Comparing the effect of lead and switch resistances on voltage- and current-fed strain-gage circuits

by Manfred Kreuzer

Strain gages are passive measuring elements and therefore require an electrical input to operate. The two main measuring circuits in use today are current-fed circuits and voltage-fed bridge circuits. This article describes both types and discusses their advantages and disadvantages. The author also introduces a new bridge circuit which permits error-free measurement with strain gages in quarter-, half-, and full-bridge circuits by not allowing lead or contact resistance at selector switches to falsify readings. The new circuit permits the use of single strain gages with long leads.

In modern scanning systems, a large number of transducers are connected via suitable intermediate circuitry to a central measuring amplifier and signal conditioning unit. Transducers may either be selfgenerating signal sources such as thermocouples or voltage and current signal sources, or passive sensors such as strain gages, inductive pick-ups, resistance thermometers or other resistive pick-ups in quarter-, half- or full-bridge circuits.

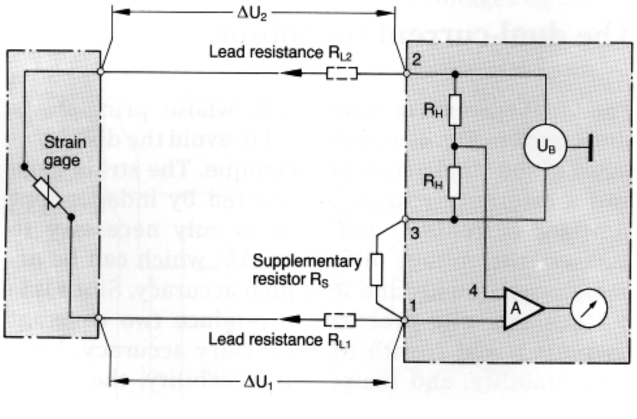


Fig. 1: Two-wire bridge circuit with 100-m lead length produces zero shift of over 100,000 µm/m

Intermediate circuitry plays an extremely important role in passive-sensor scanning. It's two primary functions are to provide excitation and ensure that the measurements are taken as error-free as possible for the period that the sensor is on-line. One prime example is the temperature-compensation circuitry often necessary for thermocouples. The intermediate circuitry must also be designed so that voltage drops across the measuring leads and switching elements cannot falsify the readings.

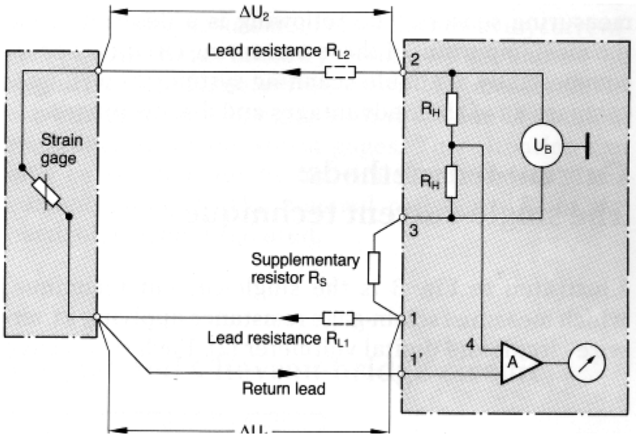


Fig. 2: Three-wire bridge circuit reduces zero shift, but still exhibits significant temperature dependence

This feat is especially difficult to accomplish when applying strain gages in half- and quarter-bridge circuits because changes in measured resistance are extremely small. For example, a strain of $\varepsilon = 1$ µm/m results in a change in resistance of only $\Delta R = 0.00024$ Ohm for a metal strain gage having a nominal resistance of $R_0 = 120$ Ohm. From these data, it's easy to understand how changes in resistance in the measuring circuit can cause enormous errors.

Take, for example, the case of a single strain gage connected with standard 0.14-mm^2 (0.0002-in^2) cross section copper leads over a distance of 100 m (ca. 330 ft) in

Manfred Kreuzer *is Manager of Research and Development, Electronic Systems, for HBM in Darmstadt*

a simple two-wire bridge circuit as shown in **Fig. 1**. The lead resistance *alone* produces a zero shift of over 100,000 µm/m, a zero temperature dependence of over 400 µm/m per 1 K (222 µm/m per 1°F), and a reduction in sensitivity of more than 10 %. Although the zero shift can be reduced by using the three-wire circuit shown in **Fig. 2**, zero temperature dependence will still be 4 µm/m per 1K (2 µm/m per 1°F) even if the resistance of the supply and return differ by only 1 %. Sensitivity reduction is unaffected by the three-wire circuit and remains at 10 % as before.

Controlling switch resistance in the circuit is equally critical. If the strain gages are switched by means of field-effect transistors having a forward resistance of around 20 Ohms and a temperature coefficient of 0.4 % per 1 K (0.2 % per 1°F), zero shift is greater than 160,000 µm/m and zero temperature dependence greater than 640 µm/m per 1K (356 µm/m per 1°F). This error is even larger than the measured signal itself.

These examples show quite clearly that with the normal Wheatstone-bridge circuit it is very easy to get into insoluble difficulties, especially if there are high lead resistances and the switching devices do not have an extremely low reproducibility error (< 0.00025 Ohm). Even the use of the three-wire circuit or Thomson circuit [1] bring no noticable improvement.

Modern high-feedback electronic circuits, however, permit the design of intermediate circuits which can prevent the undesirable influences described above. There are two fundamental solutions. One is again based on the Wheatstone bridge circuit but avoids its drawbacks by means of electronic control circuits; the other is based on a constant-current supply to the measuring sensors. The following is a description of the most important of these measuring circuits used in commercially-available scanning systems as well as a comparison of their advantages and disadvantages.

Current-fed methods: The single-current technique

Illustrated in **Fig. 3** is the single-current technique, which measures strain-gage resistance applying a four-wire circuit and digital voltmeter [2]. Each strain gage (SG) is fed with the constant current i_C via a switch and two leads; the voltage signal is picked up from each gage by two return leads and fed to the digital voltmeter via two switching contacts.

Since a strain of 1 µm/m only alters the voltage at the gage by 0.0002 %, the voltmeter must have a very high resolution — greater than 500,000 digital steps — in order to measure a strain of 1 µm/m. Such voltmeters are very expensive and also very slow. Since the return leads carry no current, lead and switch resistances in this circuit do not interfere with the results. Even with this circuit, however, it is possible for large errors to occur. Reason: The constant current i_C and the sensitivity of the voltmeter cannot be kept at the extremely high level of stability needed without incurring unreasonable cost.

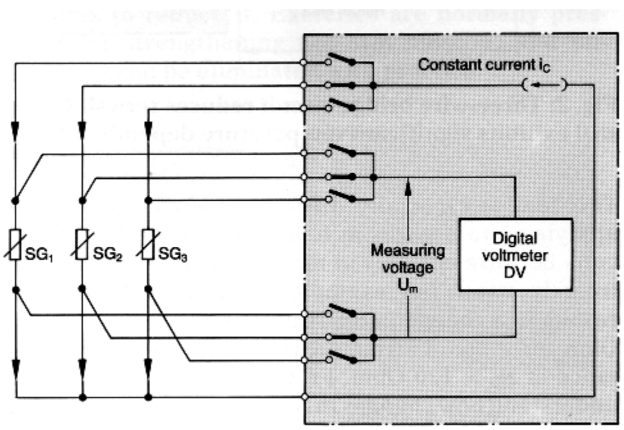


Fig. 3: Single-current-technique accuracy depends on voltmeter sensitivity and current stability. For stability, a voltmeter with high resolution is required

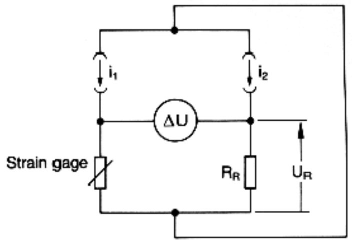


Fig. 4: In the dual-current technique, errors in the two constant currents must be minimized by check measurements and correction calculations

The dual-current technique

The dual-current technique [3], whose principle is illustrated in **Fig. 4**, is often used to avoid the disadvantages of the single-current technique. The strain gage and a reference resistor R_R are fed by independent constant currents i_1 and i_2. It is only necessary to measure the voltage difference ΔU, which can be accomplished quickly and with high accuracy. Since it is not economically practical to produce two constant currents i_1 and i_2 with the necessary accuracy, longterm stability, and temperature stability, the errors must be minimized by check measurements and correction calculations.

The most widely used version of the dual-current circuit is shown in **Fig. 5**. The circuit operates in this way: A constant current i_2 is imposed on a reference resistor R_R via switch S3, producing voltage U_R. A second constant current i_1 is imposed via the switches S2 and S1 either on the strain gage or on the supplementary resistor R_S, producing the voltages U_M and U_S, respectively. The voltages across the strain gage and the supplementary resistor R_S are picked up via two return leads, selected by means of two further contacts of switch S1 and added to voltage U_R. The voltage differences ($U_M - U_R$) and ($U_S - U_R$) are then fed to an amplifier A. By operating switches S2 and S3 it is possible to ensure that only the individual voltages U_M, U_S, or U_R are present at the input of the amplifier.

With this technique it is again necessary for all the voltages U_M, U_S, and U_R across the resistors to be measured with high precision and a resolution of approximately one-million digital steps in order to attain required accuracies of 1 µm/m.

After a number of measurements have been taken, the relative change in resistance $\Delta R/R_0$ can be calculated with the aid of the computing device MC according to equation (1):

$$\frac{\Delta R}{R_0} = \frac{U_{S0}\,U_M}{U_{M0}\,U_S} - 1 \qquad (1)$$

The values with the zero subscript are zero-balance values, U_M is the voltage across the strain gage at the time of measurement, and U_S is the voltage across the supplementary resistor shortly before or after the measurement of U_M. Since the voltage U_R does not show up in the equation, it would appear unnecessary to have a second current source i_2. This is not the case, however. Reason: Operation without i_2 would necessitate a very long integration time for analog-to-digital conversion of the total voltage U_M across the strain gage. By using the second constant-current source i_2 and the reference resistor R_R, the time required for analog-to-digital conversion is reduced considerably because only the voltage difference (U_M - U_R) must be converted:

$$\frac{\Delta R}{R_0} = \frac{U_{S0}\,[(U_M - U_R) + U_R]}{U_{M0}\,U_S} - 1 \qquad (2)$$

Thus, the circuit allows faster measurement (i.e. integration time) while maintaining high resolution. Note: Although measuring the individual voltages U_R and U_S in equation (2) involves equally long integration times (ca. one second), it is not generally a serious disadvantage because the measurements are required only at extended time intervals.

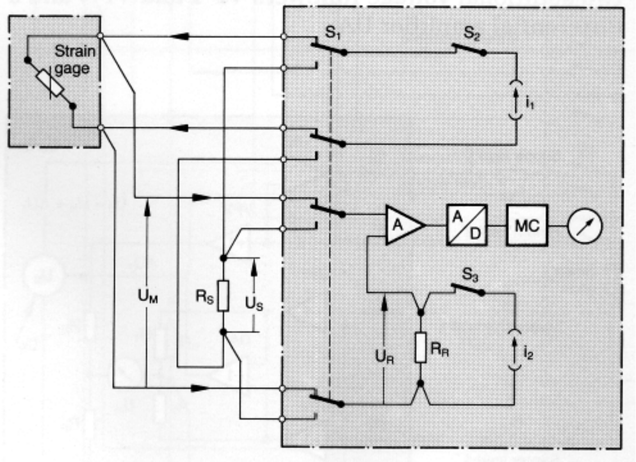


Fig. 5: In dual-current technique, voltage drops across switches and leads have no effect on readings, but resolution better than 1 µm/m is difficult to achieve

The great advantage of the dual-current technique, of course, is that the voltage drops across the switches and leads have no effect on the readings. There are several disadvantages, however. Since the dualcurrent technique is a discontinuous measuring method, it is not suitable for the continuous measurement of time-dependent processes. Also, resolution higher than 1 µm/m is very difficult to achieve because of the extreme demands made on the analog-to-digital converter. The technique is even less suitable for carrier-frequency operation; so far it has only been used in dc form.

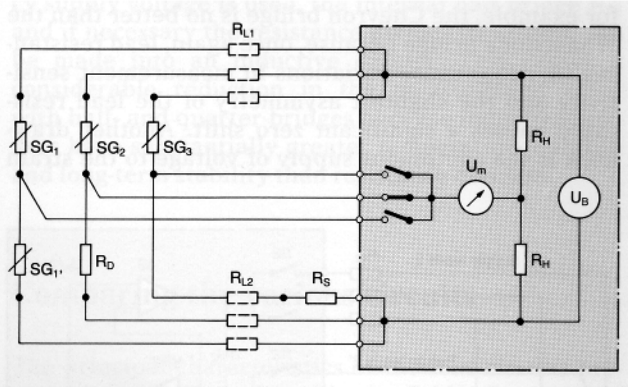


Fig. 6: Although the Chevron bridge operates without switches in the current-carrying leads, lead resistances still hamper sensitivity and slight asymmetry in lead resistance causes large zero shift

All current-fed measuring circuits possess a very linear characteristic for the measurement of change in resistance. For the measurement of strain, however, currentfed circuits have linearity errors which cannot be neglected [4]. These errors cause the measurement sensitivity to vary linearly with the value of nominal resistance R_0 of the strain gages. This situation can only be avoided only by individual adjustment of the current supply to the nominal resistance R_0 of the particular strain gage used.

Voltage-fed Chevron-bridge circuit

So far it has been shown that the three-wire circuit helps eliminate inaccuracies caused by lead resistance. But since this method is not effective when time-variable switch resistances exist in the measuring circuit, the obvious alternative is to manage without any switches in the current-carrying leads. This concept has been carried out in the Chevron bridge [2] as shown in **Fig. 6**.

Main characteristics of the Chevron bridge are:

- All measuring points must be connected individually as "half bridges".
- These half bridges are kept constantly in parallel with the bridge supply voltage.
- In order to complete the full bridge, an internal half bridge (resistors R_H) must also be connected directly to the bridge supply voltage.

Since switches exist only in the currentless return leads, no switch resistances interfere with the readings.

The Chevron bridge also possesses, however, a number of serious drawbacks which can cause large errors in certain applications. With regard to lead resistance, for example, the Chevron bridge is no better than the Wheatstone bridge because, once again, lead resistances (R_{L1}, R_{L2}) cause variations in measurement sensitivity and the slightest asymmetry of the lead resistance causes a significant zero shift. Another drawback is the continuous supply of voltage to the strain gages, which may cause undesirable effects. Example: When taking measurements on plastic models, inadequate heat dissipation can cause significant drift errors. Cost is also a factor when using single strain gages, because each measuring point requires its own supplementary resistor R_S in a half-bridge.

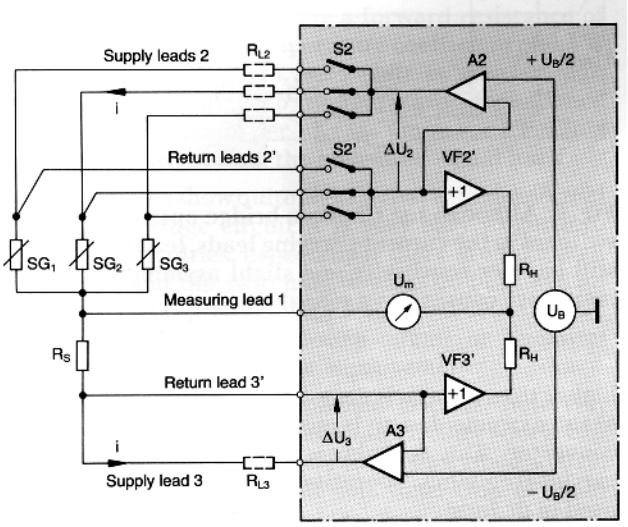


Fig. 7: New bridge circuit reduces zero and sensitivity errors significantly with internal feedback circuits which correct for voltage drops across the leads. Supplementary resistor must be placed close to gages

The HBM bridge

HBM introduced in 1976 a circuit [5] which is also based on the Wheatstone bridge but avoids its error-producing drawbacks almost completely. **Fig. 7** illustrates its principle. The bridge supply voltage U_B is no longer fed directly to the strain gages but simply acts as a reference voltage for amplifiers A2 and A3. These amplifiers correct their output voltages until the voltages picked up via the return leads and fed back via the voltage followers VF2' and VF3' are equal to voltages $+U_{B/2}$ and $-U_{B/2}$. The voltage drops ΔU_2 and ΔU_3 on the supply leads and the supply switches S2 are therefore eliminated. Since the output voltages of amplifiers A2 and A3 exceed voltages $+U_{B/2}$ and $-U_{B/2}$ by the amount of the voltage drops ΔU_2 and ΔU_3, the precise bridge supply voltage is available at the gages.

Due to the extremely high input impedances of the amplifiers and voltage followers, the return leads carry practically no current so their resistance and the resistance of the switches S2' cannot cause voltage drops, i.e. measuring errors.

The internal half-bridge (resistors R_H), which completes the full bridge, is connected to the voltages returned from the measuring points via the voltage followers. Thus the external and internal half-bridges are at precisely the same voltage. Voltage followers in combination with operational amplifiers can be made so precise that their error is small enough to be neglected. Thus, zero and sensitivity errors are eliminated almost completely.

The circuit is suitable for either dc or carrier-frequency operation. With the latter, an even greater precision is achieved because the already small offset and temperature drift of the amplifiers have no effect on the measurements. Another advantage of the circuit shown in **Fig. 7** is that only two leads and two switches are needed for each additional strain gage.

On the other hand, the circuit requires the supplementary resistor R_S to be placed close to the strain gages because voltage drops across the connecting leads between the strain gages and the resistor are not taken into account. Also, the direct connection of the strain gages to each other means that, should one of them ground, the readings from the whole group would be incorrect.

Fig. 8 shows an extended version of the previous circuit that enables single strain gages to be connected as far aways as 1000 m (ca. 3300 ft) and allows the supplementary resistor to be incorporated in the scanning instrument itself instead of in the field [6]. Compared with the circuit in **Fig. 7**, the extended circuit posesses two additional voltage followers VF 1 and VF 4 and a differential amplifier DA.

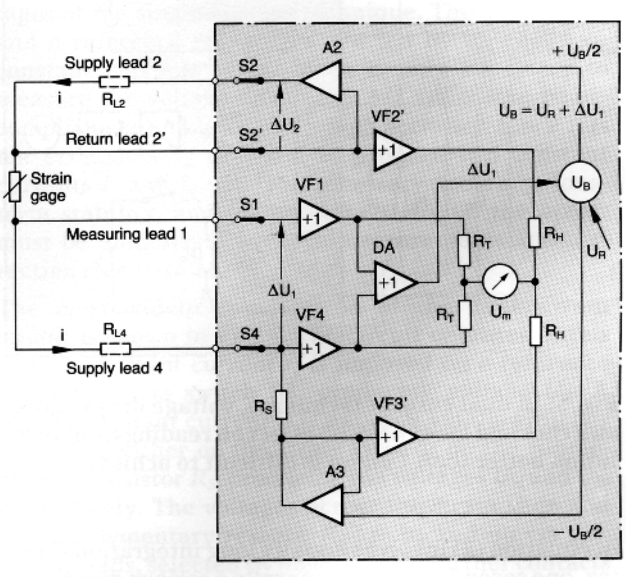


Fig. 8: This extended version of Fig. 7 enables straingage connection as far away as 1000 m (3300 ft) and the supplementary resistor to be incorporated in the scanning instrument itself instead of in the field

Here's how the circuit operates: Voltage drop ΔU_1, caused by the supply current i in supply lead 4 and the switch S4 (i.e. the connection between the strain gage and the supplementary resistor R_S), is picked up from the voltage followers VF 1 and VF 4 and halved exactly by the resistors R_T, which provides equal proportions for the two arms of the external half-bridge. This eliminates the effect of the voltage drop ΔU_1 on the zero.

The voltage difference ΔU, is then fed back to the voltage generator via differential amplifier DA and added to the bridge supply voltage in order to maintain the true voltage across the strain gage and the supplementary resistor at the constant preselected value U_R. Voltage drops ΔU_1 and ΔU_2, therefore, have no adverse effect on the measuring sensitivity. The circuit in **Fig. 8** switches each strain gage sequentially in such a way that the grounding of one strain gage will not affect the others.

Fig. 9 illustrates in greater detail the scanning technique shown in **Fig. 8** for full, half, and quarter bridges.

This circuit allows the connection of individual strain gages with physically separate dummy gages (see **Fig. 9**, measuring point M2). Single strain gages can be extended to a half bridge by means of the internal supplementary resistor $R_{S,int}$ or a common external supplementary resistor $R_{S,ext}$. When a carrier-frequency supply voltage is used, the internal half bridge R_H, and if necessary the resistance divider R_T as well, can be made into an inductive divider. This causes a considerable reduction in the measurement error with half- and quarter-bridges because inductive dividers have substantially greater temperature stability and long-term stability than resistance dividers.

Comparing the various circuits

The principal characteristics of the four circuits previously decribed are listed in the **Table**. For the pur-

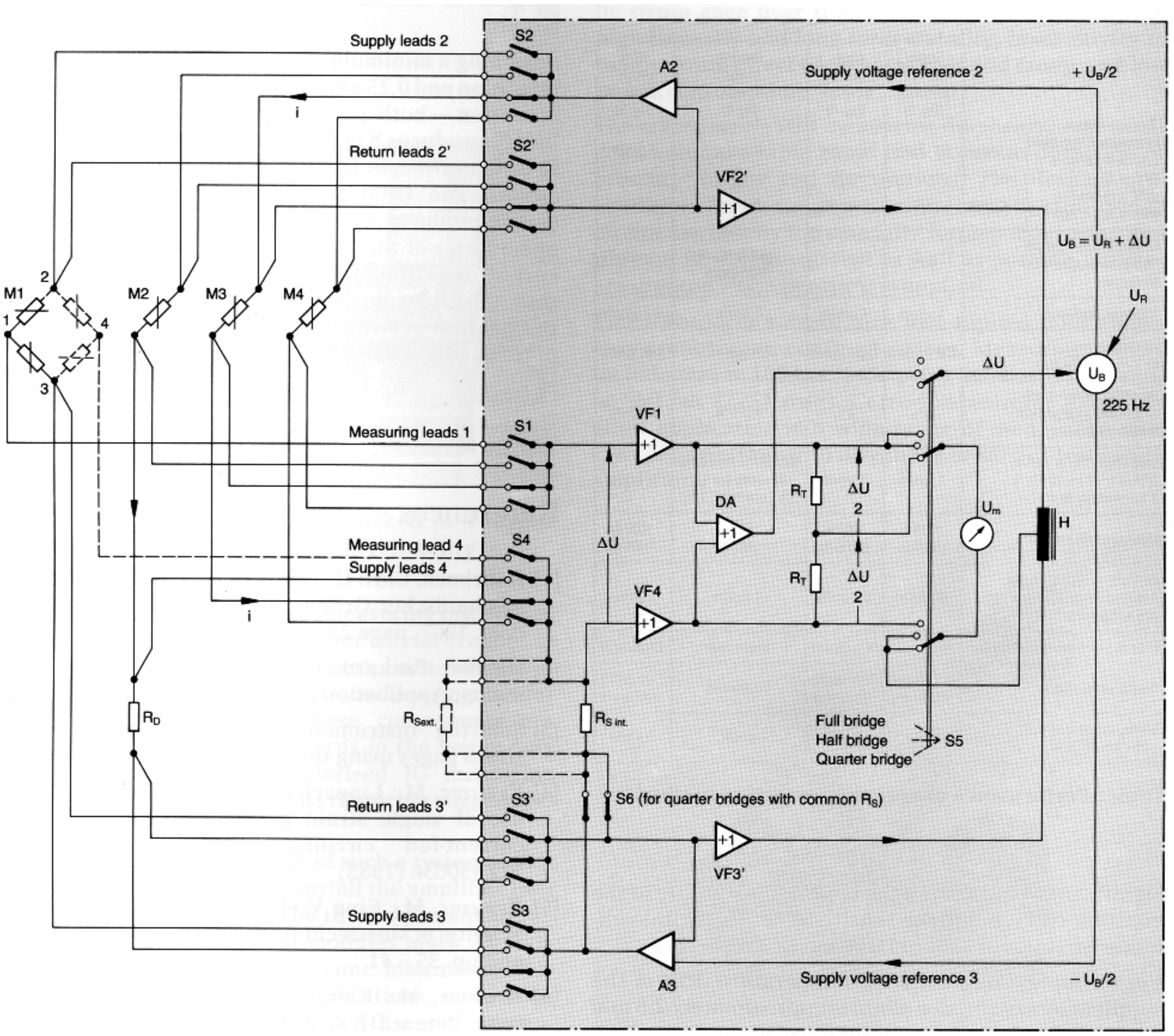
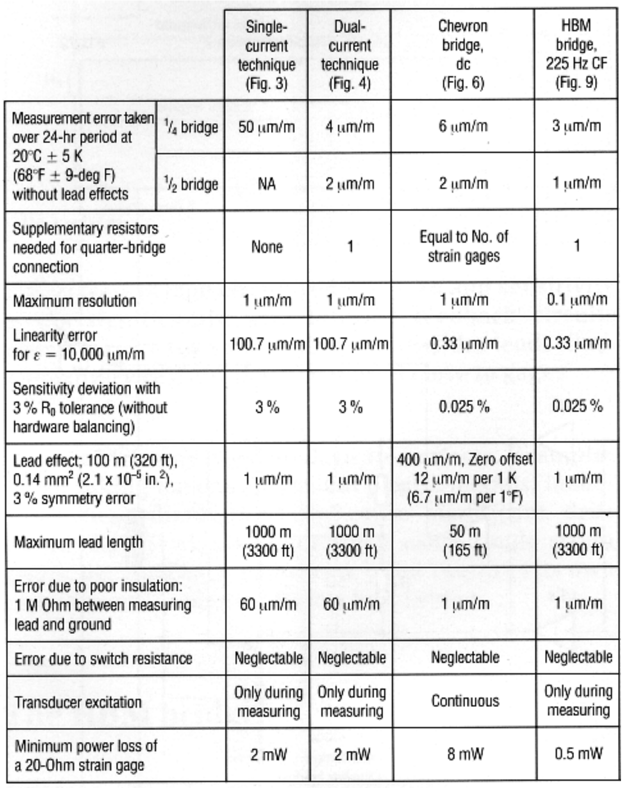


Fig. 9: Detailed bridge circuit from Fig. 8 shows connection technique for full, half, and quarter bridges. This circuit allows the connection of individual strain gages with physically separate dummy gages (see measuring point M2 above). Each gage is switched sequentially in such a way that grounding of one will not affect others

pose of comparison all circuits conform to the following conditions:

- All strain gages have a gage factor of two and are connected in quarter-bridge circuits.
- Temperature drift of the inductive internal halfbridge resistors and supplementary resistors is 10^{-6}/K ($0.6 \cdot 10^{-6}$/°F).
- Temperature drift of the inductive internal halfbridge (for carrier frequency) is 10^{-7}/K ($0.56 \cdot 10^{-7}$/ °F).
- Zero drift of the amplifiers is assumed 0.2 μm/m per 1 K (0.1 μm/m per 1°F) for dc and 0.05 μm/m per 1 K (0.03 μm/m per 1°F) for carrier frequency, with zero correction by automatic calibration in bath cases.

The first line of the table shows the anticipated error of the various methods with quarter- and half-bridge circuits. With the single-current technique, which is only suitable for quarter-bridge circuits, the errors are significant because even the slightest instability of the constant current and the sensitivity of the digital-voltmeter circuit will lead to large measuring errors.



		Single-current technique (Fig. 3)	Dual-current technique (Fig. 4)	Chevron bridge, dc (Fig. 6)	HBM bridge, 225 Hz CF (Fig. 9)
Measurement error taken over 24-hr period at 20°C ± 5 K (68°F ± 9-deg F) without lead effects	¼ bridge	50 μm/m	4 μm/m	6 μm/m	3 μm/m
	½ bridge	NA	2 μm/m	2 μm/m	1 μm/m
Supplementary resistors needed for quarter-bridge connection		None	1	Equal to No. of strain gages	1
Maximum resolution		1 μm/m	1 μm/m	1 μm/m	0.1 μm/m
Linearity error for $\varepsilon = 10{,}000$ μm/m		100.7 μm/m	100.7 μm/m	0.33 μm/m	0.33 μm/m
Sensitivity deviation with 3 % R_0 tolerance (without hardware balancing)		3 %	3 %	0.025 %	0.025 %
Lead effect; 100 m (320 ft), 0.14 mm² (2.1 x 10^{-5} in.²), 3 % symmetry error		1 μm/m	1 μm/m	400 μm/m, Zero offset 12 μm/m per 1 K (6.7 μm/m per 1°F)	1 μm/m
Maximum lead length		1000 m (3300 ft)	1000 m (3300 ft)	50 m (165 ft)	1000 m (3300 ft)
Error due to poor insulation: 1 M Ohm between measuring lead and ground		60 μm/m	60 μm/m	1 μm/m	1 μm/m
Error due to switch resistance		Neglectable	Neglectable	Neglectable	Neglectable
Transducer excitation		Only during measuring	Only during measuring	Continuous	Only during measuring
Minimum power loss of a 20-Ohm strain gage		2 mW	2 mW	8 mW	0.5 mW

Table: Performance of various connection methods

For example, the error from temperature drift of the supplementary resistor alone is approximately 2.5 μm/ m. This is practically the total error in the case of the HBM bridge. Reason: By using the carrier-frequency technique, amplifier drift can be neglected, as well as the drift of the inductive internal half bridge when one is used. As can be seen from data on the Chevron bridge, the dc technique and resistors of the internal half bridge contribute to an error twice as large as with the HBM bridge [7].

While the dual-current method possesses good stability, its dc circuit provides somewhat poorer performance than the HBM bridge, which is the only one with signal resolution of 0.1 μm/m. Both voltage-fed circuits have linearity and sensitivity errors several orders of magnitude smaller than those of the current-fed circuits. Only the Chevron bridge is seriously affected by errors originating from the measuring leads; current-fed circuits, on the other hand, are particularly affected by low insulation resistance on the measuring leads.

In the Chevron-bridge circuit, unlike the other circuits, measuring points are continuously connected to the supply voltage. This can cause overheating when taking measurements on plastic models. With such heating, the power loss in the strain gages must also be taken into account. For the same value of measured voltage at the strain gage, a current-fed circuit requires only about half the supply voltage of a voltage-fed circuit.

Assuming a minimum signal of 1 μV per 1 μm/m for dc operation and 0.25 μV per 1 μm/m for carrier-frequency operation —both quite practical values— the Chevron bridge produces 8 mW of power loss at the strain gage in dc operation, or four times as much as a current-fed circuit. The HBM bridge with a carrier-frequency supply produces a power loss of 0.5 mW, which is only one-quarter of the loss, and also the heating effect, of current-fed circuits. The figure of 0.5 mW is also so low that no serious measurement errors arise even when the smallest of strain gages are left on for long periods of time.

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