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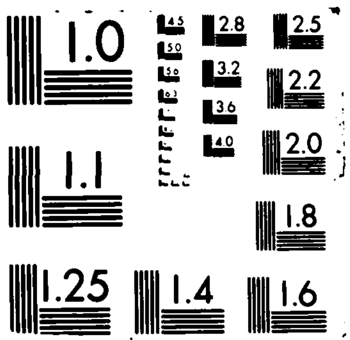
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AERODYNAMICS REPORT 170

AN IMPROVED STRAIN GAUGE TRANSDUCER AMPLIFIER FOR WIND TUNNEL USE

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N. POLLOCK

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SUMMARY

An AC excited self-balancing instrument for measuring low level strain gauge transducer outputs is described. The design of a specific circuit optimised for use with small wind tunnel force balances is presented along with sufficient information to facilitate circuit optimisation for other applications.

The instrument is intended for applications where the bridge output voltage is below the level that can be conveniently handled by conventional DC amplifier systems. Tests on a prototype instrument showed an input noise in a 0·1 Hz to 0·7 Hz bandwidth of 4 nV RMS, a zero drift of less than 0·1 μV per day and negligible sensitivity drifts.

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1. INTRODUCTION

Force transducers employing electrical resistance strain gauges bonded to metal structures have been widely used for many years. For high accuracy applications it is virtually universal practice to use metal foil gauges in a full bridge configuration.

Under carefully controlled conditions force transducers of this type have achieved a precision of $0 \cdot 5 \times 10^{-6}$ in mass comparisons [Ref. 1]. They are also widely used as secondary force standards and in this application precision levels rivaling that of the primary force standard (2×10^{-5}) are achieved [Ref. 2].

In wind tunnels, multi-component strain gauge balances are widely used to measure loads on models and parts of models. Precision levels between 10^{-3} and 10^{-4} are required and achieved with varying degrees of difficulty [Ref. 3]. The accuracy of wind tunnel force measurements is limited primarily by the geometrical constraints on the force transducers and secondarily by the non-ideal operating environment of the transducers and output processing electronics. Many different types of readout electronics have been used in wind tunnels over the years [Ref. 3–6]. Although never employed in a tunnel, the ingenious approach of Ref. 7 would be very attractive in applications where it was required to compare closely equal forces with very high precision.

At the present time most wind tunnels use regulated *DC* excitation of the strain gauge bridges on balances and measure the outputs with the aid of high input impedance *DC* amplifiers [Ref. 8]. This arrangement is simple and has adequate performance for the majority of applications. However in circumstances where the balance electrical output is unusually small, the limitations of *DC* amplifier performance can severely restrict the precision of measurement.

In this report a simple *AC* excited self-balancing electronic system for use with strain gauge transducers is described. A specific embodiment of this system, optimised for use with low output wind tunnel balances, is described in detail, but general design data to permit optimisation for other applications is included.

2. STATEMENT OF PROBLEM

Six component balances with a diameter greater than about 30 mm, suitable for mounting inside a wind tunnel model, would typically use strain gauges with a $3 \cdot 2$ mm square grid and a resistance of 350 Ω or higher. Reference 9 suggests that for high accuracy static strain measurements where the balance structure provides a good heatsink, the power dissipated in the gauge should not exceed $0 \cdot 0031$ W/mm^2. For a $3 \cdot 2$ mm square, 350 Ω gauge this infers a maximum gauge supply of $3 \cdot 3$ V or a full bridge supply of $6 \cdot 6$ V. Balances of the above type are usually designed with output strain levels of about 1000 μm/m. The maximum bridge output is given by the product of the strain, the gauge factor and the bridge supply voltage. For the commonly used constantan gauges the gauge factor is approximately $2 \cdot 0$ and the maximum bridge ouput about 13 mV. To obtain a measurement precision of 10^{-3} the output electronics will have to reliably resolve 13 μV. This is within the capabilities of *DC* systems of the type used in wind tunnels [Ref. 3]. If a precision of 10^{-4} was required an output resolution of $1 \cdot 3$ μV would be required and although this could be achieved easily by *DC* systems under laboratory conditions it is doubtful if it could be maintained in a working tunnel environment. Gauges with grid resistances of 1 kΩ and to a lesser extent 5 kΩ are available. The use of these gauges would permit an increase in the supply voltage by factors of $1 \cdot 7$ and $3 \cdot 8$ for 1 kΩ and 5 kΩ gauges respectively, with a direct effect on the output resolution problem.

Strain gauge balances mounted external to the tunnel test section are not subject to rigid size contraints and can utilise much larger gauge areas with resulting higher excitation voltage levels. One commercial load cell manufacturer products single component force balances with a recommended maximum excitation of 100 V.

Very small strain gauge balances for use in model stores (bombs, missiles etc.) and balances for measuring control surface loads pose a number of special problems. Due to their small size they require gauges of small dimensions and a grid size of 0·8 mm square is quite common and 0·4 mm square grids have been used. In these small grid sizes it is difficult to manufacture high resistance gauges and a resistance of 120 Ω is most common. Many small balances are required to resolve small forces and therefore tend to be very fragile. To facilitate manufacture and handling it is often necessary to reduce the design output strain from around 1000 μm/m to around 200 μm/m. The current state of the art in small six component store balances is about 8·0 mm to 10·0 mm diameter with a normal force capacity of 100 N and an axial force capacity of 10 N. Considering a typical small balance with 0·8 mm square, 120 Ω gauges and with the maximum power dissipation of 0·0031 W/mm^2 used previously, a permissible bridge excitation of about 1·0 V is obtained. For a maximum strain of 200 μm/m the output will be 400 μV. To obtain precisions of 10^{-3} and 10^{-4}, output voltages of 400 nV and 40 nV respectively would have to be measured. These voltages are beyond the capability of any current *DC* amplifier system suitable for use in a wind tunnel environment.

To summarise: for strain gauge balances mounted external to the tunnel test section and for the larger (≥30 mm diameter) balances mounted inside models, measuring systems based on *DC* bridge excitation and *DC* amplifiers provide adequate performance with simplicity and relatively low cost. For small balances of the type used for store and control surface load measurement the output voltage available is often reduced to the extent that a *DC* amplifier system limits the achievable measurement accuracy.

3. DESCRIPTION OF CONCEPT

The concept of a new type of high performance *AC* excited strain gauge balance readout system was described in Reference 6. This concept forms the basis of the improved system described here. For completeness the principle of operation will be briefly outlined.

A block diagram of the system is shown in Figure 1. The bridge is supplied from a sinusoidal *AC* source. In general the bridge output will be sinusoidal with a frequency equal to the excitation supply and a component in phase with the supply with an amplitude proportional to the bridge resistive unbalance and a component in quadrature (90° phase shift) with the supply with an amplitude proportional to the bridge reactive unbalance. The bridge output, amplified by A_1, and the bridge supply, amplified by A_2, are applied to the phase sensitive demodulator D_1. The output of this demodulator is a *DC* level proportional to the amplitude of the component of the bridge output in-phase with the bridge supply and second and higher harmonics of the supply. The demodulator output is applied to integrator I_1 which attenuates the *AC* components (providing the integrator time constant is sufficienty large) and produces a change of output level with a slew rate proportional to the *DC* input. The integrator output is multiplied by the bridge supply reference signal in M_1. The output voltage from M_1 forces a current through the rebalancing resistor into the bridge. It should be noted that the bridge supply must be referenced to the same ground as M_1 to provide a return path for the rebalancing current. The sign of the rebalancing current is arranged such that it opposes the in-phase component of the bridge output. A simple analysis[6] shows that the rebalancing loop is stable and that the final approach to balance is exponential with a time constant of:

$$T = \frac{4T_1R_r(R + R_{in})}{A^2G_DG_MR\,R_{in}} \tag{1}$$

where

T_1 = Integrator time constant

R_r = Rebalancing resistance

R = Strain gauge resistance

R_{in} = Error amplifier input resistance

A = Supply amplitude

D_D = Gain of demodulator

G_M = Gain of multiplier

The loop formed by D_2, I_3 and M_2 is identical to that formed by D_1, I_1 and M_1 except that it is supplied by a quadrature reference signal generated by integrator I_2. The effect of this loop is to null the bridge reactive unbalance.

The gain of the system will be derived for a four active arm bridge with initially equal arm resistances and equal resistance changes. Similar results may be obtained for other bridge configurations. From Figure 2 it can be seen that:

$$V_{ab} = E(R + \delta R)/2R \tag{2}$$

$$V_{ad} = I_2(R - \delta R) \tag{3}$$

$$I_r = I_2 - I_1 \tag{4}$$

$$E = I_2(R - \delta R) + I_1(R + \delta R) \tag{5}$$

$$V_r = G_1\, E\, V_{out} \tag{6}$$

From (4) and (5)

$$I_2 = [E + I_r(R + \delta R)]/2R \tag{7}$$

From (3) and (7)

$$V_{ad} = (ER + I_rR^2 - E\delta R - I_r\delta R^2)/2R \tag{8}$$

For balanced conditions $V_{bd} = 0 \therefore V_{ab} = V_{ad}$

Equating (2) an (8) and simplifying:

$$I_r = 2E\delta R/(R^2 - \delta R^2) \tag{9}$$

Assuming the bridge supply is symmetrical about earth potential, V_d will be zero for a balanced ($\delta R = 0$) bridge. When the bridge is disturbed and rebalanced by the injection of current I_r:

$$V_d = E\delta R/2R \tag{10}$$

$$I_r = (V_r - V_d)/R_r \tag{11}$$

From (6), (9), (10) and (11)

$$V_{out} = \frac{1}{G_1}\,\frac{4RR_r\delta R + R^2\delta R - \delta R^3}{2R^3 - 2R\delta R^2} \tag{12}$$

If $\delta R \ll R$ equation (12) can be written:

$$V_{out} = \frac{1}{G_1}\left(\frac{2R_r}{R} + \frac{1}{2}\right)\frac{\delta R}{R} \tag{13}$$

For strain levels and resistance values typical of wind tunnel balance applications the difference between equations (12) and (13) will be less than 0·001 %. Even if the non-linearity inherent in equation (12) was significant, this would not pose a problem since virtually all data acquisition and processing systems in current use have adequate computing power to apply non linear calibrations.

The major advantage of this system is that the performance is governed entirely by high level parts of the circuit. The low level error amplifier does not have any critical performance parameters. It can be seen from equation (13) that the system gain depends only on the rebalancing multiplier gain and resistances R and R_r. The system zero stability depends on the DC output drift of demodulator D_1, the drifts of integrator I_1 and the offset drifts of one of the inputs of multiplier M_1 (Fig. 1). The performance of the quadrature rebalancing loop formed by I_2, D_2, I_3 and M_3 is not critical. This loop is included to ensure a consistent state of quadrature balance since experience has shown a small interaction between quadrature and in-phase balance exists, presumably due to small phase shifts in connecting wiring etc. The quadrature loop also removes the necessity for manual preset quadrature balance adjustments.

It should be noted that this system could be used with capacitive or inductive bridges instead of resistive bridges simply by taking the output from I_3 (Fig. 1). A device based on a similar concept was proposed independently by the Royal Aircraft Establishment [Ref. 10], but this does not appear to have been widely used.

4. DESIGN OPTIMISATION

4.1 Specification

To meet the requirements of measuring the outputs from small force balances in wind tunnels the following desirable performance characteristics were identified:

Bridge excitation supply	— 1·0V RMS maximum
Full scale output	— ± 10V
Full scale input strain (4 active arm bridge)	— ± 200 μm/m
Resolution and accuracy over 1 hour period with 0·7 Hz two pole output filter	— 10^{-4} of full scale
Maximum input frequency	— 50 Hz.

Desirable attributes are simplicity, low cost and the absence of initial set-up adjustments.

4.2 Bridge excitation frequency

From considerations of information content it is necessary that the excitation frequency is at least double the maximum input frequency that is to be followed. An excitation frequency of ten times the maximum input frequency is a more practical minimum for a system of this type [Ref. 11]. Since second and higher harmonics of the excitation frequency exist on the output of integrator I_1, increasing the frequency eases the problem of arranging the output filter to adequately attenuate the excitation harmonics without restricting the system response.

The choice of excitation frequency has a direct impact on the signal to noise ratio achievable from the input amplifier. All amplifiers, and indeed most physical systems, have a noise level which increases at low frequencies due to so called $1/f$ or flicker noise. Noise voltage curves for some typical integrated circuit amplifiers are presented in Figure 3. It is obviously desirable that the signal to be amplified is in the high frequency, constant noise voltage region. Input current noise has a very similar characteristic.

Excessively high excitation frequencies will result in large phase shifts in the interconnecting wiring leading to the possibility of changes in quadrature balance producing significant outputs from the in-phase loop.

For the current application a frequency of 1 kHz was selected.

4.3 Error Amplifier

The requirements of the input amplifier are:

(*a*) Minimum noise in a bandwidth equal to double the maximum response frequency, centred on the excitation frequency i.e. 0·95 kHz to 1·05 kHz for a 50 Hz frequency range and a 1 kHz excitation frequency.

(*b*) Sufficient gain to raise the desired minimum resolvable bridge output voltage to a level greater than the noise level of the demodulator and integrator. Excessive gain makes the amplifier liable to overloading with excitation harmonics which cannot be nulled.

(*c*) Sufficiently narrow bandwidth to prevent overloading from electrical noise (e.g. mains supply frequency) and *DC* drifts. Excessive bandwidth reduction is neither necessary nor desirable since it introduces a lag in the system response which destabilises the rebalancing loop. A narrow bandwidth would also introduce undesirable phase shifts with small drifts in the excitation frequency.

(*d*) Reasonably low *DC* offset at output to minimise fundamental excitation frequency output from synchronous demodulator.

(*e*) Low phase shift at excitation frequency.

For a low source impedance like a strain gauge bridge the amplifier voltage noise will dominate over the current noise. In these circumstances a transformer before the first active gain stage would have obvious noise advantages. However since the voltage noise of readily available integrated circuit audio amplifiers ($\simeq 5\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$ at 1 kHz) is of the same order as the thermal noise of the bridge ($2{\cdot}4\ \mathrm{nV}/\sqrt{\mathrm{Hz}}$ for a 350 Ω bridge at 300 K) the extra complexity is not considered worthwhile in this application.

The final amplifier design is shown in Figure 4. The input stage uses a parallel pair of low cost audio amplifiers with their output summed to give a $\sqrt{2}$ gain in signal to noise ratio. The overall voltage gain is $4{\cdot}3 \times 10^4$ and the -3 db bandwidth about 200 Hz to 15 kHz. The passive filters formed by R and C at the amplifier input were found to greatly improve the rejection of sharp spikes caused by switching nearby electrical equipment. The unity gain output amplifier was used to minimise the *DC* offset. The shorted input, output noise of the amplifier was approximately 100 mV peak to peak.

To meet the specification of Section 4·1 the input voltage resolution must be better than 40 nV RMS which is equivalent to an output voltage of around 2 mV RMS. This figure is important in determining the performance requirements of following stages of the circuit. The fact that the desired resolvable signal level is deeply buried in the amplifier noise is not a problem since it is readily recovered by the demodulator and integrator.

4.4 Reference Amplifiers

The bridge supply must be amplified to a level suited to the remainder of the circuit and referenced to the system ground. From considerations of the multiplier specifications a voltage swing of about $\pm$ 10 V peak to peak was chosen. The circuit adopted is shown in Figure 5. The quadrature reference was generated by an integrator with $R_1C_1 = 1/(2 \times \pi \times$ Excitation frequency). The resistors R_2 and R_3, which should be as large as practical, are required for *DC* stability. The two *AC* coupled unity gain amplifiers on the in-phase and quadrature reference outputs were used to minimise the *DC* offsets. This is not a critical consideration since the only detrimental effect of offsets would be to non-productively use some of the available multiplier voltage swing and inject a small *DC* component back into the bridge.

Any departure from zero and 90 degree phase shift between the bridge excitation and the in-phase and quadrature reference signals respectively will result in an interaction between the in-phase and quadrature rebalancing loops. For the proposed application the reactive unbalance of the bridge and connecting wiring will be very nearly constant and interactions between loops will have no effect on the system accuracy. For this reason no extreme measures to minimise phase shifts were used. In an application where the resistive and reactive bridge unbalances varied significantly in an uncoupled manner (the author is unable to envisage any such situation of practical importance) care in minimising extraneous phase shifts in all parts of the circuit would be necessary.

4.5 Demodulator

The demodulator works primarily as a phase sensitive null detector. As such the only significant performance parameter is *DC* output drift. Linearity, gain stability and noise are unimportant. In these circumstances a simple switching demodulator has significant advantages. The LM 311 comparator provides the drive signal for the AD 7512 CMOS switch array which cyclically connects the error amplifier to the inverting and non inverting inputs of the LM 308 unity gain differential amplifier.

The output offset drift of this arrangement is simply the sum of the amplifier offset voltage drift, the amplifier input offset current drift flowing in the input resistors and noise rectification due to variations of the "on" resistance drifts of the individual CMOS switches. The worst case amplifier drift specifications lead to an output drift of 15 μV/°C. The contribution of the switches can be estimated from the following considerations. The "on" resistance of the switches is about 100 Ω and this appears effectively in series with the 43 kΩ input resistors. If the resistance of the switches all change together (as they do with temperature changes) the demodulator gain will change but no output drifts in its null sensing mode will result. The AD 7512 specifications indicate that the drift in the match of the "on" resistances of the switches on one chip is less than 0·01 %/°C. The total amplifier input resistance (43 kΩ plus switch resistance) match can therefore drift by 0·000023 %/°C. This will be reflected directly in a drift in the match of the inverting and non-invertin gain.

Considering the highly unlikely worst case of 1 V *DC* at the error amplifier output, this drift will produce an output change from the demodulator of 230 nV. The temperature drifts in the output are therefore dominated by the amplifier and have a magnitude ($\simeq$ 15 μV/°C) which is negligible compared with the desired resolution limit error amplifier output ($\simeq$ 2 mV).

4.6 Integrator

The integrator has the function of slewing the rebalancing voltage to null the bridge unbalance. It has the secondary function of attenuating the *AC* component of the demodulator output. If this attenuation is inadequate the resulting harmonics on the rebalancing current will produce components on the bridge output which the system can not null and may, for small rebalancing resistor values, overload the error amplifier.

From the specification of Section 4·1 the maximum output frequency is 50 Hz. For a $\pm$ 10 Volt output swing the maximum output voltage rate of change is 3140 V/sec. To allow some overload margin it would be desirable if this could be achieved without exceeding an error amplifier output of 8 V RMS. Since the demodulator has unity gain this is equivalent to an input to the integrator with a *DC* level of 8 Volt. Since for an integrator $RC = V_{in}/dV_{out}/dt$, the *RC* used in this case should not exceed 8/3140 = 0·0025. Values of *R* and *C* of 10 kΩ and 0·22 μF respectively were chosen, giving an *RC* of 0·0022. For attenuation of supply harmonics *RC* should be much larger than 1/(2 · π · Excitation frequency) which for this case is equal to 0·00016.

The only integrator performance parameter which affects the system accuracy is the input offset drift which for the arrangement shown in Figure 7 has a worst case value of 15 μV/°C. This should not have a significant effect on the system performance.

4.7 Rebalancing Multiplier

The rebalancing multiplier is the component with the major impact on the system performance. The system gain is directly proportional to the multiplier gain (Equation 13, Section 3) and the system linearity is dominated by the multiplier integrator input (i.e. the multiplier input that comes from the integrator) linearity. Any *DC* offset drifts of the integrator input will appear directly as an instrument zero drift.

Non linearities of the reference input will produce harmonic distortion of the rebalancing signal which will have no impact on the system performance provided it is not so large as to overload the error amplifier. Similarly offset drifts of the reference input will have no significant effect.

The multiplier selected was the AD 534 (Fig. 7) with the selected "L" suffix type in the in-phase loop and the lower specification, and lower cost, "*J*" type in the quadrature loop. One input of these multipliers has an order of magnitude better linearity than the other. The high quality input was used for the integrator input. The multiplier gain was 0·1 i.e. $V_{out} = V_x$ in $\times$ V_y in $\times$ 0·1. Brief specifications of the "L" type multiplier are: gain drift $\pm 0{\cdot}005\%/{}^\circ C$, X input linearily $\pm 0{\cdot}1\%$, Y input linearity $\pm 0{\cdot}005\%$, input offset voltage drift $50\,\mu V/{}^\circ C$, small signal band width 1·0 MHz. All of these specifications are consistent with achieving the design performance without the need for special environmental conditions. The input offset drift would appear to be the dominant source of zero drift in the system.

4.8 Excitation Supply

The bridge excitation supply has no critical performance parameters but, since they can be relatively simply achieved, the following specifications were adopted: frequency 1000 Hz $\pm$ 50 Hz, amplitude 1·0 V. RMS $\pm$ 100 mV and harmonic distortion $< 1\%$. The major functional requirements are that the supply should be symmetrical around the system ground and have a low resistance to ground. A supply suitable for up to 12, 120 Ω bridges is shown in Figure 8. In any single experimental facility all channels of this equipment should use the same excitation supply, or if multiple suppliers are needed, they should be synchronised to avoid low frequency beating.

4.9 Offset Nulling Network

A satisfactory offset nulling arrangement is shown in Figure 9. For 120 Ω gauges it has sufficient authority to balance a bridge constructed from unselected, $\pm 0{\cdot}5\%$ initial resistance, gauges. An offset change equivalent to a $1000\,\mu m/m$ strain level results in a sensitivity change of less than $0{\cdot}01\%$ with no other effects on the system calibration [Ref. 12]. If different resistor arrangements or higher value potentiometers are used, unacceptable changes in calibration with changes in zero offset can result.

4.10 Bridge Connections

The optimum arrangement for connecting a bridge to the equipment is shown in Figure 9. The effects of variations in the resistance of the seven connecting leads is a follows:

(*a*) Error amplifier (6 and 7), no effect.

(*b*) Excitation supply (1 and 3), no effect.

(*c*) Rebalancing line (5), sensitivity varies inversely with sum of rebalancing resistance and resistance of wire 5.

(*d*) Reference amplifier (6 and 7), sensitivity varies inversely with sum of reference amplifier input resistance and resistance of wires 6 and 7.

The effect of the resistance of the rebalancing line could be removed entirely if a precision current pump was substituted for the rebalancing resistor. The effect of the reference amplifier leads could be reduced by increasing the amplifier input impedance.

A full analysis of the effect of lead resistance if a basic four wire connection is used, is far from simple. However to a first order approximation the sensitivity will be inversely proportional to the sum of the gauge resistance and three times an individual lead resistance. This is more than two orders of magnitude more sensitive to lead resistance than the seven wire configuration.

4.11 Output Filter

A significant ripple will exist on the integrator output and a suitable low pass filter will be required prior to analogue to digital conversion. A multi-pole filter will probably be required and care should be taken with the design to ensure that it does not degrade the system performance.

5. PERFORMANCE MEASUREMENTS

Twenty channels of the equipment described here have been constructed for use in the ARL low speed and transonic wind tunnels (Fig. 10). Tests have shown very little variation in performance and the following figures are based on extensive tests on one channel (the first completed) with random checks of the remaining devices.

A quadrature rebalancing resistance of 680 kΩ was used for all tests. This gave the quadrature loop sufficient authority to balance the reactive unbalance caused by paralleling one 120 Ω bridge arm with a 3200 pF capacitor. Unless otherwise stated, a 430 kΩ in-phase rebalancing resistor was used giving an input $\Delta R/R$ of ± 0·0006 (± 300 μm/m strain for a gauge factor = 2) for a full ± 10 Volt output swing. The design maximum input strain was ± 200 μm/m but the lower sensitivity was selected to allow some headroom for unsteady inputs before the output filter. An overload detector was also provided prior to the output filter.

For performance measurements a specially constructed very stable bridge of 120 Ω gauges bonded onto a large steel block was used. Unbalance was produced with a precision resistor network [Ref. 6]. To provide data on the required specifications for the ± 15 Volt power supplies and the excitation frequency and amplitude, the effect of varying these parameters was investigated. The results presented below are expressed in terms of bridge output volts assuming a non-nulling system was used. This form of specification was selected to facilitate comparisons with the more usual *DC* amplifiers. For reference the full scale (300 μm/m strain) bridge output with 1 volt excitation would be 600 μV.

Measured performance

Linearity	better than 0·01%
Zero drift	
(*a*) with time	0·1 μV in 24 hours
(*b*) with temperature	6 nV/°C
(*c*) with ± 15 V supplies (± 14 V to ± V16)	not measurable
(*d*) with excitation frequency (900 Hz to 1·1 kHz)	1 nV/Hz
(*e*) with excitation voltage (0·5 V RMS to 1·0 V RMS)	not measurable
Sensitivity drift	
(*a*) with time	not measurable over 1 month
(*b*) with temperature	0·003%/°C
(*c*) with ± 15 V supplies (± 14 V to ± 16 V)	0·02%/Volt

(*d*) with excitation frequency (900 Hz to 1·1 kHz)	not measurable
(*e*) with excitation voltage (0·5 V RMS to 1 V RMS)	not measurable

Noise

(*a*) before output filter	0·2 μV RMS predominantly 2 kHz ripple.
(*b*) after 0·7 Hz 2 pole filter	4 nV RMS (See Fig. 11)

It is believed that the zero drift with time recorded above was dominated by drifts in the reference bridge rather than in the readout equipment. This conclusion was reached from tests where the error amplifier gain was switched between 5×10^4 and 5×10^5. Over a period of hours the difference in zero readings for the two gain settings drifted by less than 5% of the amount that the two zeros drifted. This behaviour is consistent with a real input change rather than an instrument drift.

As predicted the system noise level appears to be dominated by the input amplifier and there is good correlation between the amplifier specifications and the measured system performance.

6. LIMITATIONS AND PRECAUTIONS

For a bridge with unequal initial arm resistances the present system should always be connected to the bridge in exactly the same way. With the familiar *DC* amplifier system, the amplifier connections can be reversed with a change in sign, but no change in magnitude, of the sensitivity. With the present system such a connection reversal would produce a small but significant sensitivity change. This is due to the basically asymmetric nature of the device with the rebalancing signal being injected into one corner of the bridge.

Many load cells incorporate a small positive temperature coefficient resistance in one of the bridge supply wires to compensate for changes in material modulus, and hence cell calibration, with temperature [Ref. 13]. With the present system this type of modulus compensation will introduce a false zero shift with temperature and should therefore not be used. The problem can be overcome by splitting the required modulus compensation resistance equally between the two bridge supply wires.

The system described here is inherently best suited to low output transducers with high ratios of rebalancing resistance to bridge resistance. For low sensitivity settings the system linearity is degraded and the injection of non-nullable harmonics into the bridge becomes a problem. For high output transducers ($\Delta R/R > 0{\cdot}005$) a high input impedance amplifier would probably be more appropriate.

7. CONCLUSION

The prototype instrument easily exceeded its design specification of an accuracy of 10^{-4} for measuring the output of a 120 Ω bridge of foil gauges at a strain level of 200 μm/m with a 1·0 V excitation. There are indications that by altering the error and reference amplifier gains the 10^{-4} accuracy could be maintained with 0·1 V bridge excitation. This would allow precision strain gauge force measurements to be made with 0·4 mm gauges under very poor heatsink conditions.

The performance of the instrument described is limited primarily by the noise of the input error amplifier. The potential performance improvement available with improved amplifier designs is around a factor of 5. At that stage the thermal noise of the bridge resistors produces a fundamental performance limitation.

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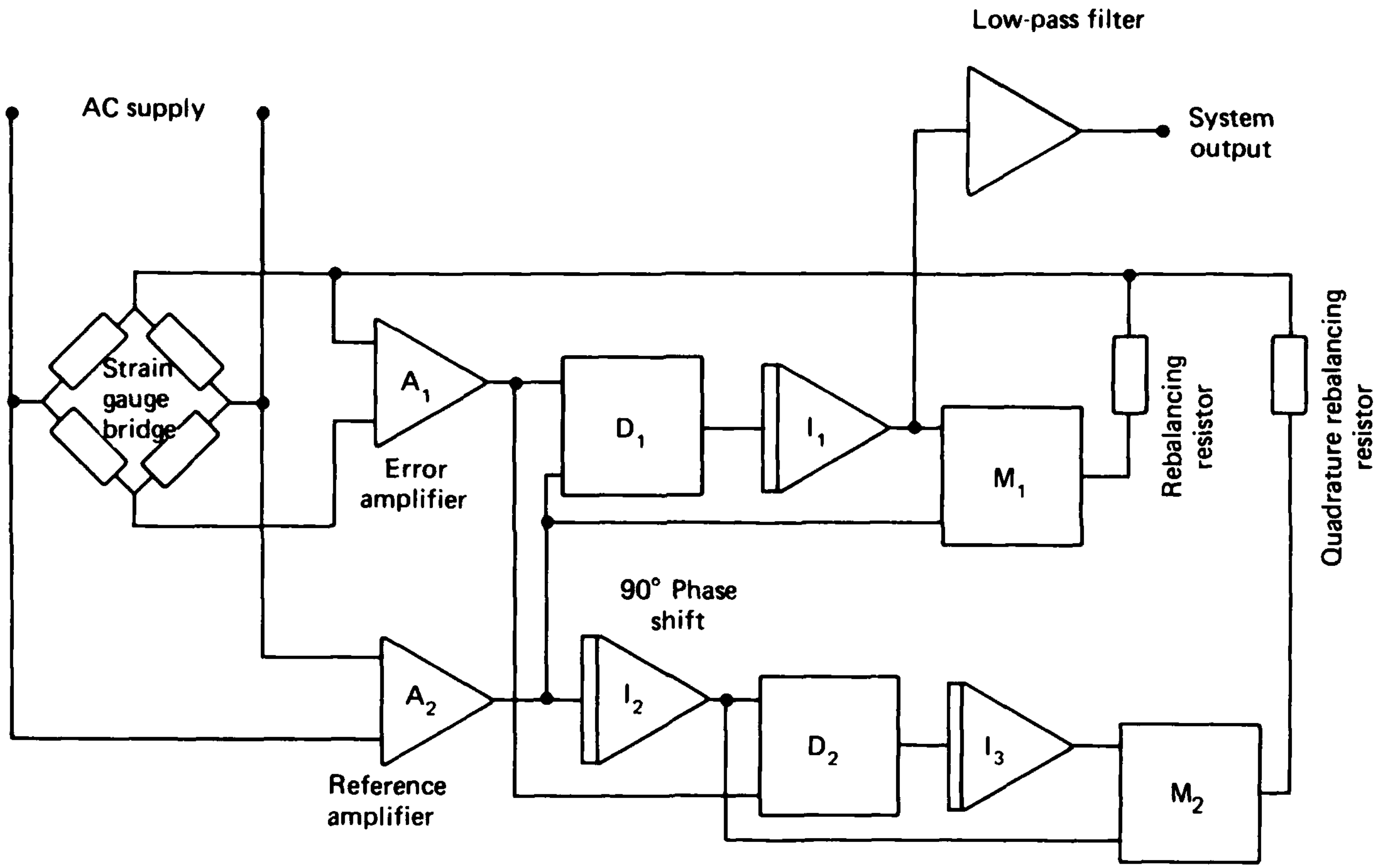


D = Phase sensitive demodulator
I = Integrator
M = Multiplier

FIG. 1 BLOCK DIAGRAM OF SYSTEM

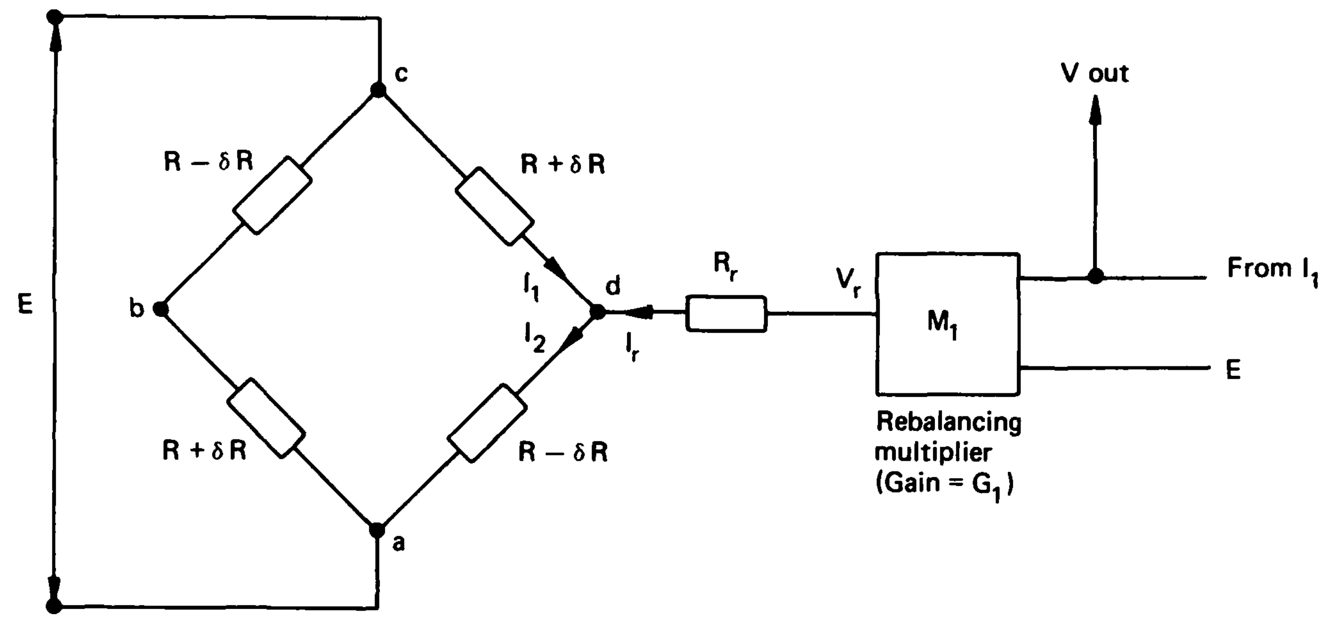


FIG. 2 REBALANCING PROCESS

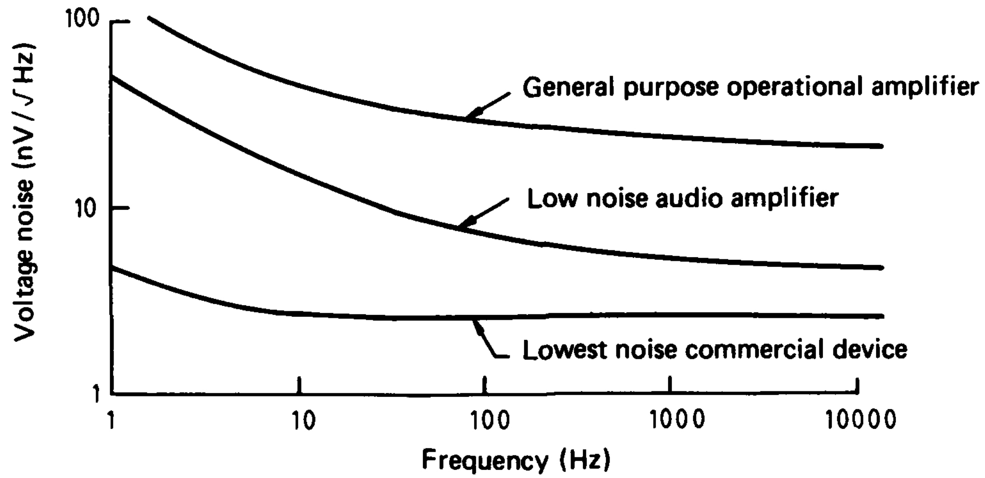


FIG. 3 NOISE VOLTAGE SPECTA OF INTEGRATED CIRCUIT AMPLIFIERS

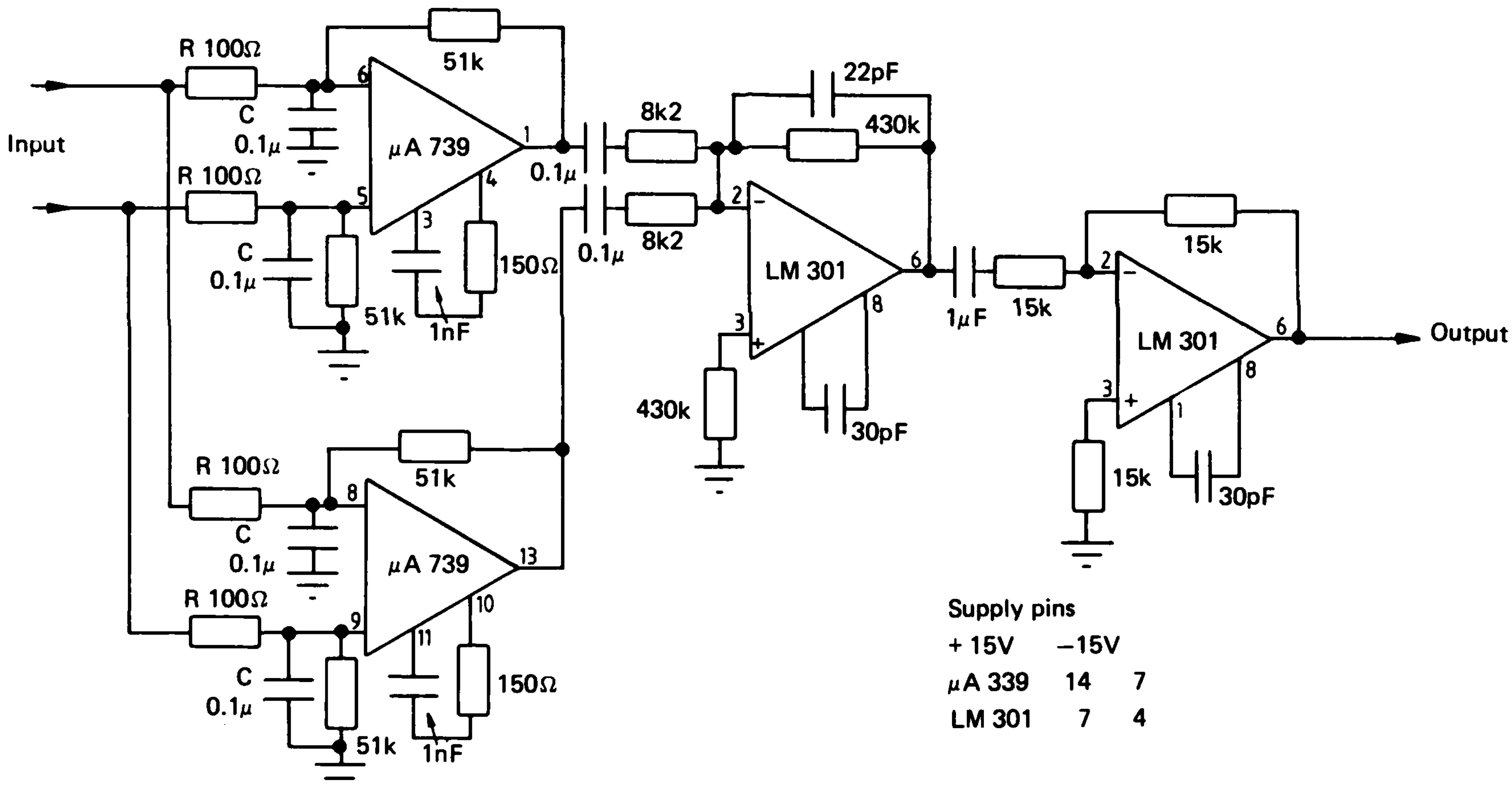


FIG. 4 INPUT AMPLIFIER CIRCUIT

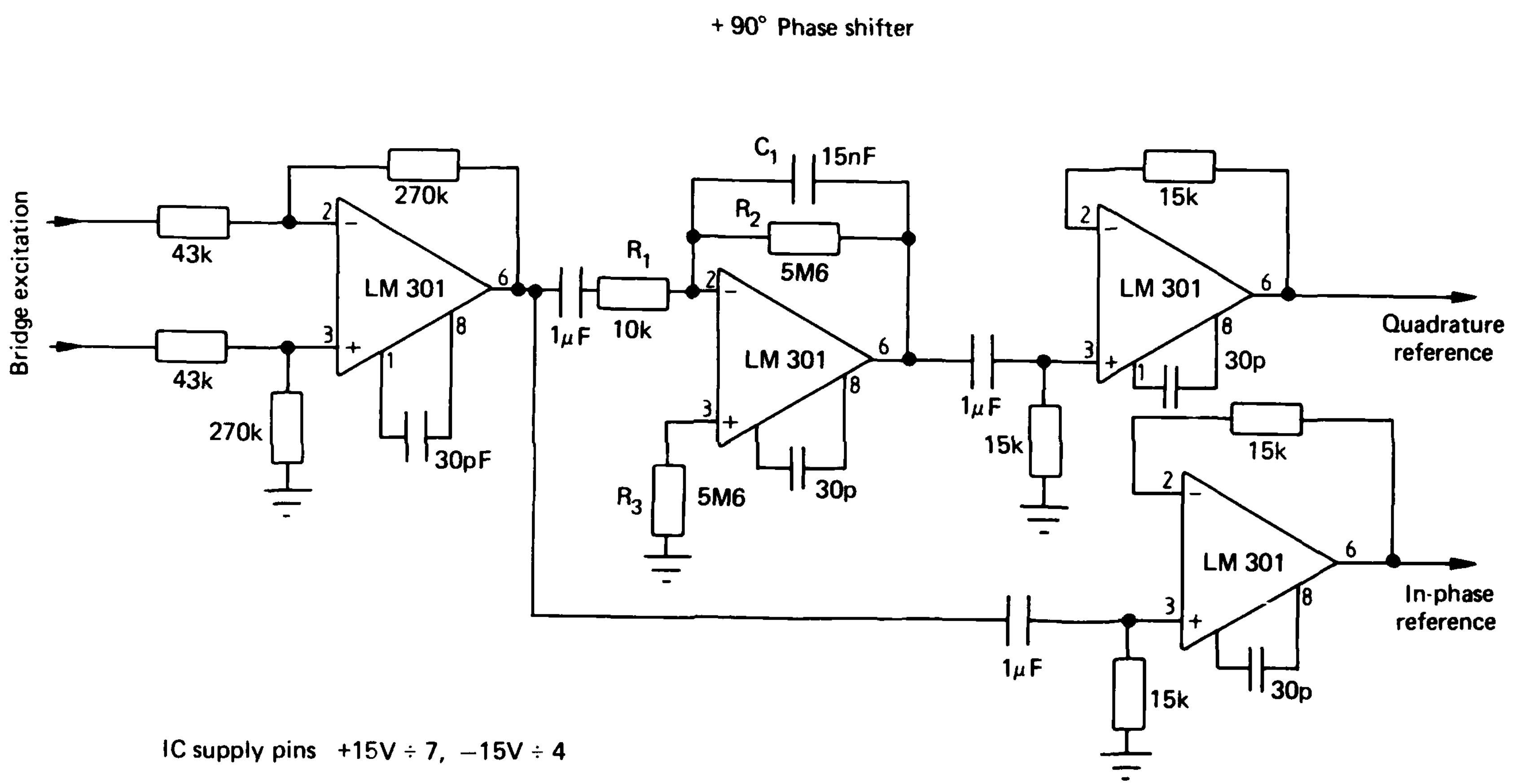
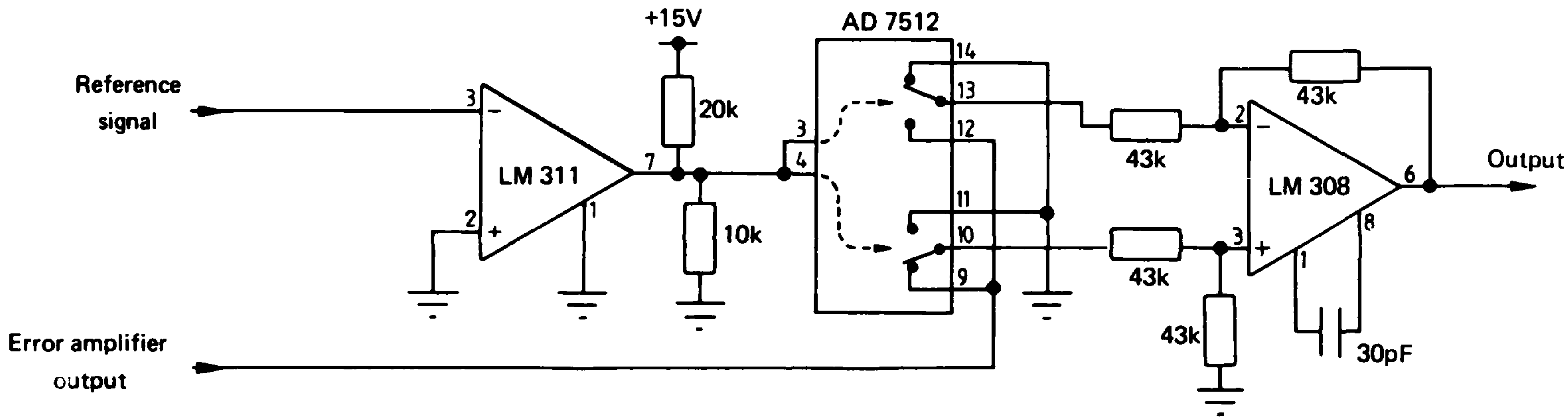


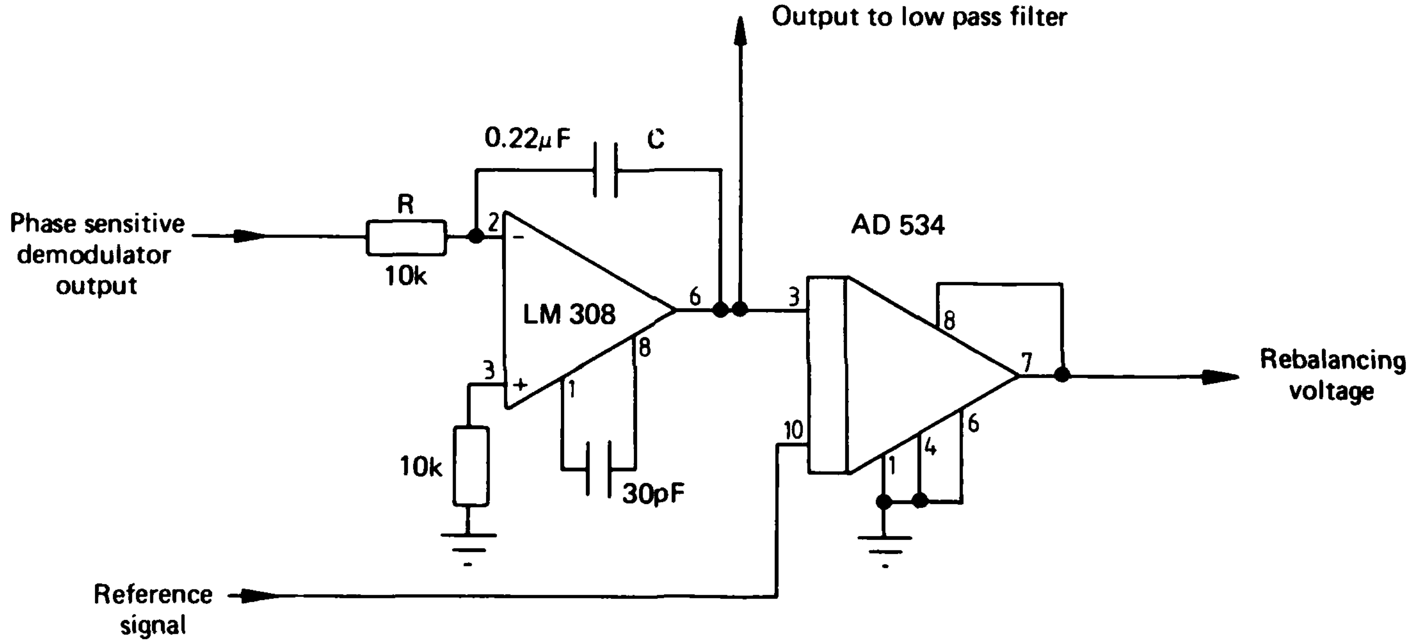
FIG. 5 REFERENCE AMPLIFIER CIRCUIT



Supply pins

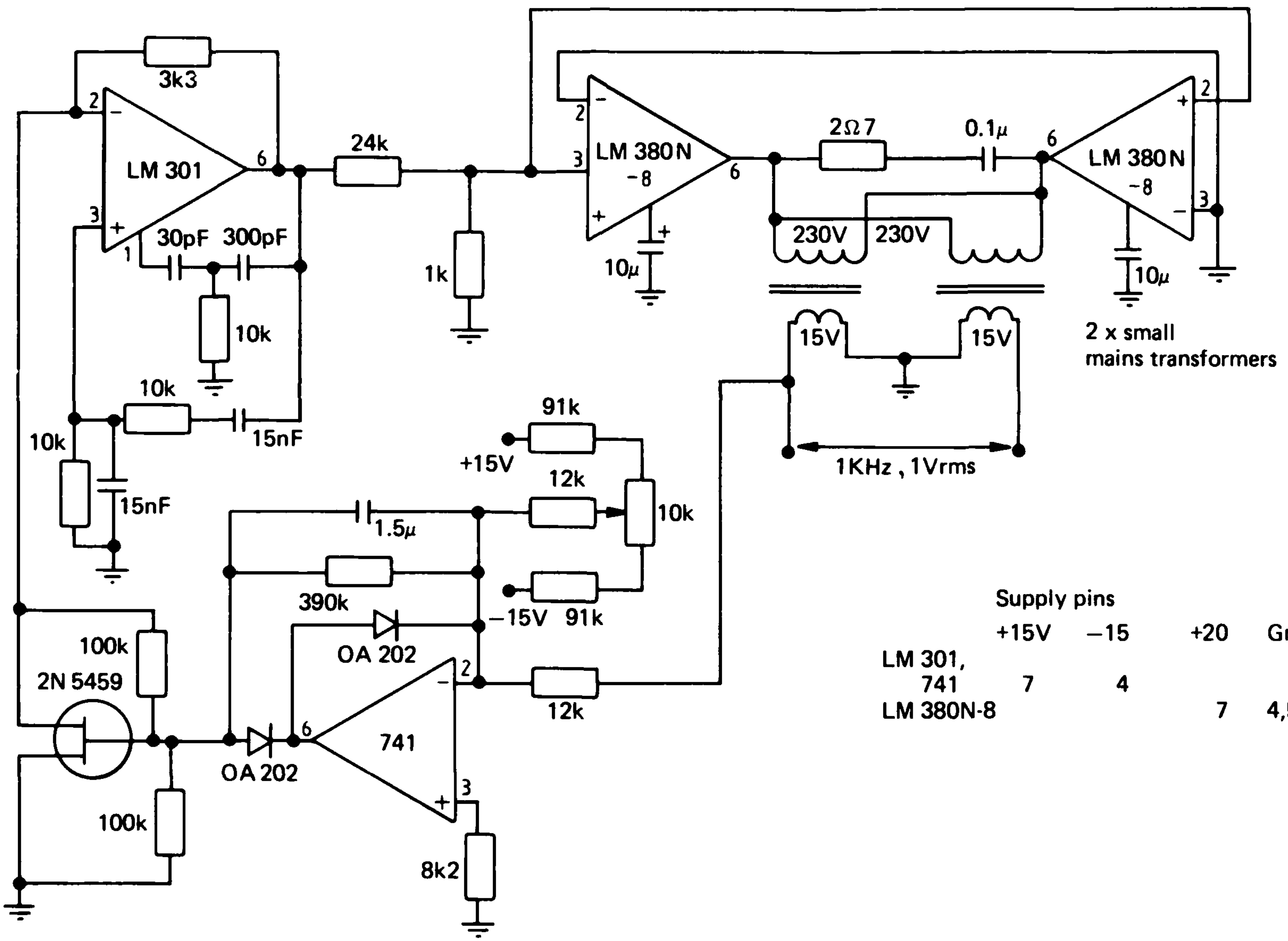
	+15V	−15V	Gnd.
LM 311	8	4	–
LM 308	7	4	–
AD 7512	7	1	2

FIG. 6 PHASE SENSITIVE DEMODULATOR CIRCUIT



	Supply pins	
	+15V	−15V
LM 308	8	4
AD 534	9	5

FIG. 7   INTEGRATOR AND REBALANCING MULTIPLIER CIRCUIT



Supply pins

	+15V	−15	+20	Gnd.
LM 301, 741	7	4		
LM 380N-8			7	4,5

FIG. 8 CIRCUIT OF EXCITATION SUPPLY

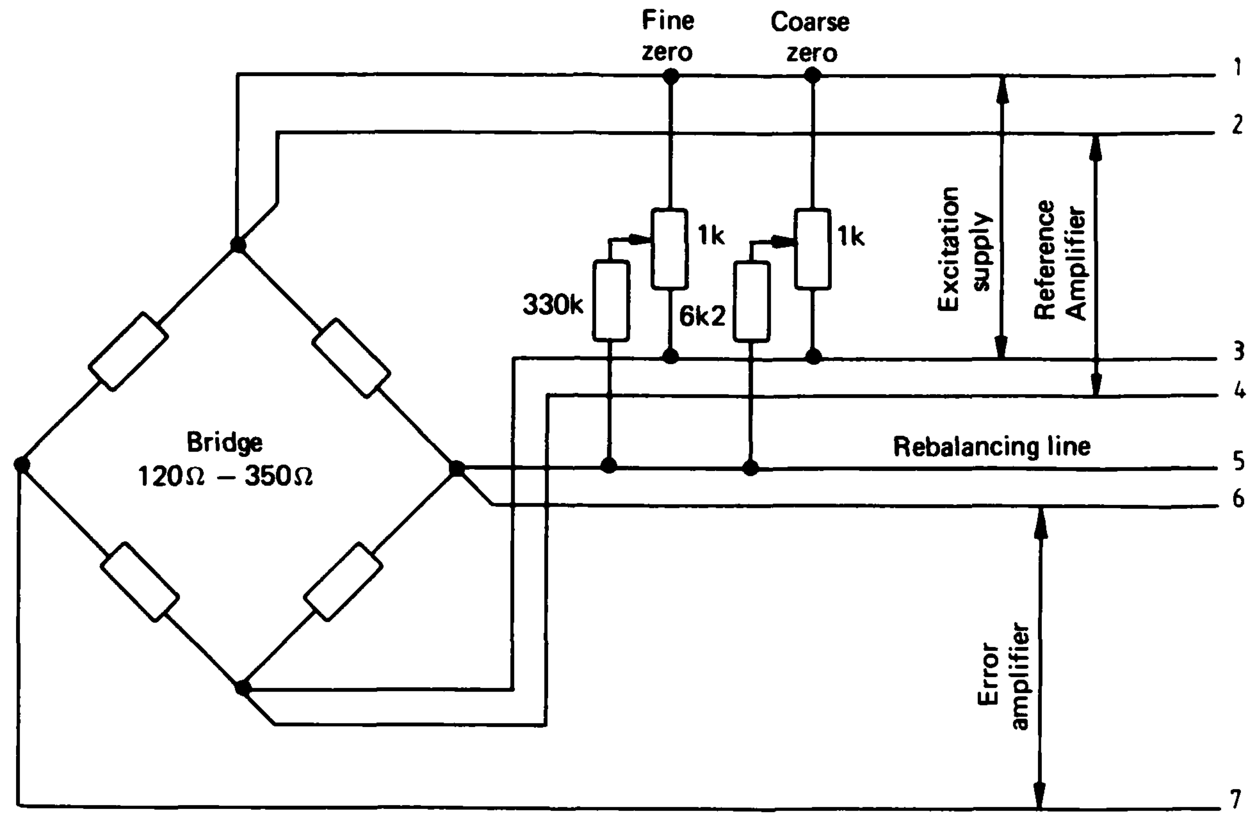


FIG. 9 BRIDGE CONNECTIONS AND ZERO OFFSET NETWORK

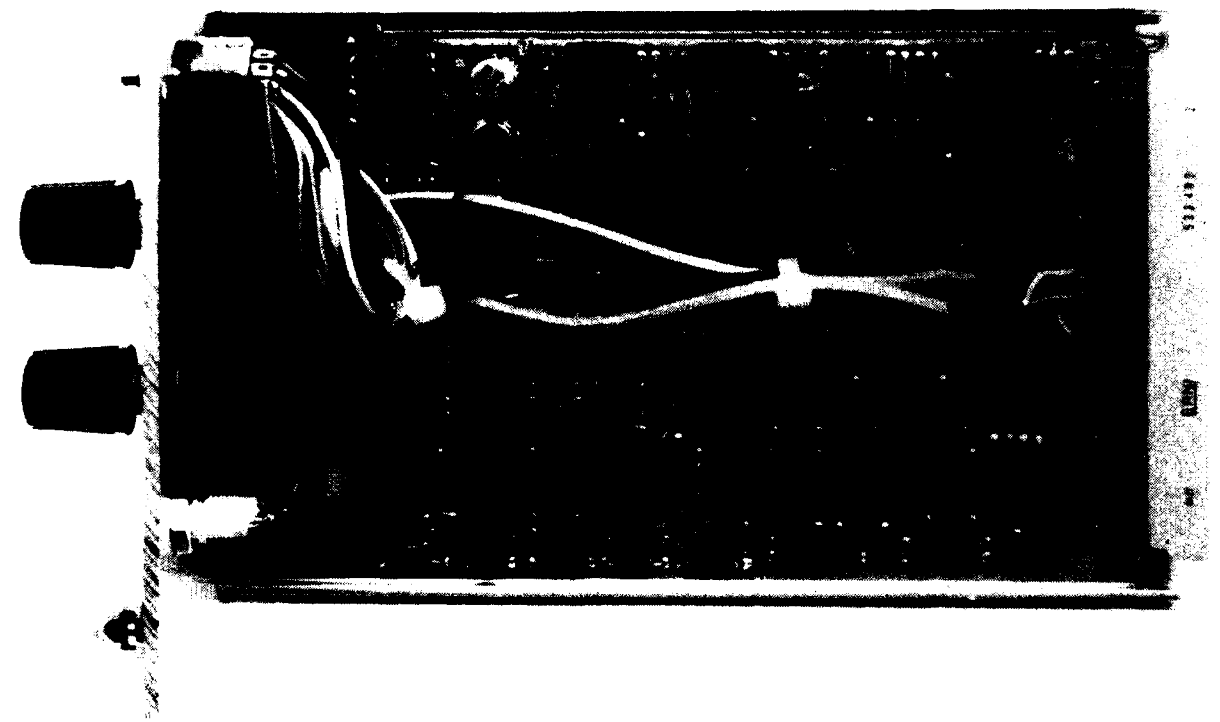


FIG. 10 PHOTOGRAPH OF COMPLETE ARL AMPLIFIER

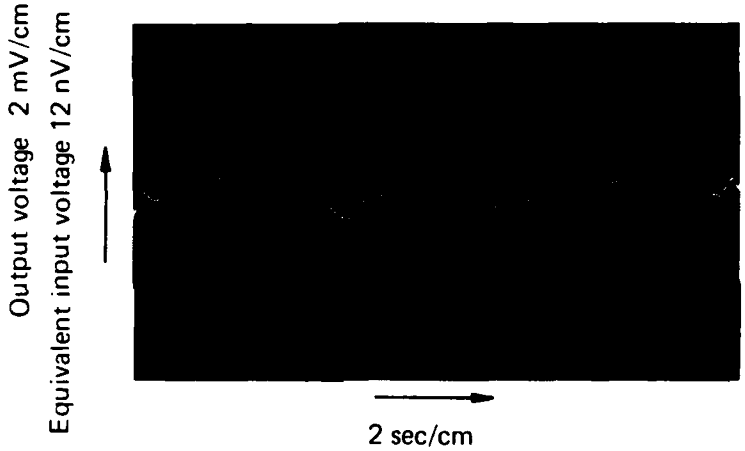


FIG. 11 OUTPUT FOLLOWING 0.7 HZ FILTER
(10 MΩ rebalancing resistor)

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16. Abstract

An AC excited self-balancing instrument for measuring low level strain gauge transducer outputs is described. The design of a specific circuit optimised for use with a small wind tunnel force balances is presented with sufficient information to facilitate circuit optimisation for other applications.

The instrument is intended for applications where the bridge output voltage is below the level that can be conveniently handled by conventional DC amplifier systems. Tests on a prototype instrument showed an input noise in a 0·1 Hz to 0·7 Hz bandwith of 4 nV RMS, a zero drift of less than 0·1 µV per day and negligible sensitivity drifts.

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