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Bob G. Mahrenholz

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Sverdrup Technology, Inc./AEDC Group
Arnold Air Force Base, TN 37389-9998

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# AN IMPROVED STRAIN-GAGE SIGNAL CONDITIONER FOR DYNAMIC STRESS MEASUREMENT*

Bob G. Mahrenholz

## ABSTRACT

A new signal conditioning circuit has been developed for single-active-arm strain gages typically used for dynamic stress measurement on turbine engine rotor blades. This new circuit uses constant-voltage excitation and corrects for lead wire resistance changes by automatically controlling resistances in series with the leads to the gage. This technique provides a balanced configuration which offers better high-frequency common-mode rejection and has less broadband noise than with constant-current excitation but still is insensitive to lead wire resistance changes. Also included is a unique ΔR calibration circuit that provides a more accurate simulation of actual gage resistance change. The circuits described may be useful in any AC-coupled strain-gage system.

## INTRODUCTION

Dynamic stresses on the compressor and turbine blades of a turbojet engine are typically measured using resistance strain gages bonded to the blades. Increased demand for accuracy and for extended data bandwidth has led to the reevaluation of the signal conditioning circuits used with single-active-arm strain gages. As may be expected, there is an increased demand for accuracy and for extended data bandwidth. Achievement of this improved performance is made more difficult by the proliferation of fast rise time digital electronic circuitry with its accompanying generation of electrical noise.

Also complicating the picture is the practice by engine manufacturers of using small-diameter, high-resistance thermocouple wire as lead wire for the strain gages. The resistance of this wire changes significantly with temperature. The resistance of the strain gage itself is normally either 120 or 350 ohms and exhibits a change of a fraction of an ohm when subjected to strain. Small-diameter wire connects the gage to a junction point located near the engine where permanent shielded cable runs to the location of the signal conditioning equipment. The small-diameter lead wire is normally unshielded and frequently has a significant resistance, especially if thermocouple extension wire is used. Since the lead wire is subjected to temperature excursions as the engine is tested, the change in lead wire resistance may be greater than the change in the gage resistance caused by stress.

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A strain-gage signal conditioner has been designed and prototyped that accommodates changing lead wire resistance without loss of accuracy and still maintains low noise and good common-mode rejection of unwanted electrical noise. This conditioner also includes a novel calibration circuit which produces a $\Delta R$ resistance change exactly equivalent to the resistance change of the strain gage and thereby reduces the errors associated with calibration by millivolt substitution.

## BACKGROUND

The error caused by gage lead wire resistance changes varies from setup to setup and is a function of the composition of the wire, the length of the wire, the temperature extremes to which the wire is exposed, and of course, the circuit used to condition the gage. Under certain conditions, errors of over 5 percent have been observed.

There are several techniques for reducing or eliminating errors attributable to lead wire resistance changes in strain-gage measurement systems.(1,2) The most accurate scheme is to use the four-wire resistance measurement circuit with separate leads for voltage and current. As shown in Figure 1, two of the wires are used to provide a current through the gage. Since the gage is fed from a constant-current source, the current through the gage will be the same regardless of the lead resistances. The other two wires are then used to measure the voltage right across the gage. If the recorder input resistance is much higher than the lead resistances ($R_{LEAD}$), there is no significant drop in the leads and the recorder (REC) shows the actual voltage across the gage, thereby eliminating any effect of lead resistance. Unfortunately, this scheme doubles the number of lead wires installed on the engine.

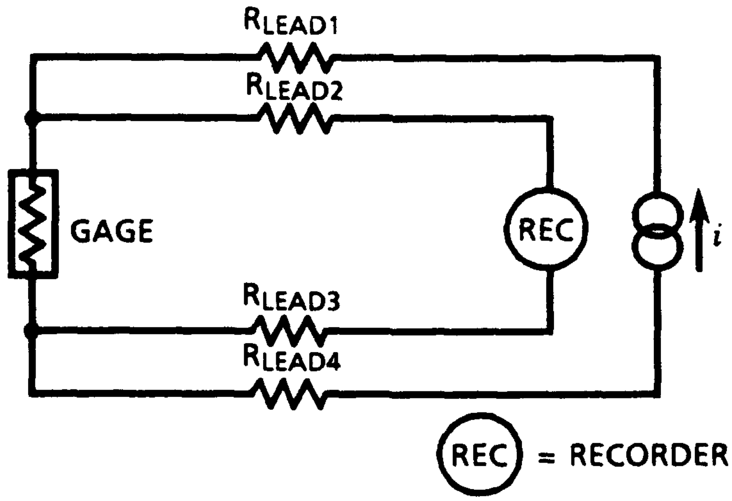


Figure 1. Four-Wire Strain-Gage Resistance Measurement.

When dynamic stress is measured, it is possible to AC or capacitively couple the recorder to the gage so that only two lead wires are needed. In the circuit shown in Figure 2, a change in the gage resistance ($\Delta R$) produces the same change in voltage ($\Delta V$) across the gage regardless of the value of the lead resistance ($R_{LEAD}$) since the current through the gage is constant. If the lead resistances ($R_{LEAD1}$, $R_{LEAD2}$) vary slowly with respect to the data, the voltage across each $R_{LEAD}$ is essentially constant, so the $\Delta V$ seen by the recorder is not affected. Normally, $R_{LEAD}$ changes due to temperature which is a very slow change.

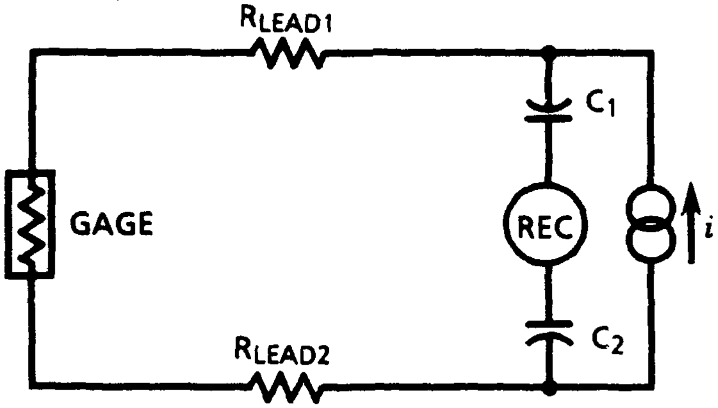


Figure 2. Dynamic Stress Measurement Using Constant-Current Excitation.

A second approach is to use a constant-voltage excitation supply with series or ballast resistances ($R_{BALLAST1}$, $R_{BALLAST2}$) in each lead in lieu of a constant-current supply (see Figure 3). If the ballast resistances are made large enough, changes in $R_{LEAD}$ will have minimal effect on the current through the gage and there will be minimum loading on the output of the gage. In this condition, the performance approaches that of the constant-current excitation circuit. However, maximum gage output for a given excitation voltage occurs when $R_{BALLAST1} + R_{BALLAST2} = R_{LEAD1} + R_{GAGE} + R_{LEAD2}$ so that large value ballast resistances result in a low output signal from the gage or require the use of a large excitation voltage.

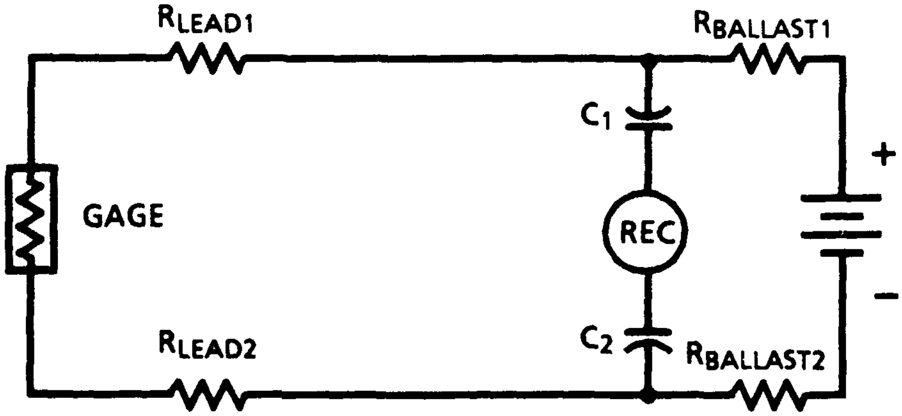


Figure 3. Dynamic Stress Measurement Using Constant-Voltage Excitation.

A common technique with strain gages is to make the gage one arm of a Wheatstone bridge and use dummy resistors in the other three arms. The bridge is then excited with a constant-voltage or, in some cases, a constant-current supply. The effects of lead wire resistance change are compensated for by using a three-wire connection to the active gage. It should be noted that this compensation scheme only corrects for the effects of lead wire resistance change on bridge unbalance and does not eliminate sensitivity changes. With an AC-coupled system bridge balance is of no importance.

There are problems associated with each of these techniques for eliminating the effects of lead wire resistance changes. The constant-current technique has several problems. First, the constant-current generator must respond fast enough to keep the current through the gage constant even with resistance changes equal to the frequency response of the measurement, e.g., 100,000 times a sec. This requires a supply with a fast slew rate capable of driving the capacitive load of the shielded cable to the gage.

Although there are constant-current supplies available that are capable of tracking high-frequency gage resistance changes, the necessary bandwidth of such a supply presents a second problem. The broadband noise on the output of an excitation supply is proportional to the square root of the bandwidth. Since a constant-current regulator has such a wide bandwidth, it follows that the noise on its output must be considerably greater than that of a constant-voltage supply whose bandwidth is typically limited by the large filter capacitor in its output. Because of the response requirement of the constant-current supply, the noise cannot be filtered.

The third problem with the constant-current circuit is not so obvious. A constant-current excitation source usually consists of a constant-voltage power supply with a constant-current regulator on its output (see Figure 4). Since an ideal constant-current source has an infinite resistance, the excitation source as seen by the recorder and gage has a low impedance to the power supply in one lead and a very high impedance to the power supply in the other lead. The common-mode rejection of the recorder is highly dependent upon the balance of the input circuitry, and this unbalanced condition diminishes the ability of the recorder to reject common-mode signals.

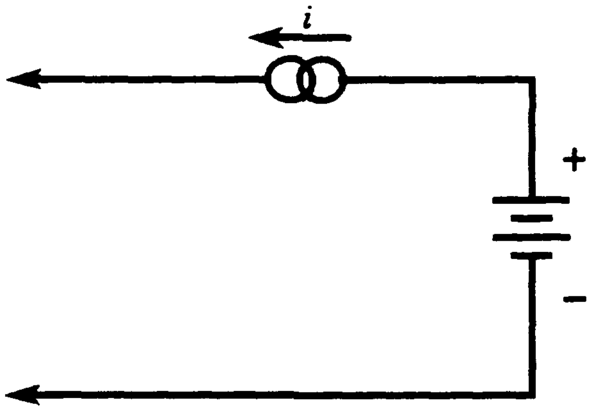


Figure 4. Constant-Current Excitation Source.

These problems of the constant-current circuit are not present with the constant-voltage excitation scheme shown in Figure 3. The circuit, however, is sensitive to changes in $R_{LEAD}$. Changes in $R_{LEAD}$ affect the output in two ways: first, a change in $R_{LEAD}$ will change the current through the gage so that a given $\Delta R$ produces a different signal ($\Delta V$); second, the output seen by the recorder is a function of the voltage divider made up of $R_{LEAD1} + R_{LEAD2} + R_{GAGE}$ and $R_{BALLAST1} + R_{BALLAST2}$. Consequently, as $R_{LEAD}$ changes, the fraction of the gage output seen by the recorder changes. Figure 3 has been redrawn in Figure 5 to show this effect. The only way to minimize the problem would be to greatly increase the values of $R_{BALLAST1}$ and $R_{BALLAST2}$. This would require a very high voltage power supply and large wattage resistors for $R_{BALLAST1}$ and $R_{BALLAST2}$. Also, as $R_{BALLAST1}$ and $R_{BALLAST2}$ are increased in value, circuit balance becomes more sensitive to stray resistances and capacitances and common-mode rejection again suffers.

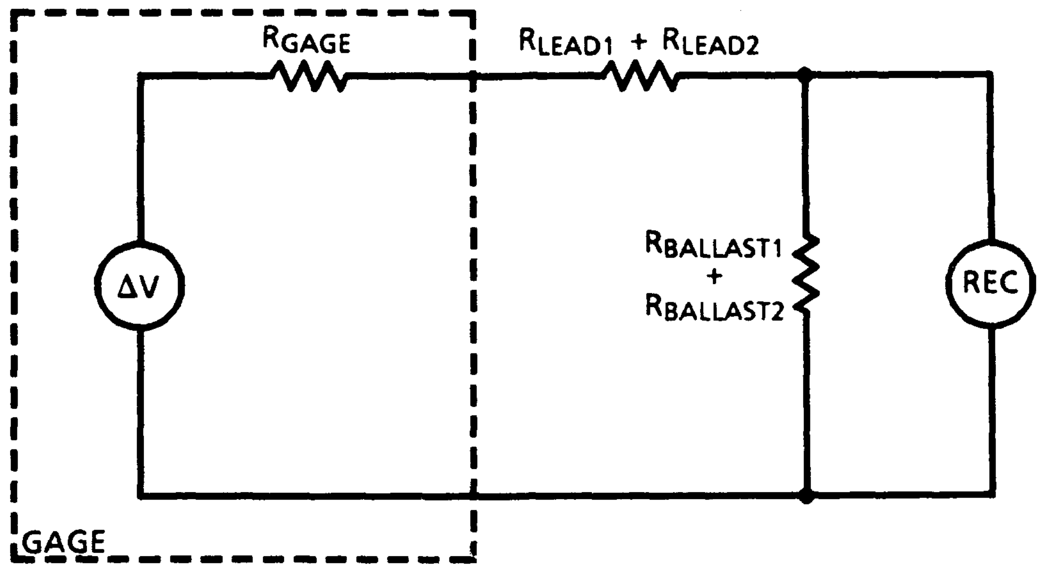


Figure 5. Loading of Gage Ouput by Excitation Supply.

In summary, constant-current excitation supplies are inherently noisier than constant-voltage supplies and have poorer common-mode noise rejection. Constant-voltage excitation supplies, on the other hand, are sensitive to changes in the resistance of the lead wires going to the gage.

## SIGNAL CONDITIONER DEVELOPMENT

### Constant-Resistance Concept Development

After some effort to develop a constant-current excitation supply having balanced impedances in its output leads, it was recognized that the best performance (as far as low internal noise and the optimum rejection of stray signals are concerned) is obtained with a conditioner using a constant-voltage excitation supply. It was then decided to use constant-voltage excitation, but to devise a means of compensating or correcting for the change in lead resistance. The solution found was to add variable resistors, in series with the lead wires, which could be controlled to keep the total resistance constant (see Figure 6). As long as the total loop resistance remained constant, the current through the gage and the loading on the gage would remain constant; therefore, the output signal would not change. Because the lead wire resistance changes slowly with time, its change may be distinguished from that of the gage, which changes very rapidly.

If the total loop resistance is constant, the voltage between points A and B in Figure 6 is constant. This voltage could then be sensed and used to control the series resistances. The resistances are split equally in each lead to keep a balanced situation.

### Variable Resistance Device Selection

Several candidate devices were considered for the variable resistances. Theoretically, motor-driven, ganged potentiometers (pots) would be ideal. The resistance of the two pots could be made to track very closely, and the response time of a motor-driven pot would be adequate since the lead wire resistance changes very slowly. Even though small (0.5-in.-diam) motors and pots are now available, it was felt that this application would require the motor to be continuously hunting or controlling and the life of the pot might be less than desirable.

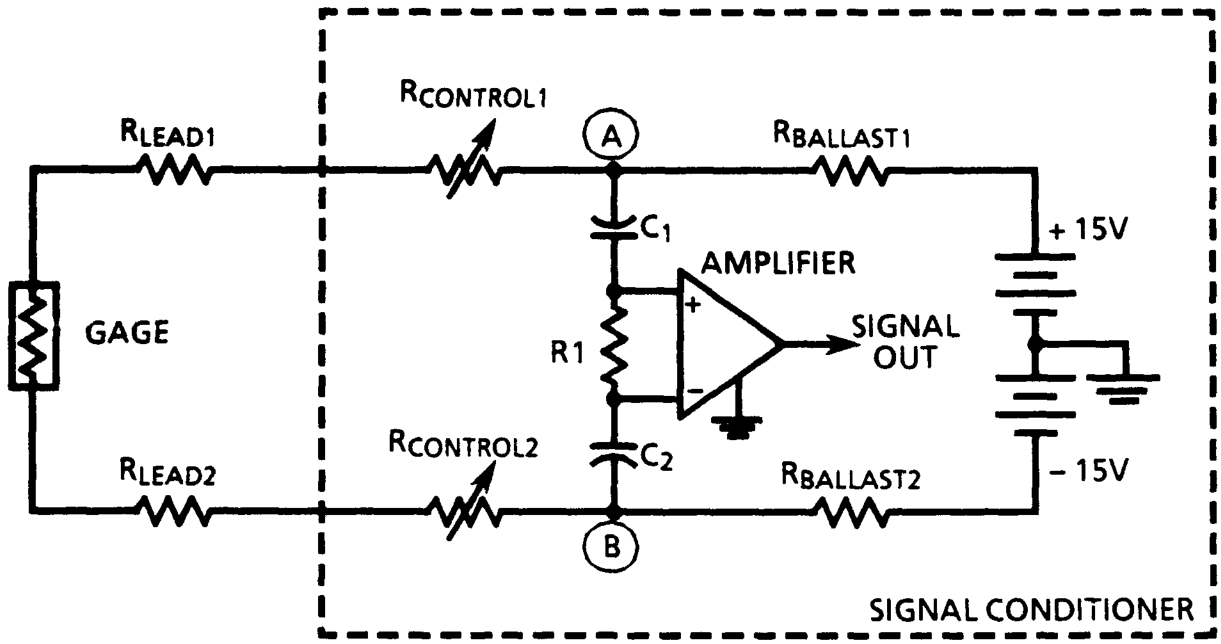


Figure 6. Scheme for Lead Wire Resistance Correction.

The next device considered was the power MOSFET. These devices have a desirably low "on" resistance and require very little power to control. An N-channel device would be required in the negative lead and a P-channel device would be used in the positive excitation lead. With these devices, isolation from the control circuit becomes involved and tracking between the devices may be a problem. Also, there were concerns about noise and the devices not acting like true resistances, particularly at high frequencies.

The device finally selected was a lamp/photoconductive cell combination such as the VACTROL® manufactured by EG&G Vactec.(3) Although several manufacturers make these devices, the Vactec Model VTL5C4 has the lowest "on" resistance for a light emitting diode (LED) lamp-type device. Devices using incandescent lamps have a lower "on" resistance but are microphonic because of the filament of the lamp.

**Variable Resistance Control**

A circuit was designed using the VTL5C4 to provide a variable resistance in each gage lead that would vary from 150 to 50 ohms to accommodate lead resistances between 0 and 100 ohms. (The total resistance in each lead would always total 150 ohms.) The resultant circuit is shown in Figure 7. The voltage difference between points A and B is measured by the control instrumentation amplifier and then compared against a reference voltage by the error amplifier. The output of the error amplifier is buffered by a current driver which drives the LEDs of the VACTROL® LED/photoconductive cells. The compensation capacitor in the error amplifier limits the error amplifier bandwidth so it will not respond to the rapid resistance changes of the gage. The signal instrumentation amplifier provides an AC-coupled strain signal to the data recording system.

To obtain a control device that would have a resistance range of 50 to 150 ohms using the available LED/photoconductive cells, it was necessary to use two of the EG&G Vactec Model VTL5C4 devices in parallel with a 150-ohm resistor (see Figure 8). One of these parallel combinations is used in each lead wire.

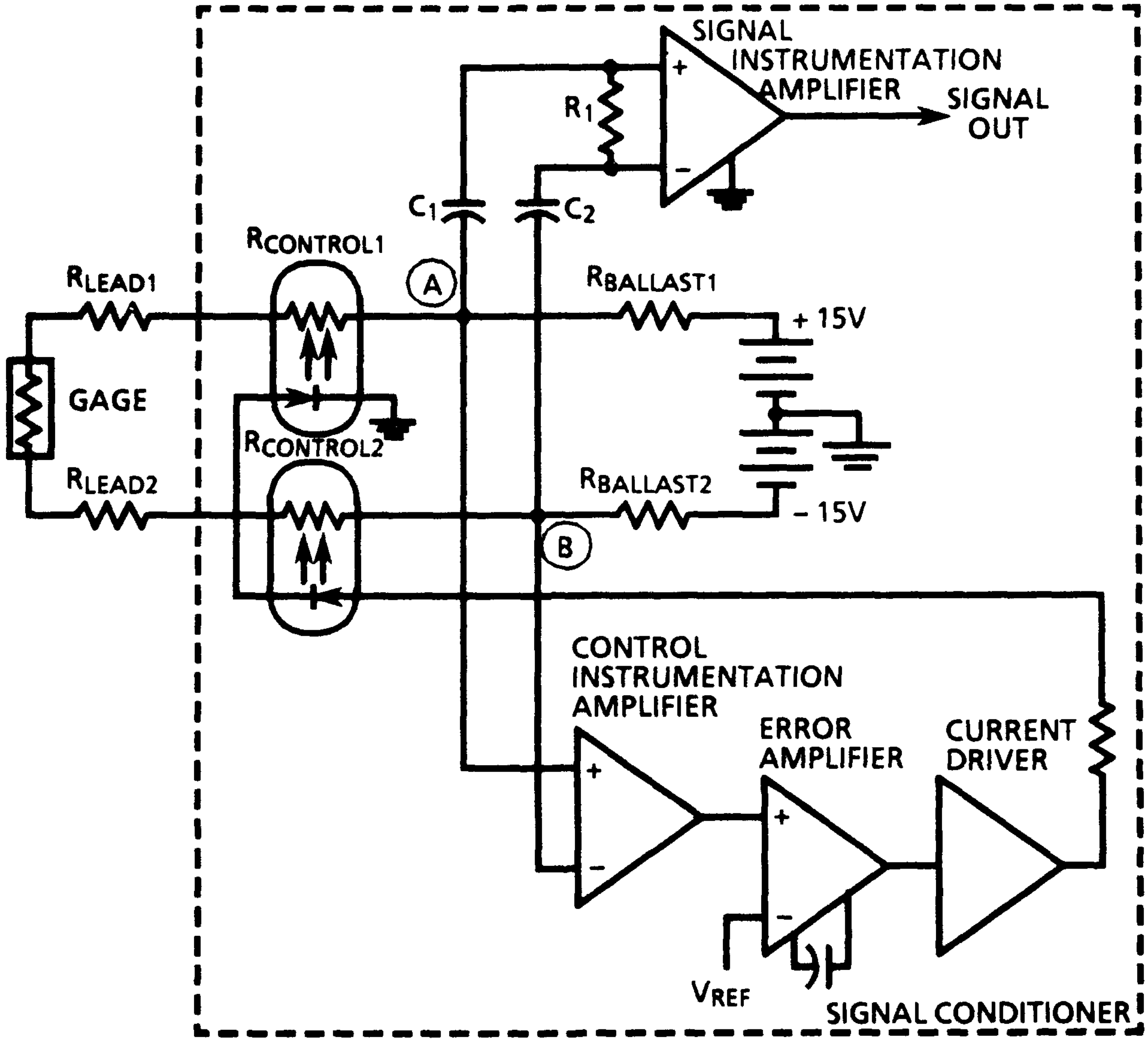


Figure. 7. Circuit to Correct for Lead Wire Resistance Change.

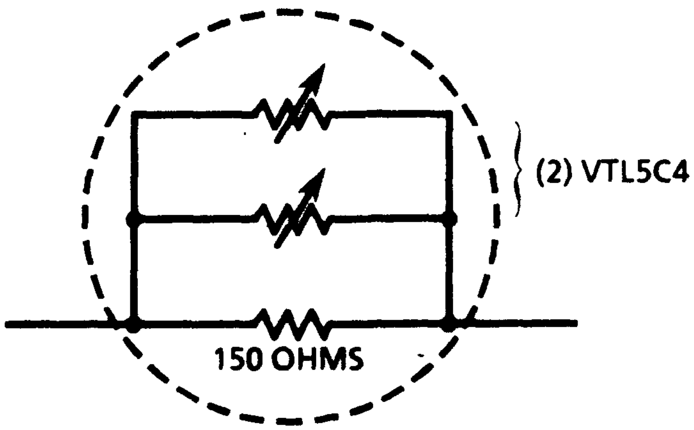


Figure. 8. Parallel Combination of VACTROLs®.

In operation, if the gage lead resistances should increase, the voltage between points A and B of Figure 7 will increase. This will result in a greater positive voltage into the error amplifier and a greater signal out. More current, then, will be driven though the VACTROL® LEDs, reducing the resistance of the photoconductive cells and correcting for the increased lead resistances. Since the circuit functions by maintaining a constant-resistance in each sensor lead, the circuit has been named "constant-resistance." An invention disclosure has been submitted to the Air Force on the constant-resistance excitation technique.

### ΔR Calibration Circuit Concept

Calibration of dynamic strain-gage systems is customarily done by disconnecting the gage and its power supply from the input to the data instrumentation amplifier by means of relays and substituting a sine wave signal from a calibration source. The amplitude of the calibration signal is related back to stress by an involved calculation involving Young's modulus, gage resistance, gage current, lead resistance, load resistance, etc. To simplify the calibration process and reduce the errors resulting from any change in lead resistance or gage excitation, a new calibration scheme was devised. This scheme uses a small resistor in series with the gage whose resistance is changed by an amount equal to the change in gage resistance produced by a given amount of strain. This relationship may be determined from manufacturer's specifications for the strain gage.

The circuit as shown in Figure 9 consists of a precision 10-ohm resistor in the negative excitation lead to the gage that is shunted by an $R_{CAL}$ resistor to produce a resistance change that is equal to a similar resistance change in the gage. Typically, $R_{CAL}$ might be 390 ohms, which will produce a 0.25-ohm change. This technique might be compared to the shunt resistance calibration of a commercial strain-gage pressure transducer. A second 10-ohm resistor is placed in the positive excitation lead to maintain circuit balance.

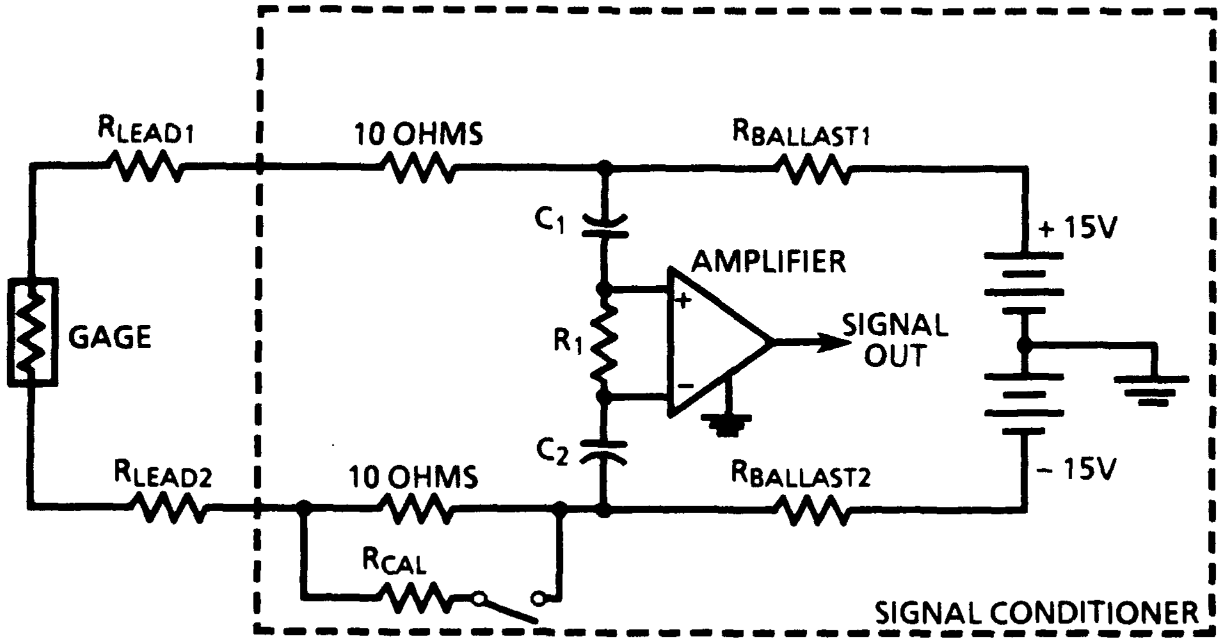


Figure 9. Calibration Scheme.

Since the recording data system is capacitively coupled and will not sustain the level resulting from a step change in resistance for any length of time, the calibration switch is alternately opened and closed for 500 μsec, producing a 1-kHz square-wave signal. The peak-to-peak value of this square wave is equivalent to a known value of peak-to-peak stress. It should be noted that this calibration scheme is usable with any type of strain-gage conditioner and will eliminate any errors caused by lead wire resistance changes.

### Calibration Switch Selection

The difficulty in implementing the calibration scheme of Figure 9 lies in the switch. There are high-speed relays that will operate at speeds approaching those required, but the life of these devices, which is quoted typically at $10^9$ operations, might soon be consumed.

JFET transistors and CMOS switches are currently used for switching applications where high-speed switching and essentially infinite lifetimes are requirements. However, the "on" resistance of these devices is 50 to 100 ohms or more, which would cause an unacceptable error.

Power MOSFETs have recently become popular for use in switching type power supplies and similar applications. To obtain maximum efficiency from the power supply, these devices have "on" resistances of less than an ohm. Although designed for switching high currents, the "on" resistance of a power MOSFET actually decreases slightly at low currents. As mentioned previously, the power MOSFET requires very little power to drive; its gate appears as a capacitor. A Type IRF511 power MOSFET (4) with an "on" resistance of 0.6 ohm was selected as the switching device.

Figure 10 shows the circuit arrangement used to drive the power MOSFET and isolate it from the rest of the circuit. The power MOSFET requires a positive pulse on its gate of at least 10 volts to turn it on rapidly. This is produced by a Hewlett-Packard type HCPL-2201 opto-isolator (5) which has a totem pole output and will operate with Vcc voltages from 4.5 to 20 volts. The DC-DC converter is a Burr-Brown Type 700 (6) and has an input/output capacitance is only 3 pF. The optical isolator is driven from a 1,000-Hz square wave source. Exact amplitude is unimportant as long as it will turn on the LED of the isolator; calibration accuracy is controlled by the 390-ohm $R_{CAL}$ resistor.

The power MOSFET used in the calibration circuit to switch the $R_{CAL}$ resistor adds small, narrow spikes to the trailing edge of the square wave calibration signal. These are a result of the MOSFET gate-driving signal being charge coupled to the output of the MOSFET. Although reduced by the low-pass filter in the amplifier, they are still somewhat noticeable on the output at the higher frequency low-pass filter settings and may have some effect on the calibration process.

### Alternate Switching Device

Optically-coupled power MOSFETs (Type AQV221 Photo-MOS) from Aromat Corporation (7) were also tried in the calibration circuit. These devices do not require a separate isolated power supply and may be driven directly from TTL signals. By placing one device in the positive signal lead and a second in the negative signal lead, the gate charge spikes may be canceled out. These devices do not have as low an "on" resistance as the IRF511 and so require two devices in parallel for a satisfactorily low "on" resistance (making a total of four devices). Two $R_{CAL}$ resistors, each of twice the value, are also required. Although more MOSFETs are used, the DC-to-DC converter is no longer required.

The main disadvantage of the Photo-MOS switching devices is their switching speeds which are measured in the tens of microseconds. They have slightly different turn-on and turn-off times which produces a square wave having other than a 50-percent duty cycle. The units evaluated produced a square wave with 49-percent "on" and 51-percent "off" times.

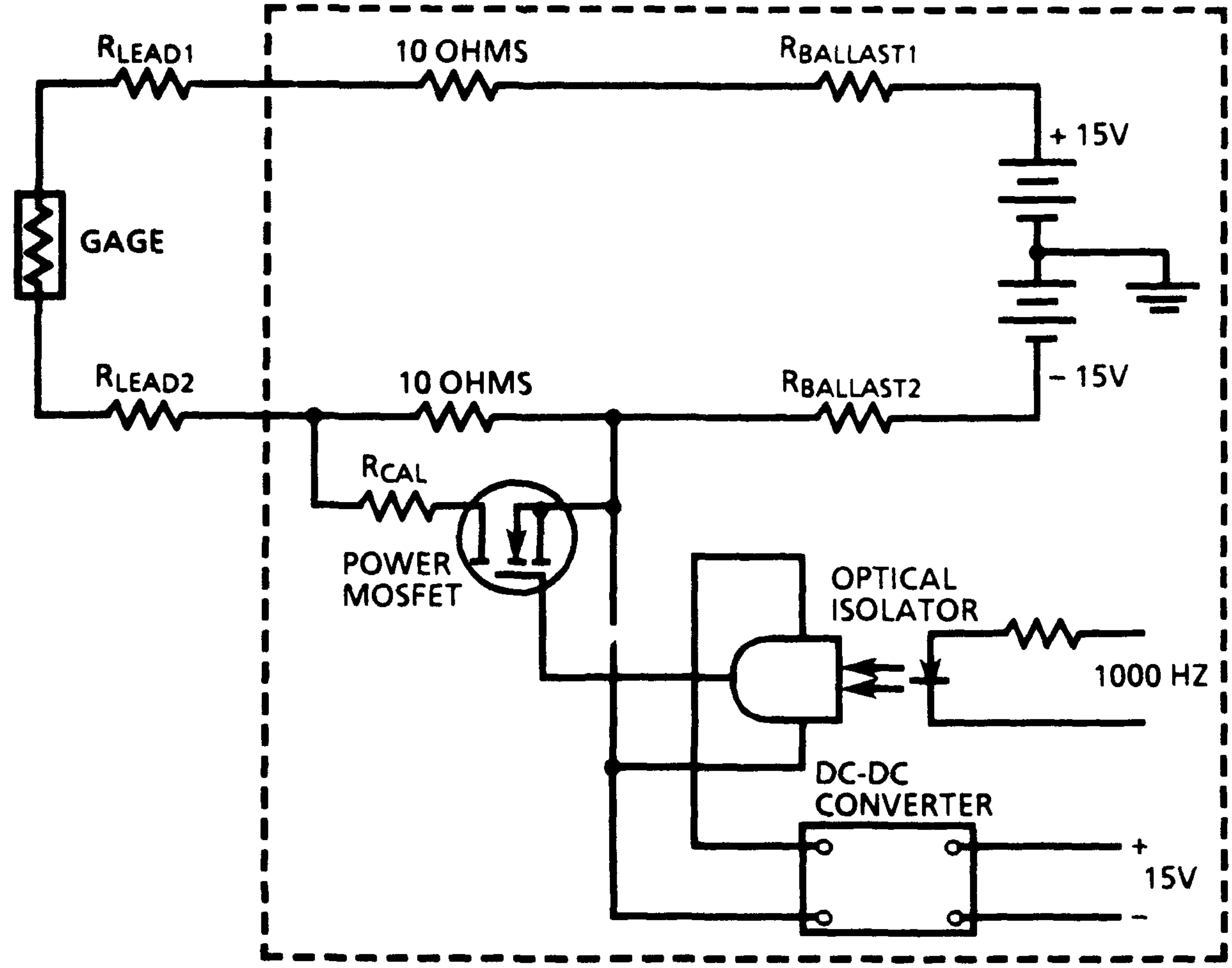

Figure 10. Calibration Circuit Using Power MOSFET.

The ΔR calibration circuit, in addition to producing a signal that is directly equivalent to gage resistance change, provides a check of the overall recording system frequency response. The square wave calibration signal reveals inadequate high-frequency response by a rounding of its leading edge. A drop or tilt in the top and bottom of the waveform indicates poor low-frequency response.

## SIGNAL CONDITIONER PERFORMANCE EVALUATION

The prototype constant-resistance signal conditioner was tested for excitation regulation accuracy, internally generated noise, and for common-mode rejection. For comparison purposes, measurements of internal noise and common-mode rejection were also made on a typical signal conditioner used at AEDC: Neff Instrument Company (Neff) signal conditioning amplifier (SCA).† Tests were conducted with the SCA in its normal constant-

† The Neff SCA is a signal conditioner for dynamic data built to government specifications for use in the Aeropropulsion Systems Test Facility (ASTF) at the AEDC. It consists of a constant-voltage/constant current excitation supply, a millivolt substitution and shunt resistance calibration system, and a programmable-gain instrumentation amplifier with programmable low-pass filter. The SCA uses plug-in mode cards for different sensor types.

voltage mode and on the same Neff SCA with a constant-current regulator built up on a plug-in mode card (see Figure 11).

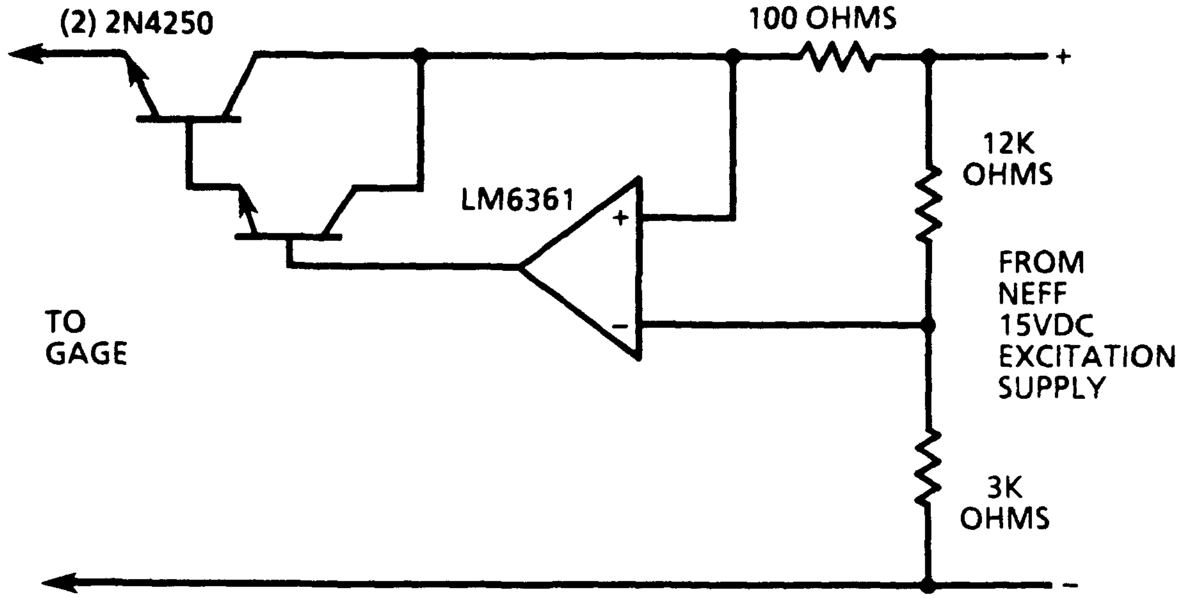


Figure 11. Constant-Current Regulator for Neff Signal Conditioner.

**Excitation Regulation**

Regulation accuracy was determined by measuring with a digital voltmeter the voltage developed across the sensor resistance for different values of lead wire resistance. These voltages were then converted to the corresponding gage currents using Ohm's law. The voltages were measured with a 120-ohm gage resistance and lead resistances (in each lead) of 0, 20, 40, 60, 80, and 100 ohms.

The results of the measurement of regulation accuracy are given in Table 1. The change in current through the gage for the full 100-ohm change in gage resistance is less than 0.05 percent. The absolute error in setting the gage current to 30 ma (0.4 percent) is caused by the 1-percent tolerance resistors used in the circuitry. The current regulation obtained using the constant-resistance approach is considered very good. The current setting accuracy could be improved by substituting resistors of higher precision but is considered to be adequate for this application.

Table 1. Excitation Regulation Accuracy for Constant-Resistance Signal Conditioner

| Lead Resistance, ohms | Voltage across Gage, volts | Current thru Gage, milliamperes |
|---|---|---|
| 0 | 3.5855 | 29.879 |
| 20 | 3.5854 | 29.878 |
| 40 | 3.5853 | 29.877 |
| 60 | 3.5852 | 29.877 |
| 80 | 3.5849 | 29.874 |
| 100 | 3.5838 | 29.865 |

## Conditioner Internal Noise

Internal noise comes from the input amplifier and the excitation supply of the conditioner and is a function of the bandwidth of the data amplifier. For comparison purposes, the noise was measured on all three conditioners using a bandwidth of 80 kHz. The low-pass filter in the Neff SCA is a 6-pole Bessel, while the prototype conditioner has only a 4-pole Bessel filter.

Internal noise was measured using the circuit shown in Figure 12. Two series-connected 120-ohm metal film resistors were connected directly across the + and − signal input pins of the conditioner and their junction connected to the guard (shield) input. The resistors represent a 120-ohm strain gage and two lead wires each having a resistance of 60 ohms. The excitation was set to give 30 ma through the gage. The peak-to-peak noise was then measured at the output of the conditioner using a Tektronik Model 2465 oscilloscope with its bandwidth limited to 20 MHz.

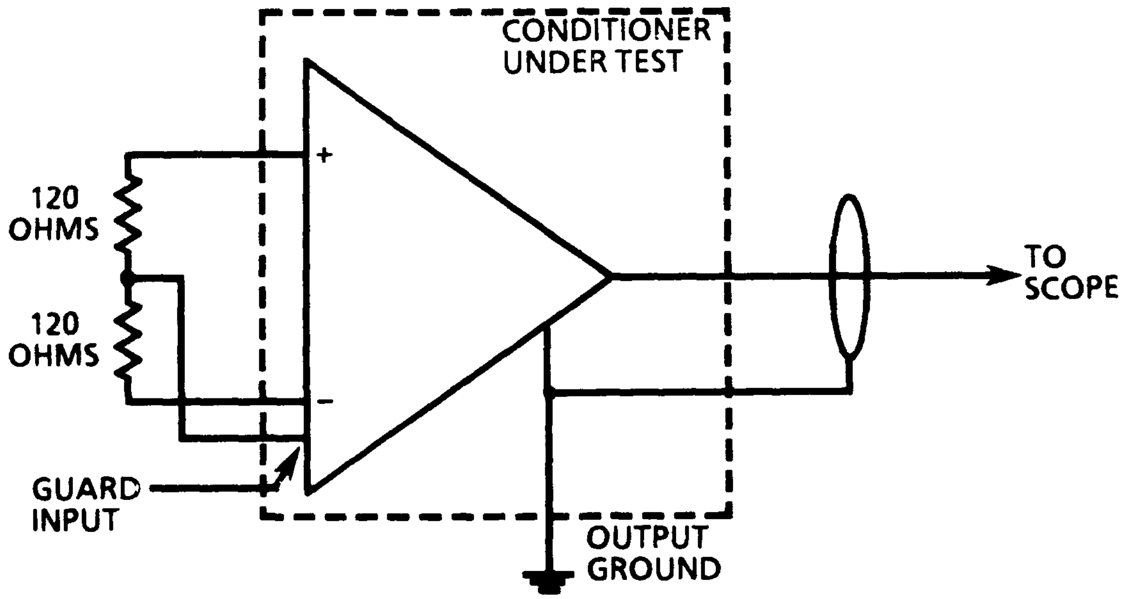


Figure 12. Measurement of Conditioner Internal Noise.

The internal noise was measured for each of the three conditioners. The values obtained were then divided by the gain of the amplifier (800X for the prototype and 256X for the Neff) to give noise in microvolts referred to the input (RTI). See Table 2.

Table 2. Comparison of Internal Noise of Signal Conditioners

| Constant-Resistance F.S. = 4.35 mv | | Neff SCA Constant-Voltage F.S. = 3.75 mv | | Neff SCA Constant-Current F.S. = 7.5 mv | |
|---|---|---|---|---|---|
| RTI | Percent Full Scale | RTI | Percent Full Scale | RTI | Percent Full Scale |
| 41 µv P-P | 0.95 | 35 µv P-P | 0.94 | 139 µv P-P | 1.85 |

NOTES:

RTI is noise at conditioner output divided by gain of amplifier.

Bandwidth of amplifiers is 80 kHz. Neff has a 6-pole Bessel LP filter and prototype constant-resistance has 4-pole Bessel.

F.S. assumes a 0.25-ohm change in a 120-ohm gage.

The percent of full-scale signal was next calculated by dividing the RTI valuє by the full-scale voltage at the input to the instrumentation amplifier and multiplying by 100. Full-scale voltage is defined as that voltage resulting from a 0.25-ohm change in a 120-ohm gage (4.35 mv for the constant-resistance circuit, 3.75 mv for the constant-voltage circuit, and 7.5 mv for the constant-current circuit). The results are given in Table 2 and illustrate that the constant-resistance circuit has a noise level less than the constant-current technique and comparable to that of the constant-voltage approach. In addition to the wide-band noise given in Table 2, the Neff constant-current conditioner had 60-Hz spikes having an amplitude of 217 µv peak-to-peak (2.9 percent).

**Common-Mode Rejection**

Common-mode rejection was measured using the circuit of Figure 13. The two 60-ohm resistors on the input represent the 120-ohm strain gage. The 80-ohm resistor and the 40-ohm resistor represent typical thermocouple wire resistances. The 100 ft of Belden 8451 shielded-pair cable represents the wiring from the sensor to the signal conditioner location. Common-mode rejection was checked at 1, 2, 5, 10, 20, 50, and 100 kHz for each of the three conditioner types, with low pass filters set for a bandwidth of 80 kHz. A second measurement of common-mode rejection was then made with the 40- and 80-ohm resistors simulating the lead wire resistances changed to 60 ohms to provide a balanced condition.

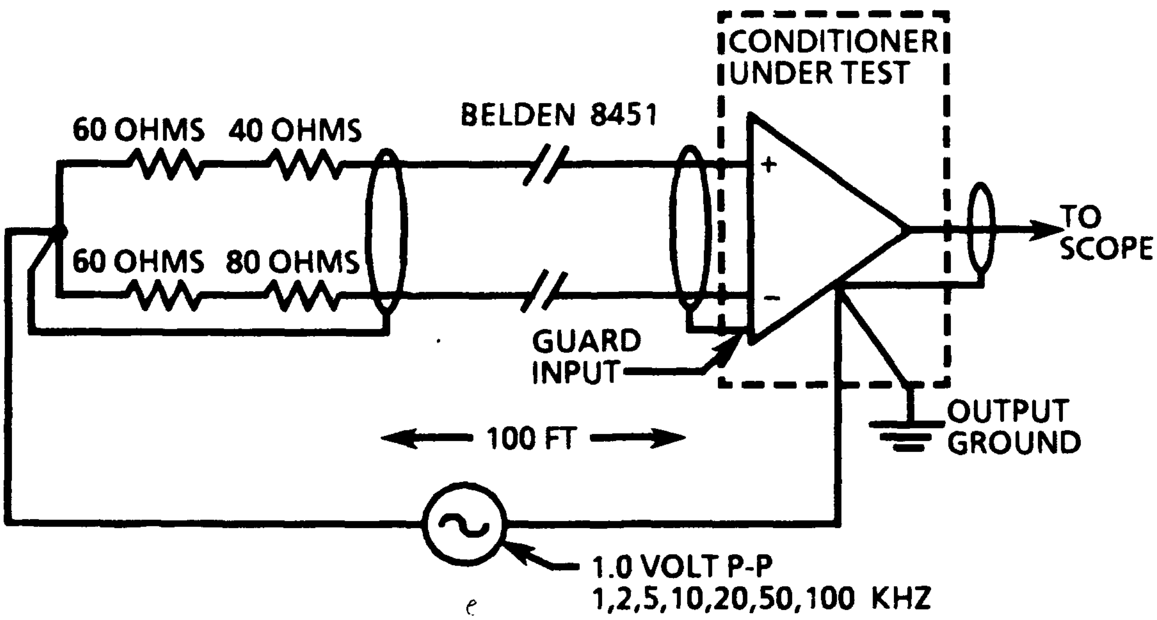


Figure 13. Measurment of Common-Mode Rejection.

The common-mode rejection measured for the each of three types of conditioners is shown in Figures 14 and 15. The plots show absolute common-mode rejection in decibels. To compare signal-to-interference performance between conditioners it is necessary to take into account the different gage output signal level obtained with each of the circuits. When the constant-resistance conditioner is compared to the constant-voltage conditioner, the improvement must be increased by 1.29 decibels. Comparison of constant-resistance and constant-current data requires a decrease of 4.73 decibels in the improvement; comparison of contant-voltage and constant-current data requires a decrease of 6.02 decibels in the difference.

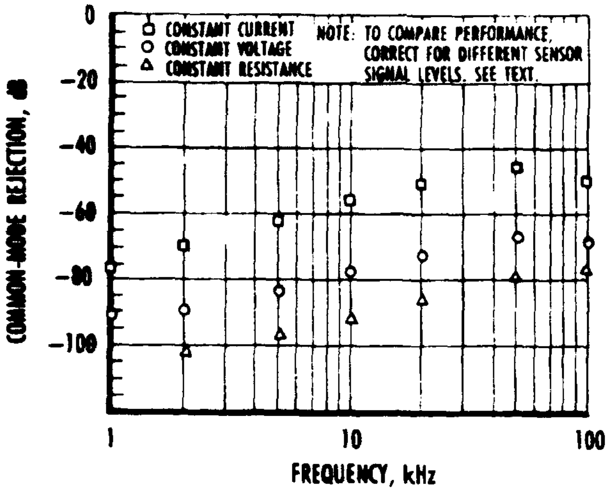


Figure 14. Common-Mode Rejection with Shield Tied to Common Mode at Sensor, Unbalanced Lead Resistances.

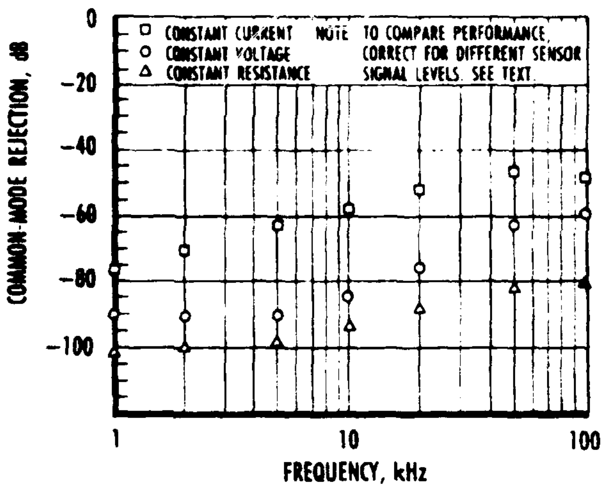


Figure 15. Common-Mode Rejection with Shield Tied to Common Mode at Sensor, Balanced Lead Resistances.

For example, as shown in Figure 14, the constant-current conditioner has a common-mode rejection at 10 kHz of −56.1 decibels and the constant-resistance conditioner has a common-mode rejection of −92.0 decibels. The difference between the two curves of 35.9 decibels should be decreased by 4.73 decibels, resulting in an actual improvement in common-mode rejection for the constant-resistance conditioner of 31.1 decibels.

Generally, the common-mode rejection deteriorates at a rate of 6 decibels per octave because of unbalanced input resistances and capacitances. Above about 50 kHz, the roll off of the low-pass filter reduces the signal at the conditioner output, thus improving the apparent common-mode rejection. At low frequencies, the measured common-mode rejection levels off when the common-mode signal level approaches the internal noise level of the conditioner. Actually, the 6 decibels per octave slope continues on to below 60 Hz and could be measured by increasing the level of the common-mode signal applied to the conditioner. The increased common-mode input level, however, would cause amplifier overload at the higher frequencies of interest.

The prototype constant-resistance signal conditioner has more than 30 decibels improvement in common-mode rejection over the constant-current circuit at all frequencies. The reason(s) for the improvement over the constant-voltage conditioner has not been investigated. The increased common-mode rejection may result from the use of a balanced ±15-volt excitation supply or possibly to the use of integrated circuit devices in the input stages of the instrumentation amplifier.

## SUMMARY

Recent advances in turbojet engine testing have placed greater demands on the data systems used to measure dynamic stress. A new signal conditioner using the constant-resistance technique has been designed and prototyped. Tests on the prototype show that the new circuit offers significant improvement in high-frequency common-mode rejection while simultaneously showing immunity to the effects of errors caused by slowly changing sensor lead-wire resistance. With this circuit, the amplitude of the signal from the strain gage is unaffected by resistance changes of up to 100 ohms.

The new conditioner includes a novel ΔR calibration circuit that provides a simple and more accurate way to simulate the resistance changes of the sensor and which is usable with any type of gage excitation circuitry. The square wave calibration signal produced by the ΔR circuit also allows verification of system frequency response.

## REFERENCES


1. "Practical Strain-gage Measurements," Application Note 290-1, Hewlett-Packard Co., Palo Alto, CA, 1987.

2. Stein, Peter K., Measurement System Engineering Short Course Lecture Notes, Stein Engineering Services, Inc., Phoenix, Arizona, 1982.

3. Optoelectronics. EG&G VACTEC, St. Louis, MO, pp. 39-48.

4. HEXFET Power MOSFET Databook. International Rectifier, EL Segundo, CA, 1984, pp. D-97 – D-102.

5. Optoelectronics Designer's Catalog 1988-1989. Hewlett-Packard, Palo Alto, CA, 1987, pp. 9-15 – 9-18.

6. Integrated Circuits Data Book, Volume 33. Burr-Brown Corporation, Tucson, AZ, 1989, pp. 4-86 – 4-87.

7. Relay Catalog. Aromat Corporation, New Providence, NJ, 1989, pp. 259-279.