Low Distortion Wideband Power Op Amp

National Semiconductor
Application Note 261

The LH0101 is a power operational amplifier capable of delivering a high-current low-distortion output. The device is conservatively rated at 2 amps continuous current. A novel design technique is used to eliminate the crossover distortion often plaguing power op amps. Additional features include a frequency response from DC to greater than 4 MHz. Excellent DC performance is attained by using FET input devices, and the unity gain frequency compensation has been performed internally. Finally, the device is hermetically sealed in a standard 8-pin TO-3 power package.

The initial LH0101 design goal was to develop an easy to use wideband operational amplifier capable of driving a variety of loads. This requirement focused a major portion of the design effort in the power output stage where considerable emphasis was placed on eliminating crossover distortion. Another consideration was to remove the ground connection typically associated with power amplifiers in order to ease the usage with single or dual power supply configurations.

Our discussion is sectioned into three subtopics where the first details the LH0101 internal circuitry, the second presents a variety of user product development/design precautions, and the third presents typical applications for the LH0101, supported by circuit diagrams.

TABLE I. LH0101 Typical Performance Characteristics at 25°C Ambient, ±15V Supply

Parameter	Conditions	Value
Output Current		2A
Input Offset Voltage		5 mV
Input Bias Current		50 pA
Input Offset Current		25 pA
Input Resistance		$10^{12}\Omega$
Large Signal Voltage Gain		200V/mV
Output Voltage Swing	$R_L = 100\Omega$	±12.5V
	$R_L = 10\Omega$	±11.6V
	$R_L = 5\Omega$	±11V
Slew Rate	$A_V = +1$	10/ $V\mu s$
Full Power Bandwidth	$A_V = +1, R_L = 10\Omega$	300 kHz
Small Signal Rise Time	$A_V = +1, R_L = 10\Omega$	10 ns
Small Signal Settling Time to 0.01%	$V_{IN} = 10V, A_V = +1$	2 μs
Gain Bandwidth		4 MHz
Harmonic Distortion	f = 1 kHz, P_O = 1W	0.005%
	$R_L = 10\Omega, A_V = +1$	
	f = 20 kHz, P_O = 1W	0.05%
	$R_L = 10\Omega, A_V = +1$	

CIRCUIT TOPOLOGY

The LH0101 consists of 3 essential stages, an operational amplifier, a buffer, and a power output stage.

Selection of a BI-FET operational amplifier was prompted by a balance between the desired AC and DC performance. This decision was made in order to take advantage of the high performance BI-FET series' slew rate, settling time, and low bias current characteristics. The added feature of internal frequency compensation aided in making it an ideal amplifier upon which to build.

The zero-crossing distortion associated with high current and high frequency conditions is an age-old problem of class B and class AB power amplifiers. In order to minimize the distortion at crossover, the amplifier must maintain a low output impedance throughout zero crossing. This requires the push-pull output transistors to smoothly alternate current sourcing and sinking duty during the crossover.

To obtain a low output resistance the output stage must constantly remain in the active region. The usual approach is to incorporate a class AB output stage similar to that shown in *Figure 1.* During no load conditions, both output transistors are biased ON thus providing a low output resistance and hence eliminating crossover distortion. Under rated current load conditions, however, a potential source of distortion can develop. Take the case of an output at a positive voltage delivering the rated current to a load. The increased base-to-emitter voltage of the drive transistor tends to bias the bottom transistor OFF.

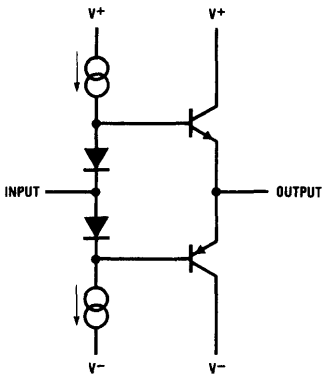

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FIGURE 1. Class AB Output Stage as a Possible Solution to Minimize Distortion

As the output swings negative, crossover distortion can be seen to depend upon how quickly the bottom transistor can turn ON to assume its drive portion of the duty cycle. This condition becomes more acute at higher frequencies.

Of even greater concern is the risk of thermal runaway. An internal rise in temperature decreases transistor junction voltages which in turn increase the collector operating current. Typically emitter degeneration resistors can be used to compensate for this effect, but prove themselves inadequate under rated current load conditions. In order to minimize the junction voltage temperature effect a large resistor value must be selected. At the same time this limits the output drive current and, hence, the output voltage swing. On the other hand, if a small resistor value is selected, the output drive current is maintained but the voltage drop across this small resistance is inadequate to compensate for a decrease in junction voltage. This result brings the problem back to one of thermal runaway and clarifies the shortcomings of *Figure 1.*

Another commonly used technique is a pseudo-class B output stage found in many integrated power op amps, (see *Figure 2*). In this configuration, the limited output swing problem of class AB amplifiers is eliminated. The output swings to within one or two volts of either supply.

The obvious problem with this type of circuit is that it has significant crossover distortion. Distortion occurs when both output transistors are biased completely OFF during zero crossing, thus exhibiting relatively high resistance at the amplifier output. In addition, the minor loop feedback between the base drive of the transistors and the output almost always induces an abrupt change in the response. This further aggravates the amplifier distortion.

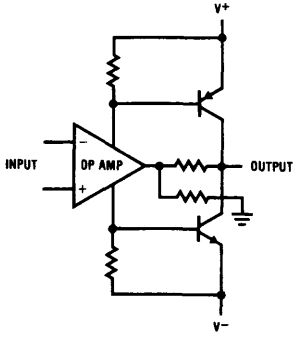

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FIGURE 2. Class B Output Stage

The new op amp successfully combines both techniques to achieve remarkably smooth crossover transitions under most demanding conditions. The complete schematic is shown in *Figure 3.*

The buffer stage, which consists of transistors Q3, Q5, Q10, and Q11 is a current amplifier with unity voltage gain. Connected as a class AB amplifier, its function is to provide distortion-free drive during zero crossing. Bandwidth is in excess of 50 MHz to ensure no bandwidth-induced distortion.

The buffer stage output is current limited by transistors Q7 and Q8 to no more than 50 mA. However, the power stage transistors Q1 and Q2 are designed to turn ON as the load current reaches about 25 mA. Any additional current demanded is sustained by these two output transistors right up to the rated output limit. Thus, the reserve drive of the buffer stage is used only to "smooth" the turn-on delay of the output Darlington transistors.

Q6 and Q9 base-to-emitter junctions are used as current limit sense to protect the output stage. Current sense resistors connected between the supply pins and the SC pins program the limit threshold. In operation, an approximate 0.6V differential turns ON either transistor Q6 or Q9, which in turn drives Q12 and Q4 respectively, starving any excess base current from driving the output beyond the preset limit.

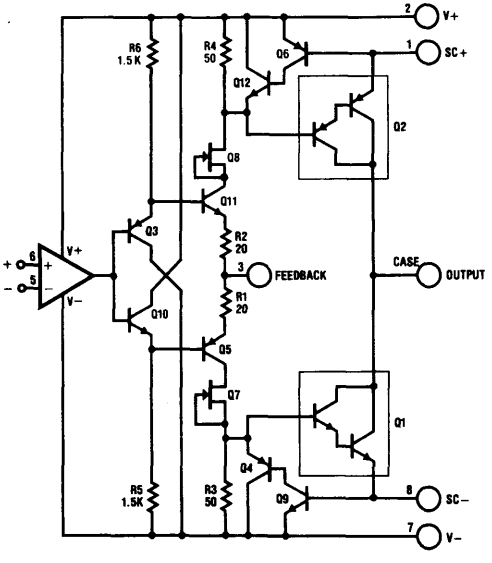

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FIGURE 3. LH0101 Complete Schematic

RESULT

The performance of the LH0101 is best demonstrated in the following photographs. *Figure 4* shows the large signal slew response of the LH0101 into a 10Ω load. No crossover distortion is evident.

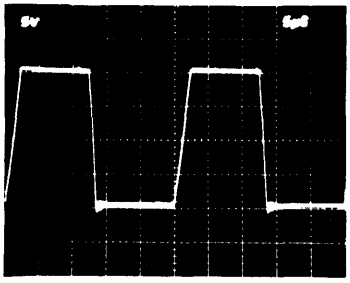

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FIGURE 4. Large Signal Pulse Response, 10Ω Load

Generally, crossover distortion occurs within a small region near zero crossing. In order to amplify its effect, a signal of small amplitude is used. *Figures 5* and *6* show a signal amplitude of 2 volts peak, and loads of 10Ω and 1Ω respectively. Notice that a slight distortion is observed in *Figure 6,* but only under the extreme condition imposed by the 1Ω load!

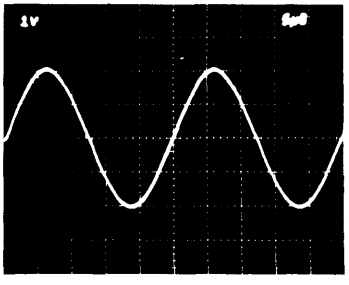

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FIGURE 5. Small Signal Response, 10Ω Load

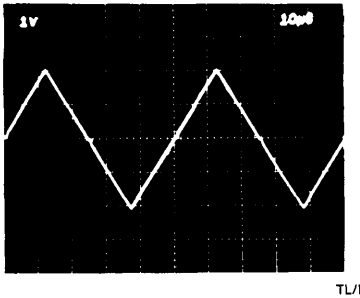

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FIGURE 6. Small Signal Triangular Wave Response, 1Ω Load

DESIGN PRECAUTIONS

Circuit Layout Considerations

In high power applications, one must pay close attention to the trace connections in which high current is carried. Critical connections should be short to minimize line drop. For example, a 10 mΩ PC trace carrying 2 amps develops 20 mV of error voltage. It is important to be aware of where this error is generated and how it impacts accuracy.

Ground connections are probably the most important, if not the most troublesome. Not only can they contribute to circuit error, but in many situations the circuit can become unstable if the layout induces excessive phase error. *Figure 7* shows one correct technique for circuit grounding. The heavy lines represent high current paths. The analog signal ground is returned to the supply common.

Output Current Limit

As described in the previous section, current sense resistors may be inserted between the supply pins and the SC pins to limit excessive load currents. A voltage of 0.6V developed across the sense resistor triggers the limiting circuit. *Figure 8* illustrates the usage.

In cases where the chosen R_{SC} is small ($<1\Omega$), the contact resistances from solder connections and socket ohmic contacts become important and must not be overlooked. A good solder joint typically exhibits 5 mΩ of resistance and socket contacts have about 10 mΩ. Even interconnecting traces will become significant if they are long.

Consider the circuit in *Figure 8*; a pair of good solder joints on the 0.3Ω current sense resistors contribute more than 3% error. Also, one can expect the current sense transistor threshold to vary as much as 10% from device to device. Furthermore, this threshold has a temperature coefficient of −2 mV/°C. In summary, the expected accuracy is on the order of 20% to 25% under all operating conditions.

When designing the current limit, the threshold should not be set too close to the worst case peak current under any normal operating conditions. Signal distortion will occur even if the threshold is intermittently exceeded for a very short duration. In the worst instance, the circuit can trigger spurious oscillation, such as in the case of driving a capacitive load during transient conditions. These occurrences are very real in nearly all op amps having similar current sense circuits. Although the current limit circuit has high enough gain to produce a sharp response, it is a good idea to allow a 20% margin above the worst case operating condition.

Safe Operating Conditions

In order to preserve the reliable performance of the LH0101, the device must not operate beyond the boundary defined in the Safe Operating Area curve in the data sheet. Because of its importance, it has been reproduced in *Figure 9.*

Power Dissipation Considerations

Probably the single most important gauge of reliability is the operating temperature of this device. The derating curve, which has again been reproduced in *Figure 10,* must be followed faithfully. Similar to the Safe Operating Area curve, under no circumstances should the boundary be exceeded.

The curves relate operating and junction temperature, power dissipation and thermal resistance. The general relationship is expresed as follows.

$$P_{DISS} = \frac{T_{J(MAX)} - T_{A(MAX)}}{R_{\theta JC} + R_{\theta CS} + R_{\theta SA}} \quad (1)$$

where: P_{DISS} = the power dissipated by the device in watts.

$T_{J(MAX)}$ = the maximum junction temperature allowed, for the LH0101, $T_{J(MAX)}$ = 150°C.

$T_{A(MAX)}$ = the maximum ambient temperature in °C under which the device must operate.

$R_{\theta JC}$ = thermal resistance from the junction to case in °C/W, for the LH0101, $R_{\theta JC}$ = 2.5°C/W.

$R_{\theta CS}$ = thermal resistance from case to surface of heat sink in °C/W.

$R_{\theta SA}$ = thermal resistance from heat sink to free air ambient in °C/W.

In simple terms, the expression is a measure of how well the internally generated heat is removed such that the power dissipated will not give rise to a maximum permissible junction temperature of 150°C. Thus, the sum of all the thermal resistance represents the thermal efficiency of the mechanical design. The lower the sum, the more efficient the thermal conductivity.

In a typical design, first and foremost is to calculate the maximum power dissipation that the device is designed to handle. There are two components, which are related by the following equation:

$$P_{DISS} = P_Q + P_O \quad (2)$$

The first part of the equation is the quiescent power at which the device operates under no load. The second term is the power dissipated by the output transistors due to the load. This is calculated as the average voltage difference between the supply voltage and the output voltage multiplied by the maximum rms load current the amplifier is required to deliver.

Once the power dissipation is calculated, the next step is to determine the maximum ambient temperature in which the device must operate.

To complete the thermal design, all contributions of thermal resistances must be summed per equation (1) above. First, the junction-to-case thermal resistance for the LH0101 is given in the data sheet; it is typically 2.5°C/W.

The metal case of the LH0101 is electrically connected to the output of the amplifier. Unless the application permits direct mounting to a heat sink, a sheet of insulation should be sandwiched between the case and the mounting surface for isolation purposes. Many types of insulators are available. The most popular of these is mica film. Its thermal resistance is listed in Table II along with other types.

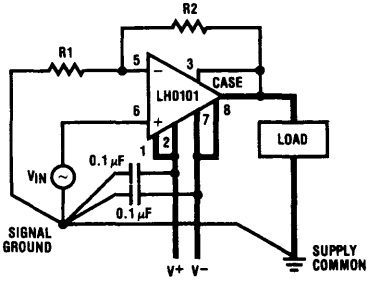

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FIGURE 7. Proper Supply Connection

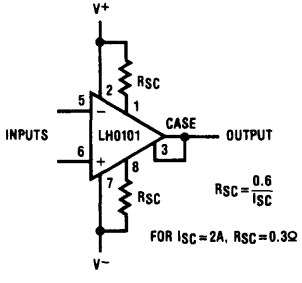

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FIGURE 8. Current Limit Protection

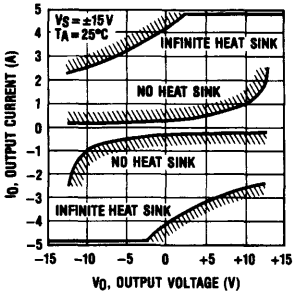

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FIGURE 9. Safe Operating Area

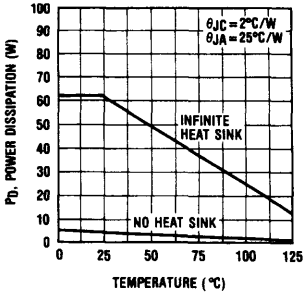

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FIGURE 10. Power Derating Curve

TABLE II. Thermal Resistance (Note 1)

Insulator Material Type	W/O Thermal Joint Compound	W/Thermal Joint Compound	Sources
Mica	1.3°C/W @ 0.003" thick	0.25°C/W	Thermalloy Inc.
	1.2°C/W @ 0.002" thick	0.33°C/W	Keystone Electronics Mod. #4658
Thermalfilm I2	1.5°C/W @ 0.002" thick	0.52°C/W	Thermalloy Inc.
Aluminum Oxide2	1.0°C/W @ 0.062°C	0.3°C/W	
Beryllium Oxide2-3	0.6°C/W @ 0.062" thick	0.15°C/W	
Insul-Cote2	0.5°C/W @ 0.002" thick		Thermalloy Inc.

Note 1: Mounting bolts torqued to 6 oz.-in.

Note 2: Consult manufacturer on availability for 8-lead TO-3 package style.

Note 3: Particle, dust, or fumes present health hazards when inhaled. Grinding, sanding, and pulverizing the material should be avoided.

In critical applications, thermal-joint compound should be used to maximize heat transfer across the case to the heat sink. With air being a poor heat conduction medium, the use of thermal joint compound eliminates air gaps between mounting surfaces, thus providing more than 3 times improvement in thermal efficiency over those cases without.

The remaining unknown, $R_{\theta SA}$, can now be determined from the proper selection of the heat sink. By itself, that is, with no heat sink, the TO-3 case has a junction-to-ambient thermal resistance ($R_{\theta JA}$ or θ_{JA}) of about 25°C/W. Consequently, a heat sink is almost always required in applications involving significant power. Most heat sink manufacturers specify the mounting-surface to ambient thermal resistance $R_{\theta SA}$. In a nutshell, the heat sink is selected such that the right hand side of equation (1) is equal to or greater than the left hand side, or total power dissipation. It is good engineering practice to allow at least a 10% safety margin.

Design Consideration Driving Inductive Load

The LH0101 is suitable for driving most inductive loads including voice-coils and motors. However, in many situations the device should be protected from the harmful effects of energy stored in the inductor. Such a condition exists when power is removed from the circuit at an instant when a high current is flowing through the inductor. A back-emf may have energy high enough to forward bias internal junctions at a current density level sufficient to destroy the device. *Figure 11* illustrates a simple way to prevent this.

Theoretically, an inductive load does not cause amplifier loop instability. However, if the circuit Q is high enough and stray capacitances are within a critical range, the load circuit can break out into oscillation. A series RC damping circuit of 10Ω and a 0.01 μF capacitor across the inductor as shown in *Figure 12* usually alleviates the problem.

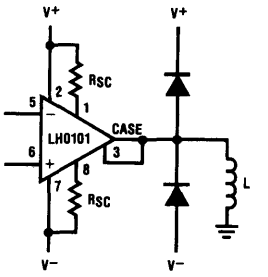

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FIGURE 11. Back EMF Suppression Technique

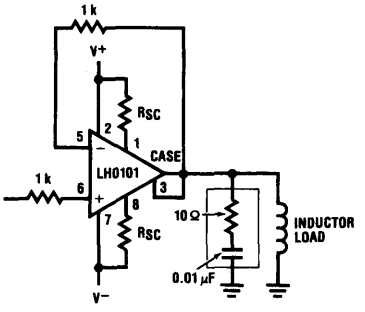

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FIGURE 12. RC Damping to Compensate Inductor Load

In some applications where it is desirable to prevent power-on surges from actuating the load, for example a motor valve actuator or a disk drive read/write head servo loop, the same RC damping circuit provides an alternate conductive path to suppress surge current.

Design Considerations Driving Capacitor Load

Capacitive loads tend to create an unwanted pole at the tail end of the frequency response where the open loop gain approaches unity gain frequency. The effect is a net reduction of phase margin. For example, a 500 pF load capacitor reduces the phase margin of the amplifier from a no load of 58° to 45°. A 1000 pF capacitor pushes it down to 40°. With a large 0.01 μF capacitor, the amplifier has a mere 22° phase margin. The latter cases are susceptible to oscillation. *Figure 13* shows a compensation technique to restore stability. The value of the lead capacitor C1 should be such that the capacitive reactance is one-fifth the resistance of R2 at the unity-gain crossover frequency of the amplifier, or 4 MHz.

It is interesting to note that there is a critical value for the load capacitor above which oscillation cannot occur. That value is approximately 0.1 μF. Under such a condition, the time constant is so large that the heavy damping effectively suppresses any chance for the circuit to oscillate.

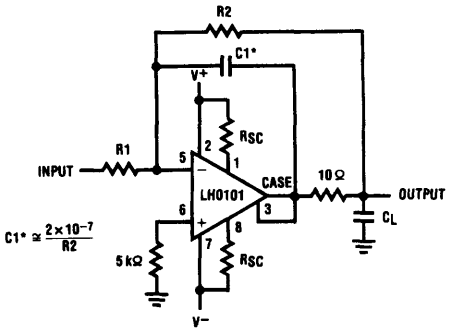

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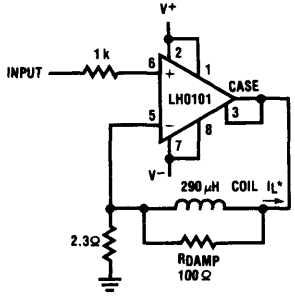
FIGURE 13. Compensation for Capacitance Load

TYPICAL APPLICATIONS

CRT Yoke Driver

One of the most natural applications for the LH0101 op amp is the deflection yoke driver for high resolution CRTs. The low distortion characteristics allow virtually unrestricted use in any circuit configuration. A typical design is shown in *Figure 14.*



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*Coil Current I_L Measured with
Tektronix Current Probe Model P6042.

FIGURE 14. CRT Yoke Driver Circuit

A 500 mV peak-to-peak triangular waveform about ground is input to the amplifier, giving rise to a 100 mA peak current to the inductor. As shown in *Figures 15* and *16,* the responses were recorded at 60 Hz and 20 kHz respectively. At higher frequencies, R_{DAMP} becomes important. The value should be selected to yield the cleanest waveform.

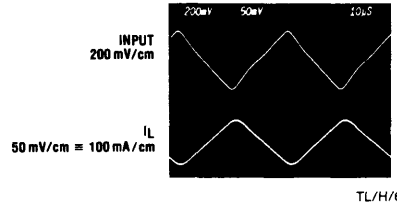

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FIGURE 15. 60 Hz Current Drive Waveform of CRT Deflection Coil

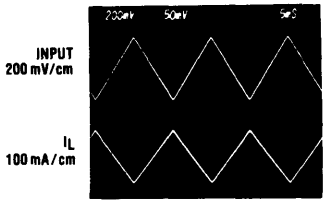

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FIGURE 16. 20 kHz Current Drive Waveform of CRT Deflection Coil

Servo Motor Amplifier

A typical motor driver circuit is shown in *Figure 17.* The amplifier will deliver the rated current into the motor. Again, care should be taken to keep power dissipation within the permitted level.

A variation of the same servo design is shown in *Figure 18.* This precision speed regulation circuit employs rate feedback for constant motor current at a given input voltage.

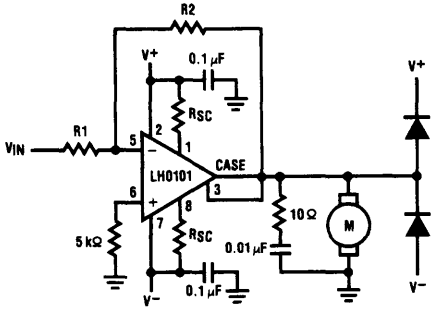

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FIGURE 17. Servo Motor Amplifier

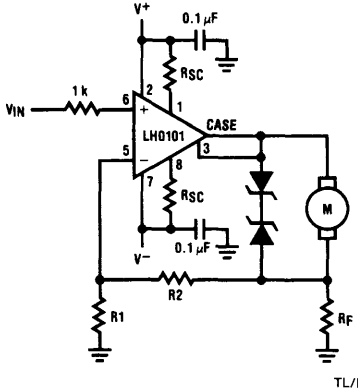

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FIGURE 18. Rate Feedback Servo Motor Amplifier

Digitally Programmable Power Source

Designing precision voltage and current sources is made simple using the LH0101. Adding a digital-to-analog converter provides a tremendous amount of flexibility in speed and control. Applications range from DC precision power supplies to sophisticated programmable waveform generators. The design of the voltage source is relatively straightforward, whereas the programmable current source is a bit more involved. Such a circuit is shown in *Figure 19.* The DAC is configured to operate in a bipolar mode with an output range of ±10.000V. With 12 bits, the DAC outputs an equivalent of 4.88 mV per bit-weight. Consequently, the resolution at the current source is 0.488 μA/bit.

The output sources and sinks current only to ground referenced loads. A negative full-scale code (all digital inputs low) effects a negative (source) 1 amp current output. A zero scale (MSB low and all other bits high) gives zero current. And a positive full scale code (all digital inputs high) forces a positive (sink) 1 amp current at the output.

The versatility of this circuit configuration is not without limitation. Because the output voltage is dependent upon the ground referred load, one must be aware of the potentially destructive power dissipation level the LH0101 must sustain. For example, 1A current into a 5Ω grounded load generates 10W of power in the amplifier. This level is high enough to destroy the device unless an appropriate heat sink is used to keep the device junction temperature from exceeding the 150°C limit.

Coaxial Cable Driver

The LH0101 makes an ideal cable driver of any type. It has adequate bandwidth for most audio and sub-video applications. The high current, distortion free output can easily interface any termination required. Large line capacitance does not present a problem for the LH0101. It has adequate reserve current capability to charge the capacitance without seriously degrading bandwidth. However, current limit protection against cable shorts is recommended. A typical interface circuit is shown in *Figure 20.* The op amp can drive up to 6 coaxial lines without the use of a heat sink.

Low Distortion Audio Amplifier

At this juncture, it would be of great interest to see how well the LH0101 performs in the audio high fidelity arena. The intent here is not to set a new standard but merely to use the stringent requirements of the audio specifications as an ideal yardstick for comparison.

The complete design is illustrated in *Figure 21.* The circuit is configured as a bridge amplifier to maximize available output power for a given set of supply voltages.

The result is an impressive set of specifications summarized in Table III. Although not earth-shaking, 0.14% total harmonic distortion at the rated 40W output, within the full audio frequency spectrum, is very respectable.

Transient slew rate of greater than 10V/μs extends the full power bandwidth to beyond 200 kHz, and the distortion response is plotted over the entire audio spectrum in *Figure 22.* This would satisfy all but a handful of audio purists.

About the only difficulty encountered was finding a heat sink that was good enough for convection cooling. By following the previous section on Power Dissipation Considerations, the heat sink thermal resistance required is a maximum of 3.5°C/W at an ambient temperature of 25°C. The calculation included the use of mica insulator and liberal use of thermal-coat compound. As it turned out, a large extruded heat sink with fins similar to the Thermalloy type 6141 did an excellent job of keeping the junctions cool.

TABLE III. Bridge Audio Amplifier Specifications

A_V (Voltage Gain)	3
Z_{IN} (Input Impedance)	10 kΩ
I_q (Quiescent Current)	60 mA
P_O (Output Power)	40 Watts into 8Ω
—RMS Continuous 20 Hz–20 kHz	28 Watts into 16Ω
Full Power Bandwidth	DC to >100 kHz
THD (Total Harmonic Distortion)	
1W 20–20,000 Hz	<0.8%
40W 20–20,000 Hz	<0.15%
IMD (Intermodulation Distortion)	
1W 60 Hz/100 kHz 4:1	<0.01%
40W 60 Hz/100 kHz 4:1	<0.002%
Peak Output Current	3.125A into 8Ω
Supply Voltage	±18V
Maximum Output Voltage Swing	21.2V rms
	30V Peak

REFERENCES

1. National Semiconductor, *Special Functions Databook.*
2. National Semiconductor, *Linear Applications Handbook.*
3. J. Wong, J. Sherwin, "Applications of Wide-Band Buffer Amplifier", National Semiconductor AN-227, October 1979.
4. National Semiconductor "LH0101 Power Operational Amplifier" data sheet.

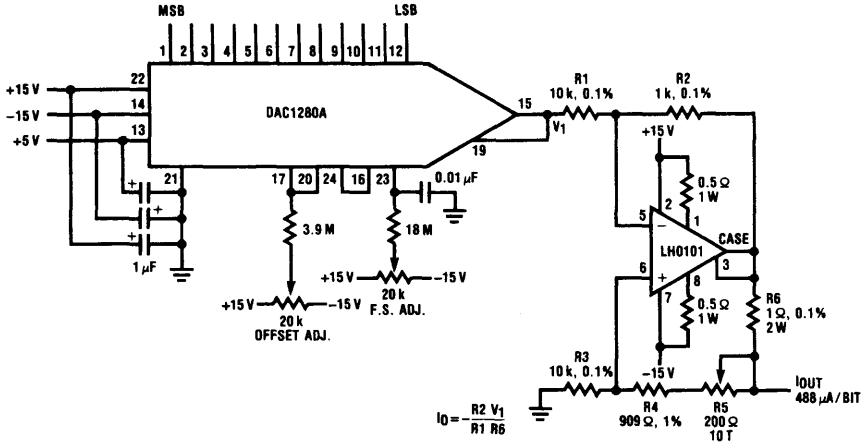

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FIGURE 19. Digitally Programmable Current Source

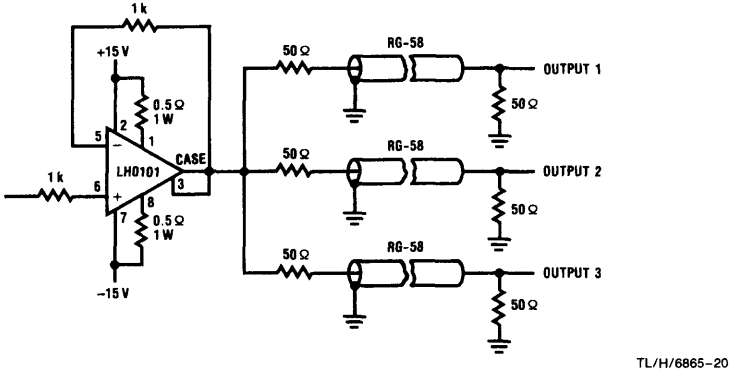

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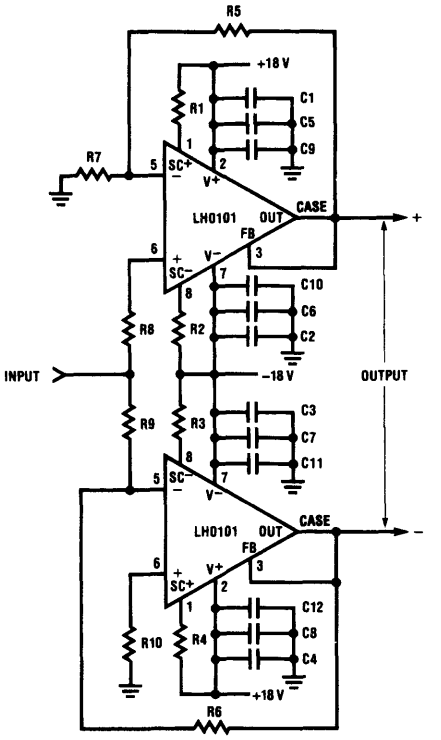
FIGURE 20. Multi-Line Coaxial Cable Driver



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R1–R4	Current Limit Resistor	0.15Ω, 2W
R5	Feedback Resistor	5 kΩ
R6	Feedback Resistor	15 kΩ
R7–R10	Input Resistors	10 kΩ
C1–C4	Bypass Capacitors	47 μF, 25V Electrolytic
C5–C8	Bypass Capacitors	10 μF, 25V Tantalum
C9–C12	Bypass Capacitors	0.1 μF, 25V Ceramic

FIGURE 21. LH0101 Bridge Audio Power Amplifier

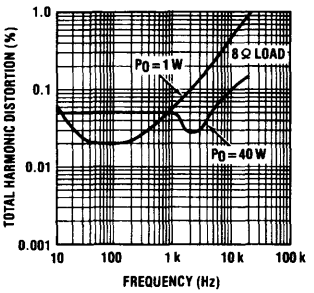

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FIGURE 22. Total Harmonic Distortion vs Frequency of Bridge Power Amplifier