Design linear circuits for 5V operation

Although you often have to include linear circuit functions in predominantly digital systems, you no longer need to power the linear circuits from separate supplies. New components and design techniques let you power linear circuitry from the 5V logic rail.

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When you incorporate linear circuit functions in your digital system, you can avoid adding an extra supply just to power them. Instead, you can use analog components that furnish high performance when connected to the system's 5V logic supply.

Consider, for example, the LT1014 quad (or the LT1013 dual) op amps, which can operate from 5V supplies but whose specs compare favorably with those of the best ±15V op amps, and the LT1017/1018 Series dual comparators, which combine low power and high dc precision with speeds adequate for most applications. The common-mode range of these devices includes ground, which makes it easy to operate them from a single supply. (The nearby **table** summarizes the devices' specs.) In addition, the op amp's outputs can swing very close to the ground rail.

You can use such components to design signal-conditioning circuitry that operates from a 5V supply. **Fig 1** illustrates a scheme that provides complete, linearized signal conditioning for a platinum RTD that has a highly linear positive temperature coefficient. One side of the RTD sensor is grounded to minimize noise problems. The LT1004 2.5V reference establishes A_1's operating point, and A_1's output serves as a reference for the current source (Q_1 and Q_2).

The RTD is a constant-current device, so its voltage drop varies as a function of its resistance. A slight nonlinearity in the sensor's resistance curve causes several degrees of error over the circuit's 0 to 400°C operating range. A_2 amplifies R_P's output and simultaneously provides nonlinearity correction by feeding a portion of its (A_2's) output back to A_1's input via the 10-kΩ/250-kΩ resistive divider. Consequently, the current-source output shifts with R_P's operating point, compensating for sensor nonlinearity to within ±0.05°C.

A_3, which is also referenced to the LT1004, conditions the offset signal at A_2's inverting input so that 0V at A_2's output is equal to 0°C. The resistive divider at A_4's noninverting input establishes circuit gain.

You can calibrate this circuit by substituting a precision decade resistance box (eg, GenRad's 1432K) for R_P. You set the box for 100Ω (0°C value) and adjust the offset trim for a 0V output. Then you set the box for 154.26Ω (140°C), and adjust the gain trim for a 1.4V output. Finally, you set the decade box at 249Ω (400°), and trim the linearity adjustments until the circuit output equals 4V. By repeating this sequence until you fix all three points, you'll establish a total error range of

BASIC SPECS FOR THE LT1014 AND LT1017/1018 COMPONENTS

PARAMETER	LT1014	LT1017/1018
E_{OS} (μV)	150	500
E_{OS} TC (μV/°C)	2	5
BIAS CURRENT (nA)	20	100
SUPPLY CURRENT (μA)	500	60 (1017) 240 (1018)
GAIN	1.5×10^6	10^6
COMMON-MODE RANGE (V)	0 TO (V_S-1.5V)	0 TO (V_S-1.4V)
SUPPLY VOLTAGE (V_S)	3.4 TO 40V	1.2 TO 36V
NOISE (0.1 TO 10 Hz)	0.55 μV p-p	—
RESPONSE TIME (μSEC)	—	1 (1018)
OUTPUT CURRENT	—	25 mA PULL DOWN 60 μA PULL UP
OUTPUT SWING (V) NO LOAD 600Ω LOAD	 0.025 TO (V_S-1V) 0.01 TO (V_S-1.6V)	—

±0.05°C max over the entire temperature range.

Although the resistance values in **Fig 1** are for a sensor with a nominal 0°C resistance of 100Ω, you can use sensors with different nominal resistance values by factoring in the deviation from 100Ω. This deviation, specified by the manufacturer for each individual sensor, is an offset term that's caused by winding tolerances during RTD fabrication. The gain slope of platinum is primarily fixed by the purity of the material, and it represents a very small error term.

Fig 2 illustrates a second signal-conditioning circuit, this one for a thermocouple. The design features cold-junction compensation; one leg of the thermocouple is grounded, which minimizes noise problems. One switch section of the LTC1043 combines the compensation network's differential output with the thermocouple's output at the input of the LTC1052.

The 1043's second switch section generates a small negative potential, which allows the 1052's output stage to operate as a class A amplifier for low-level outputs, permitting a swing to zero volts.

The table in **Fig 2** lists optimum resistance values for various thermocouple types. By adjusting the R_F/R_I divider ratio, you can set output scaling to whatever slope you desire. Over a 0 to 60°C range, cold-junction compensation holds to within ±1°C.

Condition a pressure-transducer output

You can also signal-condition the bridge output of a strain-gauge pressure transducer (**Fig 3**). Despite the 2-amplifier circuit's simplicity, it provides an auxiliary

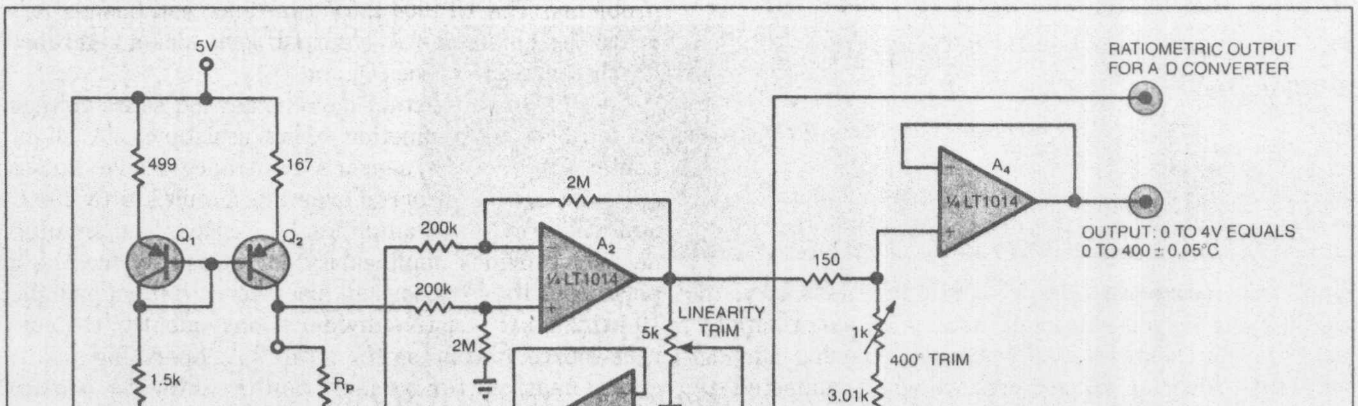


Fig 1—To minimize noise problems, this signal-conditioning design scheme grounds one side of the RTD sensor.

Theoretically, transducers with single-ended outputs don't require a differential-input instrumentation-type amplifier.

ratio output for a monitoring A/D converter, and it minimizes the need for precision resistors.

A_1 provides bias for the LT1044 positive-to-negative converter. As the 1044's output pulls the bridge output negative, it balances A_1's input at 0V. Thanks to this local-loop action, a single-ended amplifier (A_2) can handle the bridge's differential output.

A_2's gain is set to provide the desired output scale factor. The RC network at the amplifier's input provides noise filtering. Because the LT1004 reference provides the bridge drive, power-supply shifts have no effect on A_2's output.

To calibrate this circuit, apply (or electrically simulate) 0 psi, and trim the zero adjustment for 0V output. Next, apply (or electrically simulate) 350 psi, and trim the gain for 3.5V output. Repeat these adjustments until you fix both points.

Taking care of detection needs

The circuit in **Fig 4a** may look like just another scheme for signal conditioning; however, besides signal conditioning, it performs a more complex mathematical operation: It monitors the methane level detected by the specified transducer (the circuit's frequency output is directly proportional to the methane level). The transducer output varies approximately as

$$\frac{1}{\sqrt{\text{CONCENTRATION}}}, \qquad (1)$$

and the circuit linearizes this function.

A_1 converts the sensor's resistance value (vs methane concentration) to a voltage and feeds A_2. The LT1004 serves as a reference. The exponential relationship between a transistor's V_{be} and its collector current is used to generate a current in Q_3's collector that's proportional to the square of A_2's input current. This operation compensates for the sensor's square root term. Q_3's collector current establishes the operating point for the oscillator (A_3, A_4).

A_3, an integrator, generates a positive-going linear ramp (trace A in **Fig 4a**). The summing point at A_4 (trace B) compares the ramp with Q_3's current. A_4 is configured as a current-summing comparator (the diode-bound feedback network minimizes delay caused

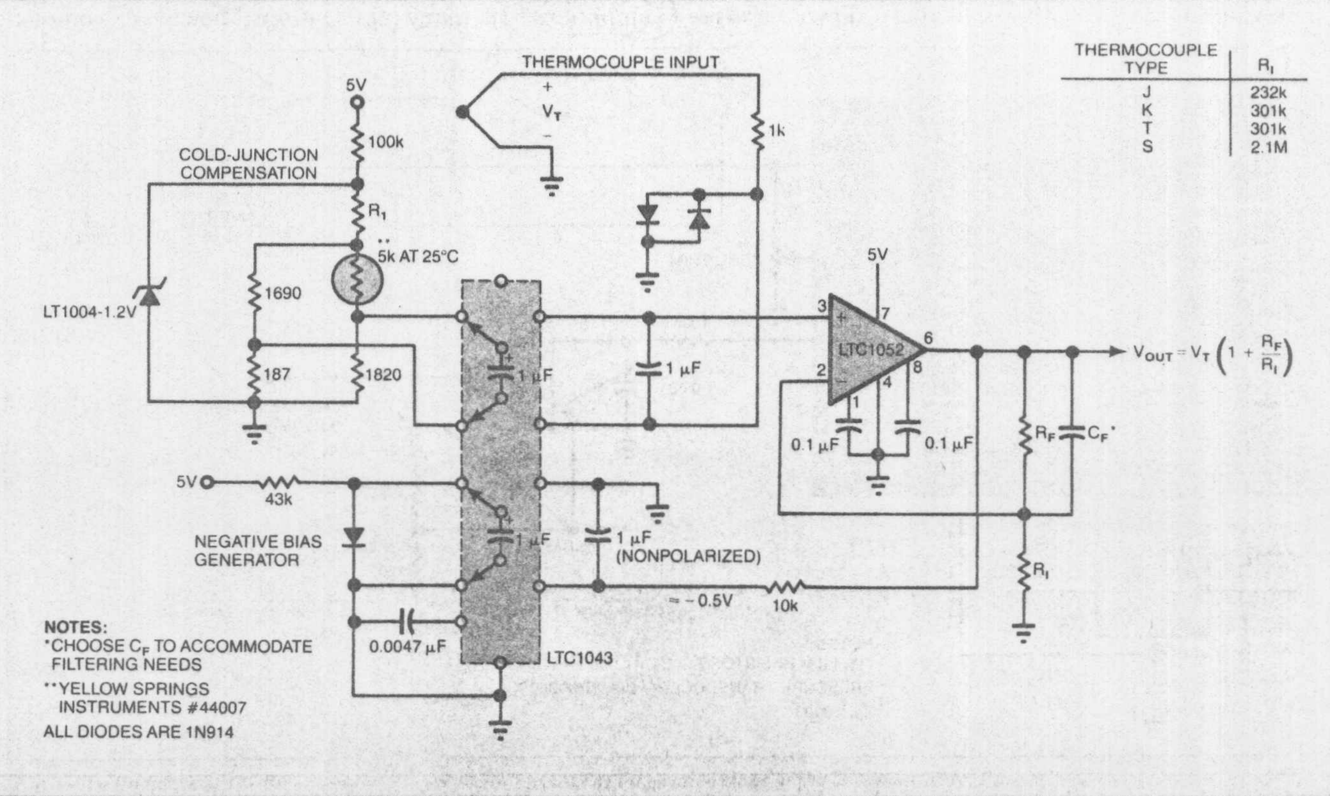


Fig 2—Featuring cold-junction compensation, this signal conditioner works with a variety of thermocouple types.

Clean supplies are not the rule in the digital world, they are the exception.

by output slew time). When the ramp forces the summing point positive, A_4's output (trace C) swings negative. The output of CMOS inverter A (trace D) goes high, turning on the C4016 switch to reset the integrator.

At the same time, the output of inverter B (trace E) goes low, supplying positive ac feedback to A_4's noninverting input (trace F). When the positive feedback decays, A_4's output goes high, enabling the integrator, and the entire cycle repeats.

Q_3's collector current determines how long A_3 generates the ramp before it's reset by A_4. The ramp time is directly proportional to Q_3's collector current; thus, the oscillation frequency is inversely related to the current.

The transfer function for the overall circuit takes the form $1/X^2$, which, in theory, should linearize the sensor's output. In practice, the sensor response deviates slightly from the ideal, and is actually

$$\frac{1}{\sqrt[1.9]{\text{CONCENTRATION}}}. \quad (2)$$

The reset time constants at A_4's input introduce enough oscillator down time to partially compensate for the deviation. Because the oscillator's down time retards frequency shifts, it affects the oscillator's high frequencies, providing a first-order compensation. The overall linearization achieved (**Fig 4b**) is within the sensor's manufacturing tolerances. The slight bump in the circuit's response curve is caused by the mismatch between the denominator terms of the sensor-response expression (**Eq 1**) and the circuit's transfer function. The down-time correction in the oscillator smoothes this error out above 4000 ppm.

The LT1044 voltage converter generates a negative supply voltage directly from the 5V rail. The sensor's heater is powered from the 5V rail, as the manufacturer specifies. To calibrate the circuit, place the sensor in a 1000-ppm-methane environment and adjust the 5-kΩ trimmer to develop a 100-Hz output. Circuit accuracy from 500 to 10,000 ppm is limited by the sensor's 10% specification.

Taking care of noise problems

Although many transducer-based analog functions are compatible with 5V-power-supply operation, you shouldn't overlook a source of possible trouble: noise. Theoretically, transducers with single-ended outputs don't require differential-input instrumentation-type amplifiers. In many applications, however, common-

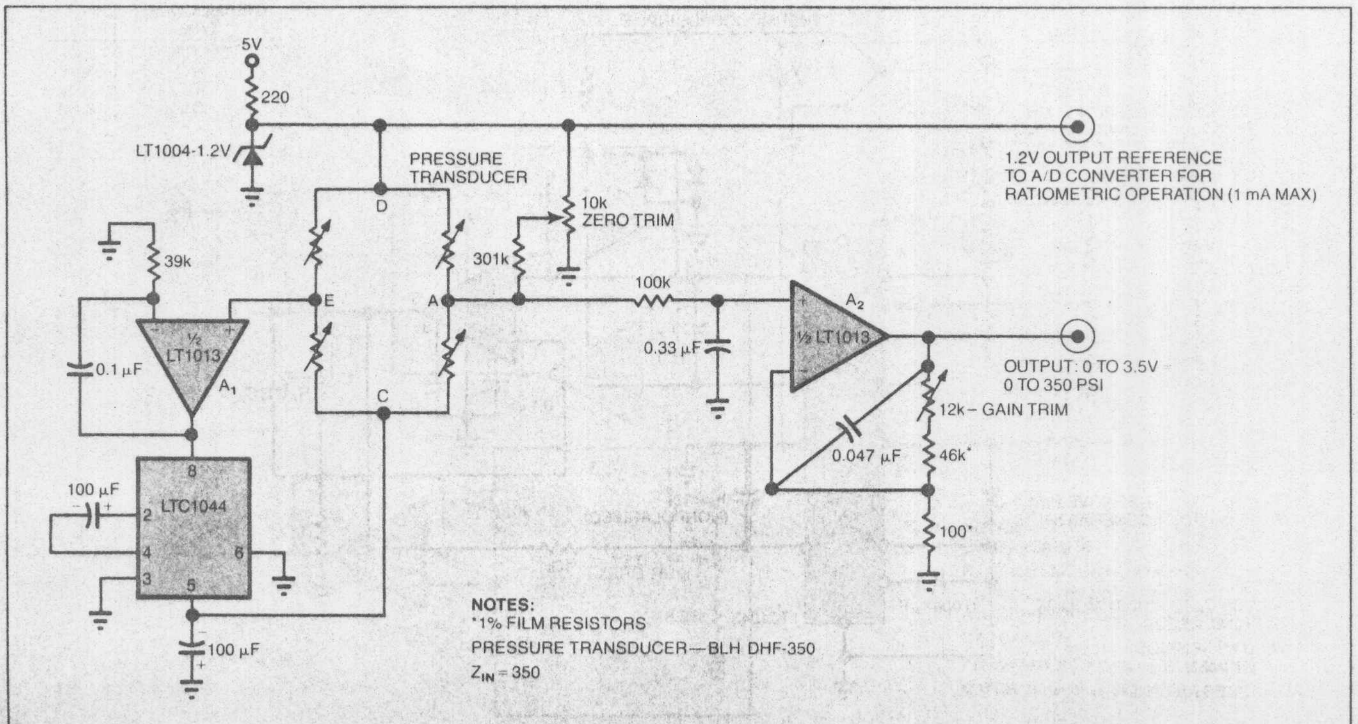
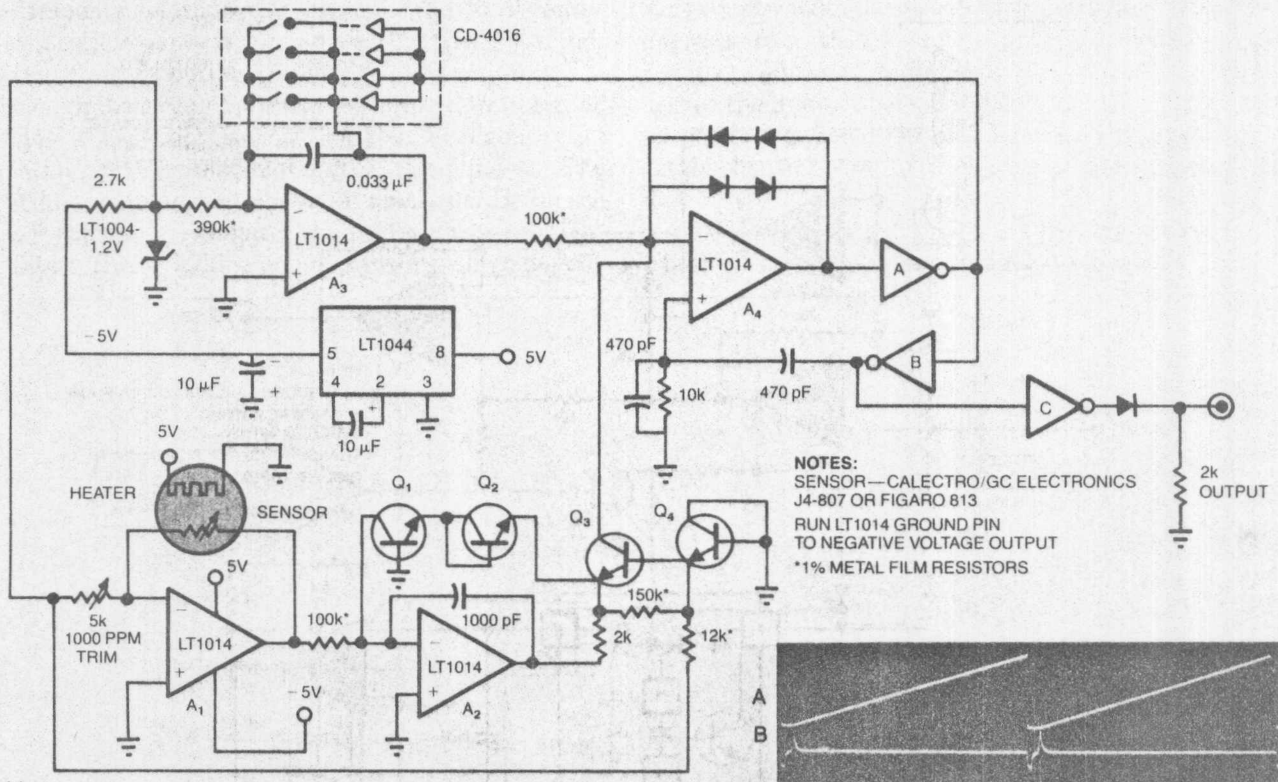


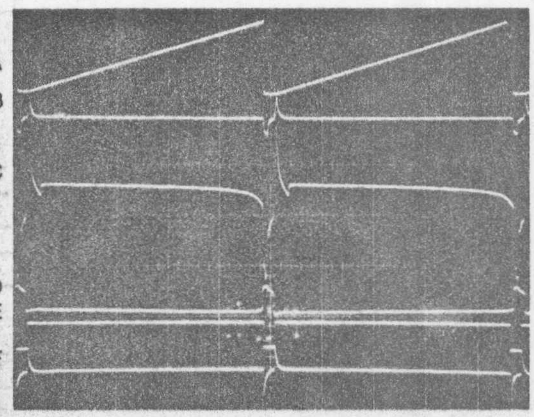
***Fig 3—Despite its simplicity,** this circuit, which conditions a strain-gauge pressure transducer's output, minimizes the need to use precision resistors; in fact, you don't have to worry about resistor matching at all.*

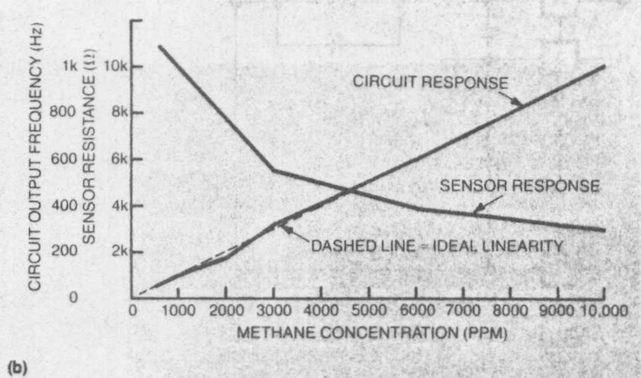
mode noise is often larger than the signal of interest, so most transducer-based systems do employ instrumentation-type amplifiers.

If you need such an amplifier, you'll have to build your own, because no commercially available instrumentation amplifier will function from a 5V supply. You can use the circuits in **Fig 5** to build an instrumentation amplifier. The circuits feature input protection, filtering capability, and a shield-driver output.

In **Fig 5a,** A_1, A_2, and A_3 provide the differential-input to single-ended-output conversion; R_G sets the gain. Because of the offset and drift performance of this







TRACE	VERTICAL	HORIZONTAL
A	200 mV/DIV	200 µSEC/DIV
B	5V/DIV	200 µSEC/DIV
C	2V/DIV	200 µSEC/DIV
D	20V/DIV	200 µSEC/DIV
E	20V/DIV	200 µSEC/DIV
F	10V/DIV	200 µSEC/DIV

***Fig 4—To linearize transducer output,** this detector circuit **(a)** takes advantage of the exponential relationship between a transistor's V_{be} and its collector current. Overall circuit linearization **(b)** is within the sensor's manufacturing tolerances.*

One key to achieving good design results is to consider power bus routing an integral part of the signal-processing chain.

circuit, the circuit will accommodate low-level transducers such as thermocouples and strain gauges.

The RC networks at the inputs filter out noise and 60-Hz pickup—the LT1014 is never exposed to high-frequency common-mode noise. The transistors and Schottky clamp diodes combine with the 100-kΩ resistors to prevent high voltage spikes or faults, which are common in industrial environments, from doing any damage to the circuit. To reduce the effects of input-cable capacitance, you can use A_4 to drive the cable shield at the input common-mode voltage level, which is derived from the output of the input amplifiers.

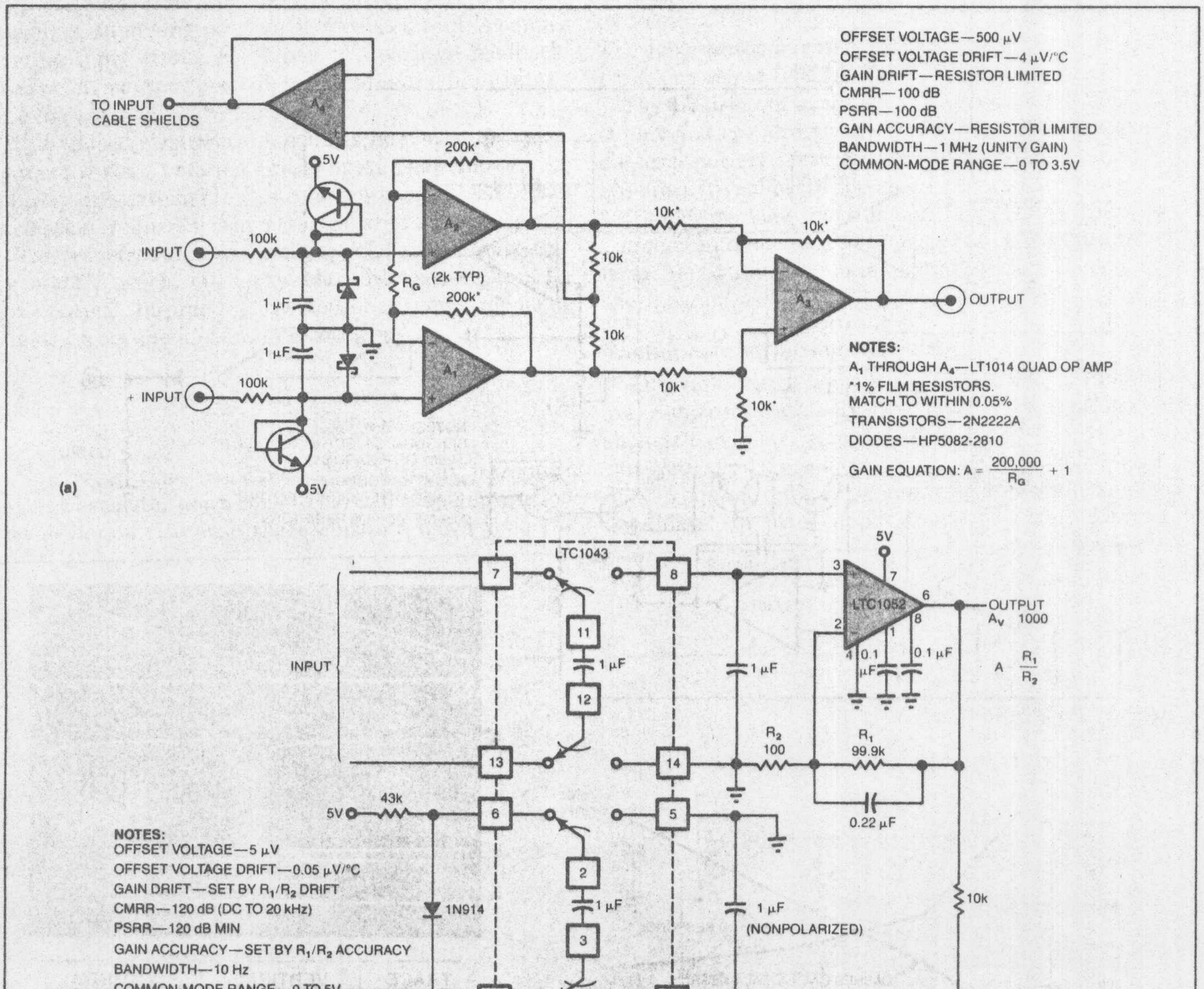


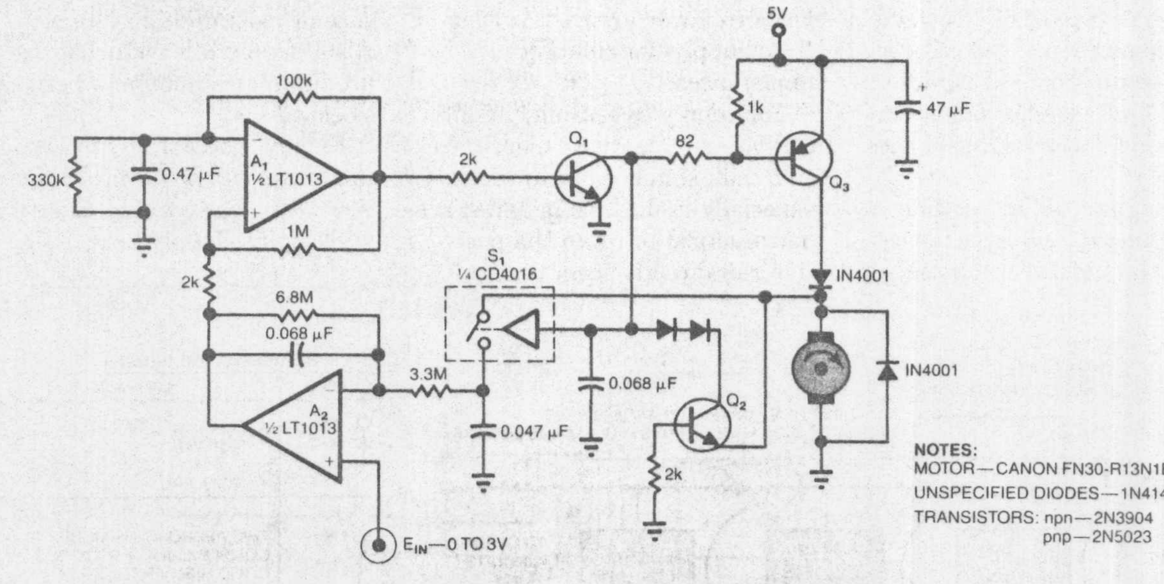
***Fig 5—Input protection and filtering capability** are key features of these instrumentation-type amplifiers. What circuit you select depends on your application: You can opt for wide bandwidth (a) or trade bandwidth for high dc precision (b).*

If you need a wider bandwidth, use the circuit in **Fig 5a**. If you need higher dc precision, use the circuit in **Fig 5b**, in which one section of an LTC1043 switched-capacitor building block provides the differential-to-single-ended conversion by alternately commutating a 1-μF capacitor between the circuit input and the LTC1052's input. This scheme allows the 1052 to make measurements referenced to ground. The 1043's other switch section generates a small negative potential, allowing the 1052 output to swing all the way to 0V.

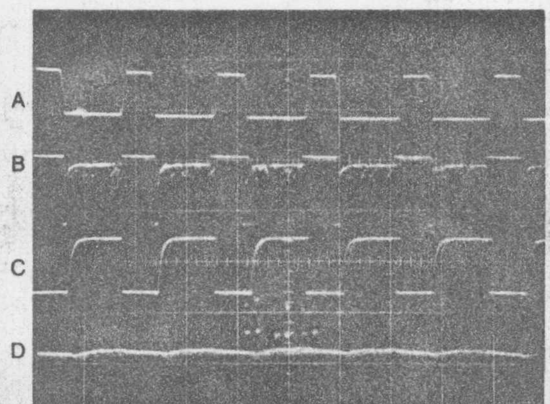
Although the circuit bandwidth is limited to 10 Hz, dc precision is excellent, surpassing that available in all monolithic ±15V instrumentation amplifiers. The LTC1043's switching action, set at about 400 Hz by the 0.01-μF capacitor, simulates the performance of a low-pass filter. The switching action provides a high degree of noise rejection—CMRR specs in excess of 120 dB at 20 kHz.

You can also power circuits like motor controllers, current-loop transmitters, dc/dc converters, limit comparators, and A/D converters from the 5V logic rail. Of course, these applications will require novel designs. The circuit in **Fig 6**, for instance, provides a means of servo-controlling the speed of a dc motor. Because it operates from the 5V logic supply, this design doesn't require additional motor-drive supplies. The circuit senses the motor's back EMF to determine the motor's speed. It uses the difference between the speed and an established set point to close a sampling loop around the motor.

Specifically, A_1 generates a pulse train (trace A in **Fig 6**). When A_1's output is high, it turns on Q_1, and Q_3



NOTES:
MOTOR—CANON FN30-R13N1B
UNSPECIFIED DIODES—1N4148
TRANSISTORS: npn—2N3904
pnp—2N5023



TRACE	VERTICAL	HORIZONTAL
A	5V/DIV	1 mSEC/DIV
B	5V/DIV	1 mSEC/DIV
C	5V/DIV	1 mSEC/DIV
D	5V/DIV (AC COUPLED)	1 mSEC/DIV

Fig 6—Offering a 20-rpm to full-speed control range, this circuit senses a motor's back EMF to determine the motor's speed.

Galvanically isolating the linear circuits is probably the most effective way to deal with digital-supply noise.

Using logic supplies

Operating linear components from logic supplies can be difficult because of the fast clocking and transient high currents characteristic of digital systems. One key to achieving good design results when you're running linear components from 5V supplies is to consider power-bus routing as an integral part of the signal-processing chain.

As **Fig A** shows, supply-rail impedances cause both dc and ac errors at various points in a system. This is true of any power distribution scheme, but it's especially troublesome in digitally oriented systems where fast current spiking and clock harmonics are present.

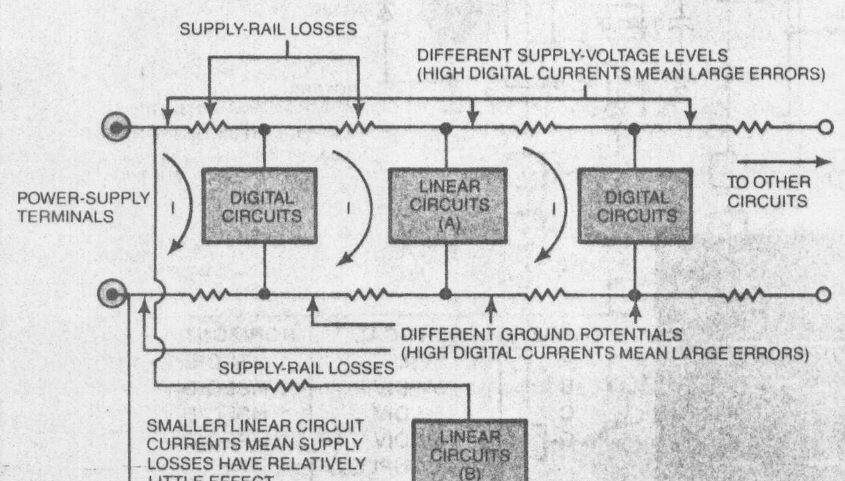
Circuitry located at position A, for example, will experience appreciable positive rail noise, and the relatively high currents returning through conductor impedances will corrupt the ground potential. Although you can reduce positive rail noise by bypassing the supply, ground-potential uncertainty can still cause unacceptable errors.

Locating linear circuits as shown at position B eliminates the digitally related currents and reduces both positive- and ground-rail problems. The linear device's lower operating current leads to fewer errors resulting from supply-distribution impedances.

For supply bypassing, LC filters are substantially more effective than simple capacitors are, especially in designs in which it's not practical to route the positive rail directly from the supply. When you use RC filtering, voltage drops across the resistor, but this is often acceptable because most linear components have low power requirements.

In many cases it's impossible to develop a clean power supply for the linear components. In such circumstances, it may be possible to have all noise-sensitive linear circuit operations occur between system clock pulses. This approach takes advantage of the synchronous nature of most digital systems. Also, supply bus disturbances are often at a minimum between clock pulses.

Finally, galvanically isolating the linear circuits is probably the most effective way to deal with digital-supply noise.



BYPASSING TECHNIQUES

TYPE	COMMENTS
	SIMPLE
	THIS APPROACH IS MORE COMPLEX, BUT IT PROVIDES GOOD HIGH-FREQUENCY REJECTION WITH LOW DC LOSSES. INAPPROPRIATE FOR USE WITH FAST LINEAR CIRCUITRY.
	THE RC FILTER IS EFFECTIVE FOR HIGH-FREQUENCY REJECTION, BUT IT HAS DC LOSSES. YOU CAN USE THIS APPROACH WITH LOW-CURRENT LINEAR CIRCUITS. DON'T USE THIS APPROACH WITH FAST LINEAR CIRCUITRY.

*Fig A—**Errors caused by supply-rail impedances** are a major problem in digitally oriented systems where fast current spiking and clock harmonics are present.*

drives the motor's ungrounded terminal (trace B). When A_1 goes low, Q_3 turns off, and the motor's back EMF appears at the motor terminals after the inductive flyback ceases. During this period, S_1's input (trace C) turns on, and the 0.047-μF capacitor charges to the value of the back EMF. A_2 compares this value with the setpoint level, and the amplified difference (trace D) changes A_1's duty cycle to control motor speed. A_2 has the desirable characteristic of assuming unity gain when there's no feedback signal; as a result, start-up or input overdrive, which causes the loop to lose control of motor speed, cannot force the sampling loop to experience servo lock-up. The loop is self-restoring; that is, it will re-establish control over the motor speed when abnormal conditions cease.

For the circuit to operate properly, you must carefully control S_1's input signal. Q_2 prevents switch closure until the negative-going flyback interval is over, and the 0.068-μF capacitor slows the switch's turn-on speed. These measures ensure that the signal the circuit applies to the 0.047-μF storage capacitor will be clean. The diodes in Q_2's collector compensate for the motor's clamp-diode drops, ensuring that no destructive negative voltages appear at S_1's input. This circuit's control range extends from 20 rpm to full speed.

Note that the gain and roll-off terms in A_2's feedback loop are optimal for the motor shown; if you use a different motor, you have to change the values of the components in the feedback loop.

Generally, process-control circuitry used in industrial environments must transmit standard 4- to 20-mA current-loop signals to valves and other actuators. Because of resistive line losses and actuator impedances, industrial current-loop transmitters must be able to develop at least a 20V drive capability. Systems that operate from 5V usually don't meet this requirement.

The circuit in **Fig 7**, however, does meet industrial current-loop-transmitter requirements. The circuit's design uses a servo-controlled dc/dc converter to generate the compliance voltage necessary to meet current-loop requirements. The circuit can drive 4- to 20-mA signals into loads as high as 2200Ω (44V compliance), and it's inherently short-circuit protected.

Fig 7's amplifier A_2 accepts the circuit input and biases the noninverting input of A_1 via the offsetting

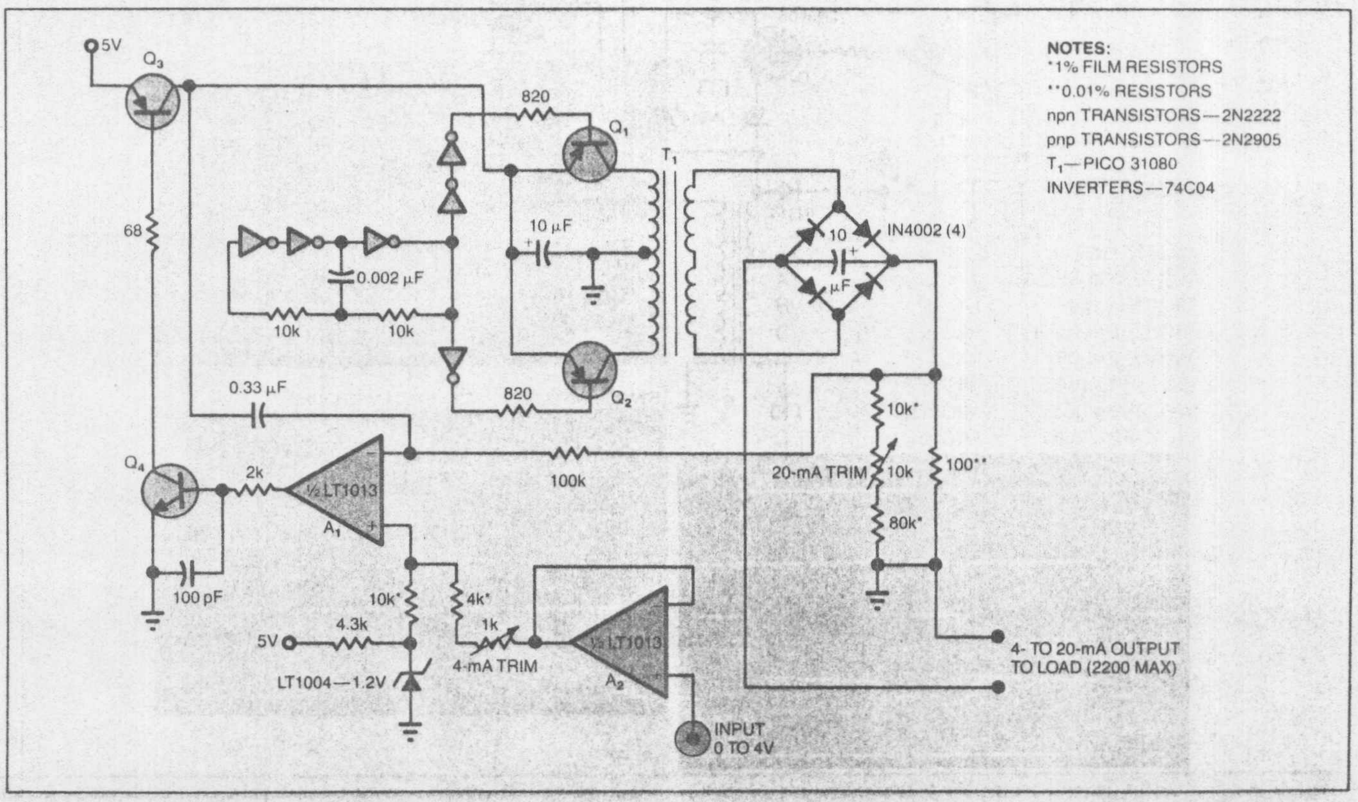


Fig 7—Using a dc/dc converter to generate the required compliance voltage, this loop transmitter, which operates from a single 5V supply, can drive 4- to 20-mA current signals into loads as high as 2200Ω.

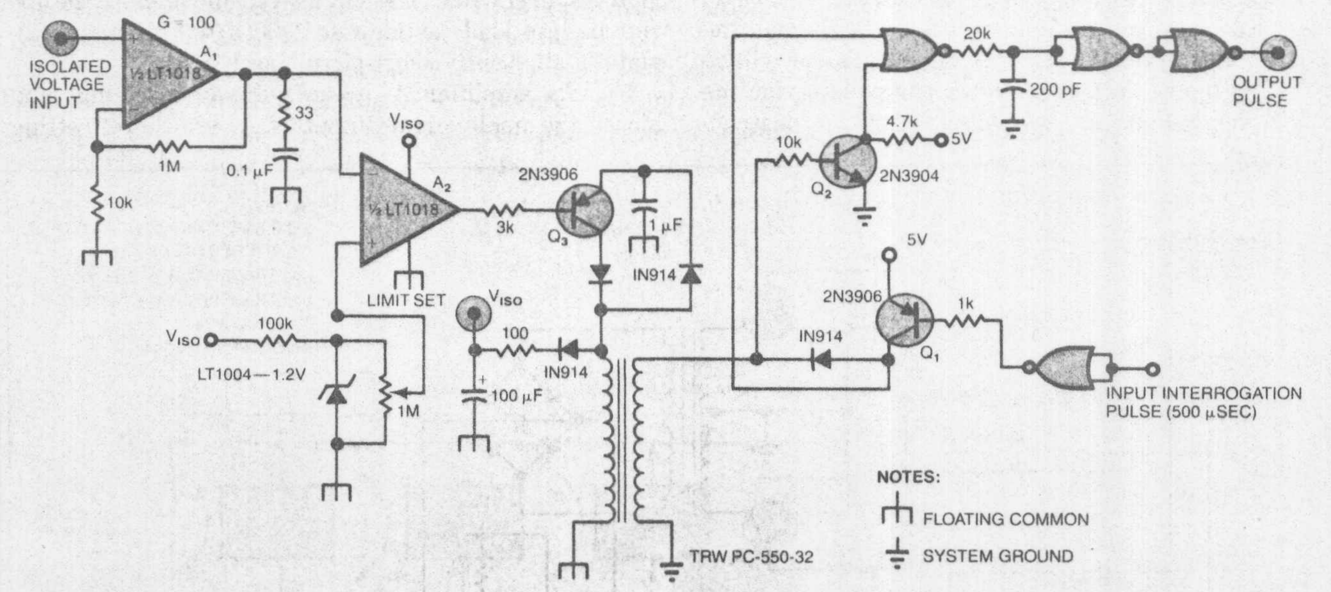
A move to lower power-supply voltages for digital circuits underscores the need for low-voltage, high-performance linear ICs.

network. A_1's output goes high, turning on Q_4 and Q_3. Q_3's collector drives the dc/dc converter (T_1, Q_1, Q_2), which the RC-gate oscillator clocks. T_1 steps up the voltage, and after rectification and filtering, secondary current flows through the 100Ω resistor and the load. A_1's inverting input measures the voltage across the 100Ω resistor, completing a current-control loop around T_1. The 0.33-μF capacitor provides stable loop roll-off, and the 100-pF capacitor suppresses local oscillation at Q_4. A_1 maintains constant output current within the compliance limit, regardless of load impedance shifts or supply changes.

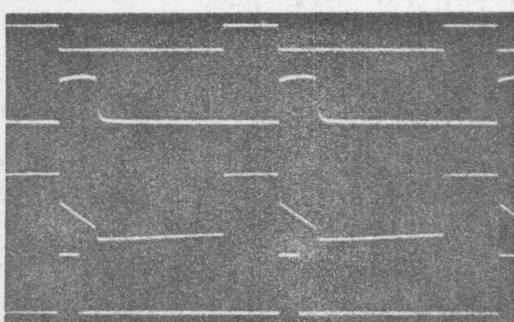
Calibrating this circuit is a straightforward process. First, you short the output, apply 0V to the input, and adjust the 4-mA trim to develop 0.3996V across the 100Ω resistor. Next, shift the input level to 4V and adjust the 20-mA trim for 1.998V across the 100Ω resistor. Repeat this procedure until you fix both points. The odd voltage-trim target values result because the gain trim network shunts the 100Ω resistor.

Another difficulty with operating analog circuits from 5V supplies is that such supplies are not often clean ones. The circuit in **Fig 8** minimizes the noise problems by performing a fully isolated limit comparison on low-level signals. It produces a digital output that indicates whether the input is above or below a preset limit. The circuit is well suited for process-control applications in which transducers operate at high common-mode voltages or in which large ground loops exist. The damper network in A_1's output allows it to function as an op amp for low-level signals, so this circuit will accommodate thermocouples and other low-level sources.

Fig 8's circuit functions by echoing an interrogation

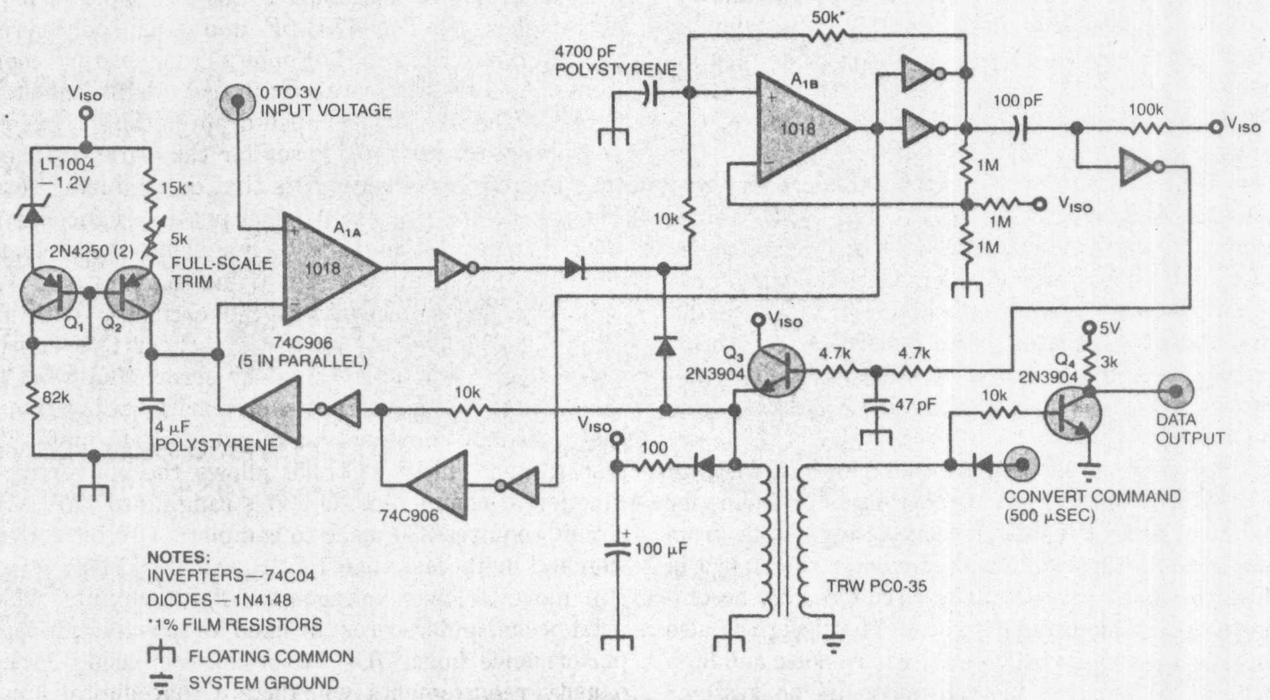


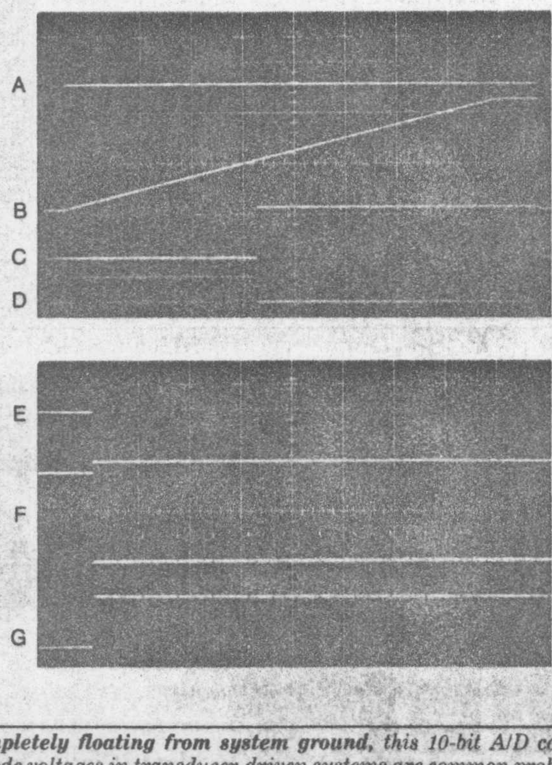
A
B
C
D



TRACE	VERTICAL	HORIZONTAL
A	5V/DIV	500 μSEC/DIV
B	10 mA/DIV	500 μSEC/DIV
C	5V/DIV	500 μSEC/DIV
D	5V/DIV	500 μSEC/DIV

***Fig 8—Suitable for process-control** applications where transducers operate at high common-mode voltages, this circuit produces a digital output that indicates whether the input is above or below a preset limit.*





TRACE	VERTICAL	HORIZONTAL
A	5V/DIV	50 mSEC/DIV
B	2V/DIV	50 mSEC/DIV
C	5V/DIV	50 mSEC/DIV
D	10V/DIV	50 mSEC/DIV
E	5V/DIV	500 µSEC/DIV
F	2V/DIV	500 µSEC/DIV
G	5V/DIV	500 µSEC/DIV

Fig 9—Completely floating from system ground, this 10-bit A/D converter is useful in industrial environments, where noise and high common-mode voltages in transducer-driven systems are common problems. Scope-photo traces A, B, C, and D are described in the text; trace E shows the convert-command pulse; trace F, the transformer primary; and trace G, the TTL-compatible pulses at Q_4's collector.

Linear components should not sacrifice performance when operating from logic supplies.

pulse only when the input is above the preset level. The transformer helps to establish a 2-way, galvanically isolated signal path; the energy contained in the interrogation pulse serves to power the circuits' floating elements.

When you apply an input interrogation pulse, Q_1's collector drives the transformer primary (trace A in **Fig** 8). If the input level is above the preset limit, comparator A_2's output will be low, biasing Q_3 on and allowing Q_3 to drive current into the transformer secondary (trace B). This current reflects into the transformer primary (trace C), where it's detected by a demodulator (Q_2 and its associated gate circuitry) that produces an output pulse (trace D). If A_2's output is high, the transformer receives no secondary drive and there's no output pulse.

The high common-mode noise that's characteristic of predominantly digital systems also poses problems for data converters. Because it completely floats from system ground, the 10-bit A/D converter shown in **Fig 9** solves the noise problem. The circuit readily accommodates high common-mode noise. The design is also useful in industrial environments, where noise and high common-mode voltages in transducer-driven systems are common problems.

You initiate circuit operation by applying a pulse to the convert-command input (trace A in **Fig 9**). The pulse appears at the transformer secondary and charges the 100-μF capacitor. This capacitor potential provides the supply voltage for the floating A/D conversion circuitry. The secondary's pulse provides two additional functions—it biases the inverter/open-drain buffer combination to discharge the 4-μF capacitor (trace B), and it biases a diode to disable the oscillator's (A_2's) 3-kHz output (trace D).

At the same time, A_1's output goes high, forcing the inverter in its output line to switch to a low state (trace C). When the convert-command pulse ceases, the current source (Q_1 and Q_2) charges the 4-μF capacitor with a linear ramp and enables the oscillator. When the ramp voltage exceeds the value of the input voltage, A_1's output goes low, and the inverter in its output line switches to a high state, disabling the oscillator.

The number of oscillator pulses that occur during this interval is proportional to the value of the input voltage. The differentiator at Q_3's base differentiates the oscillator pulses and feeds them to transformer driver Q_3. The spike transformer-drive scheme eases the power drain on the 100-μF energy-storage capacitor. The RC delay in Q_3's base, in conjunction with the inverter/buffer combination at A_{1B}'s output, prevents Q_3's emitter pulses from triggering a ramp reset.

Several factors contribute to the 10-bit performance of this circuit. The 4700-pF and 4-μF polystyrene capacitors both have −120-ppm/°C temperature coefficients. As a result, overall circuit gain drift is about 25 ppm/°C. The five 74C906 open-drain buffers in parallel provide an effective 0V reset for the 4-μF capacitor, thus minimizing offset errors that occur during reset. The parallel inverters in A_{1B}'s output line reduce errors related to saturation, thereby stabilizing the oscillator against shifts in supply voltage and temperature.

Finally, by synchronizing the oscillator with the conversion sequence, the diode path at Q_3's emitter prevents a ±1-count uncertainty error. The 5-kΩ potentiometer in the current source trims calibration so that a 3V input develops an output of 1024 counts. The transformer (TRW PCO-35) allows the converter to function at common-mode levels ranging to 175V. The circuit requires 330 msec to complete a 10-bit conversion and drifts less than 1 LSB over 0 to 50°C.

A move to lower voltages for digital circuits, which must occur, underscores the need for low-voltage, high-performance linear ICs. After all, increasing circuit-density requirements will dictate that digital-supply output levels decrease, lowering the IC breakdown requirements. And users will require new equipment to be portable, so its circuitry will have to be directly compatible with battery potentials.

However, linear components must not sacrifice performance to function in this low-voltage, digitally driven environment. Despite their narrower dynamic operating range, low-voltage linear circuits must still be able to provide the same precision that higher-voltage devices provide. **EDN**

Author's biography

Jim Williams, staff scientist at Linear Technology Corp (Milpitas, CA), specializes in analog-circuit and -instrumentation design. He has served in related capacities at National Semiconductor Corp, Arthur D Little Inc, and the Instrumentation Development Lab at the Massachusettts Institute of Technology. Jim is a former student of psychology at Wayne State University, and he enjoys tennis, art, and collecting antique scientific instruments.

Article Interest Quotient (Circle One)
High 482 Medium 483 Low 484