LM12,LM318

Application Note 446 A 150W IC Op Amp Simplifies Design of Power Circuits

A 150W IC Op Amp Simplifies Design of Power Circuits

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Abstract: *A power op amp capable of driving ±35V at ±10A has been fabricated on a single silicon chip. Peak power ratings to 800W allow it to handle reactive loads. The IC incorporates internal management circuitry to insure smooth turn on and automatic protection from a variety of fault conditions; this includes instantaneous peak-temperature limiting within the power transistors. The op amp is described briefly, but emphasis is placed on the practical problems encountered in designing with power amplifiers. Numerous application examples are also given.*

introduction

Advances in IC technology have produced a power amplifier that is an order of magnitude more powerful than its predecessors. Unlike other IC's, its peak dissipation rating is many times higher than continuous, as is required for handling reactive loads. Protection circuitry is also more effective. The performance of the new IC, the LM12, puts it in the same class as discrete and hybrid amplifiers. However, it offers far more effective control of turn on, fault and overload conditions in addition to the economies of monolithic construction.

In the late 1960's, the availability of low cost IC op amps prompted their use in rather mundane applications, replacing a few discrete components. This power op amp now promises to extend this to high-power designs. Replacing single power transistors with an op amp may become cost-effective because of improved performance, simplification of attendant circuitry, vastly improved fault protection, greater reliability and the reduction in design time.

Some applications are given here to illustrate op amp design principles as they relate to power circuitry. Unusual design problems that have cropped up in using the LM12 in a wide variety of situations with all sorts of fault conditions are identified along with solutions.

the op amp

The performance of the LM12 is summarized in *Table 1*. The input common-mode range extends to within a volt of the positive supply and to three volts above the negative supply. No input-polarity reversal is experienced should the input-voltage range be exceeded, and no damage results should the inputs be driven beyond the supplies.

The IC is compensated for unity-gain feedback, with a small-signal bandwidth of 700 kHz. Slew rate is 9V/µs, even as a follower. This translates to a 60 kHz power bandwidth under load with a ±35V output swing. The op amp is stable with or without capacitive loading; the maximum load capacitance depends upon loop gain. There are no spurious output stage oscillations, and a series-RC snubber is not required on the output.

The IC delivers ±10A output current at any output voltage yet is completely protected against output overloads, including shorts to the supplies. Dynamic safe-area protection is provided by peak-temperature limiting within the power tran-

sistor array. The turn-on characteristics are controlled by keeping the output open-circuited until the total supply voltage reaches 15V. The output is also opened should the case temperature exceed 150°C or as the supply voltage approaches the BV_{CEO} of the output transistors. The IC withstands overvoltages to 100V.

The LM12 is supplied in a steel TO-3 package with four through leads, plus case. A gold-eutectic die attach to a molybdenum interface is used to avoid thermal fatigue problems with power cycling. Two voltage grades are available; both are specified for either the military or industrial temperature range.

TABLE 1. Some typical characteristics of the LM12 for V_s = ±40V and T_C = 25°C.

parameter	conditions	value
input offset voltage	$V_{CM} = 0$	2 mV
input bias current	$V_{CM} = 0$	150 nA
voltage gain	$R_L = 4\Omega$	50V/mV
output voltage swing	$I_{OUT} = \pm 1.5A$	±38V
	±10A	±35V
peak output current	$V_{OUT} = 0$	±13A
continuous dc dissipation	$T_C = 25°C$	90W
	100°C	55W
pulse dissipation	$t_{ON} = 10$ ms	120W
	1 ms	240W
	0.2 ms	600W
power output	$R_L = 4\Omega$	150W
total harmonic distortion	$R_L = 4\Omega$	0.01%
bandwidth	$A_V = 1$	700 kHz
slew rate	$R_L = 4\Omega$	9V/µs
supply current	$I_{OUT} = 0$	60 mA

general advice

Power op amps are subject to many of the same problems experienced with general-purpose op amps. Excessive input or feedback resistance can cause a dc offset voltage on the output because of bias-current drops, or it can combine with stray capacitances to cause oscillations. Improper supply bypassing and capacitive loading, alone or in combination, can also result in oscillations. Many hours spent tracking down incomprehensible design problems could have been saved by monitoring the op amp output with a wide-band oscilloscope.

With low impedance loads and current transients above 10A, the inductance and resistance of wire interconnects can become important in a number of ways. Further, an IC op amp rated to dissipate 90W continuously will not do so unless it is properly mounted to an adequate heat sink.

The management and protection circuitry of the LM12 can also affect operation. Should the total supply voltage exceed ratings or drop below 15V, the op amp shuts off completely. Case temperatures above 150°C also cause complete shut

down until the temperature drops to 145˚C. This may take several seconds, depending on the thermal system. Activation of dynamic safe-area protection causes both the main feedback loop to lose control and a reduction in output drive current, with possible oscillations. In ac applications, the dynamic protection will cause waveform distortion.

supply bypassing

All op amps should have their supply leads bypassed with low-inductance capacitors having short leads and located close to the package terminals to avoid spurious oscillation problems. Power op amps require larger bypass capacitors. The LM12 is stable with good-quality electrolytic bypass capacitors greater than 20 µF. Other considerations may require larger capacitors.

The current in the supply leads is a rectified component of the load current. If adequate bypassing is not provided, this distorted signal can be fed back into internal circuitry. Low distortion at high frequencies requires that the supplies be bypassed with 470 µF or more, at the package terminals.

lead inductance

With ordinary op amps, lead-inductance problems are usually restricted to supply bypassing. Power op amps are also sensitive to inductance in the output lead, particularly with heavy capacitive loading. Feedback to the input should be taken directly from the output terminal, minimizing common inductance with the load. Sensing to a remote load must be accompanied by a high-frequency feedback path directly from the output terminal. Lead inductance can also cause voltage surges on the supplies. With long leads to the power source, energy stored in the lead inductance when the output is shorted can be dumped back into the supply bypass capacitors when the short is removed. The magnitude of this transient is reduced by increasing the size of the bypass capacitor near the IC. With 20 µF local bypass, these voltage surges are important only if the lead length exceeds a couple feet (>1 µH lead inductance). Twisting together the supply and ground leads minimizes the effect.

ground loops

With fast, high-current circuitry, all sorts of problems can arise from improper grounding. In general, difficulties can be avoided by returning all grounds separately to a common point. Sometimes this is impractical. When compromising, special attention should be paid to the ground returns for the supply bypasses, load and input signal. Ground planes also help to provide proper grounding.

Many problems unrelated to system performance can be traced to the grounding of line-operated test equipment used for system checkout. Hidden paths are particularly difficult to sort out when several pieces of test equipment are used but can be minimized by using current probes or the new isolated oscilloscope preamplifiers. Eliminating any direct ground connection between the signal generator and the oscilloscope synchronization input solves one common problem.

output clamp diodes

When a push-pull amplifier goes into power limit while driving an inductive load, the energy stored in the inductance can drive the output beyond the supplies. *Figure 1* shows the overload response of the LM12 driving ±36V at 40 Hz into a 4Ω load in series with 24 mH to illustrate the point.

The IC has internal supply-clamp diodes, but these clamps have a parasitic current that dissipates roughly half the clamp current across the total supply voltage. This dissipation cannot be controlled by the internal protection circuitry and will result in catastrophic failure if sustained. Therefore, the use of external diodes to clamp the output to the power supplies is strongly recommended.

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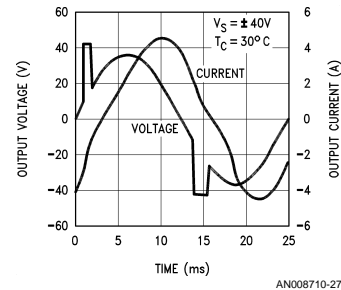
FIGURE 1. Output voltage and current waveforms with dynamic safe-area protection activated on an inductive load. Stored energy in the inductor drives the output beyond the supplies.

Experience has demonstrated that hard-wire shorting the output to the supplies can induce random failures if these external clamp diodes are not used. Therefore, it is prudent to use output clamp diodes even when the load is not obviously inductive. Failure is particularly violent when operating from low-impedance supplies: the V⁺pin can vaporize, with a hole being blown through the top of the can. If there are failures, install diodes before proceeding.

Heat sinking of the clamp diodes is usually unimportant in that they only clamp current transients. Forward drop with 15A transients is of greater concern. The clamp to the negative supply can have somewhat reduced effectiveness should the forward drop exceed 0.8V. Mounting this diode to the op amp heat sink improves the situation. Although the need has not been demonstrated, including a third diode, D_3 in *Figure 2*, will eliminate any concern about the clamp diodes. This diode, however, must be capable of dissipating continuous power as determined by the negative supply current of the op amp.

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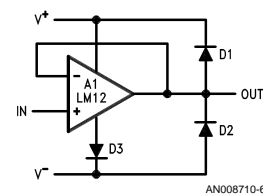
FIGURE 2. Output clamp diodes, D_1 and D_2, dump inductive-load current into the supplies when op amp goes into power limit. A third diode, D_3, may be required if the forward drop of D_2 is excessive.

reactive loading

The LM12 is normally stable with resistive, inductive or smaller capacitive loads. Larger capacitive loads interact with the open-loop output resistance (about 1Ω) to reduce the phase margin of the feedback loop, ultimately causing

oscillation. The critical capacitance depends upon the feedback applied around the amplifier; a unity-gain follower can handle about 0.01 µF, while more than 1 µF does not cause problems if the loop gain is ten. With loop gains greater than unity, a speedup capacitor across the feedback resistor will aid stability. In all cases, the op amp will behave predictably only if the supplies are properly bypassed, ground loops are controlled and high-frequency feedback is derived directly from the output terminal, as recommended earlier.

So-called capacitive loads are not always capacitive. A high-Q capacitor in combination with long leads can present a series-resonant load to the op amp. In practice, this is not usually a problem; but the situation should be kept in mind.

Large capacitive loads (including series-resonant) can be accommodated by isolating the feedback amplifier from the load as shown in *Figure 3*. The inductor gives low output impedance at lower frequencies while providing an isolating impedance at high frequencies. The resistor kills the Q of series resonant circuits formed by capacitive loads. A low inductance, carbon-composition resistor is recommended. Optimum values of L and R depend upon the feedback gain and expected nature of the load, but are not critical. A 4 µH inductor is obtained with 14 turns of number 18 wire, close spaced, around a one-inch-diameter form.

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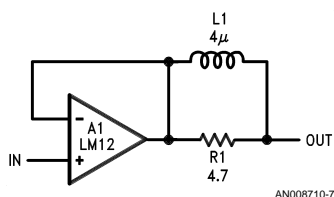
FIGURE 3. Isolating capacitive loads with an inductor. The non-inductive resistor avoids resonance problems with load capacitance by dropping Q.

The LM12 can be made stable for all loads with a large capacitor on the output, as shown in *Figure 4*. This compensation gives the lowest possible closed-loop output impedance at high frequencies and the best load-transient response. It is appropriate for such applications as voltage regulators.

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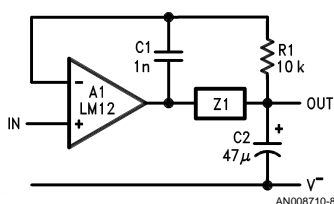
FIGURE 4. Using a large output capacitor to stabilize for all capacitive loads. The impedance, Z_1, is the wire connecting the IC output to the load-capacitor terminal.

A feedback capacitor, C_1, is connected directly to the output pin of the IC. The output capacitor, C_2, is connected at the output terminal with relatively short leads. Single-point grounding to avoid dc and ac ground loops is advised.

The impedance, Z_1, is the wire connecting the op amp output to the load capacitor. About 3 inches of number-18 wire (70 nH) gives good stability and 18-inches (400 nH) begins to

degrade load-transient response. The minimum load capacitance is 47 µF, if a plastic film or solid-tantalum capacitor with an equivalent series resistant (ESR) of 0.1Ω is used. Electrolytic capacitors work as well, although capacitance may have to be increased to 200 µF to bring ESR below 0.1Ω.

Loop stability is not the only concern when op amps are operated with reactive loads. With time-varying signals, power dissipation can also increase markedly. This is particularly true with the combination of capacitive loads and high-frequency excitation.
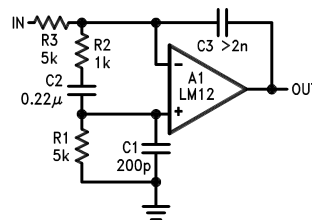
input compensation

The LM12 is prone to low-amplitude oscillation bursts coming out of saturation if the high-frequency loop gain is near unity. The voltage follower connection is most susceptible. This glitching can be eliminated at the expense of small-signal bandwidth using input compensation. Input compensation can also be used in combination with LR load isolation to improve capacitive load stability.

An example of a voltage follower with input compensation is shown in *Figure 5 a*. The R_2C_2 combination across the input works with R_1 to reduce feedback at high frequencies without greatly affecting response below 100 kHz. A lead capacitor, C_1, improves phase margin at the unity-gain crossover frequency. Proper operation requires that the output impedance of the circuitry driving the follower be well under 1 kΩ at frequencies up to a few hundred kilohertz.

Extending input compensation to the integrator connection is shown in *Figure 5 b*. Both the follower and this integrator will handle 1 µF capacitive loading without LR output isolation.


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a) follower

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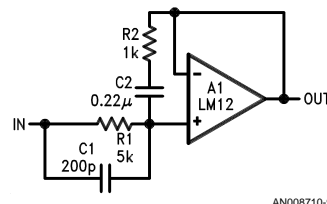
b) inverter

FIGURE 5. Using input compensation to reduce bandwidth and increase stability with capacitive loads a) for a voltage follower and b) for an integrating inverter.

parallel operation

Load current beyond the capability of one power amplifier can be obtained with parallel operation as shown in *Figure 6*. The power op amps, A_2 and A_3 are wired as followers and

connected in parallel with the outputs coupled through equalization resistors, R_4 and R_5. More output buffers, with individual equalization resistors, may be added to meet even higher drive requirements. A standard, high-voltage op amp is used to provide voltage gain. Overall feedback compensates for the voltage dropped across the equalization resistors.

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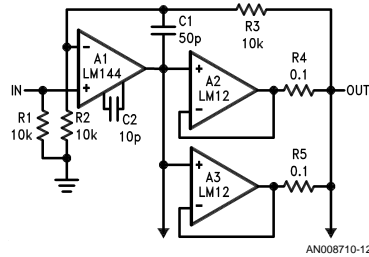
FIGURE 6. Paralleling the outputs of two op amps. The power amplifiers, A_2 and A_3, are wired as followers and connected in parallel with the outputs coupled through equalization resistors.

With parallel operation there will be an increase in unloaded supply current related to the offset voltage of A_2 and A_3 across the equalization resistors. In some cases, it may be desirable to use input compensation on the followers for increased stability. It is important that the source resistance introduced by input compensation not increase the offset voltage overmuch.

A method of paralleling op amps that does not require a separate control amplifier is shown in *Figure 7*. The output buffer, A_2, provides load current through R_5 equal to that supplied by the main amplifier, A_1, through R_4. Again, more output buffers can be added.

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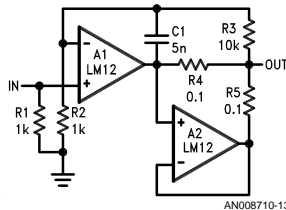
FIGURE 7. Two power op amps can be paralleled using this master/slave arrangement, but high frequency performance suffers.

The cross-supply current between the outputs of paralleled amplifiers can be affected by gain error as the power-bandwidth limit is approached. In the first circuit, the operating-current increase will depend upon the matching of the high-frequency characteristics. In the second circuit, however, the entire input error of A_2 appears across R_4 and R_5. The supply current increase can cause the power limiting to be activated as the slew limit is approached. This will not damage the LM12. It can be avoided in both cases by connecting A_1 as an inverting amplifier and restricting bandwidth with C_1.

current drive

The circuit in *Figure 8* provides an output current proportional to the input voltage. Current drive is sometimes preferred for servo motors because it aids in stabilizing servo loops by reducing phase lag caused by motor inductance. In applications requiring high output resistance, such as operational power supplies running in the current mode, matching of the feedback resistors to 0.01 percent or better is required. Alternately, an adjustable resistor, R_3, can be used for trimming. Offsetting R_3 from its optimum value will give decreasing positive or negative output resistances.

The current source input is actually differential. It can be driven as shown, or from the bottom of R_3 to obtain the opposite output sense. Both inputs should be connected to a low source impedance like ground or an op amp output. Otherwise, the source resistance will imbalance the feedback, changing output resistance. Alternately, an input can be driven by a known source resistance, like a voltage divider, if this resistance is made part of the feedback network.

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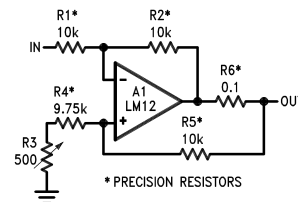
$$I_{out} = \frac{R2\, V_{in}}{R1\, R6}$$

FIGURE 8. This voltage/current converter requires excellent resistor matching or trimming to get high output resistance. Bandwidth can be reduced by the inductance of R_6.

The frequency characteristics of the current source can be expressed in terms of an equivalent output-load capacitance given by

$$C_{eq} = \frac{R_1 + R_2}{2\pi f_o R_1 R_6}, \tag{1}$$

where f_o is the extrapolated unity gain bandwidth of the op amp (in this case about 2 MHz for the LM12). The equation is only valid for $Z_L \gg R_6$.

This output capacitance can resonate with inductive loads such as motors, causing some peaking. Inductive loads can oscillate should the feedback network be imbalanced to give sufficient negative output resistance.

Inductance of the current sense resistor, R_6, can affect operation. With a 0.1Ω resistor, 3 µH series inductance will re-

duce the maximum obtainable bandwidth to 5 kHz. Proper supply bypassing and connecting R_2 directly to the output pin of the op amp are important with this circuit.

single-supply operation

Although op amps are usually operated from dual supplies, single-supply operation is practical. The bridged amplifier in *Figure 9* supplies bi-directional current drive to a servo motor while operating from a single positive supply. One op amp, A_1, is a voltage/current converter with a differential input. The second is a unity-gain inverter driven from the output of the first. It has its non-inverting input referred to half the supply voltages so that the two outputs swing symmetrically about this voltage.

Either input may be grounded, with bi-directional drive provided to the other. It is also possible to connect one input to a positive reference, with the signal to the other input varying about this voltage. If this reference voltage is above 5V, R_2 and R_3 are not required.

The output is easily converted to voltage drive by shorting R_6 and connecting R_7 to the output of A_2, rather than A_1. Although not shown, clamp diodes to V^+ and ground on the output of both amplifiers are recommended for motor loads.
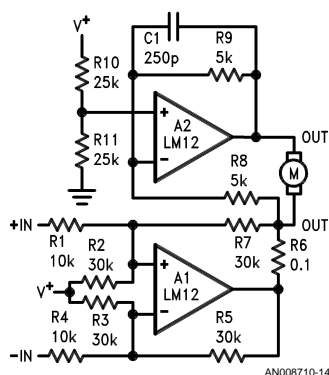

FIGURE 9. The output current of this bridged amplifier is proportional to differential input voltage. Although not shown, output clamp diodes are recommended with a motor load.

high voltage amplifiers

Using two amplifiers in a bridge connection also doubles the voltage swing delivered to the load. The configuration in *Figure 10* gives good results with split supplies. One op amp is an inverting amplifier while the other is a non-inverting amplifier with equal gain. A load connected between the outputs sees twice the swing of either amplifier. Understandably, the output slew rate doubles while the full-power bandwidth stays the same.

The current limit of two op amps cannot be expected to be the same. Therefore, a short between the outputs of a bridge amplifier can result in one amplifier saturating while the output transistor of the second handles the overload at the full supply voltage. Not all power amplifiers can take this kind of treatment; the LM12 will.
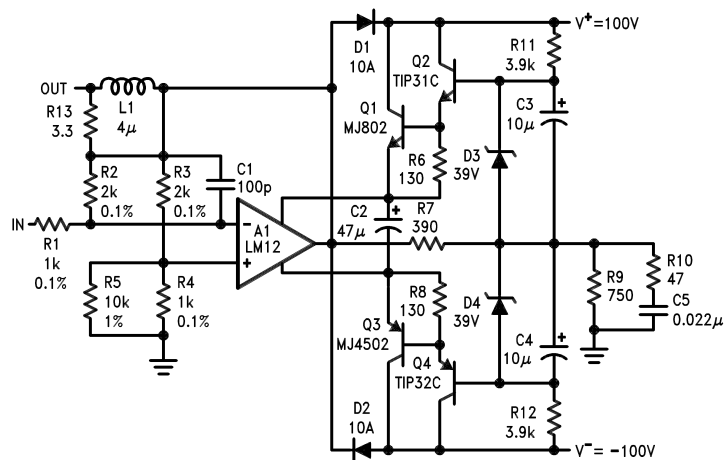

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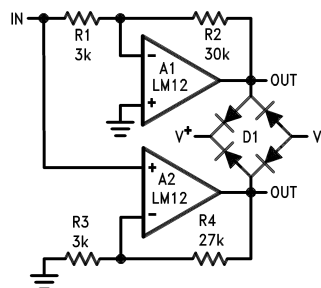
FIGURE 10. Bridge connection gives differential output approaching twice the total supply voltage. Diode bridge clamps outputs to supplies.

Figure 10 shows how a bridge-rectifier module can be used to provide output clamping for both outputs.

The LM12 can be operated in cascode with external transistors to get output swings several times higher than the basic op amp. The design in *Figure 11* drives ±90V at ±10A. Significantly, the IC provides current and power limiting for the external transistors.

The transistors and zener diodes form a simple voltage regulator that is driven at 70 percent of the output swing from the R_7/R_9 divider. Thus, the total supply voltage of the IC stays constant while the voltage to ground swings some ±60V.

The supply terminals of the LM12 swing both above and below ground at full output. Therefore, the input terminals must be bootstrapped to the output to keep them within the common-mode range. The R_1–R_4 bridge does this. The bridge is unbalanced by R_5 to set the gain near 30. Naturally, R_4 and R_5 can be combined.

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FIGURE 11. This amplifier can drive ±90V at ±10A, more than twice the output swing of the LM12. The IC provides current and power limiting for the discrete transistors.

Bootstrapping the power supplies reduces the voltage swing across the internal frequency compensation capacitors of the LM12. The effectiveness of the capacitors is in proportion to the output swing across them. If the voltage swing between the output and V⁻ terminals of the IC is one-third the actual output swing, the slew rate and gain-bandwidth product of the complete amplifier will be three times that listed for the IC. The minimum loop gain must be increased accordingly.

Distortion on the bootstrapped supplies can show up on the output because the op amp has limited supply rejection at high frequencies. If R_6 and R_8 are not low enough, these power followers cannot track high frequency waveforms and performance will suffer.

This circuit is more sensitive to capacitive loading than the basic op amp because the supply terminals of the IC cannot be bypassed directly to ground. The effects of this can be mitigated by using an appropriate LR network in the output.

When the IC goes into power limit, current will likewise be cut back in the external transistors. The voltage on these external transistors is not necessarily regulated, so the discrete transistors must be enough stronger than the IC transistors to handle the extra voltage. The IC can handle orders of magnitude more power cycling than commercial power transistors with soft-solder die attach. Cycling in and out of power limit at low frequencies could be a problem and should not be ignored.

The output swing can be increased using more-conventional circuitry if floating supplies are available. *Figure 12* shows a bridged amplifier that drives a ground-referred load. A differential input is provided, but one input can be grounded and the other driven from a low-impedance source. If the non-inverting input is grounded, R_7 and R_8 can be replaced

by a single resistor. Operation is like a standard bridge, except that it is a bit more sensitive to capacitive loading. Output swing is ±70V at ±10A.
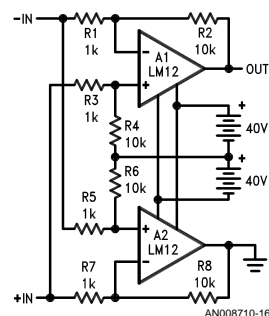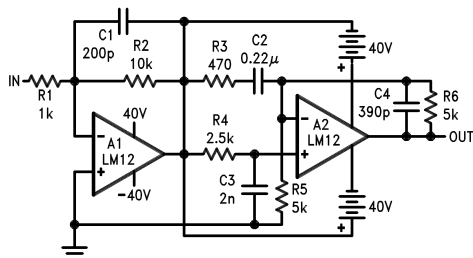

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FIGURE 12. Bridge amplifier with a single-ended output uses floating supply. Either input can be grounded.

A final circuit in *Figure 13* shows how two op amps can be stacked to double output swing. A third op amp with a gain of 0.5 added to the output will triple the basic swing. Any number of op amps can be cascaded, adding to the swing, but a floating supply is required for each.

With two stages, clamp diodes from each amplifier output to its supply terminals are recommended. With three or more stages, the diodes are required to avoid supply reversals.

The bandwidth is limited by R_4 and C_3. This isolation also prevents load transients on the output, reflected back to the

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output of A_1, from regenerating (this really shows up as capacitive load sensitivity). Like the other designs, an output LR will help reduce sensitivity to capacitive loading.
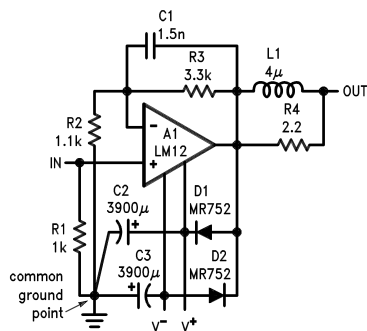

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FIGURE 13. Cascading two op amps doubles output swing. Output may be increased by any number of stages, but a separate floating supply is required for each.

audio amplifiers

High quality audio amplifiers are wideband, power op amps with tight distortion specifications. The performance of the LM12 puts it in this class.

A practical design for an audio power amplifier is shown in *Figure 14*. Output-clamp diodes are mandatory because loudspeakers are inductive loads. Output LR isolation is also used because audio amplifiers are usually expected to handle up to 2 µF load capacitance. Large, supply-bypass capacitors located close to the IC are used so that the rectified load current in the supply leads does not get back into the amplifier, increasing high-frequency distortion. Single-point grounding for all internal leads plus the signal source and load is recommended to avoid ground loops that can increase distortion.
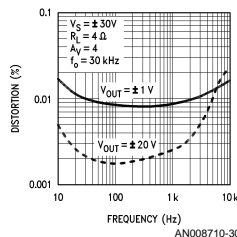

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FIGURE 14. As an audio amplifier, the LM12 has better distortion, transient response and saturation recovery than most power op amps.

The total harmonic distortion measured for this circuit is plotted in *Figure 15*. The increase at high frequencies is due to crossover distortion of the class-B stage. That at low frequencies is caused by thermal feedback within the LM12.

The effect of thermal feedback on the response of the LM12 is indicated in *Figure 16*. The offset voltage change is plotted as a function of time after the application of an output load that dissipates 50W in the source and sink transistors.

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FIGURE 15. Total harmonic distortion of the circuit in *Figure 14* **is plotted here for both low- and high-level outputs.**
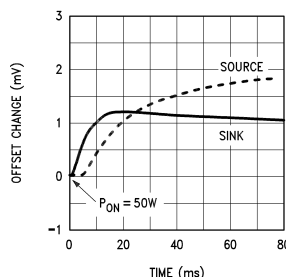
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FIGURE 16. The offset voltage change after the application of a load that dissipates 50W in each output transistor is plotted here. This thermal feedback causes increased distortion below 100 Hz.

Unlike crossover distortion, the low-frequency distortion can be virtually eliminated by using the LM12 as a buffer inside the feedback loop of a low-level op amp. However, the low-frequency harmonic distortion, being generated thermally, is slow and does not cause the more objectionable intermodulation distortion. The latter measured 0.015 percent with ±10V into a 4Ω load under the standard 60 Hz/7 kHz, 4:1 test conditions.

The transient response of the circuit in *Figure 14* is clean; and saturation characteristics are glitch-free even at high frequencies. In addition, the 9V/µs slew rate of the LM12 virtually eliminates transient intermodulation distortion.

The availability of a low-cost power amplifier that is suitable as a high-quality audio amplifier can be expected to generate interest in using a separate amplifier to drive each speaker. Not only does this eliminate high-level crossover networks and attenuators, but also it prevents overloading at low frequencies from causing intermodulation distortion at high frequencies. With separate amplifiers, such clipping is far less noticeable.

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servo amplifiers

When making servo systems with a good power op amp, there is a temptation to use it for frequency shaping to stabilize the servo loop. Sometimes this works; other times there are better ways; and occasionally it just doesn't fly. Usually it's a matter of how quickly and to what accuracy the servo must stabilize. A couple of examples should make the point.

With fast motor-driven servos, it is best to make the motor current proportional to the servo amplifier drive. With current drive, motor response is basically unaffected by the series inductance of the motor windings. At higher frequencies, current drive can give 90 degrees less phase shift in the motor transfer function when compared to voltage drive. Should the servo loop go through unity gain at a frequency at which motor inductance is unimportant, the advantage of current drive is lost.

The motor/tachometer speed control shown in *Figure 17* gives an example of optimizing performance using a current drive that is supplied by A_2, connected as a voltage/current converter. The tachometer, on the same shaft as the dc motor, is simply a generator. It gives a dc output voltage proportional to the speed of the motor. A summing amplifier, A_1, controls its output so that the tachometer voltage equals the input voltage, but of opposite sign.

With current drive to the motor, phase lag to the tachometer is 90 degrees, before second order effects come in. Compensation on A_1 is designed to give less than 90 degrees phase shift over the range of frequencies where the servo loop goes through unity gain. Should response time be of less concern, a power op amp could be substituted for A_1 to drive the motor directly. Lowering break frequencies of the compensation would, of course, be necessary.

The circuit in *Figure 17* could also be used as a position servo. All that is needed is a voltage indicating the sense and magnitude of the motor shaft displacement from a desired position. This error signal is connected to the input, and the servo works to make it zero. The tachometer is still required to develop a phase-correcting rate signal because the error signal lags the motor drive by 180 degrees.
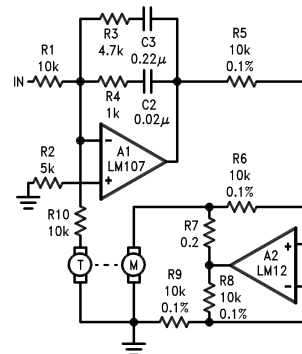
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FIGURE 17. Motor/tachometer servo gives an output speed proportional to input voltage. Using current drive to motor reduces loop phase shift due to motor inductance.

The concept of a rate signal can be understood from a simple example. The problem is to rotate a radar antenna to acquire a target from a large angle off point. When the motor has limited power and the antenna has mass, the quickest path into point is to run full bore toward point; pick the correct instant to reverse at full power before getting there; and shut down in just the right place. In a servo, the rate signal added to the error signal is what tells it when to reverse in order to acquire the target without overshooting.

With a fixed target, a tachometer on the drive motor will give the rate signal. If the target is moving across the antenna, it does not: it produces the rate signal plus or minus the angular velocity of the target. This disrupts acquisition and generates a pointing error.

The rate signal can be obtained by differentiating the error signal. A design that gives the required error plus rate signal

at the output is shown in *Figure 18 a*. Neither op amp should saturate under any condition, no matter how far off point or how fast the error changes. If it saturates, a proper rate signal is not developed; and acquisition will be degraded. This can degenerate to where the servo will oscillate continuously once a certain tracking error is exceeded.

Acquiring from large errors quickly and to great accuracy requires an extremely wide dynamic range. In *Figure 18a*, it is necessary to make R_1 and R_3 so low to keep the amplifiers from saturating that chopper stabilization may be required to preserve accuracy.

In *Figure 18a*, R_3 can be raised to any value if back-to-back zeners are put across it. The waveform below the clamp level will be unchanged from the case where A_2 has unbounded output swing. Should the clamp levels be large enough to saturate the motor drive, operation is unimpaired.

This principle is developed further in *Figure 18b*. It gives identical response, except that the resistor in series with C_1 breaks back the differentiator above the unity gain frequency. Off point, the voltage at the junction of R_1 and R_2 should not get so large that the output of A_1 cannot saturate A_2 without the clamps conducting.

operational power supply

External current limit can be provided for an op amp as shown in *Figure 19*. The positive and negative limiting currents can be set precisely and independently down to zero with potentiometers R_3 and R_7. Alternately, the limit can be programmed from a voltage supplied to R_2 and R_6. The input controls the output when not in current limit. This is just the set-up required for an operational power supply or voltage-programmable power source.
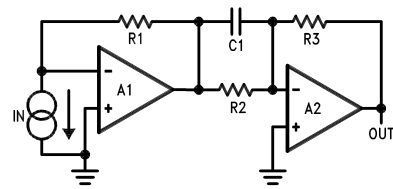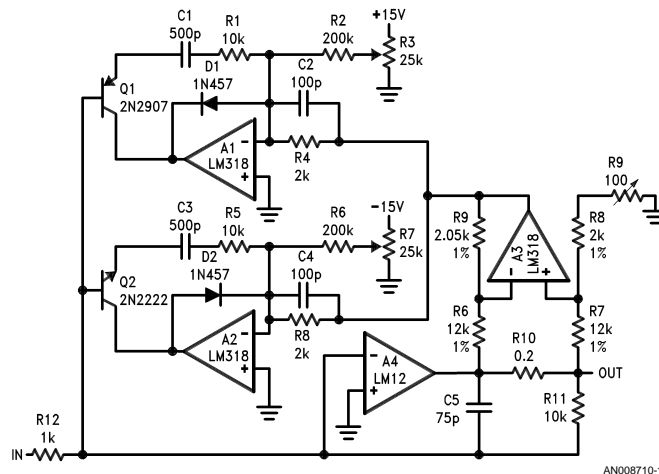

a) generating rate signal electrically
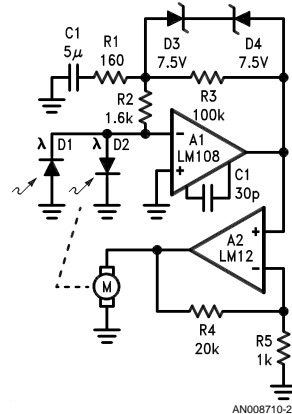
b) compressing dynamic range

FIGURE 18. When electrical rate signals must be developed with large error signals well beyond saturation of motor drive, a linear approach a) requires wide dynamic range and great precision. More practical design b) uses feedback clamps to increase effective dynamic range.

Note: Supply voltages for the LM318s are ±15V

FIGURE 19. Bi-directional limiting currents of the power op amp, A_4, are set independently by R_3 and R_7. Fast response is ensured by clamp diodes, D_1 and D_2.

The power op amp, A_4, is connected as an inverting amplifier. Its output current is sensed across R_{10}. This sense voltage is level shifted to ground by A_3, a differential amplifier that is made insensitive to the op amp output level by trimming R_9.

With current below preset levels, the outputs of A_1 and A_2 are clamped by D_1 and D_2 with Q_1 and Q_2 turned off. When the current threshold is reached, the relevant amplifier will come out of clamp, saturate the transistor on its output and take over control of the summing node. The clamp diodes limit the swing on the outputs of the current-control amplifiers while the transistors disconnect frequency compensation until the summing node is engaged. This ensures fast activation of current limit. Recovery back to voltage mode is also fast. The LM318 wideband amplifier is required for A_1 through A_3.
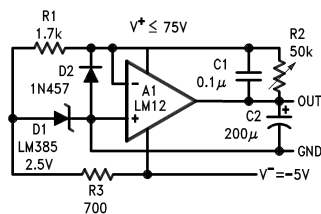
voltage regulators

An op amp can be used as a positive or negative regulator with equal ease. Unlike most dedicated voltage regulators, the output can both source or sink to absorb energy dumped back into the supply and prevent overvoltage with certain fault conditions. Output transient response is also improved, especially overshoots.

A particular reason for using the LM12 as a regulator is its exceptional high-voltage capability. This not only gives output voltages to 70V, but also ensures startup under worst-case full-load conditions.

Compared to conventional IC regulators, using an op amp with an external reference has better accuracy: an optimum reference can be selected and thermally isolated from the power circuitry. Better regulation, temperature drift and long term stability result. Remote, output-voltage sensing at the load to further reduce errors is also practical.

A positive regulator with a 0–70V output range is shown in *Figure 20*. The op amp has one input at ground and a reference current drawn from its summing junction. With this arrangement, output voltage is proportional to the setting resistor, R_2.

FIGURE 20. Positive regulator with 0–70V output range. Output will source or sink current, and start-up capability with higher input voltages is superior to standard IC regulators.

A negative supply is used to operate the op amp within its common-mode range, provide zero output with sink current and power a low-voltage bandgap reference, D_1. Current drawn from this supply is under 150 mA, except when sinking load current.

The output load capacitor, C_2, is part of the op amp frequency compensation. This requires that C_1 be connected directly at the op amp output and C_2 at the load, as de-

scribed earlier. The reference noise is filtered by C_1, which also controls the start-up rate. The clamp diode, D_2, resets C_1 when the output is shorted and keeps the op amp input from being driven below V^-.

Dual supplies are not required to use an op amp for a regulator, as can be seen from the 4V to 70V adjustable regulator shown in *Figure 21*. This regulator also has overvoltage protection. Should an overvoltage condition exceed the current or power capabilities of the LM12, a comparator will trigger a SCR, crowbarring the output.
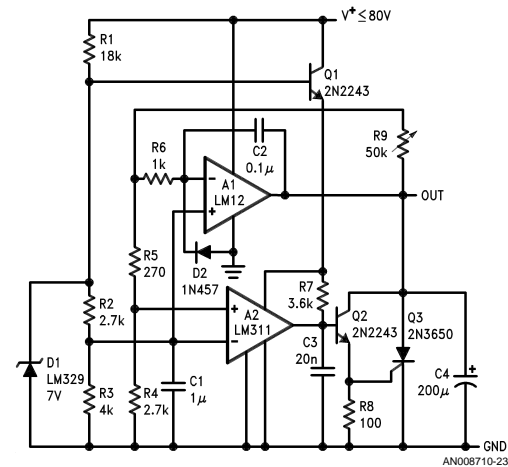
FIGURE 21. This 4V to 70V regulator operates from a single supply. Should the op amp not be able to control an overvoltage condition, the SCR will crowbar the output.

The reference is a low drift zener, D_1, powered from V^+ through R_1. The reference voltage is dropped to 4V and fed to the non-inverting input of the op amp, A_1, with zener noise attenuated by C_1. Thus, the output will be this 4V plus a voltage that is proportional to the resistance of R_9. As before, D_2 is a clamp while C_2 makes sure the IC input is ac coupled directly to its output terminal.

With overvoltage, a comparator, A_2, fires the SCR through a buffer, Q_2, after about a 20 μs delay from C_3 to eliminate spurious transients. The comparator receives its power from Q_1 so that V^+ can be increased above the rating of the LM311.

Should the feedback terminal of the op amp rise more than 0.4V above the regulating value for longer than 20 μs, the comparator will provide the signal required to fire the SCR. Since this can only happen if the considerable current and energy capabilities of the LM12 are exhausted, nuisance tripping is unlikely. The output trip threshold will be 0.4V above nominal as long as it happens quickly enough that the voltage across C_2 does not change appreciably. For a slow overvoltage condition, it is 10 percent above nominal.

remote sensing

With the current running at 10A, a foot of 0.1-inch-diameter copper drops 10 mV. Obviously, sizeable voltage drops will have to be accepted to run this kind of current over any distance without expensive and cumbersome cables.

Remote sensing, illustrated in *Figure 22*, can help the situation considerably. It uses a pair of small wires, in addition to the main power cables, to sense the voltage at the load. A feedback amplifier can then correct for the drop in the main cables.

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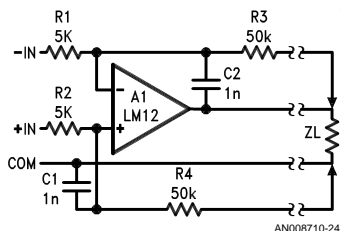
FIGURE 22. Remote sensing allows the op amp to correct for dc drops in cables connecting the load. Normally, common and one input are hooked together at the sending end.

The cables can cause delays in the feedback signal returning from the remote sense. This delay can make the feedback loop unstable unless the remote signal is ignored at higher frequencies. Thus, cable drop can be compensated but only at a limited rate; transient response suffers most. Heavy cables, closely spaced (or twisted) to minimize inductance, give fewest problems here.

The schematic in *Figure 22* shows a differential-input amplifier that has dc feedback from the remotely-sensed load. The ac feedback is directly from the op amp output and the signal common at the sending end. There is no feedback from the load at high frequencies. The optimum capacitance depends upon the cable delay.

For single-ended input, the unused input terminal in the schematic would be strapped to the common. Feedback resistors should be reasonably matched to avoid second-order errors and the feedback resistors should be made enough greater than the sense line resistance to avoid gain errors.

Sometimes provision is made to control the circuit should the sense lines be disconnected. With a regulator, an imbalance current could be put into the sense lines to bring the regulator output to zero should one line go open. With bi-directional op amps, it is not obvious whether limiting the error with back-to-back diodes between the power-out and sense-in is better than having it go open loop.
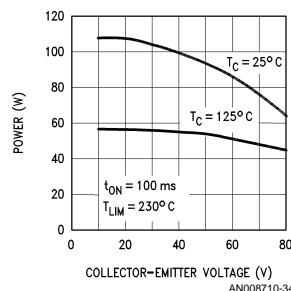
power capabilities

The output transistors of the LM12 will dissipate power until their peak junction temperature reaches 230˚C (±15˚C). When this temperature is reached, internal limiting circuitry takes over to regulate peak temperature. How this works is illustrated in *Figure 23*, which gives the peak output current waveform with the output instantaneously shorted to ground. Conventional current limiting holds the short-circuit current near 13A for a few hundred microseconds, then temperature limiting takes over as junction temperature tries to rise above 230˚C. The response time of the temperature limiter is well under 100 μs.

With this type of protection, the power capabilities will depend on case temperature, transistor operating voltages and how the dissipation varies with time. *Figure 24* shows the amplitude of a power pulse required to activate power limiting in 100 ms as a function of collector-emitter voltage on the output transistors for two case temperatures. The continuous dissipation limit is about 15 percent less than the 100 ms limit.

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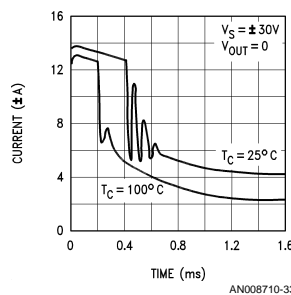
FIGURE 23. Output short-circuit current is reduced when power transistor junction temperature reaches 230˚C and power limit takes over.

AN008710-34

FIGURE 24. The power required to activate power limit is less at higher voltage, but this is not so pronounced at higher case temperatures.

11

The pulse capabilities of the output transistors are shown in *Figure 25*. The curves give the amplitude of a constant-power pulse required to activate power limiting in the indicated time. With pulse widths longer than 1 ms, the pulse capability decreases with collector voltages above 40V as indicated in *Figure 24*.
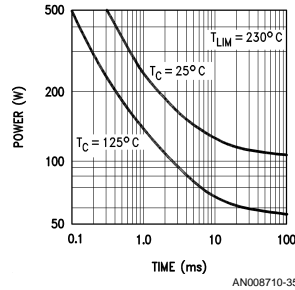

FIGURE 25. The peak-dissipation capabilities of the power transistor are shown here. For times greater then 100 ms, the external heat sink will determine ratings.

a) safe area curve

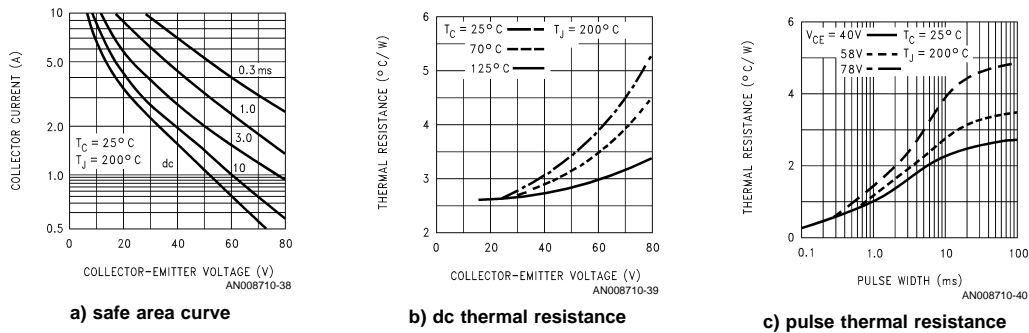**b) dc thermal resistance**

c) pulse thermal resistance

FIGURE 26. The worst-case power ratings of the output transistors are described by a) a safe area curve, b) the change in dc thermal resistance with temperature and operating voltage and c) the pulse thermal resistance.

power ratings

The guaranteed power ratings of the LM12 are based on a peak junction temperature of 200˚C rather than the 230˚C limiting temperature. Test accuracy, guard bands and unit-to-unit variations are also taken into account. The result is that the guaranteed ratings are about 40 percent less than the power required to activate thermal limit.

The worst-case, safe-area curves for a peak junction temperature of 200˚C with a 25˚C case temperature are shown in *Figure 26 a*. The guaranteed-maximum, dc thermal resistance is given as a function of collector-emitter voltage in *Figure 26b*. It can be seen from this figure that the increase in thermal resistance with voltage is much less at higher case temperatures. Finally, the equivalent thermal resistance for power pulses is given in *Figure 26c*. Again, these are worst-case numbers. The voltage dependency of thermal resistance in *Figure 26c* is for a 25˚C case temperature. At higher case temperatures this dependency will be moderated as shown in *Figure 26b*.

The guaranteed power ratings are not established by statistical methods from sample tests. Instead, they are interpolated from actual measurements of power capability into thermal limit: these are standard production tests.

With ac loading, both power transistors share the dissipation; and the worst-case thermal resistance can drop to 1.9˚C/W. However, it is necessary that the frequency be sufficiently high that the peak ratings of neither output transistor are exceeded.

thermal derating

It is not unusual to derate the maximum junction temperature of semiconductors below the manufacturer's specified value in worst-case design. The derating is often dictated by unpredictable operating conditions and design uncertainties. An equipment manufacturer does not want his product failing because of some obscure stress that is not apparent to the customer.

Company policies, equipment requirements and individual preferences vary as to what constitutes appropriate derating. When pressure is on for the best performance at lowest cost, a 200˚C junction temperature for power semiconductors has been accepted, although this might well be influenced by whether hermetic or plastic packaging is used. Continuous operation at 200˚C should also be treated differently than infrequent excursions to this temperature. Nonetheless, reducing temperature is a recognized method for increasing reliability; and ultra-reliable military and space applications have required that maximum junction temperatures be under 125˚C.

The protection circuitry of the LM12 brings a new dimension to derating. Such conditions as out-of-spec line voltage or lack of air circulation cause the equipment to stop working temporarily; excessive stress or catastrophic failure does not result. It should be recognized, however, that there are certain applications where a temporary misfunction can be the same as a permanent one, definitely recommending derating.

Derating also reflects the user's faith in the ability of the manufacturer to adequately test the parts. Dynamic safe-area protection helps out here. Should a die-attach void or other defect produce hot spots in the power transistor, it will be rejected during production testing as having reduced dissipation capability, rather than being passed on as a reliability risk. This cannot be done with conventional power semiconductors.

No short-term failure mode has been found with modern IC power transistors even with peak junction temperatures of 300˚C. However, power cycling can cause problems. Die-attach failures at 3×10^4 cycles with a 70˚C temperature rise are possible with power transistors having a soft-solder die attach. The LM12 avoids this by using a gold-eutectic die attach to a molybdenum spacer. Even so, metalization failures have been experienced with the LM12 at 10^6 cycles from 50˚C to power limit at 230˚C with 200W dissipation.

Thermal derating is more applicable to the control circuitry of the LM12. Operating the control circuitry above 150˚C can be expected to affect reliability. Fortunately, the control circuitry is exposed to only a fraction of the temperature rise in the power transistor. Derating may be based on a thermal resistance of 0.9˚C/W independent of operating voltage. With ac loading, where power is being dissipated in both power transistors, this thermal resistance drops, finally approaching 0.6˚C/W.

package mounting

The ratings of the LM12 are based on the case temperature as measured on the bottom of the TO-3 package near the center. Proper mounting is required to minimize the thermal resistance between this region and the heat sink.

A good thermal compound such as Wakefield type 120 or Thermalloy Thermacote should be used when mounting the package directly to the heat sink. Without this compound, thermal resistance will be no better than 0.5˚C/W, and possibly much worse. With the compound, thermal resistance will be 0.2˚C/W or less, assuming under 0.005-inch combined flatness run-out for the package and the heat sink. Proper torquing of the mounting bolts is important. Four to six inch-pounds is recommended.

Should it be necessary to isolate V⁻ from the heat sink, an insulating washer is required. Hard washers like beryllium oxide, anodized aluminum and mica require the use of thermal compound on both faces. Two-mil mica washers are most common, giving about 0.4˚C/W interface resistance with the compound. Silicone-rubber washers are also available. A 0.5˚C/W thermal resistance is claimed without thermal compound. Experience has shown that these rubber washers deteriorate and must be replaced should the IC be dismounted.
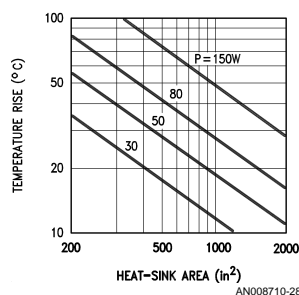
heat sinking

With no heat sink, the internal temperature rise of the LM12 can be as high as 160˚C with ±40V supplies and no load. A heat sink is required. Heat sinks are commercially available,

with data on their power rating and temperature rise supplied by the manufacturer. The types most suitable for dissipation in the order of 50W are made from extruded aluminum channel equipped with multiple fins. It is important that the heat sink have enough metal under the package to conduct heat from the center of the package bottom to the fins without excessive temperature drop.

The power rating of a multi-finned heat sink is determined largely by the surface area subject to convection cooling and the allowable temperature rise above ambient. Heat loss due to radiation can also be important with simple heat sinks. However, with multiple fins radiating toward each other, the significance of the radiation term drops. Nonetheless, heat sinks are usually black anodized to maximize radiation losses.

The surface area required for a given temperature rise and power dissipation can be estimated with fair accuracy from *Figure 27*. The area efficiency is affected by heat sink orientation, length and fin spacing. The figure assumes that the surfaces are located in a vertical plane. With the surfaces horizontal, temperature rise is increased by perhaps 20 percent. Vertical dimensions longer than 4 inches are less efficient. Commercial heat sinks are normally designed so that fin spacing is not so close as to affect the results of the figure.

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FIGURE 27. A heat sink is required to cool the IC package. This curve gives the rise in case temperature as a function of heat-sink fin area with convection cooling.

It is not possible to specify an unqualified thermal resistance for a convection or radiation cooled heat sink. Both mechanisms will give a lower thermal resistance with increasing temperature rise, while heat losses to radiation also increase with absolute temperature. Since radiation losses are not dominant with multi-finned heat sinks, power dissipation and temperature rise should characterize performance. Heat sink size can be drastically reduced by forced air cooling, should it be available.

determining dissipation

It is a simple matter to establish the power that an op amp must dissipate when driving a resistive load at frequencies well below 10 Hz. Maximum dissipation occurs when the output is at one-half the supply voltage with high-line conditions. The individual output transistors must be able to handle this power continuously at the maximum expected case temperature.

If there is ripple on the supply bus, it is valid to use the average value in worst-case calculations as long as the peak rat-

ing of the power transistor is not exceeded at the ripple peak. With 120 Hz ripple, the peak rating is 1.5 times the continuous power rating.

Dissipation requirements are not so easily established with time-varying output signals, especially with reactive loads. Both peak- and continuous-dissipation ratings must be taken into account, and these depend on the signal waveform as well as load characteristics.

With a sine wave output, analysis is fairly straightforward. With supply voltages of $\pm V_S$, the maximum average power dissipation of both output transistors is

$$P_{MAX} = \frac{2V_S^2}{\pi^2 Z_L \cos \theta}, \quad \theta < 40°;$$

(2)

and

$$P_{MAX} = \frac{V_S^2}{2Z_L}\left[\frac{4}{\pi} - \cos\theta\right], \quad \theta \geq 40°,$$

(3)

where Z_L is the magnitude of the load impedance and θ its phase angle. Maximum average dissipation occurs for a peak output swing, E_P, given by

$$E_P = \frac{2V_S}{\pi \cos \theta}, \quad \cos \theta > \frac{2V_S}{\pi V_P};$$

(4)

or

$$E_P = V_P, \quad \cos \theta \leq \frac{2V_S}{\pi V_P};$$

(5)

where V_P is the maximum available output swing.
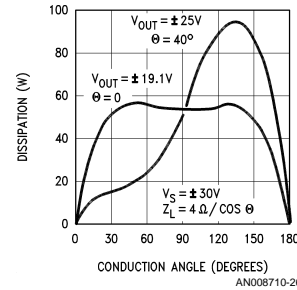
The instantaneous power dissipation is

$$P = \frac{E_P}{Z_L} \cos \omega t \left[V_S - E_P \cos (\omega t + \theta)\right].$$

(6)

For $E_P = V_S$, the power peak occurs for $\omega t = \frac{1}{3} (\pi - \theta)$.

With practical amplifiers $E_P < V_S$, and a numerical solution is required.

The instantaneous power dissipation over the conducting half cycle of one output transistor is shown in *Figure 28*. Power dissipation is near zero on the other half cycle. The output level is that resulting in maximum peak and average dissipation. Plots are given for a resistive and a series R-L load. The latter is representative of a 4Ω loudspeaker operating below resonance and would be the worst-case condi-

tion in most audio applications. The peak dissipation of each transistor is about four times average. In ac applications, power capability is often limited by the peak ratings of the power transistor.

AN008710-26

FIGURE 28. Instantaneous power dissipation of one output transistor over its conducting half cycle with a resistive and an inductive load.

The pulse thermal resistance of the LM12 is specified for constant power-pulse duration. Establishing an exact equivalency between constant-power pulses and those encountered in practice is not easy. However, for sine waves, reasonable estimates can be made at any frequency by assuming a constant power pulse amplitude given by:

$$P_{PK} \cong \frac{V_S^2}{2Z_L}\left[1 - \cos (\phi - \theta)\right],$$

(7)

where $\phi = 60°$ and θ is the absolute value of the phase angle of Z_L. Equivalent pulse width is $t_{ON} \cong 0.4\tau$ for $\theta = 0$ and $t_{ON} \cong 0.2\tau$ for $\theta > 20°$, where τ is the period of the output waveform.

With the LM12, the peak junction-temperature rise for any given waveform can actually be measured. This is done by raising case temperature until power limiting is activated. Temperature rise is then computed from the measured case temperature based on a power-limit temperature of 230°C. Alternately, the power-limit temperature may be determined directly by measuring the dc dissipation at a collector-emitter voltage of 20V required to activate power limit. If this is done over a range of case temperatures, the results can be plotted and extrapolated to the power-limit temperature.

This procedure was used to give the peak temperature rise as a function of frequency for an op amp driving a resistive load under conditions of worst-case dissipation. The results are plotted in *Figure 29*.

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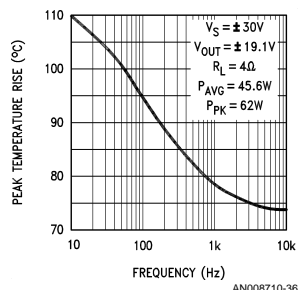
FIGURE 29. Peak junction-temperature rise as a function of frequency with an op amp driving a resistive load under conditions of worst-case dissipation.

dissipation driving motors

A motor with a locked rotor looks like an inductance in series with a resistance, for purposes of determining driver dissipation. With slow-response servos, the maximum signal amplitude at frequencies where motor inductance is significant can be so small that motor inductance does not have to be taken into account. If this is the case, the motor can be treated as a simple, resistive load as long as the rotor speed is low enough that the back emf is small by comparison to the supply voltage of the driver transistor.

A permanent-magnet motor can build up a back emf that is equal to the output swing of the op amp driving it. Reversing this motor from full speed requires the output drive transistor to operate, initially, along a loadline based upon the motor resistance and total supply voltage. Worst case, this loadline will have to be within the continuous dissipation rating of the drive transistor; but system dynamics may permit taking advantage of the higher pulse ratings. Motor inductance can cause added stress if system response is fast.

Shunt-and series-wound motors can generate back emfs that are considerably more than the total supply voltage, resulting in even higher peak dissipation than a permanent-magnet motor having the same locked-rotor resistance.

voltage regulator dissipation

The pass transistor dissipation of a voltage regulator is easily determined in the operating mode. Maximum continuous dissipation occurs with high line voltage and maximum load current. As discussed earlier, ripple voltage can be averaged if peak ratings are not exceeded; however, a higher average voltage will be required to ensure that the pass transistor does not saturate at the ripple minimum.

Conditions during start-up can be more complex. If the input voltage increases slowly such that the regulator does not go into current limit charging output capacitance, there are no problems. If not, load capacitance and load characteristics must be taken into account. This is also the case if automatic restart is required in recovering from overloads.

Automatic restart or start-up with fast-rising input voltages cannot be guaranteed unless the continuous dissipation rating of the pass transistor is adequate to supply the load current continuously at all voltages below the regulated output voltage. In this regard, the LM12 performs much better than IC regulators using foldback current limit, especially with high-line input voltages above 20V.

power supplies

Power op amps do not require regulated supplies. However, the worst-case output power is determined by the low-line supply voltage in the ripple trough. The worst-case power dissipation is established by the average supply voltage with high-line conditions. The loss in power output that can be guaranteed is the square of the ratio of these two voltages.

Relatively simple off-line switching power supplies can provide voltage conversion, line isolation and 5-percent regulation while reducing size and weight. The regulation against ripple and line variations may provide a substantial increase in the power output that can be guaranteed under worst-case conditions. In addition, switching power supplies can convert low-voltage power sources such as automotive batteries up to regulated, dual, high-voltage supplies optimized for powering power op amps.

checking thermal design

Thermal design margins can be established by determining how far the part can be pushed beyond nominal worst-case before power limiting is activated. This extra stress can be applied by increasing case temperature, supply voltage or output loading.

Raising case temperature with worst-case electrical conditions has the advantage of giving results that are easily interpreted in terms of thermal design margins. If the case temperature, as measured at the center of the package bottom, must be raised 50°C above the maximum design value to activate power limiting, the worst-case, peak-junction temperature is 180°C, or 50°C below the power-limit temperature.

With this technique, it is important that the case temperature be kept below 140°C. At 150°C, the case-temperature limit activates, shutting the IC down completely.

conclusions

A new concept for power limiting and advances in IC design have been combined to produce a monolithic, high-power op amp that challenges the best hybrid and discrete designs. Impressive power ratings are obtained along with better control of fault conditions. The part is easy to use, quite tolerant of abuse and has few disagreeable characteristics.

Design problems peculiar to power op amps have been discussed. They present no serious difficulty if kept in mind. Methods of increasing output capabilities beyond those of the basic part were also shown to demonstrate its flexibility. A number of conventional applications for power op amps were detailed along with others that would not normally use an op amp. It is expected that this power IC will recommend itself for a wide range of currently obscure, general-purpose applications.

One of the more onerous tasks with power devices is establishing that a design operates components within ratings. Guidelines were given for determining continuous and transient dissipation. In addition, the basic problems of providing adequate heat sinking were outlined. Dynamic safe-area protection is of particular help here in that design margins can actually be measured, rather than inferred from published data.

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2. A critical component in any component of a life support device or system whose failure to perform can be reasonably expected to cause the failure of the life support device or system, or to affect its safety or effectiveness.

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