Application Miscellany

Chapter 15

In this chapter, we discuss some circuit techniques and hardware that go beyond the basics covered in Part One, are useful for more than just one type of sensor, and embody some interesting ideas.

4-TO-20mA TRANSMISSION REVISITED

Circumstances can arise in which it is useful to tap into an existing current loop and transmit up-to-1500-volt-isolated 4-to-20mA information via a second loop. Figure 15-1 is a circuit for interfacing a process loop (having an arbitrary 5:1 current range, for example 1 to 5mA, 4 to 20mA, 10 to 50mA) with an isolated 4-20mA loop, using a 2B22 isolated V/I converter. Since the current range is proportional to that of the output, it is necessary only to transform the input current (I_{IN}) to voltage with a resistance, R_C, and to scale the input voltage range (V_{INFS}) to the out-

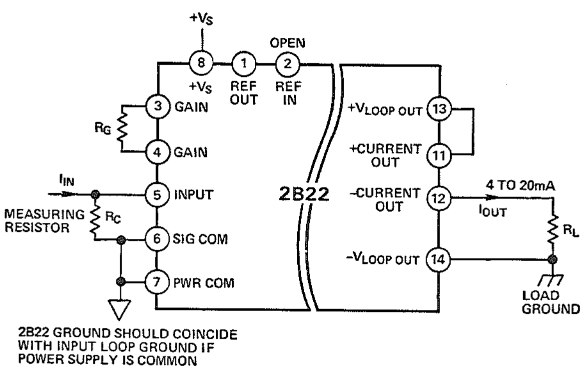


Figure 15-1. Process signal current isolator

put full-scale range, via resistance R_G. The formulas given for R_C and R_G are:

$$R_C = V_{INFS}/I_{INFS} \tag{15.1}$$

and

$$R_G = \frac{6.31 V_{INFS}}{12.62 - V_{INFS}} k\Omega \tag{15.2}$$

The value of R_G determined in (15.2) has a tolerance of ±5%. Therefore, R_G should consist of a smaller-than-nominal value in series with a trim resistance. V_{INFS} can be any value in the range 1V to 10V, as permitted by the primary loop.

For example, if I_{INFS} is 20mA and V_{INFS} is to be 2V, R_C = 100Ω and $R_G \cong$ 1.19kΩ, which might consist of a 1.13kΩ resistor and a 100Ω trim potentiometer.

If the input current is zero-based, e.g., 0 to 50mA, (15.2) becomes

$$R_G = \frac{6.31 V_{INFS}}{10.1 - V_{INFS}} k\Omega \tag{15.3}$$

It is also useful to convert a 4-to-20mA current to frequency for noise immunity, isolation, and ease of going digital. If the output frequency is to be scaled directly (say 1kHz to 5kHz), the circuitry, using IC V/f converters, such as the AD537, is straightforward. Isolation can be combined with self-powering from the

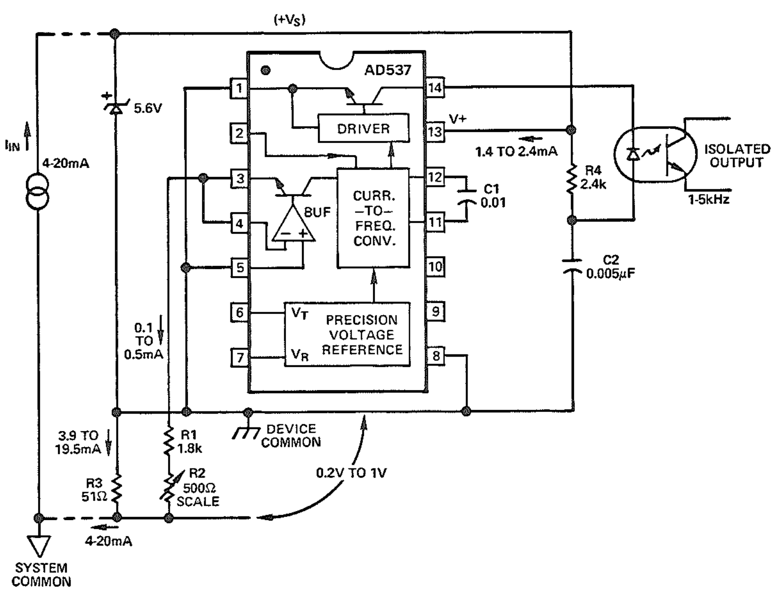


Figure 15-2. Isolated 4-20mA-to-frequency converter

loop current, as Figure 15-2 shows. All required power is provided by a portion of the minimum loop current of 4mA, flowing through the AD537's internal circuitry, with the excess flowing through the 5.6V Zener diode, which establishes the supply voltage. As will be shown, the output is optically coupled.

Note that there are two return paths for the signal current: through the main load resistor, R3, and through the timing resistance, $(R_1 + R_2)$. Since the timing current is theoretically proportional to the load current in R3, linearity is preserved (even though the current gets used in a variety of ways by the rest of the circuit).

The supply voltage, provided by the Zener diode, together with the 1V full-scale voltage across R3, adds up to a worst-case compliance of 7V.

Each cycle, capacitor C2 is discharged through the LED at a peak current of approximately 25mA, internally limited in the AD537. The value of C2 is chosen to produce an optical pulse of about 1μs duration. The maximum average current through the diode is limited to about 2mA by R4, which is calculated to insure that, even when the signal current is at its minimum value of 4mA, there is at least 0.5mA flowing through the Zener; thus the optical output persists at a lower level for a half-cycle. During the next half-cycle, C2 recharges via R4, with a time-constant of 12μs, allowing full recharging in the minimum half-cycle time of 100μs.

In many cases, the only calibration required will be to set the frequency to 5kHz at the full-scale input of 20mA. However, the V_{OS} of the AD537 may also be nulled for more exacting applications, using standard techniques.[1]

MORE THOUGHTS ON ISOLATION

One of the most important considerations about using an isolation amplifier is the manner in which it is hooked up. The following guidelines, if observed, will help a user to realize the full performance capability of the isolator and minimize spurious noise and pickup. Since the more-common sources of electrical noise arise from ground loops, electrostatic coupling and electromagnetic pickup (Chapter Three), these guidelines concern the guarding of low-level millivolt signals in hostile environments.

[1] For more information, see the Application Note: "Applications of the AD537 IC Voltage-to-Frequency Converter," by Doug Grant, available from Analog Devices.

- Use twisted shielded cable to reduce inductive and capacitive pickup.

- Drive the transducer cable shield, S, with the common-mode signal source, E_G, where possible, to reduce the effective cable capacitance, as Figure 15-3 shows. This is accomplished by connecting to the signal low point, B. However, this may not always be possible. In some cases, for example, the shield must be separated from signal low by a portion of the medium being measured, causing a common-mode signal, E_M, to appear between the shield and signal low. The CMR capability between the input terminals (HI IN and LO IN) and GUARD will work to suppress that common-mode signal, E_M.

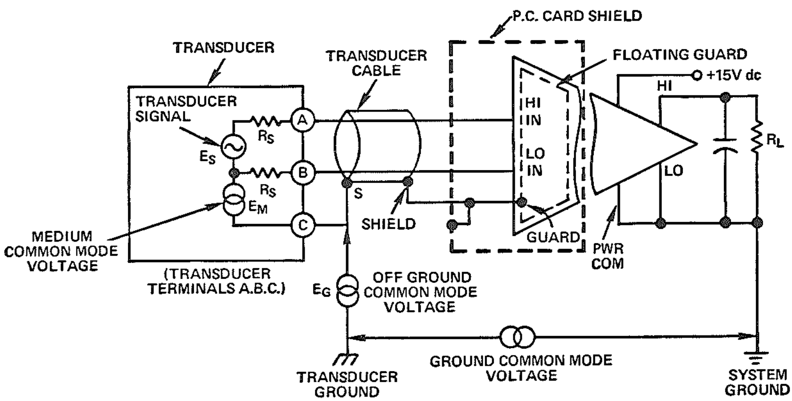


Figure 15-3. Transducer-amplifier interconnection

- To avoid ground loops and excessive hum, signal low, B, or the transducer cable shield, S, should never be grounded at more than one point.

- Dress unshielded leads short at the connection terminals and reduce the area formed by these leads to minimize inductive pickup.

Multi-channel isolators have some uses that have not been discussed elsewhere in this volume. Two of them are depicted in Figures 15-4 and 15-5.

In Figure 15-4, synchronized isolators (288K) and their driver (947) are shown as the three-channel front end of a data-acquisition system which provides up to 850V of both input-to-output isolation and channel-to-channel input isolation. Thus, the transducers can all float; ground-loop problems between the transducers, and ground loops between the inputs and output ground are elimi-

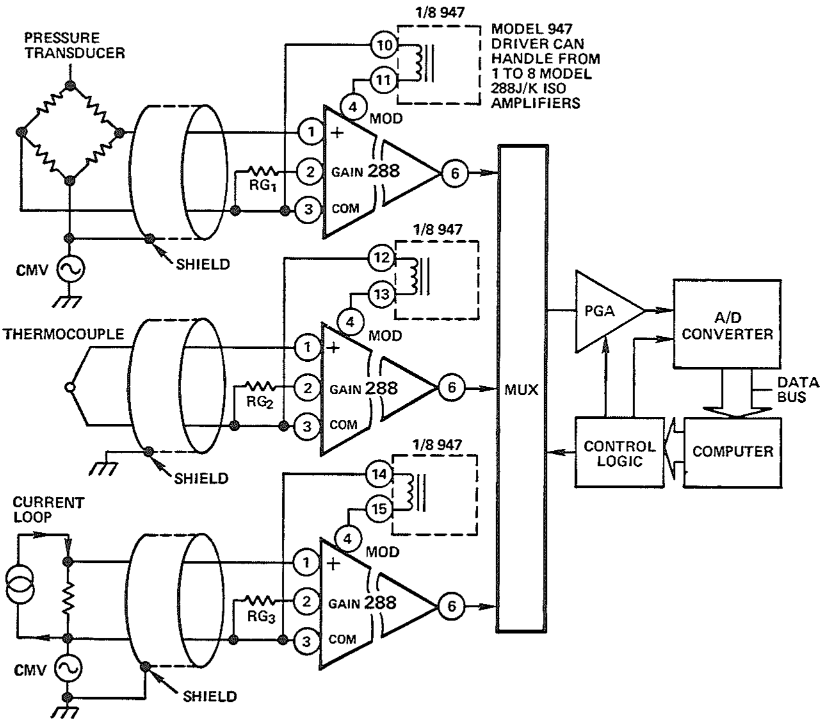


Figure 15-4. 3-channel isolated data-acquisition system front end. The 2B54 isolated multiplexer could also be used in this application. See Figure 15-14.

nated. The high CMV rating and input protection provide protection for the multiplexer against transients, fostering reliable operation in hostile environments. Other characteristics that help maintain the integrity of the data are the 288K's high CMR (92dB) and low nonlinearity (0.05%), all at low cost in multichannel applications.

In Figure 15-5, a number of strain, pressure, or temperature-measurement bridges are driven by current in a series string, to ensure that they receive uniform excitation (if their resistances are uniform), with only a single supply.

Naturally, the current-forcing scheme results in appreciable CMV levels, especially for the transducers at the high end of the chain, so an amplifier with both high tolerance of CMV and high common-mode rejection is required.

Since this application calls for a multi-channel, high-accuracy isolator with only moderately high CMV protection (less than

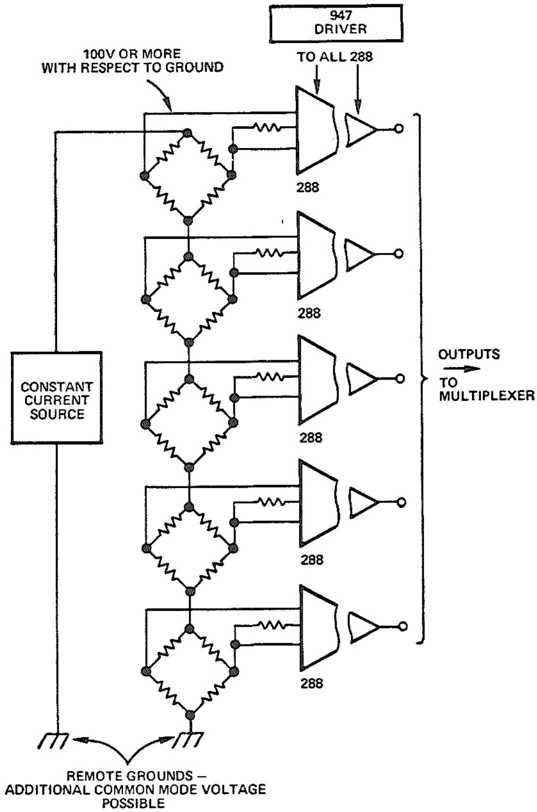


Figure 15-5. Isolation amplifiers in high-CMV, current-excited bridge scheme. Model 289, 3-port synchronizable isolators could also be used here.

850V), 288's, with their 947 driver, are an appropriate approach.

TOPICS IN FILTERING

Most of the filters dealt with in Chapter Three have fixed time constants. Variable time constants are often useful, as the following examples show.

In Figure 13-2, a voltage-controlled low-pass filter made it possible to remotely change the filtering time constant to either filter out periodic pulsations or to observe rapid changes when necessary. The same kind of filter is shown in Figure 15-6, for a somewhat more-detailed discussion. The output voltage of the multiplier is self adjusted by the feedback loop to maintain the voltage across R proportional to the product of the control input and the signal-input-less-the-capacitor-voltage. Since the capacitor voltage is proportional to the integral of the current, the response at A (with a high-impedance load) is that of a unit-lag filter with cutoff frequency proportional to E_C. In addition to that response,

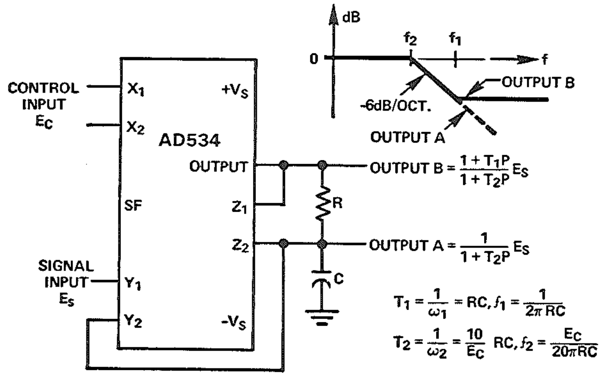


Figure 15-6. Voltage-controlled low-pass filter

output B provides a lag-lead characteristic (variable lag, fixed lead).

Output B is at low impedance and is essentially load-independent, but the voltage at output A should be buffered if its destination is not at high impedance.

Filters can be programmed digitally, too. Figure 15-7 shows an application of a multiplying d/a converter as a digitally controlled first-order low-pass filter. Cutoff frequency is proportional to D, the fractional value of the digital input ($0 \leqslant D \leqslant (1 - 2^{-n})$, for fractional binary). Frequency scale factor can be adjusted pro-

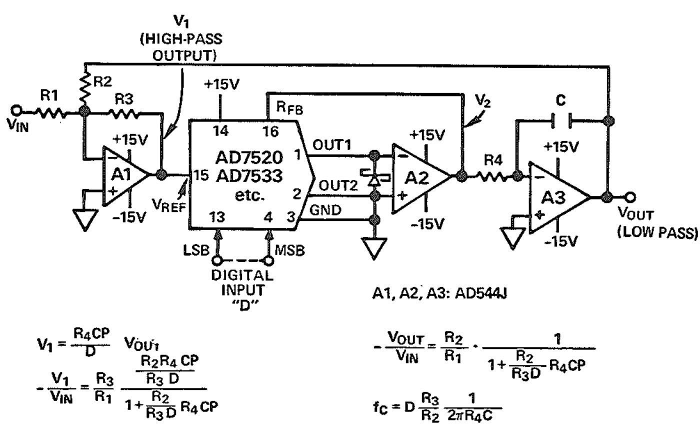


Figure 15-7. Digitally variable first-order filter

portionally with R3, time constant can be trimmed proportionally with R4, overall gain can be trimmed independently with R1.

This circuit and the one immediately preceding it both use integrators in feedback loops. As the programmed cutoff frequency is reduced, the loop gain is correspondingly reduced at low frequencies, in the same way that it is when the denominator is reduced in an analog multiplier used as a divider. Therefore, it is important to use adequately large time constants in the integrator (RC in 15-6, R_4C in 15-7), so that E_C and D approach full-scale as closely as possible at the highest cutoff frequency. For large dynamic ranges of E_C or D, it is useful to trim circuit offset to zero with E_C or D held at the low end of the range.

In Figure 15-7, the output of A1 and A2 are proportional to the derivative of the output,

$$-\frac{V_1}{V_{IN}} = \frac{R_3}{R_1} \cdot \frac{Tp}{1+Tp} = \frac{R_3}{R_1} \frac{j\frac{\omega}{\omega_C}}{1+j\frac{\omega}{\omega_C}} \qquad (15.4)$$

$$\text{where } T = \frac{R_2R_4C}{R_3D} \text{ and } \omega_C = \frac{1}{T} = \frac{R_3D}{R_2R_4C};\ f_c = \frac{R_3D}{2\pi R_2R_4C}.$$

(15.4) will be recognized as a first-order high-pass filter response, with the cutoff frequency proportional to the digital input. R_3/R_1 should be the largest value that will not cause saturation of A1 or A2 when the input is a step function (or the most-rapid change expected).

Figure 15-8 shows a simple way of changing the time-constant of a low-pass filter from one value to another. In a similar approach to that used for the scale in Figure 12-5, this circuit permits rapid response to a step function, then a longer time-constant for filtering high-frequency noise. When the switch is closed, the time-constant

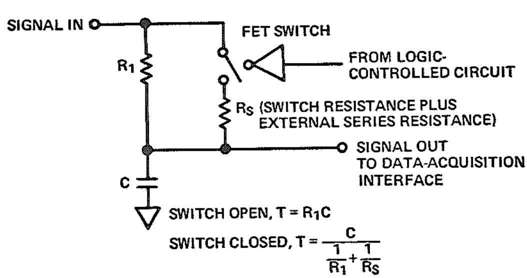


Figure 15-8. Low-pass filter with switched time constant

is short; when the switch opens, the time-constant increases. For applications where it is suitable, such as in batch processes (e.g., weighing discrete increments of load), it combines the benefits of rapid response and good filtering action.

Figures 15-9 and 15-10 depict nonlinear filters that achieve an effect similar to that of Figure 15-8, but in response to the signal itself, and with a smooth transition. They may be termed *derivative-controlled low-pass filters*.

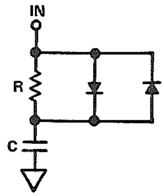


Figure 15-9. Simple derivative-controlled low-pass filter

Figure 15-9 shows how simple such a filter can be. It consists of an RC, with the resistor paralleled by a pair of diodes. If the input is a slowly varying signal with small fast noise signals riding on it, the output will be determined by the filter's RC. However, if the input signal changes rapidly by more than one diode drop (about 0.6V), the diodes will go to low impedance and bypass the resistor, providing rapid response. However, as the capacitor charges and the voltage drop across the diodes decreases, their resistance increases in approximately logarithmic fashion, more of the current flows through the resistor, and the response will increasingly depend on RC.

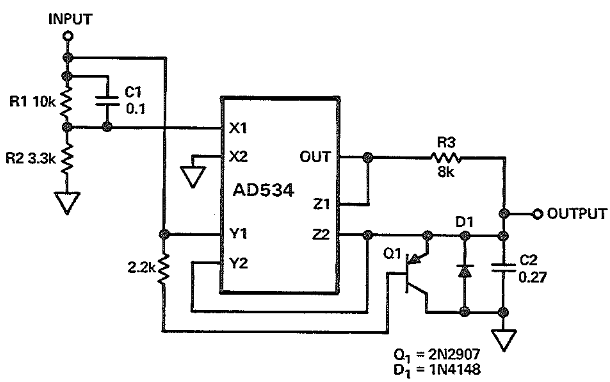


Figure 15-10. Circuit of computing-type derivative-controlled low pass filter

Figure 15-10 is a circuit having similar effect but more rapid and predictable response (and a greater number of components). It will be recognized basically as the voltage-controlled filter of Figure 15-6. The signal input goes directly to Y1. It also goes through a differentiating network, R1, R2, C1, and this *lead* signal modulates the filter time constant.

When a step voltage is applied to the input, the time constant is immediately determined by the full value of the step (which appears at X1); then the control voltage exponentially retracts until it is about 25% of the step (the divider ratio $R_2/(R_1 + R_2)$), which increases the filter time-constant fourfold.

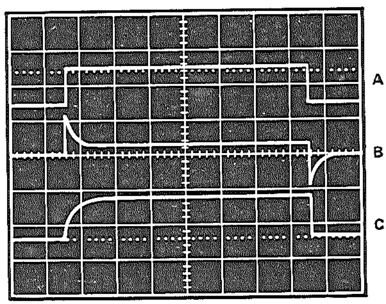


Figure 15-11. Responses of filter of Figure 15-10

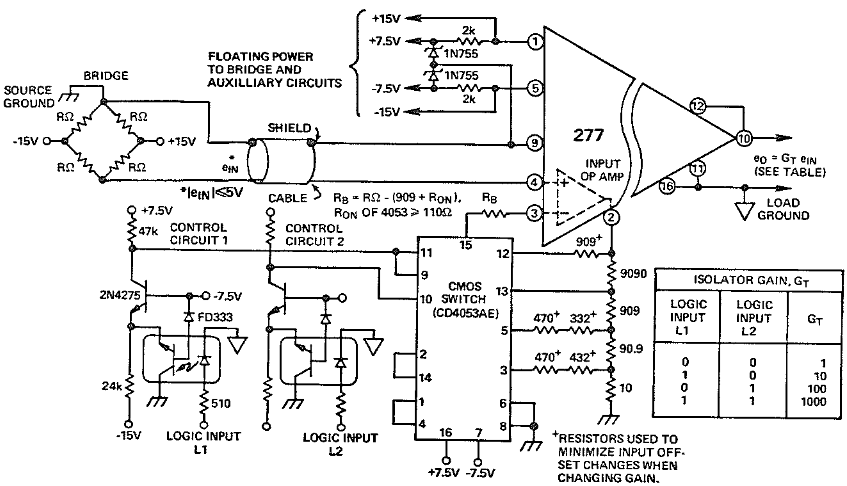
Figure 15-11 shows salient waveforms in the circuit. Trace A is the input step. Trace B is the control input, showing the immediate jump in cutoff frequency, and the decay to a cutoff frequency about 1/4 as large. Trace C is the signal output, showing a rapid response, followed by a long tail, indicating the low steady-state cutoff frequency.

When the square-wave returns, the spike applied to X1 will go in the wrong direction, tending to greatly lengthen the time constant, unless a clamping circuit (consisting of Q1 and D1) is used to rapidly reset C2's charge to zero. If derivative control of the falling edge were desired, an absolute-value amplifier could be inserted between the X1 input and the R1-R2 junction.

In this example, the steady-state time constant is determined by the height of the step and/or the ratio of the resistances R_1/R_2. If there are to be other influences on the time constant, appropriate additional voltages (of opposite polarity) could be summed in at the X2 input. This circuit will function properly only if the noise is relatively small and is not pulse-or step-type. As noted earlier, this approach to filtering has proven useful in electronic weighing applications, where long time-constants are undesirable when weighing an object, yet floor-noise has to be filtered out.

PROGRAMMABLE-GAIN ISOLATOR

Gain, too, can be switch programmed. For example, in wide-dynamic range bridge-type measurements, where gain-ranging would be a desirable feature of the system instrumentation, an isolation amplifier can provide both bridge excitation and signal conditioning. In the application shown in Figure 15-12, the uncommitted op-amp front end and isolated dual 15V supplies of the 277 Isolator permit gain to be set remotely at a value that optimizes input signal-to-noise ratio and eliminates the need for high-quality post-amplifiers at the isolator output.



ISOLATOR GAIN, G_T		
LOGIC INPUT L1	LOGIC INPUT L2	G_T
0	0	1
1	0	10
0	1	100
1	1	1000

Figure 15-12. Programmable-gain bridge transducer amplifier

Control switches are driven by TTL inputs, which are isolated from, source ground by the opto-isolators in the control switch. Control signals operate the CMOS switch network to establish the gains listed in the table. The CMOS switch network is operated in a manner that causes the resistance of the switches to be in series only with the negative input of the isolator, and not in series with the gain-setting resistors; therefore the resistance of the CMOS switches does not affect the gain. Resistor R_B is connected in series with the negative input to reduce errors due to bias-current drift. (If R_B calculates negative, the resistor should be connected in series with the + input.)

HIGH-PERFORMANCE FLOATING DATA AMPLIFIER

The amplifier configuration shown in Figure 15-13 is a chopper-stabilized isolated amplifier capable of withstanding 2500V of continuous common-mode voltage with dc CMR of 160dB (100,000,000:1), and having dc drifts better than 100nV/°C. Its gain of 200,000 permits microvolts to be measured, and volts to be delivered at the output.

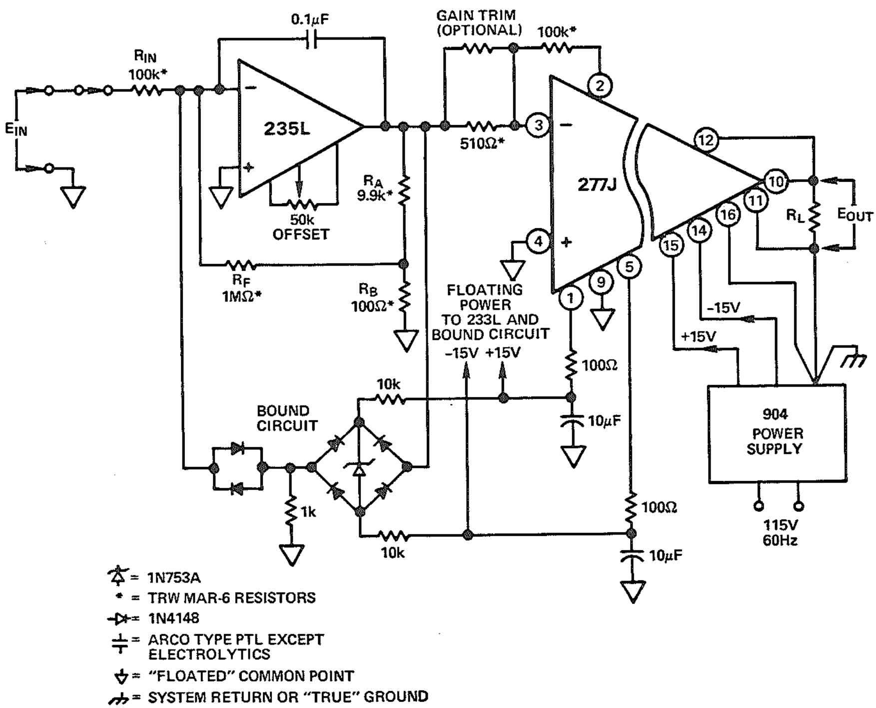


Figure 15-13. Floating input data amplifier

This high-performance amplifier is created by combining the low drift and excellent isolation characteristics of the 277 isolation amplifier with the dc stability of a 235L chopper-stabilized operational amplifier. The 235 is powered by the 277's floating supply and serves as a fully floating gain-of-1000 preamplifier. The output of the 277 is an isolated, amplified version of the 235's output. A five-microvolt signal unbalance at the 235's input will yield one volt at the 277's output.

In this example, the 235 is connected as a gain-of-1000 inverting amplifier, with an input impedance of 100kΩ. While this is certainly acceptable for most applications, a non-inverting chopper (e.g., Model 261) instead of the 235 will provide very high input impedance for applications that require it, with perhaps a small sacrifice in bandwidth. High gain is achieved by feeding 1/100th of the 235's output back to the input via a 1MΩ resistance (additional gain-ratio of 10). This produces high gain without resorting to high values of feedback resistance, which invite leakage and stability problems.

Gain adjustment may be done at the 100Ω resistor, R_G. This eliminates any possible contribution to error at the summing point and allows an inexpensive switch to be used.

The menagerie of diodes which parallel the feedback path act as a bound on the output swing of the 235 in order to ensure fast recovery from input overload. At a gain of 100, only *12 millivolts* of signal is needed to slam the amplifier into saturation. Chopper-stabilized amplifiers can require significant time to recover to zero offset after an input overload has occurred; the bound circuit lowers the gain of the amplifier, at high output levels, keeping its output from saturating and speeding recovery.

The bandwidth is deliberately limited to a few Hz by the 0.1μF capacitor, to minimize high-frequency noise transmission. For most measurements of the types discussed in this book, such response is adequate.

The drift characteristics of chopper-stabilized amplifiers are insensitive to input offset adjustments. For this reason, the offset adjustment of the 235 is used to zero the entire amplifier. If gain accuracy is important, the 510Ω resistor can be parallel-trimmed to achieve a precise value of overall gain. Since the 277 is connected as an inverting amplifier, with a gain of 200, its output will be of the same polarity as the input signal (unless the 261 non-inverting amplifier is used as the preamp).

Each amplifier should be enclosed in a "house" constructed of copper walls. The amplifiers (especially the 235) should be packed in Styrofoam or Fiberglas if thermal transients are present. The copper boxes offer protection from stray fields, from RF and static pickup, and from unwelcome conversations between the carrier circuits of the 277 and the 235. (Any coherence between the two frequencies or their harmonics can result in beat frequencies, which masquerade as dc uncertainty.)

If this seems like a great deal of trouble to go to in order to construct an amplifier, it is—but it's worth the effort. This circuit, properly embodied with respect to grounding, shielding, component choice, and good wiring practice, has yielded a powerful low-offset-error dc amplifier—higher in performance than any instrumentation amplifier commercially available to date.

Salient performance that has been achieved includes:

Gain:	200,000 (higher gains readily achievable)
Drift:	100nV/°C (typically 50nV/°C, 20° to 30°C
Time drift:	5μV/year
CMRR (dc):	160dB (100,000,000:1)
Common-mode voltage:	2500V continuous
Input noise:	1μV, peak-to-peak, in a 1Hz bandwidth

This kind of offset and common-mode performance will find use in transducer signal-conditioning where the highest grade of performance is required. The sole errors of consequence in this design are the 0.05% nonlinearity and the 50ppm/°C gain drift of the 277J, which—though quite low—should be considered if necessary.

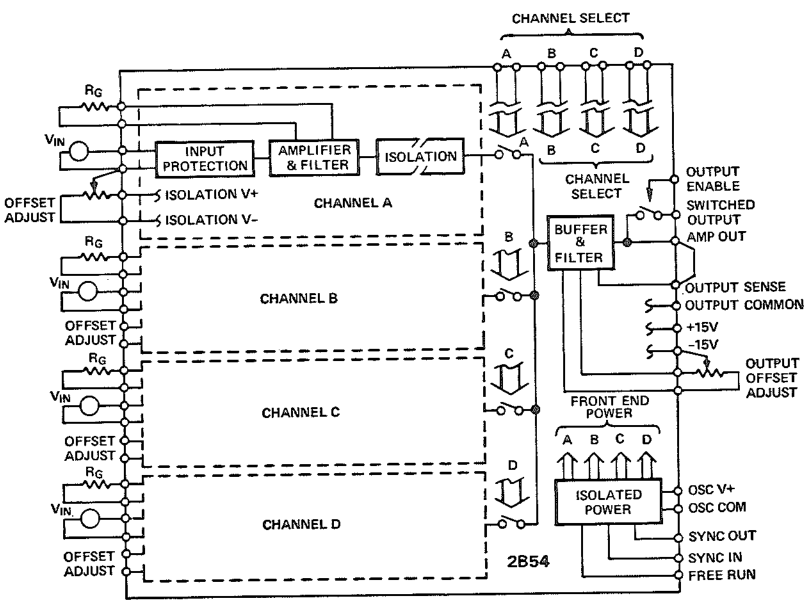


Figure 15-14. Isolated 4-channel multiplexer (2B55)

ISOLATED ALL-ELECTRONIC MULTIPLEXING

In Chapter Seven, there was a discussion of the role of a low-level multiplexer (the 2B54) in thermocouple measurement (see Figure 7-8). Figure 15-14 is a somewhat more detailed block diagram of the 2B54 and a companion high-level multiplexer, the 2B55.

As the diagram shows, there are four channels, isolated from one another, from the output, from the decoded channel-select inputs, and from the power supply for up to 750V rms. Each channel is also protected against 130V rms ac differential input voltage and has individual gain and offset adjustments, and an internal filter. Open inputs can be detected, and the output is protected against continuous short circuits to either supply or ground.

The four channels share a common oscillator, which can be synchronized to the oscillators of additional multiplexers if more than four channels are desired. Because a "three-state analog" output connection is available, additional ranks of multiplexing are not required; the outputs can be connected in common and enabled individually.

A safe, active, solid-state, *all-electronic* multiplexer, the 2B54 (2B55) is a superior alternative in many applications that heretofore have been restricted to reed-relay and flying-capacitor techniques.

PULSED-MODE BRIDGE EXCITATION

The first figure in this book concerns itself with a method for pulsing large inputs to an active transducer to obtain proportionally increased output voltages without excessive heating of the element.

Figure 15-5 is a more-informative version of the principle embodied in Figure 1-1. A bridge is suspended (electrically) between the collectors of a PNP-NPN switch referred to ±50V. For this reason, the common-mode voltage is close to zero volts, and the requirements for amplifier common-mode rejection are not severe. If the bridge were switched from ground or supply only, common-mode potentials of 50V would result, and an isolation amplifier would be required.

Although more intricate than the usual bridge-conditioning circuit, this technique is a useful way to reduce the effect of the monitoring amplifier's drift. With 100 volts being switched, instead of the usual steady ten volts, an AD521's effective drift in

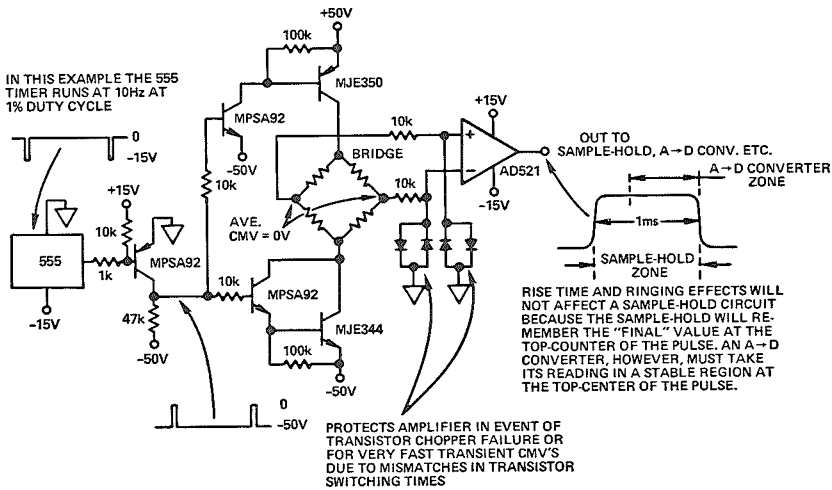


Figure 15-15. Pulsed high-voltage-bridge drive

the former case is $0.5\mu V/^\circ C$, while the low-drift AD522 would have an effective drift of $0.2\mu V/^\circ C$, and the 234 chopper-stabilized amplifier would have, in effect, *10n V/°C*!

Naturally, failures in the transistor drive-circuitry could cause disastrous consequences to the bridge circuit and the amplifier. The opto-isolator and associated components in Figure 15-16 are intended to cause the SCR crowbar to trigger and blow the fuse in the event of such a failure.

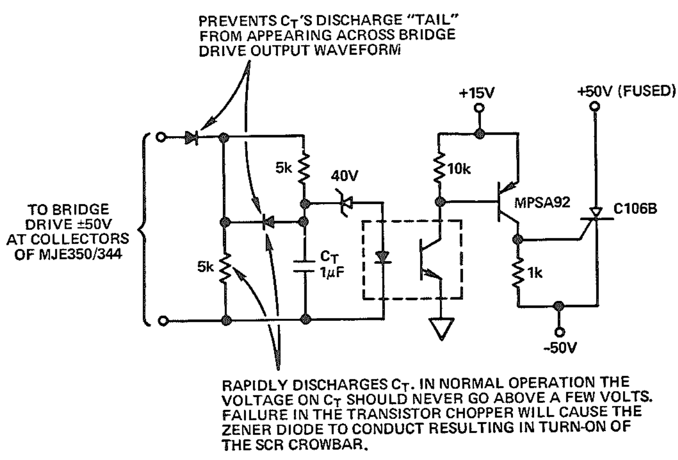


Figure 15-16. Protective circuit for high-voltage bridge

The duty cycle and the *on* time of the transistor chopper must be set with an eye towards te thermal limitations of the bridge elements, the settling-time-to-desired-accuracy of the amplifier amployed, and the characteristics of the sample-hold used to read the output of the amplifier.