc motor. You can extend this concept to an audio amplifier by replacing the motor load with a lowpass filter and a speaker and by using an audio input as the control signal. Such amplifiers are class D switching amplifiers, which are analogous to conventional class A, B, AB, and C circuits, in which each switching element is in its linear mode of conduction for a large percentage of each cycle of an audio input signal. This linear operation reduces efficiency to about 60% in the typical class AB amplifier and necessitates the use of large heat sinks.

Alternatively, the switching elements of a class D amplifier are either cut off or in saturation most of the time, allowing high efficiencies. The high efficiency translates into reduced heat sinking, smaller size, and lighter weight. Also, class D amplifiers do not suffer from crossover distortion.

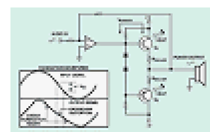
An evaluation-board design shows efficiencies of 94% and THD-plus-noise (THD+N) performance that ranges from less than 1 to 2.8%. Although the THD+N curves for this board's design are not as good as what you can obtain with a high-performance class A or class AB amplifier, the curves are more than adequate in many applications. For instance, automotive manufacturers typically require amplifiers in car stereos to have distortion below 2% at lower power levels and below 5% at peak power. Also, optimizing some of the supporting circuits should result in even higher performance than that of this evaluation board.

The concept of a class D switching amplifier has been around for about 50 years. Early attempts to develop switching amplifiers with vacuum tubes were limited by the tubes' large voltage drops and low current capabilities, which reduced the amplifiers' efficiencies and limited their output power. In the late 1960s, bipolar transistors became a practical alternative to vacuum tubes and allowed the implementation of switching amplifiers with very high efficiencies at low frequencies.

However, an audio switching amplifier requires high-frequency operation, which is generally equal to at least four or five times the bandwidth of the 20-kHz audio spectrum. Higher frequency operation makes it easier to design the filter that removes the carrier frequency before the audio signal drives the speaker. Using bipolar transistors at the required frequency of 80 kHz or greater results in excessive switching losses that eliminate the class D amplifier's efficiency advantages.

In the 1980s, MOSFETs became available that could meet both the switching-speed and conduction-loss requirements to effectively implement class D amplifiers. The first switching amplifiers using MOSFETs incorporated electrically isolated drivers to allow the use of N-channel devices. N-channel MOSFETs yield more efficient designs; these MOSFETs have approximately one-third the conduction losses of their P-channel counterparts. However, the isolated drive circuits were complex and limited the use of switching amplifiers.

Fabrication processes that allow the integration of high-voltage and logic circuits on one die made possible the first commercially available integrated predrivers for class D amplifiers. For example, the Harris HIP4080 provides monolithic isolation and allows the use of high-side N-channel MOSFETs through the use of a bootstrap circuit and an integrated charge pump. The chip integrates a PWM comparator for modulation. This full-bridge driver can switch MOSFETs at frequencies above 1 MHz, reducing the size of magnetics and simplifying the filter design.



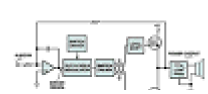
A conventional push-pull (class AB) linear amplifier (Fig 1) modulates load power by continuously varying conduction through its pass elements during most, if not all, of the conduction cycle. Q_1 conducts during the positive half-cycle, and Q_2 conducts during the negative. During each half-cycle, the conducting transistor operates in its linear region. The transistor must supply the required current to the load while reducing the voltage between the supply and the load. The power dissipated in the transistor, which equals $(V_{BUS}-V_{LOAD}) \times I_{LOAD}$, is wasted in the form of heat. Generally, you must use large heat sinks to prevent the output stage from overheating.

The class AB amplifier uses a bleeder circuit to reduce crossover distortion, which occurs during the zero-crossing of the input signal when neither transistor is on (when the input signal is below the V_{BE} of either transistor). The bleeder circuit biases both transistors on during crossover, but the circuit draws current that further reduces the efficiency of the amplifier.

Ballast resistors in class AB designs prevent the transistors from going into thermal runaway. Bipolar transistors are at risk because their V_{BE}s have negative temperature coefficients. Usually, you must mount the diodes and transistors on the same heat sink to ensure that the V_{BE}s track. The common heat sink helps minimize crossover distortion over temperature.

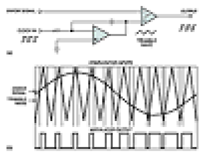
Class D audio amplifiers

Class D amplifiers convert the audio signal into high-frequency pulses that vary in width with the audio signal's amplitude. The varying-width pulses switch the power-output transistors at a fixed frequency. A lowpass filter converts the pulses back into an amplified audio signal that drives the speakers. This design approach produces an amplifier with better than 90% efficiency and that is more complex than its linear counterpart.



The amplifier requires an integrator, a duty-cycle modulator, a switch predrive circuit, and an output filter. The half-bridge class D amplifier using constant-

frequency, duty-cycle modulation (Fig 2), sums the square-wave output of the switching power transistors with the audio input to provide negative feedback. You cannot take the feedback after the lowpass filter unless you use a complicated compensation network to handle the phase shift that the filter introduces. A two-pole filter, for example, would introduce a 180° phase shift, which would cause the circuit to oscillate.



The square-wave output is synchronous with the audio input, but you must remove the carrier. The integrator sums the two signals and simulates the effect of the output filter. The circuit feeds the resultant error signal into the duty-cycle modulator, which comprises a comparator and a triangle-wave generator (Fig 3). Then, the circuit compares the triangle wave to the error signal to produce the modulated output.

The modulated output is a square wave whose duty cycle is proportional to the input signal. In this half-bridge circuit, this output drives the upper and lower power switches in antiphase; the circuit always drives one switch into saturation while it cuts the other off. The square wave causes the switches to change state as fast as possible, given the technology used to implement the switch. Fast switching limits the time that the switches spend in the linear operating region, thereby increasing efficiency and reducing heat generation. The combination of switching and conduction losses defines the upper bound of the amplifier's efficiency.

The circuit filters out the high-frequency square wave that the power switches generate, leaving only the amplified audio signal. This signal then drives a ground-referenced speaker load.

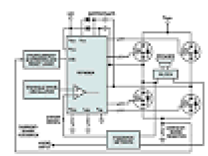


Fig 4 shows a simplified schematic of a class D audio amplifier using a full-bridge configuration. If you use a full-bridge design, in which the circuit develops a peak-to-peak load voltage equal to twice the supply voltage, you need only a unipolar supply instead of the bipolar supply required for the half-bridge design. The full-bridge circuit operation is similar to that of the half-bridge design, except the full-bridge design uses four MOSFETs instead of two, uses a differential lowpass filter in the feedback network to support the floating load, and requires two lowpass filters to attenuate the carrier.

Four N-channel MOSFETs switch high currents into a low-impedance load. These MOSFETs turn on in pairs, one high-side device with the opposite low-side device. To avoid shoot-through, the circuit ensures a dead time between turning off one pair of MOSFETs and turning on the other pair.

MOSFETs in a full bridge need to withstand only half the voltage they would see in a half-bridge circuit. This lower voltage requirement allows you to minimize conduction losses, even though two devices are in the conduction path. Given a fixed die area, two MOSFETs in series have a lower total $R_{DS(ON)}$ than does one MOSFET with twice the drain-source breakdown voltage rating (BV_{DSS}) because $R_{DS(ON)}$ increases nonlinearly with higher drain-to-source breakdown voltage.

Base your MOSFET selection on the following criteria: peak voltage and current requirements, body-diode reverse-recovery time, switching losses, and conduction losses. The peak voltage and current determine the required ratings that the MOSFET must be able to sustain. The body-diode recovery time, switching losses, conduction losses, and output-filter losses determine the output stage's efficiency. You can calculate peak voltage and current using the following equations:

$$V_{PEAK} = \sqrt{2 \cdot P_{OUT} \cdot Z_{LOAD}}$$

$$I_{PEAK} = \frac{V_{PEAK}}{Z_{LOAD}}.$$

For example, if you require an output power of 100W into an 8(ohm) speaker, V_{PEAK} and I_{PEAK} would equal 40V and 5A, respectively. Therefore, the MOSFETs must have a BV_{DSS} above 50V (using a guard band of 25%) and a rated current above 5A.

Select a MOSFET with a body diode that has a low reverse-recovery time to reduce the losses that occur when the MOSFETs switch off and on. The current state-of-the-art reverse-recovery time is 100 nsec. You can minimize switching losses by choosing a low switching frequency, a MOSFET with low gate-to-source capacitance, and a driver with high drive capability. In selecting an operating frequency, weigh the advantages of high carrier frequencies against increased switching losses and the increase in radiated EMI/RFI. The lower the switching frequency, the harder it is to filter the carrier from the audio signal.

If you effectively manage these losses, the conduction losses become the dominant factor in the amplifier's efficiency. In other words, choosing a MOSFET with a lower $R_{DS(ON)}$ increases the amplifier's efficiency. For example, a MOSFET with an $R_{DS(ON)}$ of 200 m(ohm) would reduce the efficiency 5% from ideal. This number results from

$$\Delta\eta = \frac{2 \cdot R_{DS(ON)}}{Z_{LOAD}} = \frac{0.4}{8} = 0.05,$$

multiplying by 2 in the numerator because of the full-bridge topology. However, a lower $R_{DS(ON)}$ of 80 m(ohm) would reduce the efficiency only 2%. That is, you are simply trading silicon for efficiency.

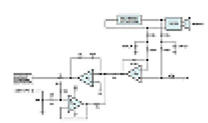
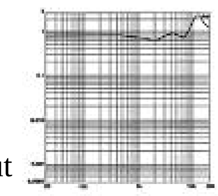


Fig 5 shows the internal circuitry of Fig 4's feedback-network block. In this circuit, the differential lowpass filter comprising IC_{1C} and associated components attenuates the output of the half-bridge by a factor of 11. Buffer amplifier IC_{1B} ac-couples the audio input. An integrating error amplifier, IC_{1A}, sums the filtered feedback signal and the buffered audio input. The circuit feeds the error signal into the inverting input of the comparator in Fig 4's driver IC, which compares the error signal with a sawtooth waveform to generate the PWM signal. A simple triangle-wave oscillator generates the sawtooth waveform.

Another feedback signal in Fig 4 comes from the current-sense resistor on the ground leg of the half-bridge. The sense resistor should have as low a value as possible to reduce its contribution to conduction losses. Simple circuitry in the overcurrent-shutdown and reset-logic block compares the current-sense output to a voltage reference. If the current rises above the maximum rated current of the MOSFETs, the voltage across the current-sense resistor exceeds the reference, causing a comparator to change state and disable the output drivers. You could also use the current-sense resistor to implement current limiting. This would protect the MOSFETs, but the amplifier's output would distort during an overcurrent condition.

The class D amplifier uses two Butterworth filters (Fig 6) to supply the amplified audio signal to the floating load. Butterworth filters provide flat response in the passband, an important criterion in an audio system because a flat response improves dynamic performance. Each filter in Fig 6 has four poles with 30-kHz cutoff frequencies. The clock rejection of each is 74 dB at 250 kHz, which is the carrier frequency for this evaluation-board circuit. You can improve the rejection by adding poles or lowering the cutoff frequency. An RC filter across the load provides impedance stabilization at high frequencies. Stabilization is necessary because the speaker becomes inductive as frequency increases, and a Butterworth filter requires a resistive load.



When driving 100W into a 4(ohm) load, this amplifier has a THD+N across the audio spectrum of less than 1% to 8 kHz (Fig 7a). Above 8 kHz, THD+N increases to about 2.8% before rolling off. THD increases because the amplifier introduces more nonlinearity as the audio input frequency rises. As the frequency approaches the high end of the audio spectrum (above 12 kHz), THD rolls off because the output filters begin to attenuate the harmonics.

However, distortion at lower power levels-that is, performance at less than peak output-is a better measure of an audio amplifier. An audio amplifier delivers average power that is typically a fraction of its maximum output power, depending on the dynamic nature of the music. At a 10W output level, this class D amplifier's THD+N is less than 1.2% over all the frequencies of interest (Fig 7b). THD increases at higher power levels because the amplifier introduces nonlinearities as the output waveform approaches clipping.

THD is primarily a function of filter selection and the feedback network, and you can further reduce distortion to suit the application's requirements. Using a higher grade op amp in the feedback circuit, modifying the compensation network, and improving the linearity of the triangle waveform should all help to improve THD and reduce residual noise. In particular, matching the output filter to the speaker's impedance reduces peaking in the amplifier's closed-loop response and improves the distortion characteristics of the amplifier.

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