

Applying a New Precision Op Amp

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Abstract: A new bipolar op amp design has advanced the state of the art by reducing offset voltage and bias current errors. Its characteristics are described here, indicating an ultimate input resolution of 10 μV and 1 pA under laboratory conditions. Practical circuits for making voltmeters, ammeters, differential instrumentation amplifiers and a variety of other designs that can benefit from the improved performance are covered in detail. Methods of coupling the new device to existing fast amplifiers to take advantage of the best characteristics of both, even in follower applications, are explored.

INTRODUCTION

A low cost, mass-produced op amp with electrometer-type input currents combined with low offset voltage and drift is now available. Designated the LM11, this IC can minimize production problems by providing accuracy without adjustments, even in high-impedance circuitry. On the other hand, if pushed to its full potential, what has been impossible in the past becomes entirely practical.

Significantly, the LM11 is not restricted to commercial and industrial use. Devices can be completely specified over a -55°C to $+125^\circ\text{C}$ range. Preliminary data indicates that reliability is the same as standard ICs qualified for military and space applications.

The essential details of the design along with an introduction to the peculiarities of high-impedance circuits have been presented elsewhere.* This will be expanded here. Practical circuitry that reduces effective bias current for those applications where performance cannot be made dependent on offset current are described. In addition, circuits combining the DC characteristics of the new part with the AC performance of existing fast amplifiers will be shown. This will be capped with a number of practical designs to provide some perspective into what might be done.

DC ERRORS

Barring the use of chopper or reset stabilization, the best offset voltage, drift and long-term stability are obtained using bipolar transistors for the op amp input stage. This has been done with the LM11. On-wafer trimming further improves performance. Typically, a 100 μV offset with 1 $\mu\text{V}/^\circ\text{C}$ drift results.

Transistors with typical current gains of 5000 have been used in the manufacture of the LM11. The input stage employs a Darlington connection that has been modified so that offset voltage and drift are not degraded. The typical input currents, plotted in *Figure 1*, demonstrate the value of the approach.

Note: *R. J. Widlar, "Working with High Impedance Op Amps", National Semiconductor AN-241.

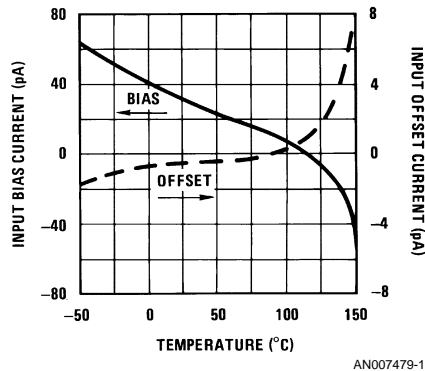


FIGURE 1. Below 100°C , bias current varies almost linearly with temperature. This means that simple circuitry can be used for compensation. Offset current is unusually low.

The offset current of this op amp is so low that it cannot be measured on existing production test equipment. Therefore, it probably cannot be specified tighter than 10 pA. For critical applications, the user should have little difficulty in selecting to a tighter limit.

The bias current of the LM11 equals that of monolithic FET amplifier at 25°C . Unlike FETs, it does not double every 10°C . In fact, the drift over a -55°C to $+125^\circ\text{C}$ temperature range is about the same as that of a FET op amp during normal warm up.

Other characteristics are summarized in *Table 1*. It can be seen that the common-mode rejection, supply-voltage rejection and voltage gain are high enough to take full advantage of the low offset voltage. The unspectacular $0.3\text{V}/\mu\text{s}$ slew rate is balanced by the 300 μA current drain.

TABLE 1. Typical characteristics of the LM11 for $T_j = 25^\circ\text{C}$ and $V_S = \pm 15\text{V}$. Operation is specified down to $V_S = \pm 2.5\text{V}$.

Parameter	Conditions	Value
Input Offset Voltage		100 μV
Input Offset Current		500 fA
Input Bias Current		25 pA
Input Noise Voltage	$0.01 \text{ Hz} \leq f \leq 10 \text{ Hz}$	8 μVpp
Input Noise Current	$0.01 \text{ Hz} \leq f \leq 10 \text{ Hz}$	1 pApp
Long Term Stability	$T_j = 25^\circ\text{C}$	10 μV
Offset Voltage Drift	$-55^\circ\text{C} \leq T_j \leq 125^\circ\text{C}$	1 $\mu\text{V}/^\circ\text{C}$
Offset Current Drift	$-55^\circ\text{C} \leq T_j \leq 125^\circ\text{C}$	20 fA/ $^\circ\text{C}$
Bias Current Drift	$-55^\circ\text{C} \leq T_j \leq 125^\circ\text{C}$	500 fA/ $^\circ\text{C}$

TABLE 1. Typical characteristics of the LM11 for $T_j = 25^\circ\text{C}$ and $V_S = \pm 15\text{V}$. Operation is specified down to $V_S = \pm 2.5\text{V}$. (Continued)

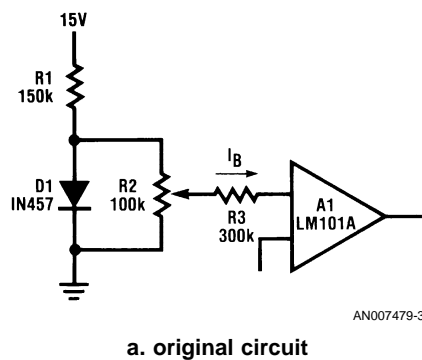
Parameter	Conditions	Value
Voltage Gain	$V_{OUT} = \pm 12\text{V}$,	1,200V/mV
	$I_{OUT} = \pm 0.5\text{ mA}$	
	$V_{OUT} = \pm 12\text{V}$,	300V/mV
	$I_{OUT} = \pm 2\text{ mA}$	
Common-Mode Rejection	$-12.5\text{V} \leq V_{CM} \leq 14\text{V}$	130 dB
Supply-Voltage Rejection	$\pm 2.5\text{V} \leq V_S \leq \pm 20\text{V}$	118 dB
Slew Rate		0.3V/ μs
Supply Current		300 μA

As might be expected, the low bias currents were obtained with some sacrifice in noise. But the low frequency noise voltage is still a bit less than a FET amplifier and probably more predictable. The latter is important because this noise cannot be tested in production. Long term measurements have not indicated any drift in excess of the noise. This is not the case for FETs.

It is worthwhile noting that the drift of offset voltage and current is low enough that DC accuracy is noise limited in room-temperature applications.

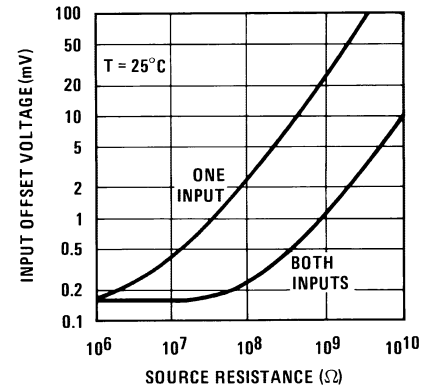
BIAS CURRENT COMPENSATION

The LM11 can operate from $M\Omega$ source resistances with little increase in the equivalent offset voltage, as can be seen in *Figure 2*. This is impressive considering the low initial offset voltage. The situation is much improved if the design can be configured so that the op amp sees equal resistance on the two inputs. However, this cannot be done with all circuits. Examples are integrators, sample and holds, logarithmic converters and signal-conditioning amplifiers. And even though the LM11 bias current is low, there will be those applications where it needs to be lower.



AN007479-3

Referring back to *Figure 1*, it can be seen that the bias current drift is essentially linear over a -50°C to $+100^\circ\text{C}$ range. This is a deliberate consequence of the input stage design. Because of it, relatively simple circuitry can be used to develop a compensating current.

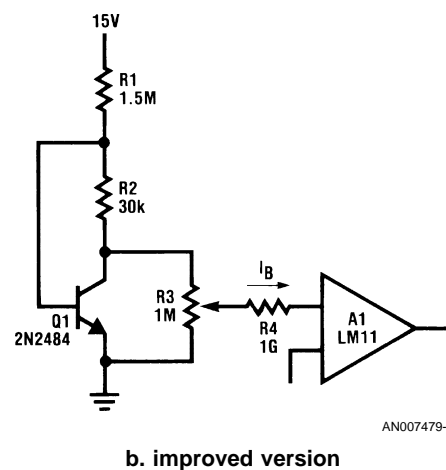


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FIGURE 2. The LM11 operates from $M\Omega$ source resistances with little DC error. With equal source resistances, accuracy is essentially limited by low frequency current noise.

Bias current compensation is not new, but making it effective with even limited temperature excursions has been a problem. An early circuit suggested for bipolar ICs is shown in *Figure 3(a)*. The compensating current is determined by the diode voltage. This does not vary as rapidly with temperature as bias current nor does it match the usual non-linearities.

With the improved circuit in *Figure 3(b)*, the temperature coefficient can be increased by using a transistor and including R_2 . The drop across R_2 is nearly constant with temperature. The voltage delivered to the potentiometer has a 2.2 mV/ $^\circ\text{C}$ drift while its magnitude is determined by R_2 . Thus, as long as the bias current varies linearly with temperature, a value for R_2 can be found to effect compensation.



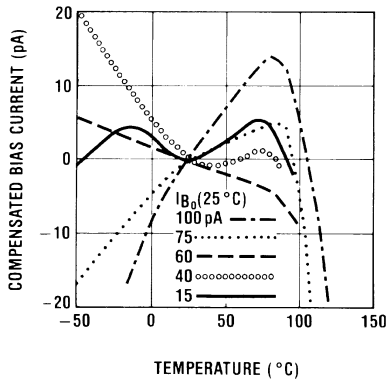
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FIGURE 3. Bias-current compensation. With the improved version, the temperature coefficient of the compensating current can be varied with R_2 . It is effective only if bias current has linear, negative temperature coefficient.

In production, altering resistors based on temperature testing is to be avoided if at all possible. Therefore, the results that can be obtained with simple nulling at room temperature

and a fixed value for R_2 are of interest. *Figure 4* gives this data for a range of parts with different initial bias currents. This was obtained from pre-production and initial-production

runs. The bias current variations were the result of both h_{FE} variations and changes in internal operating currents and represent the worst as well as best obtained. They are therefore considered a realistic estimate of what would be encountered among various production lots.



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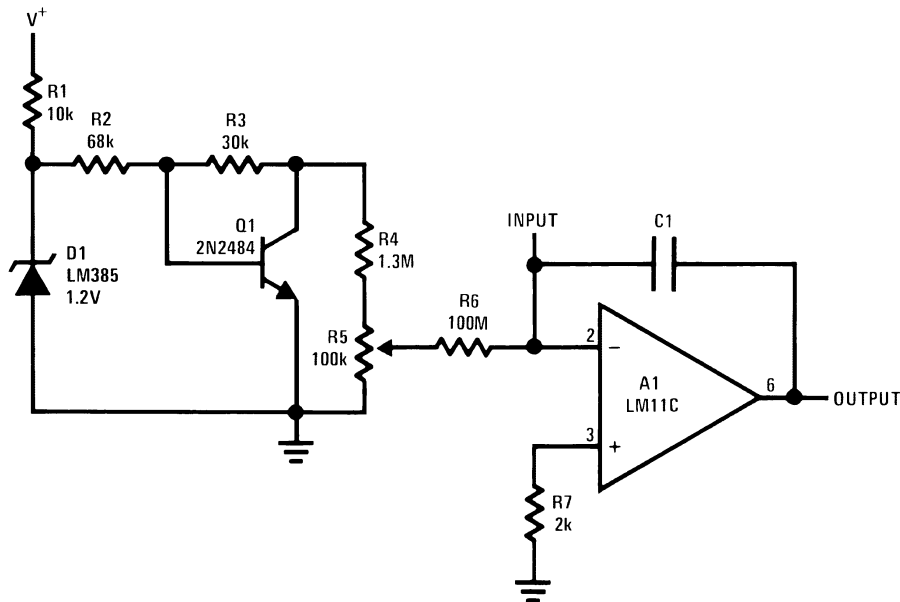
FIGURE 4. Compensated bias current for five representative units with a range of initial bias currents. The circuit in Figure 3(b) was used with balancing at 25°C. High drift devices could be improved further by altering R2.

Little comment need be made on these results, except that the method is sufficiently predictable that another factor of five reduction in worst case bias current could be made by altering R2 based on the results of a single temperature run.

One disadvantage of the new circuit is that it is more sensitive to supply variations than the old. This is no problem if the supplies are regulated to 1%. But with worst regulation it suffers because, with R2, the transistor no longer functions as a regulator and because much tighter compensation is obtained.

The circuit in Figure 5 uses pre-regulation to solve this problem. The added reference diode has a low breakdown so that the minimum operating voltage of the op amp is unrestricted. Because of the low breakdown, the drop across R3 can no longer be considered constant. But it will vary linearly with temperature, so this is of no consequence. The fact that this reference can be used for other functions should not be overlooked because a regulated voltage is frequently required in designs using op amps.

In Figure 5 a divider is used so that the resistor feeding the compensating current to the op amp can be reduced. There will be an error current developed for any offset voltage change across R6. This should not be a problem with the LM11 because of its low offset voltage. But for tight compensation, mismatch in the temperature characteristics of R4 and R5 must be considered.



AN007479-6

FIGURE 5. Bias current compensation for use with unregulated supplies. Reference voltage is available for other circuitry.

Bias current compensation is more difficult for non-inverting amplifiers because the common-mode voltage varies. With a voltage follower, everything can be bootstrapped to the output and powered by a regulated current source, as shown in Figure 6. The LM334 is a temperature sensor. It regulates against voltage changes and its output varies linearly with temperature, so it fits the bill.

Although the LM334 can accommodate voltage changes fast enough to work with the LM11, it is not fast enough for the high-speed circuits to be described. But compensation can still be obtained by using the zener diode pre-regulator bootstrapped to the output and powered by either a resistor or

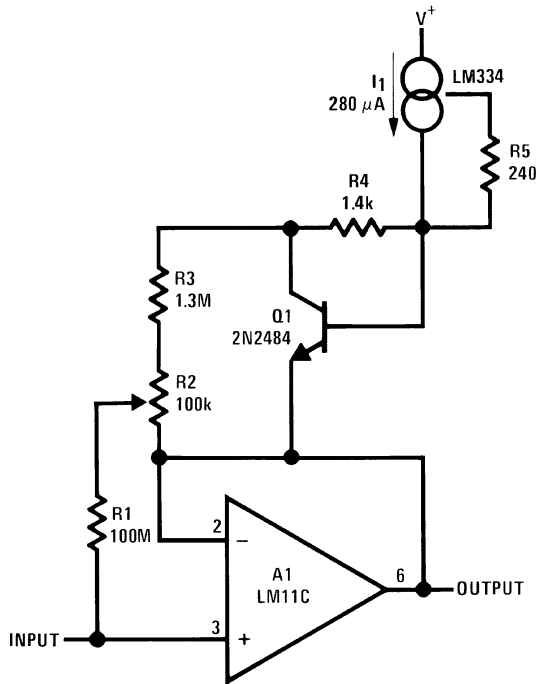
FET current source. The LM385 fits well here because both the breakdown voltage and minimum operating current are low.

With ordinary op amps, the collector base voltage of the input transistors varies with the common-mode voltage. A 50% change in bias current over the common-mode range is not unusual, so compensating the bias current of a follower has limited value. However, the bootstrapped input stage of the LM11 reduces this to about 2 pA for a $\pm 20V$ common-mode swing, giving a $2 \times 10^{13}\Omega$ common-mode input resistance.

FAST AMPLIFIERS

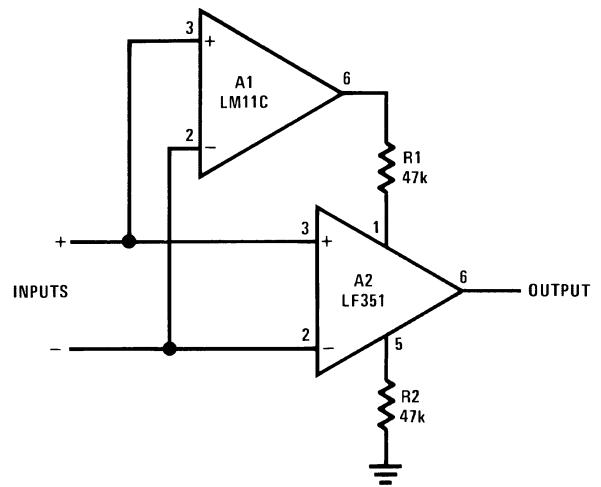
A precision DC amplifier, although slow, can be used to stabilize the offset voltage of a less precise fast amplifier. As shown in *Figure 7*, the slow amplifier senses the voltage across the input terminals and supplies a correction signal to the balance terminals of the fast amplifier. The LM11 is particularly interesting in this respect as it does not degrade the input bias current of the composite even when the fast amplifier has a FET input.

Surprisingly, with the LM11, this will work for both inverting and non-inverting connections because its common-mode slew recovery is a lot faster than that of the main loop. This was accomplished, even with circuitry running under 100 nA, by proper clamping and by bootstrapping of internal stray capacitances.



AN007479-7

FIGURE 6. This circuit shows how bias current compensation can be used on a voltage follower.



AN007479-8

FIGURE 7. A slow amplifier can be used to null the offset of a fast amplifier.

An optimized circuit for the inverting amplifier connection is shown in *Figure 8*. The LM11 is DC coupled to the input and drives the balance terminals of the fast amplifier. The fast amplifier is AC coupled to the input and drives the output. This isolates FET leakage from the input circuitry.

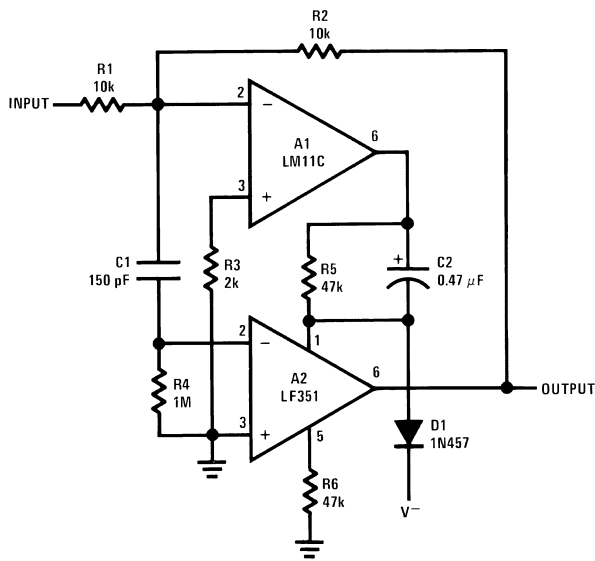
As can be seen, the method of coupling into the balance terminals will vary depending on the internal configuration of the fast amplifier. If the quiescent voltage on the balance terminals is beyond the output swing of the LM11, a differential coupling must be used, as in *Figure 8(a)*. A lead capacitor, C2, reduces the AC swing required at the LM11 output. The clamp diode, D1, insures that the LM11 does not overdrive the fast amplifier in slew.

If the quiescent voltage on the balance terminals is such that the LM11 can drive directly, the circuit in *Figure 8(b)* can be used. A clamp diode from the other balance terminal to internal circuitry of the LM11 keeps the output from swinging too far from the null value, and a resistor may be required in series with its output to insure stability.

Measurements indicate that the slew rate of the fast amplifier is unimpaired, as is the settling time to 1 mV for a 20V output excursion. If the composite amplifier is overdriven so that the output saturates, there will be an added recovery delay because the coupling capacitor to the fast amplifier takes on a charge with the summing node off ground. Therefore, C1 should be made as small as possible. But going below the values given may introduce gain error.

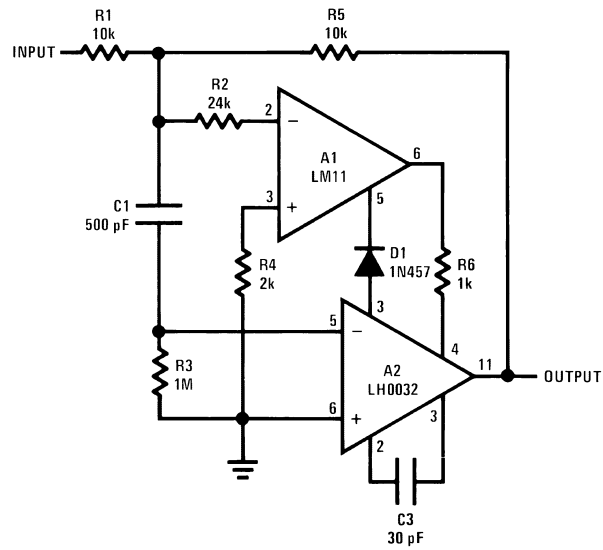
If the bias current of the fast amplifier meets circuit requirements, it can be direct-coupled to the input. In this case, offset voltage is improved, not bias current. But overload recovery can be reduced. The AC coupling to the fast-amplifier input might best be eliminated for limited-temperature-range operation.

This connection also increases the open-loop gain beyond that of the LM11, particularly since two-pole compensation can be effected to reduce AC gain error at moderate frequencies. The DC gains measured showed something in excess of 140 dB.



AN007479-9

a. with standard BI-FET



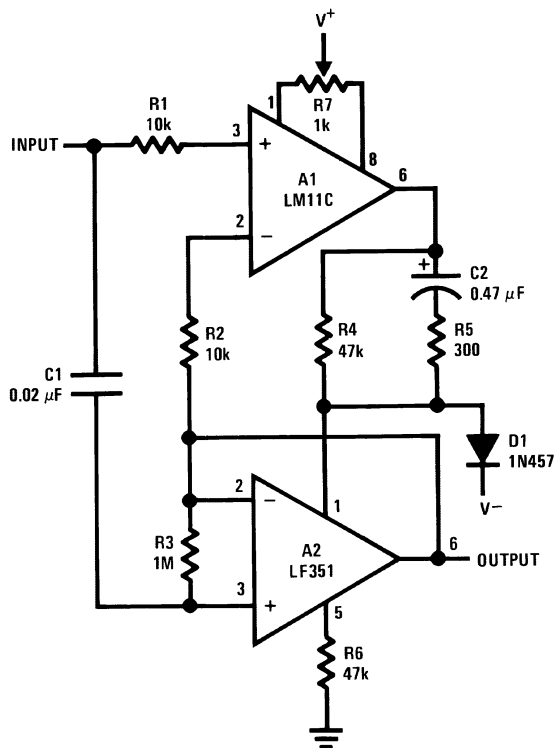
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b. with fast hybrid

FIGURE 8. These inverters have bias current and offset voltage of LM11 along with speed of the FET op amps. Open loop gain is about 140 dB and settling time to 1 mV about 8 μ s. Excess overload-recovery delay can be eliminated by directly coupling the FET amplifier to summing node.

A voltage-follower connection is given in *Figure 9*. The coupling circuitry is similar, except that R5 was added to eliminate glitches in slew. Overload involves driving the fast amplifier outside its common-mode range and should be avoided by limiting the input. Thus, AC coupling the fast amplifier is less a problem. But the repetition frequency of the input signal must also be limited to 10 kHz for ± 10 V

swing. Higher frequencies produce a DC error, believed to result from rectification of the input signal by the voltage sensitive input capacitance of the FET amplifier used. A fast bipolar amplifier like the LM118 should work out better in this respect. To avoid confusion, it should be emphasized that this problem is related to repetition frequency rather than rise time.



AN007479-11

FIGURE 9. Follower has 10 μ s settling to 1 mV, but signal repetition frequency should not exceed 10 kHz if the FET amplifier is AC coupled to input. The circuit does not behave well if common-mode range is exceeded.

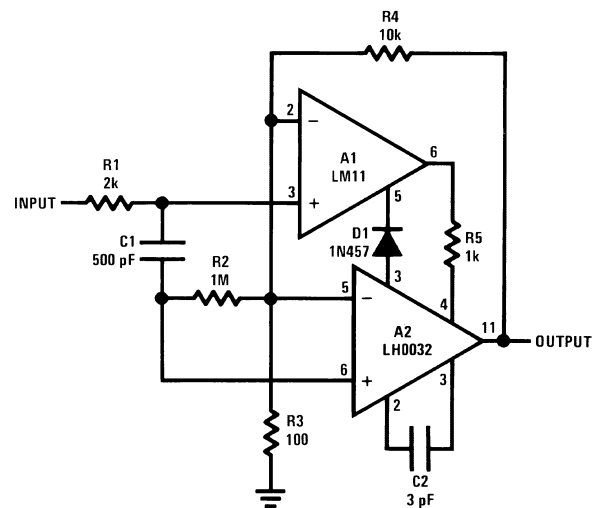
A precision DC amplifier with a 100 MHz gain-bandwidth product is shown in *Figure 10*. It has reasonable recovery ($\sim 7 \mu$ s) from a 100% overload; but beyond that, AC coupling to the fast amplifier causes problems. Alone, the gain error and thermal feedback of the LH0032 are about 20 mV, input referred, for ± 10 V output swing. Adding the LM11 reduces this to microvolts.

PICOAMMETER

Ideally, an ammeter should read zero with no input current and have no voltage drop across its inputs even with full-scale deflection. Neither should spurious indications nor inaccuracy result from connecting it to a low impedance. Meeting all these requirements calls for a DC amplifier, and one in which both bias current and offset voltage are controlled.

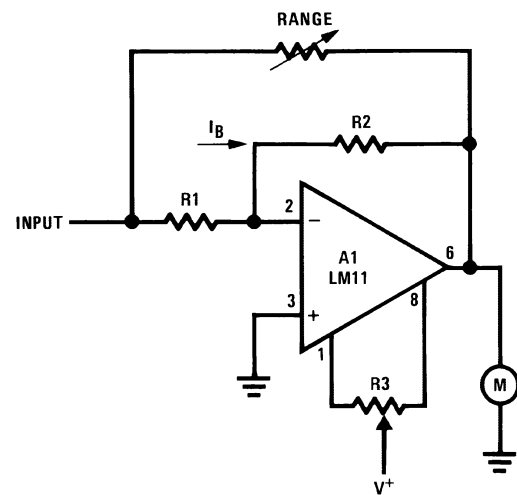
The summing amplifier connection is best for measuring current, because it minimizes the voltage drop across the input terminals. However, when the inputs are shorted, the output state is indeterminate because of offset voltage. Adding degeneration as shown in *Figure 11* takes care of this problem. Here, R2 is the feedback resistor for the most sensitive range, while R1 is chosen to get the meter deflection out of the noise with a shorted input. Adding the range

resistor, as shown, does not affect the degeneration, so that there is minimal drop across the input for full-scale on all ranges.



AN007479-12

FIGURE 10. This 100X amplifier has small and large signal bandwidth of 1 MHz. The LM11 greatly reduces offset voltage, bias current and gain error. Eliminating long recovery delay for greater than 100% overload requires direct coupling of A2 to input.



AN007479-13

FIGURE 11. An ammeter that has constant voltage drop across its input at full-scale, no matter what the range. It can have a reasonably-behaved output even with shorted inputs, yet a maximum drop of ten times the op amp noise voltage.

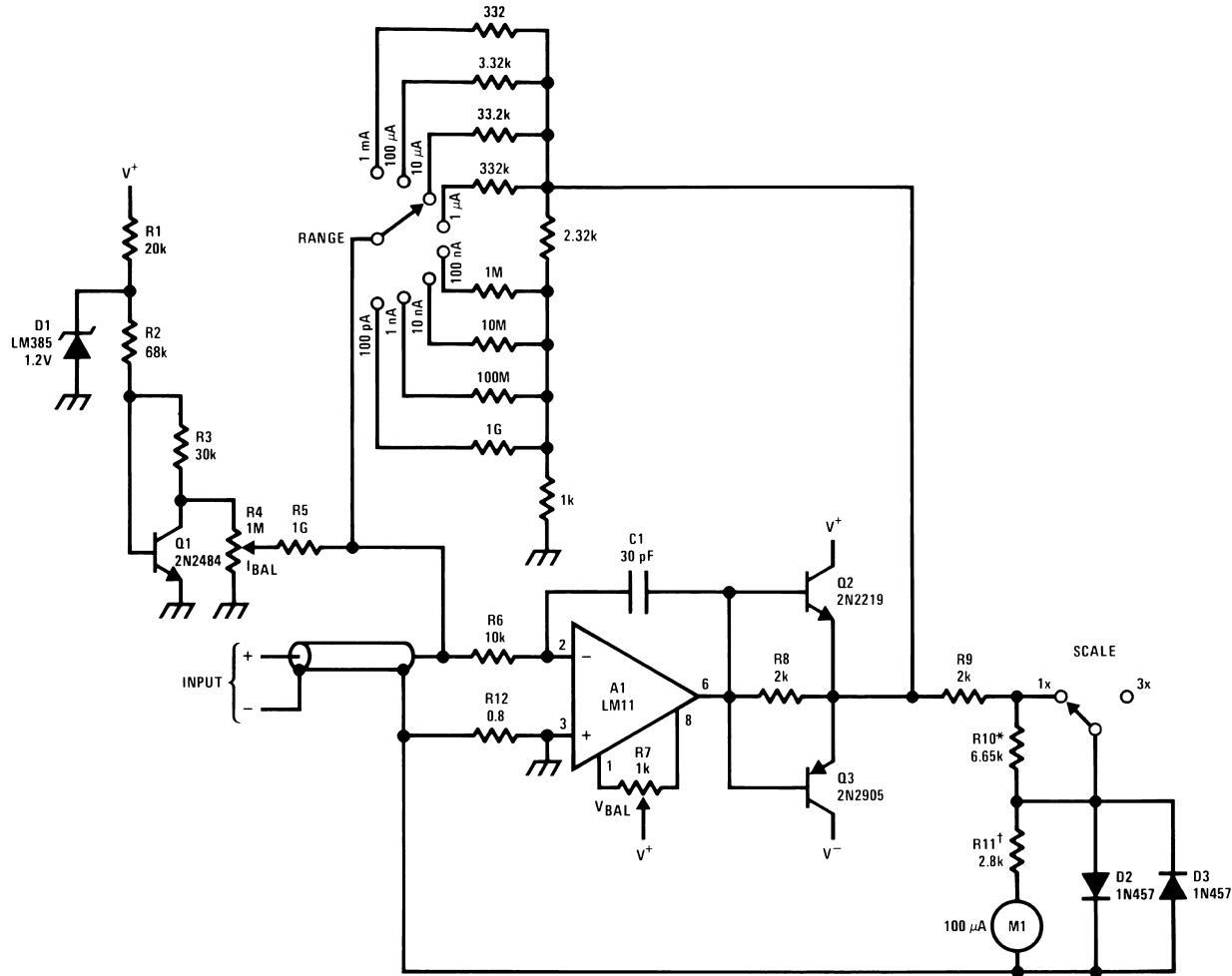
The complete meter circuit in *Figure 12* uses a different scheme. A floating supply is available so that the power ground and the signal ground can be separated with R12. At full-scale, the meter current plus the measured current flow

through this resistor, establishing the degeneration. This method has the advantage of allowing even-value range resistors on the lower ranges but increases degeneration as the measured current approaches the meter current.

Bias-current compensation is used to increase the meter sensitivity so there are two zeroing adjustments; current balancing, that is best done on the most sensitive range where it is needed, and voltage balancing that should be done with the inputs shorted on a range below 100 μA , where the degeneration is minimal.

With separate grounds, error could be made dependent on offset current. This would eliminate bias-current compensation at the expense of more complicated range switching.

The op amp input has internal, back-to-back diodes across it, so R6 is added to limit current with overloads. This type of protection does not affect operation and is recommended whenever more than 10 mA is available to the inputs. The output buffers are added so that input overloads cannot drag down the op amp output on the least-sensitive range, giving a false meter indication. These would not be required if the maximum input current did not approach the output current limit of the op amp.



AN007479-14

FIGURE 12. Current meter ranges from 100 pA to 3 mA, full-scale. Voltage across input is 100 μV at lower ranges rising to 3 mV at 3 mA. Buffers on op amp are to remove ambiguity with high-current overload. Output can also drive DVM or DPM.

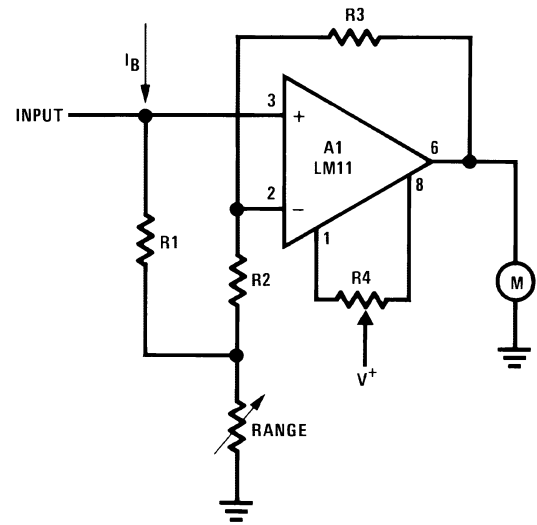
MILLIVOLTMETER

An ideal voltmeter has requirements analogous to those discussed for the ammeter, and *Figure 13* shows a circuit that will satisfy them. In the most-sensitive position, the range resistor is zero and the input resistance equals R1. As voltage measurement is desensitized by increasing the range resistor, the input resistance is also increased, giving the maximum input resistance consistent with zero stability

with the input open. Thus, at full-scale, the source will be loaded by whatever multiple of the noise current is required to give the desired open-input zero stability.

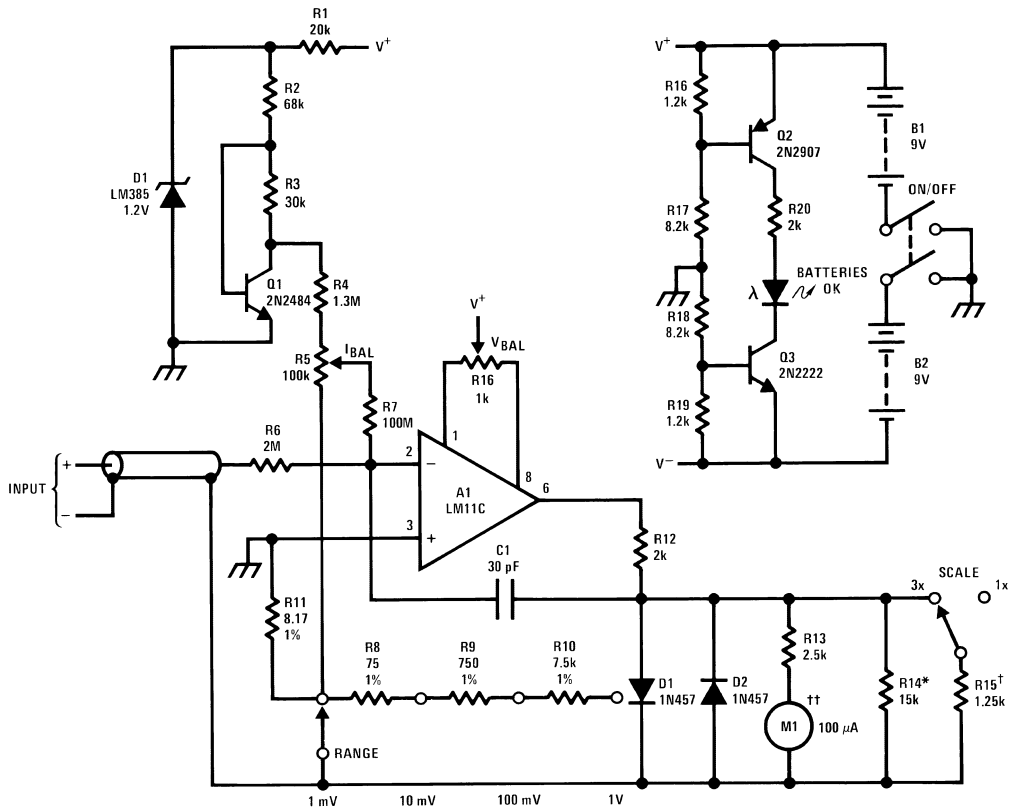
This technique is incorporated into the voltmeter circuit in *Figure 14* to give a 100 M Ω input resistance on the 1 mV scale rising to 300 G Ω on the 3V scale. The separation of power and signal grounds has been used here to simplify

bias-current compensation. Otherwise, a separate op amp would be required to bootstrap the compensation to the input.



AN007479-15

FIGURE 13. This voltmeter has constant full-scale loading independent of range. This can be only ten times the noise current, yet the output will be reasonably behaved for open input.



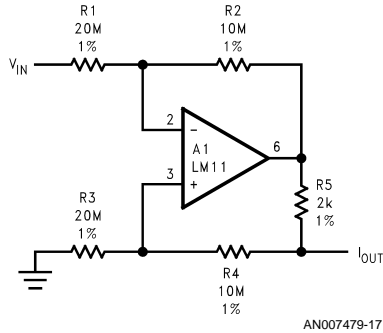
AN007479-16

*1x scale calibrate
 †3x scale calibrate
 ††includes reversing switch

FIGURE 14. High input impedance millivoltmeter. Input current is proportional to input voltage, about 10 pA at full-scale. Reference could be used to make direct reading linear ohmmeter.

The input resistor, R6, serves two functions. First, it protects the op amp input in the event of overload. Second, it insures that an overload will not give a false meter indication until it exceeds a couple hundred volts.

Since the reference is bootstrapped to the input, this circuit is easily converted into a linear, direct-reading ohmmeter. A resistor from the top of D1 to the input establishes the measurement current so that the voltage drop is proportional to the resistance connected across the input.



$$R1 = R3; R2 = R4$$

$$I_{OUT} = \frac{R2 V_{IN}}{R1 R5}$$

FIGURE 15. Output resistance of this voltage/current converter depends both on high value feedback resistors and their matching.

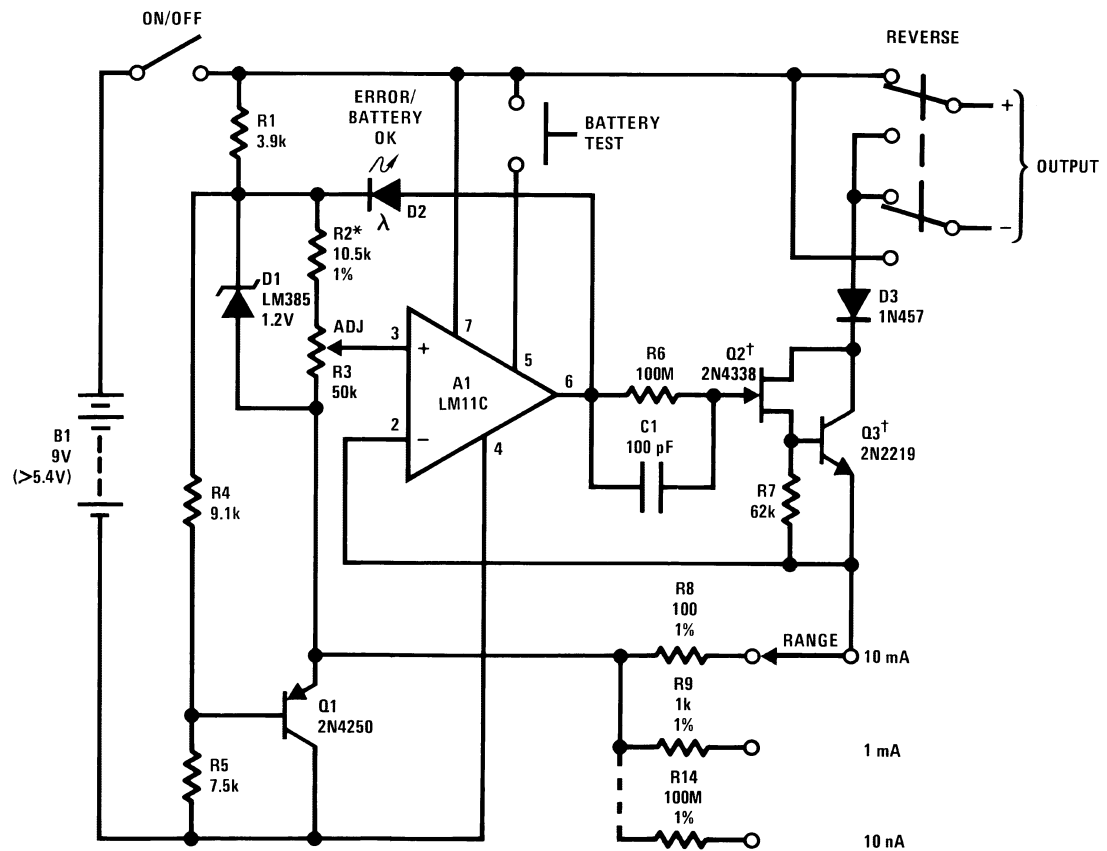
CURRENT SOURCES

The classical op amp circuit for voltage-to-current conversion is shown in *Figure 15*. It is presented here because the output resistance is determined by both the matching and the value of the feedback resistors. With the LM11, these resistors can be raised while preserving DC stability.

While the circuit in *Figure 15* can provide bipolar operation, better performance can be obtained with fewer problems if a unipolar output is acceptable. A complete, battery-powered current source suitable for laboratory use is given in *Figure 16* to illustrate this approach. The op amp regulates the voltage across the range resistors at a level determined by the voltage on the arm of the calibrated potentiometer, R3. The voltage on the range resistors is established by the current through Q2 and Q3, which is delivered to the output.

The reference diode, D1, determines basic accuracy. Q1 is included to insure that the LM11 inputs are kept within the common-mode range with diminishing battery voltage. A light-emitting diode, D2, is used to indicate output saturation. However, this indication cannot be relied upon for output-current settings below about 20 nA unless the value of R6 is increased. The reason is that very low currents can be supplied to the range resistors through R6 without developing enough voltage drop to turn on the diode.

If the LED illuminates with the output open, there is sufficient battery voltage to operate the circuit. But a battery-test switch is also provided. It is connected to the base of the op amp output stage and forces the output toward V^+ .



AN007479-18

*calibrate range

†select for $I_{CBO} \leq 100 \text{ pA}$

FIGURE 16. Precision current source has 10 nA to 10 mA ranges with output compliance of 30V to -5V. Output current is fully adjustable on each range with a calibrated, ten-turn potentiometer. Error light indicates output saturation.

Bias current compensation is not used because low-range accuracy is limited by the leakage currents of Q2 and Q3. As it is, these parts must be selected for low leakage. This should not be difficult because the leakage specified is determined by test equipment rather than device characteristics. It should be noted in making substitutions that Q2 was selected for low pinch-off voltage and that Q3 may have to dissipate 300 mW on the high-current range. Heating Q3 on the high range could increase leakage to where the circuit will not function for a while when switched to the low range.

LOGARITHMIC CONVERTER

A logarithmic amplifier that can operate over an eight-decade range is shown in *Figure 17*. Naturally, bias current compensation must be used to pick up the low end of this range. Leakage of the logging transistors is not a problem as long as Q1A is operated at zero collector-base voltage. In the worst case, this may require balancing the offset voltage of A1. Non-standard frequency compensation is used on A1 to obtain fairly uniform response time, at least at the high end of the range. The low end might be improved by optimizing C1. Otherwise, the circuit is standard.

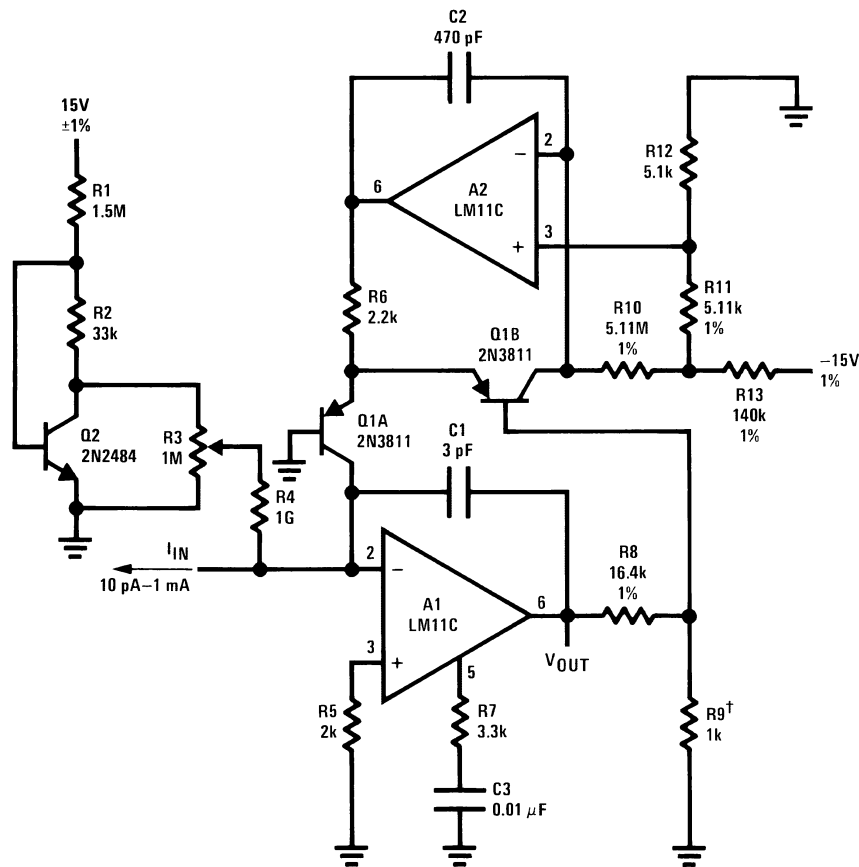
LIGHT METER

This logging circuit is adapted to a battery-powered light meter in *Figure 18*. An LM10, combined op amp and reference, is used for the second amplifier and to provide the regulated voltage for offsetting the logging circuit and powering the bias current compensation. Since a meter is the output indicator, there is no need to optimize frequency compensation. Low-cost single transistors are used for logging since the temperature range is limited. The meter is protected from overloads by clamp diodes D2 and D3.

Silicon photodiodes are more sensitive to infrared than visible light, so an appropriate filter must be used for photography. Alternately, gallium-arsenide-phosphide diodes with suppressed IR response are becoming available.

DIFFERENTIAL AMPLIFIERS

Many instrumentation applications require the measurements of low-level signals in the presence of considerable ground noise. This can be accomplished with a differential amplifier because it responds to the voltage between the inputs and rejects signals between the inputs and ground.

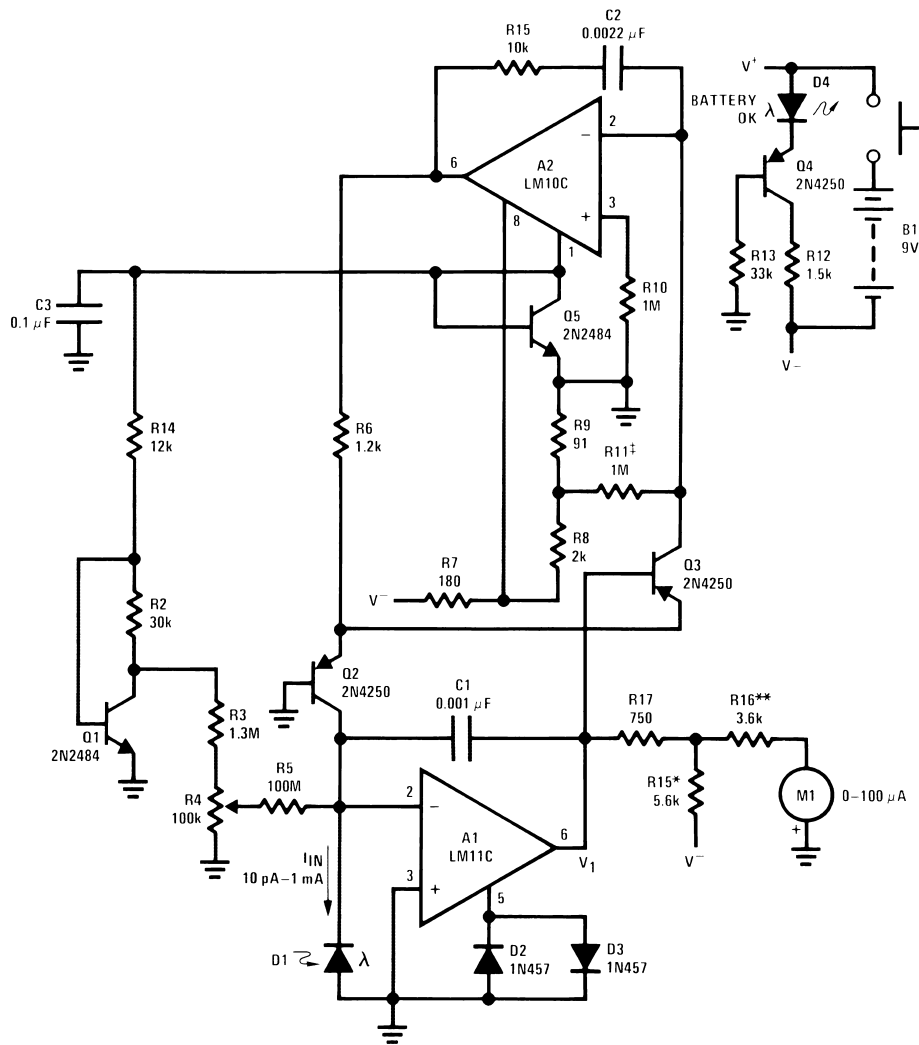


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330 ppm/ $^{\circ}$ C. Type Q209 available from Tel Labs, Manchester, N.H.Inc.,

- set R11 for $V_{OUT} = 0$ at $I_{IN} = 100$ nA
- set R8 for $V_{OUT} = 3$ V at $I_{IN} = 100$ μ A
- set R3 for $V_{OUT} = 4$ V at $I_{IN} = 10$ pA

FIGURE 17. Unusual frequency compensation gives this logarithmic converter a 100 μ s time constant from 1 mA down to 100 nA, increasing from 200 μ s to 200 ms from 10 nA to 10 pA. Optional bias current compensation can give 10 pA resolution from -55° C to $+100^{\circ}$ C. Scale factor is 1V/decade and temperature compensated.



AN007479-20

$V_1 = 0 @ I_{IN} = 100 \text{ nA}$
 $\dagger V_1 = -0.24 \text{ V} @ I_{IN} = 10 \text{ pA}$
 $*M_1 = 0 @ I_{IN} = 10 \text{ pA}$
 $**M_1 = \text{f.s.} @ I_{IN} = 1 \text{ mA}$

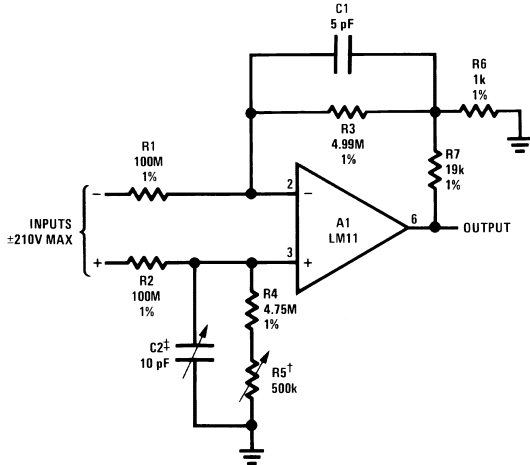
FIGURE 18. Light meter has eight-decade range. Bias current compensation can give input current resolution of better than $\pm 2 \text{ pA}$ over 15°C to 55°C .

Figure 19 shows the classic op amp differential amplifier connection. It is not widely used because the input resistance is much lower than alternate methods. But when the input common-mode voltage exceeds the supply voltage for the op amp, this cannot be avoided. At least with the LM11, large feedback resistors can be used to reduce loading without affecting DC accuracy. The impedances looking into the two inputs are not always the same. The values given equalize them for common-mode signals because they are

usually larger. With single-ended inputs, the input resistance on the inverting input is R_1 , while that on the non-inverting input is the sum of R_2 , R_4 and R_5 .

Provision is made to trim the circuit for maximum DC and AC common-mode rejection. This is advised because well matched high-value resistors are hard to come by and because unbalanced stray capacitances can cause severe deterioration of AC rejection with such large values. Particu-

lar attention should be paid to resistor tracking over temperature as this is more of a problem with high-value resistors. If higher gain or gain trim is required, R6 and R7 can be added.



AN007479-21

$V_S = \pm 15V$

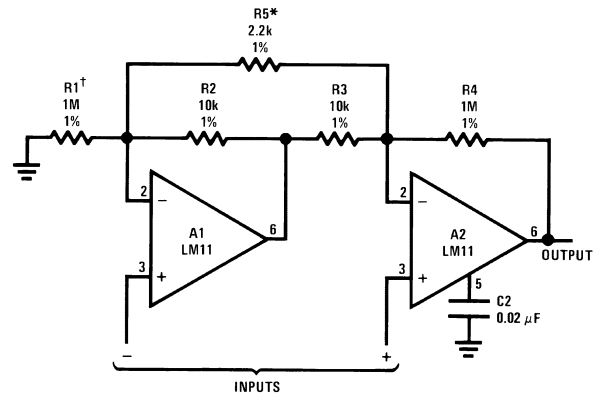
$$V_{CM(MAX)} = \frac{R1}{R3} V_{OUT(MAX)}$$

$$A_V = \frac{R3}{R1} \left(\frac{R6 + R7}{R6} \right)$$

ttrim for DC CMRR
trim for AC CMRR

FIGURE 19. This differential amplifier handles high input voltages. Resistor mismatches and stray capacitances should be balanced out for best common-mode rejection.

The simplest connection for making a high-input-impedance differential amplifier using op amps is shown in *Figure 20*. Its main disadvantage is that the common-mode signal on the inverting input is delayed by the response of A1 before being delivered to A2 for cancellation. A selected capacitor across R1 will compensate for this, but AC common-mode rejection will deteriorate as the characteristics of A1 vary with temperature.



AN007479-22

$f_2 \approx 10 \text{ Hz}$
*gain set
ttrim for DC CMRR

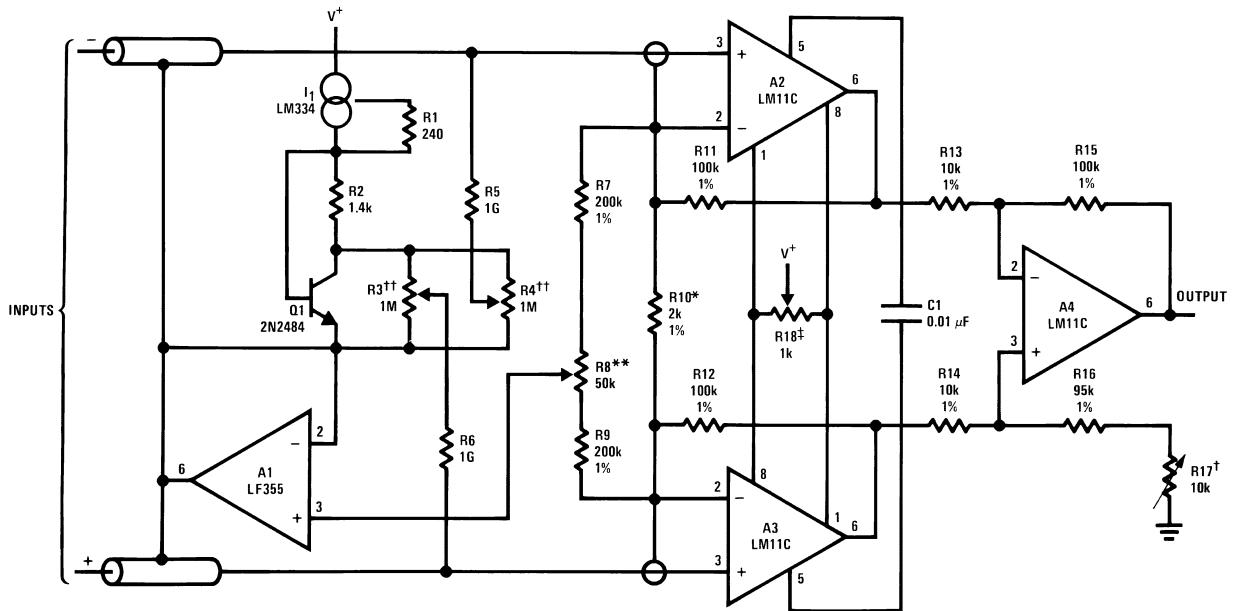
FIGURE 20. Two-op-amp instrumentation amplifier has poor AC common-mode rejection. This can be improved at the expense of differential bandwidth with C2.

When slowly varying differential signals are of interest, the response of A2 can be rolled off with C2 to reduce the sensitivity of the circuit to high frequency common-mode signals. If single-resistor gain setting is desired, R5 can be added. Otherwise, it is unnecessary.

A full-blown differential amplifier with extremely high input impedance is shown in *Figure 21*. Gain is fixed at 1000, but it can be varied with R10. Differential offset balancing is provided on both input amplifiers by R18.

The AC common-mode rejection is dependent on how well the frequency characteristics of A2 and A3 match. This is a far better situation than encountered with the previous circuit. When AC rejection must be optimized, amplifier differences as well as the effects of unbalanced stray capacitances can be compensated for with a capacitor across R13 or R14, depending on which side is slower. Alternately, C1 can be added to control the differential bandwidth and make AC common-mode rejection less dependent on amplifier matching. The value shown gives approximately 100 Hz differential bandwidth, although it will vary with gain setting.

A separate amplifier is used to drive the shields of the input cables. This reduces cable leakage currents and spurious signals generated from cable flexing. It may also be required to neutralize cable capacitance. Even short cables can attenuate low-frequency signals with high enough source resistance. Another balance potentiometer, R8, is included so that resistor mismatches in the drive to the bootstrapping amplifier can be neutralized. Adding the bootstrapping amplifier also provides a connection point, as shown, for bias-current compensation if the ultimate in performance is required.



AN007479-23

- ††current zero
- voltage balance
- *gain
- †DC CMRR
- **AC CMRR

FIGURE 21. High gain differential instrumentation amplifier includes input guarding, cable bootstrapping and bias current compensation. Differential bandwidth is reduced by C1 which also makes common-mode rejection less dependent on matching of input amplifiers.

As can be seen in *Figure 23*, connecting the input amplifiers as follows simplifies the circuit considerably. But single resistor gain control is no longer available and maximum bandwidth is less with all the gain developed by A3. Resistor matching is more critical for a given common-mode rejection, but AC matching of the input amplifier is less a problem. Another method of trimming AC common-mode rejection is shown here.

INTEGRATOR RESET

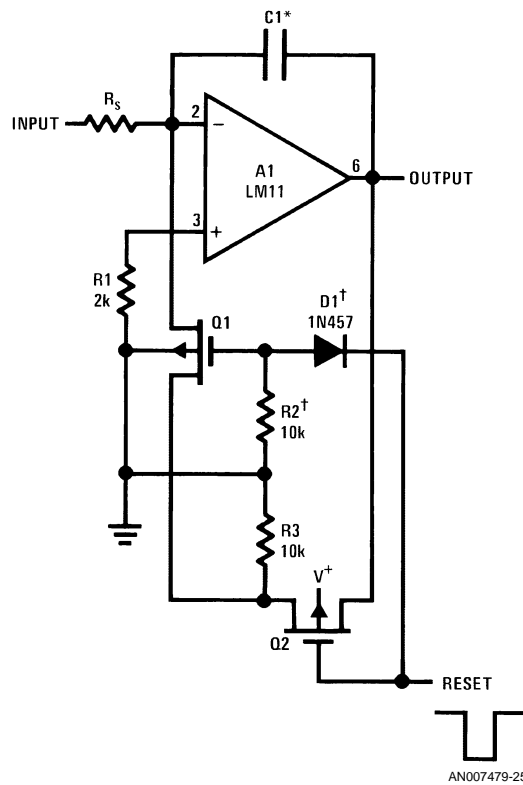
When pursuing the ultimate in performance with the LM11, it becomes evident that components other than the op amp can limit performance. This can be the case when semiconductor switches are used. Their leakage easily exceeds the bias current when elevated temperatures are involved.

The integrator with electrical reset in *Figure 22* gives a solution to this problem. Two switches in series are used to

shunt the integrating capacitor. In the off state, one switch, Q2, disconnects the output while the other, Q1, isolates the leakage of the first. This leakage is absorbed by R3. Only the op amp offset appears across the junctions of Q1, so its leakage is reduced by two orders of magnitude.

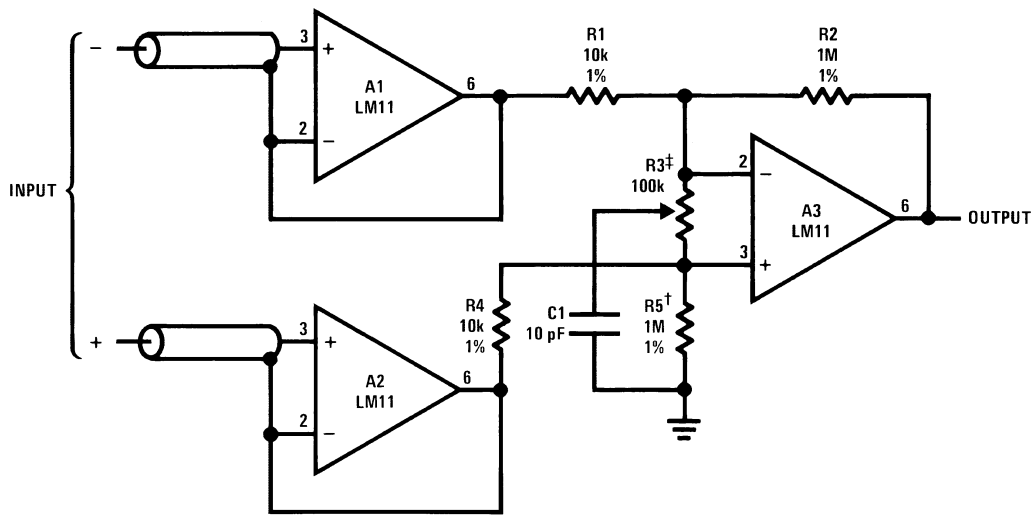
A junction FET could be used for Q1 but not for Q2 because there is no equivalent to the enhancement mode MOSFET. The gate of a JFET must be reverse biased to turn it off and leakage on its output cannot be avoided.

MOS switches with gate-protection diodes are preferred in production situations as they are less sensitive to damage from static charges in handling. If used, D1 and R2 should be included to remove bias from the internal protection diode when the switch is off.



*polystyrene recommended
 †required if protected gate switch is used

FIGURE 22. Reset is provided for this integrator and switch leakage is isolated from the summing junction. Greater precision can be provided if bias-current compensation is included.



AN007479-24

R1 = R3; R2 = R4

$$A_v = \frac{R_2}{R_1}$$

†trim for DC CMRR
 set for AC CMRR

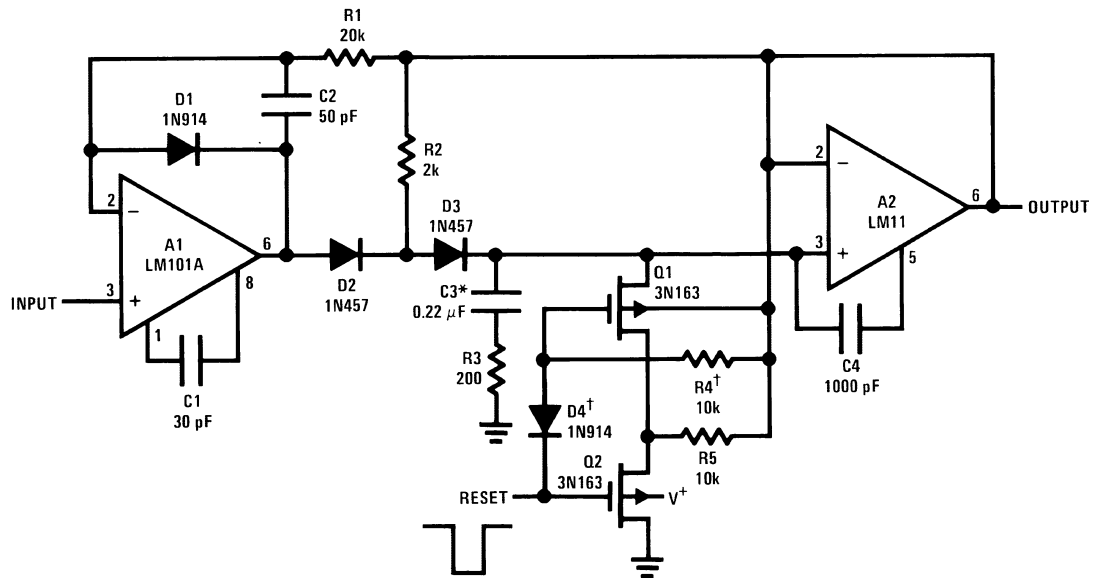
FIGURE 23. For moderate-gain instrumentation amplifiers, input amplifiers can be connected as followers. This simplifies circuitry, but A3 must also have low drift.

PEAK DETECTOR

The peak detector in *Figure 24* expands upon this idea. Isolation is used on both the peak-detecting diode and the reset switch. This particular circuit is designed for a long hold interval so acquisition is not quick. As might be expected from an examination of the figure, frequency compensation of an op amp peak detector is not exactly straightforward.

OVEN CONTROLLER

The LM11 is quite useful with slow servo systems because impedance levels can be raised to where reasonable capacitor values can be used to effect loop stabilization without affecting accuracy. An example of this is shown in *Figure 25*. This is a true proportional controller for a crystal oven.



AN007479-26

300 μ s min single pulse

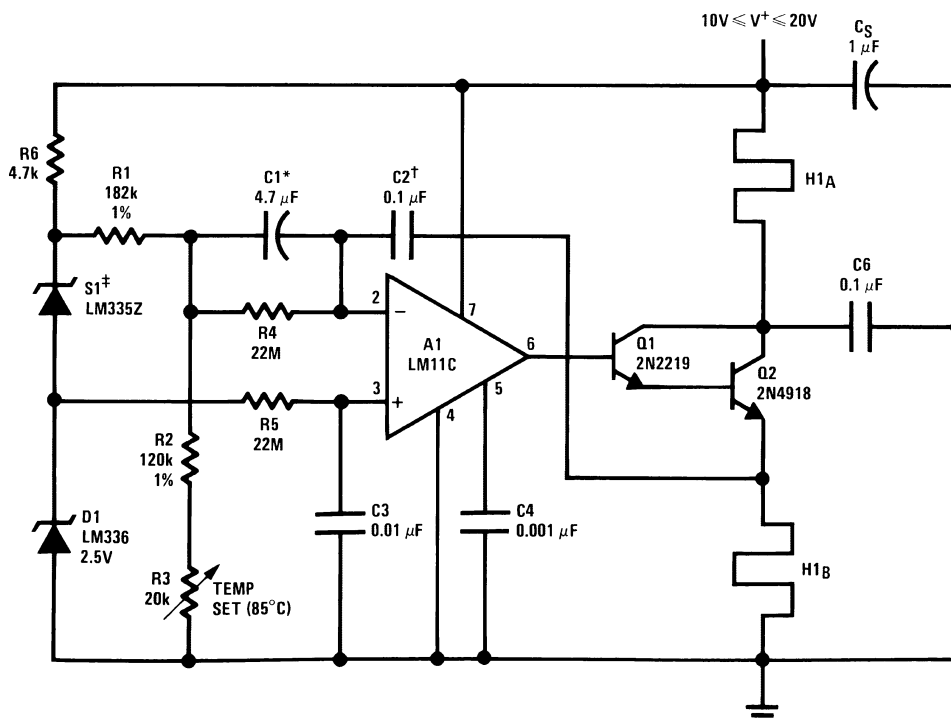
200 μ s min repetitive pulse

300 Hz max sine wave error < 5 mV

†required if Q1 has gate protection diode

*polystyrene or Teflon

FIGURE 24. A peak detector designed for extended hold. Leakage currents of peak-detecting diodes and reset switch are absorbed before reaching storage capacitor.



AN007479-27

*solid tantalum

†mylar

close thermal coupling between sensor and oven shell is recommended

FIGURE 25. Proportional control crystal oven heater uses lead/lag compensation for fast settling. Time constant is changed with R4 and compensating resistor R5. If Q2 is inside oven, a regulated supply is recommended for 0.1°C control.

Temperature sensing is done with a bridge, one leg of which is formed by an IC temperature sensor, S1, and a reference diode, D1. Frequency stabilization is done with C2 providing a lag that is finally broken out by C1. If the control transistor, Q2, is put inside the oven for maximum heating efficiency, some level of regulation is suggested for the heater supply when precise control is required. With Q2 in the oven, abrupt supply changes will alter heating, which must be compensated for by the loop. This takes time, causing a small temperature transient.

Because the input bias current of the LM11 does not increase with temperature, it can be installed inside the oven for best performance. In fact, when an oven is available in a piece of equipment, it would be a good idea to put all critical LM11s inside the oven if the temperature is less than 100°C.

AC AMPLIFIER

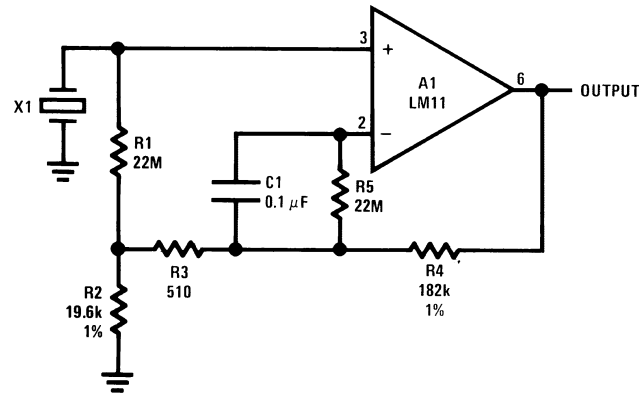
Figure 26 shows an op amp used as an AC amplifier. It is unusual in that DC bootstrapping is used to obtain high input resistance without requiring high-value resistors. In theory, this increases the output offset because the op amp offset voltage is multiplied by the resistance boost.

But when conventional resistor values are used, it is practical to include R5 to eliminate bias-current error. This gives less output offset than if a single, large resistor were used. C1 is included to reduce noise.

STANDARD CELL BUFFER

The accuracy and lifetime of a standard cell deteriorate with loading. Further, with even a moderate load transient, recovery is measured in minutes, hours or even days. The circuit in Figure 27 not only buffers the standard cell but also disconnects it in the event of malfunction.

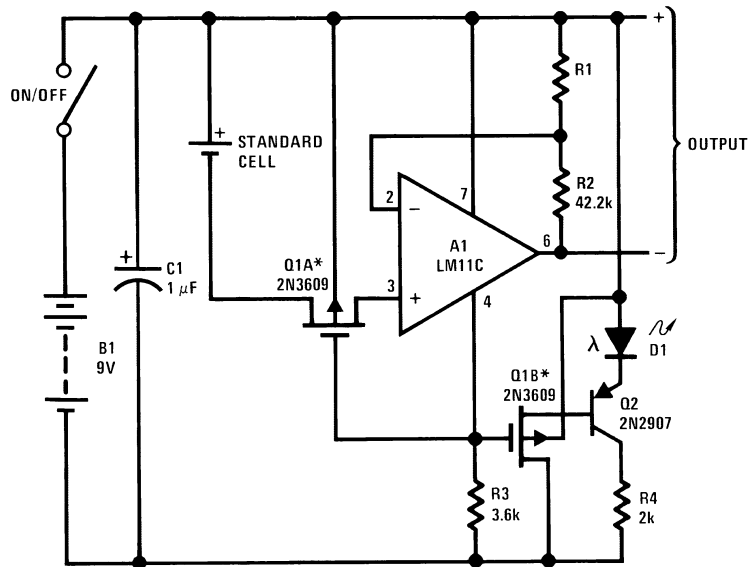
The fault threshold is determined by the gate turn-on voltage of Q1. As the voltage on the gate approaches the threshold either because of low battery voltage or excessive output loading, the MOS switch will begin to turn off. At the turn off threshold, the output voltage can rise because of amplifier bias current flowing through the increasing switch resistance. Therefore, a LED indicator is included that extinguishes as the fault condition is approached. The MOS threshold should be higher than the buffer output so disconnect and error indicator operates before the output saturates.



AN007479-28

$$R_{IN} = R1 \left(1 + \frac{R2}{R3} \right); \quad A_V = \frac{R2 + R3 + R4}{R2 + R3}$$

FIGURE 26. A high input impedance AC amplifier for a piezoelectric transducer. Input resistance of 880 MΩ and gain of 10 is obtained.



AN007479-29

*cannot have gate protection diode; $V_{TH} > V_{OUT}$

FIGURE 27. Battery powered buffer amplifier for standard cell has negligible loading and disconnects cell for low supply voltage or overload on output. Indicator diode extinguishes as disconnect circuitry is activated.

CONCLUSIONS

Although the LM11 does not provide the ultimate in performance in either offset voltage or bias current for nominal room temperature applications, the combination offered is truly noteworthy. With significant temperature excursions, the results presented here are much more impressive. With full-temperature-range operation, this device does represent the state of the art when high-impedance circuitry is involved.

Combining this new amplifier with fast op amps to obtain the best features of both is also interesting, particularly since the composite works well in both the inverting and non-inverting modes. However, making high-impedance circuits fast is no simple task. If higher temperatures are not involved, using the LM11 to reduce the offset voltage of a FET op amp without significantly increasing bias current may be all that is required.

An assortment of measurement and computational circuits making use of the unique capabilities of this IC were presented. These circuits have been checked out and the results should be of some value to those working with high impedances. These applications are by no means all-inclusive, but they do show that an amplifier with low input current can be used in a wide variety of circuits.

Although emphasis was on high-performance circuits requiring adjustments, the LM11 will see widest usage in less demanding applications where its low initial offset voltage and bias current can eliminate adjustments.

ACKNOWLEDGEMENT

The authors would like to thank Dick Wong for his assistance in building and checking out the applications described here.

*See Addendum that follows this Application Note.

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