Force-Transducer Interfacing

Chapter 12

SPRING-DRIVEN RHEOSTAT

Some force-measuring transducers use a length of spring steel as the sensor. In the case of Figure 12-1, a spring is coupled to a rheostat. The resistance is proportional to the force applied to the spring, changing from 100Ω to 500Ω as the force increases from 0 to 20 pounds.

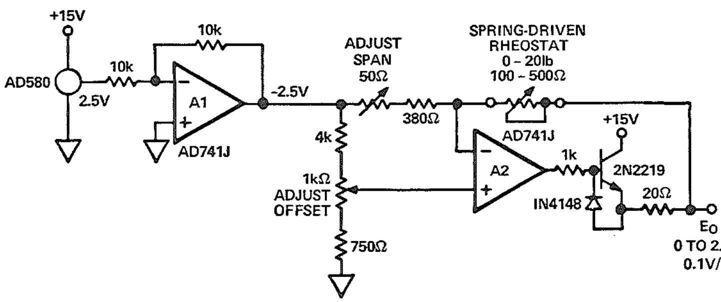

Figure 12-1. Spring-driven rheostat application

The circuit is very much like that used for the RTD in Figure 8-1. The variable resistance is connected in the feedback loop of an operational amplifier and driven by a constant current of 5mA. The 0V to 2V output range provides a scale factor of 10 pounds per volt. The reference signal is derived from an AD580 2.5V reference source, inverted in polarity in an AD741J op amp, and inverted once again in the output amplifier, to provide a positive output. The current-boost transistor provides for a load to be driven by the output amplifier.

The circuit is calibrated by adjusting the output span for 2V, for an input change of 20 pounds, then adjusting the offset for 0V at 0 pound-force.

STRAIN-GAGE LOAD-CELL INTERFACE

In this application (Figure 12-2), the sensor is a BLH Electronics T3P1 350Ω strain-gage-based load cell. The output of the bridge is amplified by an AD522 differential instrumentation amplifier. The bridge is excited by a current-boosted op amp, A1, the input to which is the circuit's 10V precision reference, an AD581. The entire circuit is powered by a Model 925 power supply, which is capable of supplying 350mA, enough for half-a-dozen similar transducers.

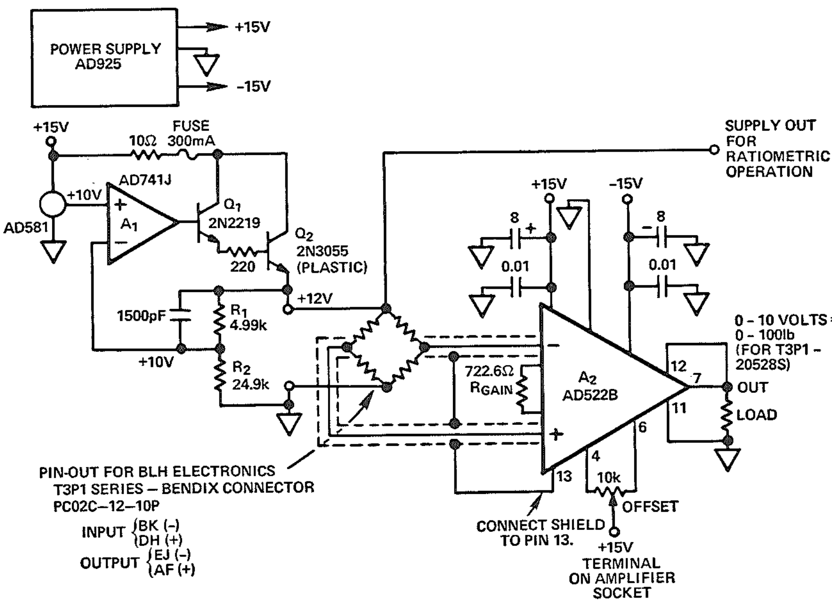

Figure 12-2. Load cell application with differential instrumentation amplifier

The op-amp circuit is connected for a closed-loop gain of 1.2 $\left(= 1 + \frac{R_1}{R_2}\right)$, so that the voltage at the output terminals is 12.0V.

The voltage at the bridge terminals is made available for referencing the bridge output to its actual drive voltage ratiometrically at a later stage (a/d converter, panel meter, processor, etc.)

In the booster circuit, the 10Ω resistor provides current limiting in the event of a short circuit. Since the follower is inside the feedback loop, variations of its gain do not affect the precision of the output voltage.

The sensitivity of the T3P1 load cell is 3mV per volt of excitation; with 12.0V applied, the full-scale output is 36mV. If the full-scale amplifier output is to be 10V, corresponding to 100 pounds (10 lb/V), the gain must be 10/0.036 = 277.8V/V. Using the AD522 gain equation,

$$G = 1 + \frac{200{,}000}{R_G} \tag{12.1}$$

hence, the nominal value of R_G is 722.6 ohms. Since the gain equation is accurate to within 0.2% for the AD522B at 25°C, a nearby fixed value, a fixed-value-plus-fixed-trim, or fixed-value-plus-variable-trim may be used, depending on the gain accuracy required.

The AD522 has a guard-drive terminal (13) that follows the common-mode level. It may be used to drive short lengths of shielding on the input leads and the gain resistor. If the leads to be shielded are lengthy, the impedance at pin 13 should be stiffened by using it to drive a follower, which in turn drives the shield circuitry.

This circuit may be used to drive 120Ω bridges (e.g., BLH type V-1), which require lower drive potentials, for example 6.0V. For such applications, the AD581 output is attenuated to 6V, and the feedback from the bridge-drive voltage is unattenuated. The 2N3055 will have to dissipate about 0.5W, in this instance, vs. about 0.1W in the circuit of Figure 12-2.

STRAIN GAGE AND SIGNAL CONDITIONER

Figure 12-3 shows how a system-solution signal conditioner and transducer power-supply are used to provide excitation for and readout from a strain-gage bridge. In this case, the excitation for both the 2B30 signal conditioner and the bridge are furnished from mains power by a 2B35 transducer power supply. A single active gage (120Ω, G.F. = 2) is used in a bridge configuration with a dummy gage (for temperature compensation) and two precision 120Ω resistors.

The adjustable power supply is set for 3V of bridge excitation, to avoid errors due to self-heating of the gage and the bridge elements. The sense terminals ensure that the voltage *at the bridge* is precisely 3V.

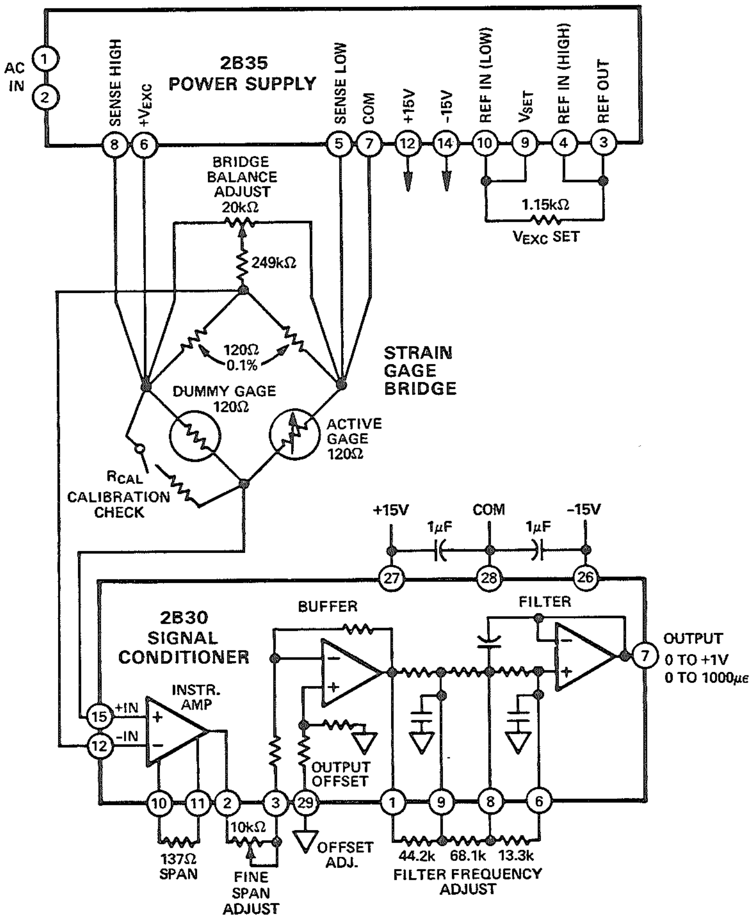

Figure 12-3. Strain measurement using a half-bridge and the 2B30 bridge signal-conditioner

The signal conditioner provides sufficient gain for an output of 1 volt, corresponding to 1000 microstrains. It also provides 3-pole Bessel filtering with a cutoff frequency of 100Hz.

The bridge-balance adjustment pot provides offset adjustment for the whole circuit, and fine span-adjustment is provided by a variable resistance connected between pins 2 and 3 of the module. A calibration-check resistance is shown across one leg of the bridge. It is of appropriate value to obtain a change in reading of about 75% of full scale when the switch is closed and opened.

Figure 12-4 shows a similar circuit employing the 2B31 signal conditioner, which contains its own adjustable bridge excitation and runs from system dc power. The circuit shows a 350Ω bridge, and a recommended shielding and grounding technique to help preserve the signal information.

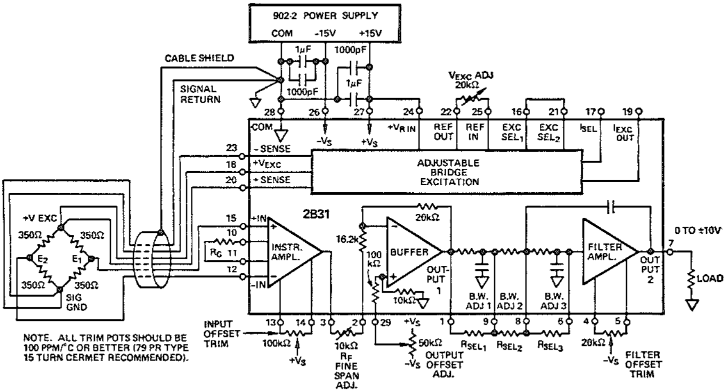

Figure 12-4. Typical strain-gage bridge transducer application using signal conditioner

Because these devices are direct coupled (i.e., not isolated), a ground-return path for bias current must be provided, generally through the bridge, as shown, unless the bridge is excited from a separate floating supply. If the inputs are floating, resistance up to 1MΩ can be used to provide leaks from the supply ground to conditioner *common*.

For best performance, especially at high gains, the sensitive input and gain-setting terminals should be shielded from noise sources. To avoid ground loops, the signal return or cable shield should never be grounded at more than one point. The device should be decoupled from the power supplies, using 1µF tantalum and 1000pF ceramic capacitors as close to the amplifier as possible.

The 2B30 and 2B31 signal conditioners have been conservatively specified, using min-max and typical values, to allow the designer to develop accurate error budgets for predictable performance. An error budget for a typical transducer application, based on calculations for the circuit of Figure 12-4 using data-sheet specifications, is given here.

The circuit employs a 350Ω bridge, with 10V drive, and 1mV/V full-scale sensitivity (i.e., 10mV full-scale output). The assumptions used for the analysis are: Gain is 1000, ambient temperature range is ±10°C, source unbalance is 100Ω, and common-mode noise at

60Hz is 0.25V (25 times as large as the full-scale signal) on the ground return.

Absolute gain and offset errors can be trimmed to zero. The remaining error sources and their effect on system accuracy (worst case) have been calculated. They are listed here as percentages of full-scale output (10V).

Error Source	Effect on Absolute Accuracy % of F.S.	Effect on Resolution % of F.S.
Gain Nonlinearity	±0.0025	±0.0025
Gain Drift	±0.025	
Voltage Offset Drift	±0.05	
Offset Current Drift	±0.004	
CMR	±0.00025	±0.00025
Noise (0.01 to 2Hz)	±0.01	±0.01
Total Amplifier Error	±0.092 max	±0.013 max
Excitation Drift	±0.15 (±0.03 typ)	
Total Output Error (Worst Case)	±0.24 max (±0.1 typ)	±0.013 max

The total worst-case effect on absolute accuracy over ±10°C is less than ±0.25%, and this circuit is capable of 1/2 LSB resolution in a 12-bit low-level-input data-acquisition system. Since the errors are added directly, and the computation is based on conservative specifications, the typical overall accuracy error would be less than ±0.1% of full scale.

In a computer or microprocessor-based system, automatic recalibration can nullify gain and offset drifts; this leaves noise, nonlinearity, and common-mode error as the only significant error sources. Transducer-excitation drift error is greatly reduced in many applications by ratiometric operation with the system's a/d converter.

HIGH-RESOLUTION LOAD-CELL PLATFORM INTERFACE

Figure 12-5 is the schematic of the high-resolution metabolic scale discussed in Chapter Six. The bridge is driven by a fully floating reference supply, employing a dc-to-dc converter. The ground-referenced 261K chopper-stabilized amplifier takes a highly stable true-differential measurement, with respect to the bridge output. Since the grounds of the two supplies are separated only by the bridge resistance, the 261K's bias current has a low-impedance ground return.

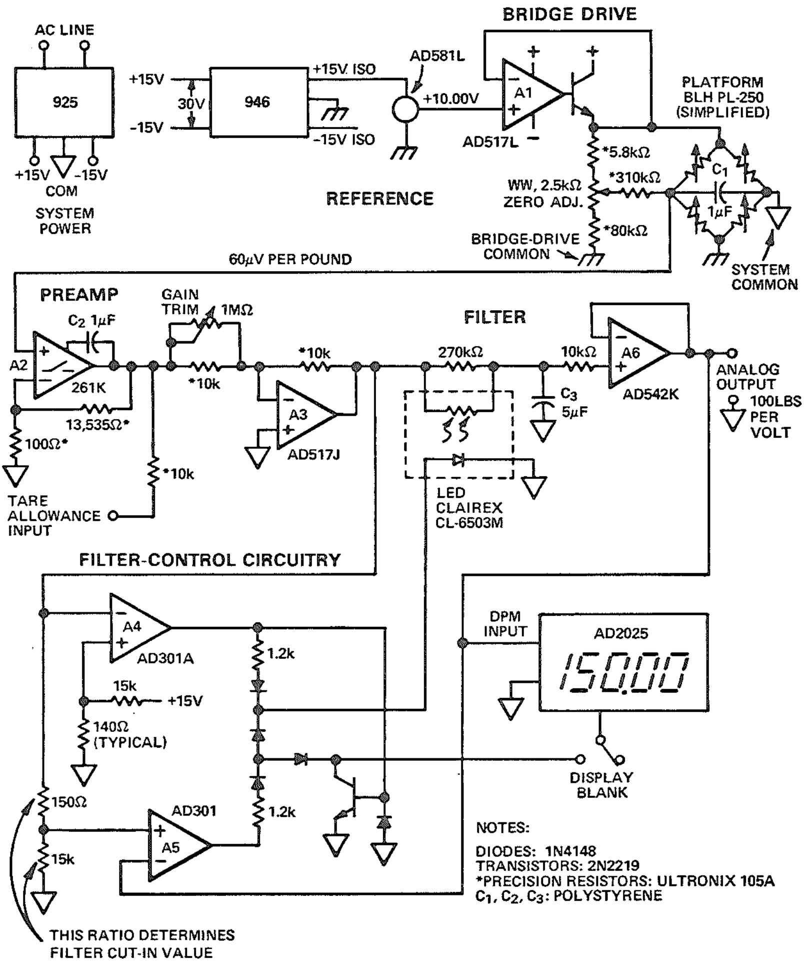

Figure 12-5. Precision platform balance circuit

The filter, in this interface, must eliminate very low frequencies (heartbeat, floor vibration), but must also respond quickly when someone gets on the scale. A simple low-pass filter with a 0.2Hz time constant will successfully eliminate noise artifacts, but it requires an unacceptable length of time to settle to five-place resolution. Higher-order linear filters will help, but they are more complex and are difficult to build and test at low frequencies. A simple solution is to employ an RC low-pass filter with a short time constant, and switch it to a long time constant after the scale has settled to within 99% of its final value. Since the time-varying filter used in this application is unusual, it will be described at some length.

With no weight on the scale, the output of A3 is at zero. The voltage at the + input of A5 is also at zero, and the voltage at its minus input is also very near zero because it is derived, via follower A6, from the same source as the + input voltage. Under these conditions, the output state of open-loop A5 is indeterminate. However, the output of A4, which is biased positive, is used to essentially ground any output from A5, via the transistor and diodes.

When a person or a weight of more than two pounds arrives on the platform, A4's output goes low (as A3's output goes high), and A5's output goes high, activating the light-emitting diode (LED). This permits rapid charging of the 5μF capacitor through the photoresistor. When the output of the AD542L is greater than the + input of A5, A5 goes to its negative limit, which shuts off the LED-photoresistor switch, after which the time constant of the filter is determined by the 270kΩ resistor and the 5μF capacitor. This effectively filters body motion and heartbeat noise without the attendant long waiting time for a simple RC to settle.

When the subject steps off the scale, A4 goes high, the LED is activated, and the capacitor is drained rapidly towards zero. While the scale is empty, the output of A4 maintains the LED in the *on* condition, insuring rapid measurement response for lightweight objects or persons placed on the platform (i.e., those weighing less than 2 lbs). The LED-photocell switch provides drive-isolated, essentially errorless, switching for the application.

The requirements for the scale and the discussion of the low-level signal handling are discussed in detail in Chapter Six. To summarize, however, the scale is capable of reading 300.00 pounds full-scale (100 lb/V), with a resolution of 0.01 lb, and an absolute accuracy to within 0.05 lb. Developed at the Massachusetts Institute of Technology, it has been described in the technical literature.[1]

[1] Williams, J., "This 30ppm Scale Proves that Analog Designs Aren't Dead Yet," *EDN* Magazine, October 5, 1976, Vol. 21, No. 18

PIEZOELECTRIC TRANSDUCER SHAKER-TABLE CONTROL

In the application diagrammed in Figure 12-6, a servo loop drives a shaker table at constant amplitude at 400Hz. The acceleration/force/amplitude of a shaker table is measured with a piezoelectric accelerometer. The resulting signal is amplified, and its rms value is computed and compared with a reference signal. The output of the low-frequency amplifier modulates the amplitude of a tuning-fork oscillator, and its output drives a transformer-isolated audio power amplifier, which excites the shaker table. If the table's amplitude tends to increase, the feedback signal tends to decrease it; if the amplitude tends to decrease, the result is increased drive, tending—in both cases—to maintain constant amplitude.

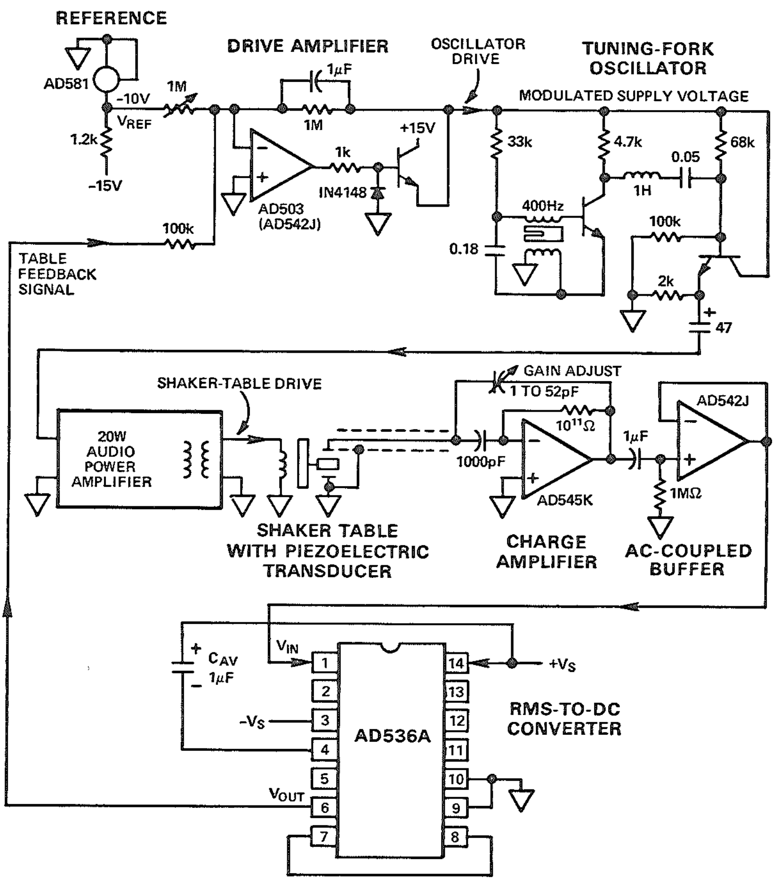

Figure 12-6. Constant-amplitude shaker-table drive

The output of the piezoelectric transducer can be thought of as a 400Hz voltage source in series with a small capacitor. Variations in capacitance due to physical deformation will cause charge to be dumped into the summing junction, and the amplifier will manipulate the output voltage to maintain the charge in the feedback capacitor equal to the charge dumped into the input, so that the voltage at the summing point tends to be at a null. The 1000pF capacitor and the large feedback resistor provide (1) ac coupling to avoid placing a dc bias voltage on the transducer and (2) a dc feedback leakage path for bias current to prevent the output from being driven into saturation.

The output voltage, as a function of Q(t) is

$$E_{OUT} = Q(t)/C_F \tag{12.2}$$

where C_F is the feedback capacitance and Q(t) is the ac variation of charge developed by the piezoelectric crystal. The output of the charge amplifier is ac-coupled to the AD536 rms-to-dc converter, buffered by a follower-connected op amp. The output power of the driver op amp is boosted by the 2N2219 follower-connected transistor inside the local feedback loop.

STRAIN-GAGE TO FREQUENCY CONVERSION

The AD537 voltage-to-frequency converter, described in Chapter Seven and elsewhere, can be used for strain-gage-to-frequency conversion. Figure 12-7 shows an application in which the AD537's 1.0V reference source is used as the excitation for the bridge,

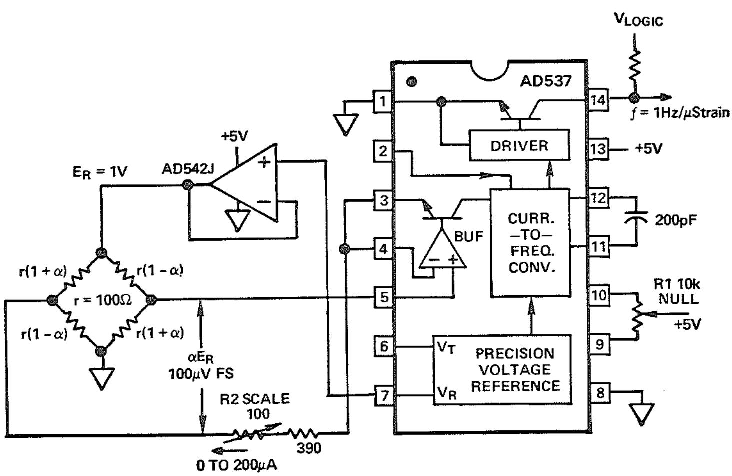

Figure 12-7. Typical connections for strain-gage-to-frequency converter

buffered by an AD542J FET-input op amp. In such applications, it is worth noting that, since the AD537's reference source is also the internal reference for the V/f converter, the conversion to frequency is ratiometric, hence quite stable with temperature and independent of the supply voltage.

The circuit of Figure 12-7 is calibrated to generate a scale of 1Hz per microstrain (100kHz at the assumed full-scale value of $\alpha = 0.1$). The assumption is that the strain is always of the same polarity (positive at terminal 5) and that the balanced set of 4 variable elements form a linear bridge. The timing resistor is returned to one side of the bridge, to both double the sensitivity and effect offset operation. Although this connection can introduce a nonlinear error term, for $R \geqslant 5r$ and $\alpha \leqslant 0.1$, the error is less than 0.1% absolute; if the system is calibrated at full-scale strain, this error is reduced to approximately +0.025% peak.

SCANNING STRAIN METER

The AD2037 scanning digital voltmeter, mentioned in previous chapters, is capable of providing excitation and readout for the broad range of transducers based on the ubiquitous strain gage. Applications include direct connection to a one-, two-, or four-gage bridge, and any of the transduction functions which use it, such as force, acceleration, level, weight, pressure, and flow. An example of its use in a bridge application is seen in Figure 11-5.

ISOLATORS AND TRANSMITTERS

As earlier chapters have noted, the outputs of interface circuits can be easily transformed to 4-to-20mA signals for transmission in industrial-process environments, with either common system wiring or isolation. Also, isolators and isolated op amps are available for two- or three-port voltage-to-voltage isolation. The AD2037 scanner, mentioned above, is powered by ac mains power and also has an isolated front end.