



Nuclear Control Systems

Sensor Interface Circuitry Design for Nuclear Applications

| | |
|-------------|---------------------------------|
| Issue | 1 |
| Date | February 2012 |
| Publication | Nuclear Future Volume 8 Issue 1 |

Contents

| | |
|---|-----------|
| 1. INTRODUCTION | 3 |
| 2. RATIOMETRIC DESIGN..... | 4 |
| 3. PHASE QUADRANT DEMODULATION | 5 |
| 4. PRACTICAL REALISATION OF A RATIOMETRIC INTERFACE CIRCUIT FOR A DC VOLTAGE DRIVEN SENSOR | 7 |
| 5. PRACTICAL REALISATION OF A RATIOMETRIC INTERFACE CIRCUIT FOR A DC CURRENT DRIVEN SENSOR..... | 9 |
| 6. PRACTICAL REALISATION OF A RATIOMETRIC INTERFACE CIRCUIT FOR A SENSOR WHICH IS NOT DRIVEN | 11 |
| 7. PRACTICAL REALISATION OF A RATIOMETRIC INTERFACE CIRCUIT FOR AN AC CURRENT DRIVEN SENSOR | 13 |
| 8. BENEFITS OF USING RATIOMETRIC DESIGN | 14 |
| 9. BENEFITS OF USING PHASE QUADRANT DEMODULATION | 16 |
| 10. ACCURACY ANALYSIS | 17 |
| 11. CONCLUDING REMARKS..... | 18 |
| 12. REFERENCES..... | 19 |

1. Introduction

This paper examines the use of ratiometric design and phase quadrant demodulation in sensor interface circuitry design for nuclear applications.

Practical realisation of ratiometric design and phase quadrant demodulation for a range of different sensors is described. These include pressure, temperature PRT, thermocouple and inductive rod position sensors.

The advice and insights offered in the sections on practical realisation comes from the authors' experiences of designing such circuitry for use in nuclear protection systems.

Typical accuracy of the sensor interface circuitry that can be achieved using ratiometric design is quoted for the full ambient temperature range and the lifetime of the equipment.

It is concluded that the use of ratiometric design and phase quadrant demodulation can result in several benefits. These include increased channel accuracy, no calibration for the life of the equipment, increased on-line fault coverage (diagnostics) and only simple digital signal processing required.

2. Ratiometric Design

Referring to figure 2, the main principles of ratiometric design are as follows:

1. A single measurement chain of differential instrumentation amplifier, filtering and Analogue to Digital Converter (ADC) is used to measure the outputs from the sensor (VS), reference high (VH) and reference low (VL). Normally VH is full scale output from the sensor and VL is the zero of the sensor output scale.
2. The sensor VS, VH and VL signals are all measured in a short time span (about 200ms typically), such that amplifier and ADC gains and offset voltages do not drift significantly with time or temperature. In order to switch quickly enough between these three signals a CMOS multiplexer is used.
3. The three ADC readings VS, VH and VL are then digitally processed to yield the sensor parameter. For example the pressure P is given by the formula:

$$P = k \times \left(\frac{V_S - V_L}{V_H - V_L} \right)$$

Where k = constant equal to the full scale pressure in bars, for example.

Subtracting of VL from VS and VH removes all errors due to DC offset voltages in the amplifier, anti-aliasing filter and ADC.

Dividing VS - VL by VH - VL removes any gain errors in the amplifier, anti-aliasing filter and ADC.

4. Where sensor excitation is required, the circuit configuration is arranged such that the same sensor drive voltage or current is used to generate VH and VL as well as VS. This means that any error in the sensor drive voltage or current is removed when VS - VL is divided by VH - VL.

This has the advantage that the sensor drive voltage or current does not need to be highly accurate, only stable over about 200ms.

5. In some cases, the errors due to finite common mode voltage rejection of the differential instrumentation amplifier can also be eliminated. For example in the case of a strain gauge pressure sensor excited by split DC supplies (e.g. +5V and -5V). Instead of connecting the VL multiplexer inputs to zero volts, connect them to the centre of a resistive divider (of two equal resistances) between the split supplies. By this means any common mode voltage, due to imbalance of the split supplies, will be applied to each of the differential voltages representing VS, VH and VL. The finite common mode rejection capability of the differential instrumentation amplifier will result in an additional offset voltage which adds to VS, VH and VL. This additional common mode offset voltage will then be eliminated in the digital processing.

3. Phase Quadrant Demodulation

This is a scheme applicable to the processing of signals from AC excited sensors, for example variable inductance or variable transformer sensors.

The sensor is excited by a digitally generated sine wave of suitable frequency. According to their characteristics, some types of sensor may be driven by a controlled current, e.g. by an “Improved Howland Current Pump” (Sheingold, 1964), while others may need a simpler voltage drive.

The sensor output and the excitation signal are measured in turn, via a multiplexer but through a common gain path and ADC. Four equally spaced blocks of ADC samples are taken per cycle of the drive waveform, as shown in figure 1. Each block of N samples is accumulated by a simple adder to obtain a digital measure of the sensor output amplitude in that quadrant. We will denote these blocks of ADC samples as P1, P2, P3 and P4 for the excitation signal, and S1, S2, S3 and S4 for the sensor output.

The measuring system is synchronous to the drive waveform, although the phase relationship is arbitrary, provided it is stable for the duration of a full measuring sequence.

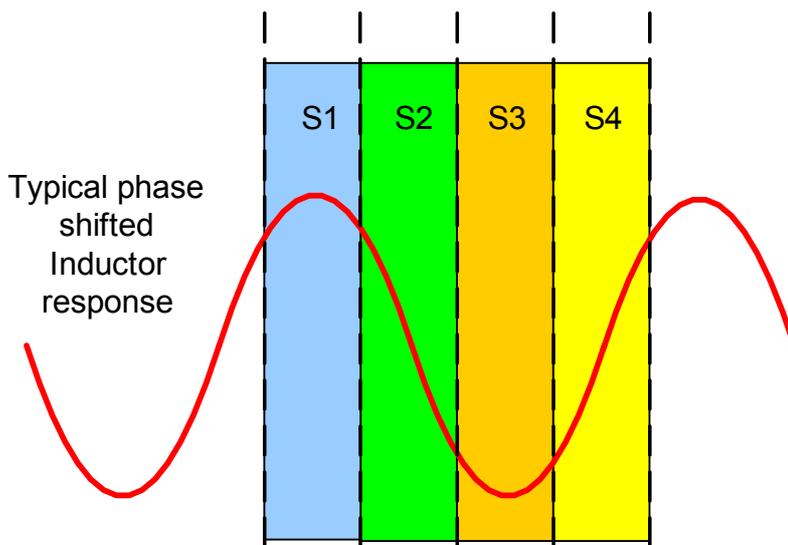


Figure 1: Phase Quadrant Sampling Scheme for Sensor Excitation Signal

Figure 1: Phase Quadrant Sampling Scheme for Sensor Excitation Signal

The accumulated values may be summed over a number of cycles of the excitation signal to obtain better mains frequency rejection. Typically the number of cycles chosen is the lowest common multiple of the period of the mains frequency and the excitation frequency. A number of cycles greater than the lowest common multiple may be utilised if more noise rejection is desired, however this will increase the response time.

Then the accumulated quadrant values are processed into notional (relative to the timing of the measuring channel, not the phase of the excitation) real and quadrature values as follows:

$$P_{REAL} = P1 + P2 - P3 - P4$$

$$P_{QUAD} = P1 - P2 - P3 + P4$$

$$S_{REAL} = S1 + S2 - S3 - S4$$

$$S_{QUAD} = S1 - S2 - S3 + S4$$

In the case of a simple variable transformer sensor (e.g. rod at bottom), the parameter of interest is the effective ratio of secondary voltage to primary current. The secondary measurement being voltage, the parameter which is measured is transimpedance, and its magnitude is given by:

$$|Z| = \frac{V_s}{I_p} = R_{REF} \sqrt{\frac{S_{REAL}^2 + S_{QUAD}^2}{P_{REAL}^2 + P_{QUAD}^2}}$$

where: $|Z|$ = magnitude of the transimpedance

V_s = voltage on secondary of variable transformer sensor

I_p = current in primary of variable transformer sensor

R_{REF} = current sensing resistor in primary circuit

Neglecting residual circuit errors, the value of $|Z|$ is ratiometric to, and dependent only on, the accuracy of the precision reference resistor.

In the case of a variable inductance sensor (e.g. rod position), the reference measurement is made by sampling the voltage across a precision resistor R_{REF} in series with the inductor, which is current driven. If the excitation frequency is F , and the sensor has inductance L and resistance R , it can be shown that:

$$R = R_{REF} \times \frac{S_{REAL} \times P_{REAL} + S_{QUAD} \times P_{QUAD}}{P_{REAL}^2 + P_{QUAD}^2}$$

$$L = \frac{R_{REF}}{2 \times \pi \times F} \times \frac{S_{QUAD} \times P_{REAL} - S_{REAL} \times P_{QUAD}}{P_{REAL}^2 + P_{QUAD}^2}$$

These calculations are carried out in the digital processing, providing accurate values for both L and R .

4. Practical Realisation of a Ratiometric Interface Circuit for a DC Voltage Driven Sensor

A strain gauge bridge pressure sensor is an example of a sensor usually driven by a DC voltage.

In Figure 2, a schematic diagram for a practical realisation of ratiometric sensor interface circuitry is shown for such a sensor.

A resistive divider is used to reduce the sensor drive voltage of +5V and -5V to a voltage the same as the full scale output from the sensor, about 30mV typically. The voltage output from this resistive divider is the reference voltage V_H .

The precision resistors chosen for this resistive divider will typically be leaded, 0.005% tolerance, 2.5ppm/°C temperature coefficient, change due to soldering 0.01% and 'shelf life' drift of 0.0025% over 1 year at 25°C.

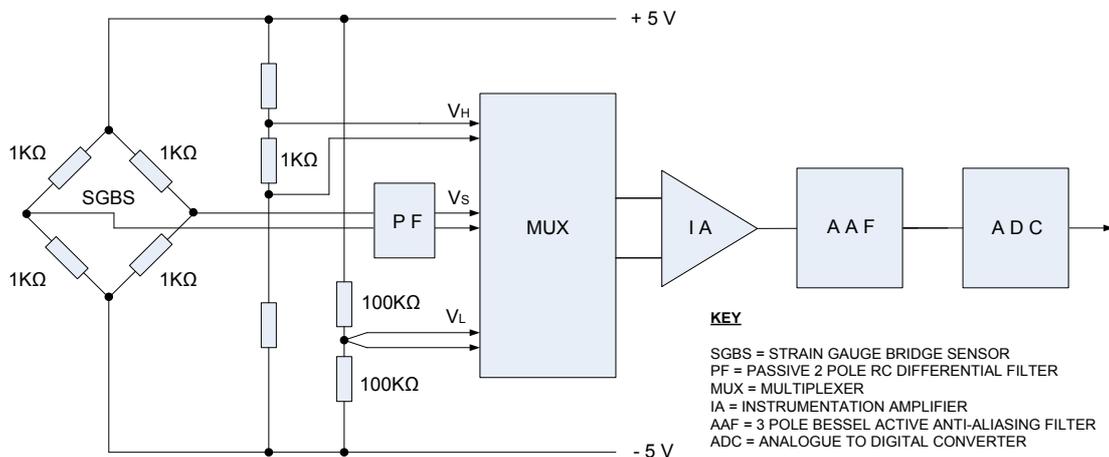


Figure 2: DC Voltage Driven Sensor Interface Schematic

The reason for choosing precision resistors with leads is that the error due to soldering is less than for a surface mount package.

If the strain gauge bridge has an output resistance of 1kΩ, then it is advantageous to arrange for the same output impedance for the resistive divider. This minimises errors due to instrumentation amplifier bias currents.

A strain gauge bridge pressure sensor will typically have an output near to zero millivolts when the applied pressure is zero. So the V_L inputs to the multiplexer could be connected to the zero volts of the power supplies. However it is advantageous instead to connect the V_L multiplexer inputs to the centre of a resistive divider (of two equal value precision resistances) between the split supplies. By this means any common mode voltage, due to imbalance of the split supplies will be applied to each of the differential voltages representing V_S , V_H and V_L .

The finite common mode rejection capability of the differential instrumentation amplifier will result in an additional offset voltage which adds to V_S , V_H and V_L . This additional common mode offset voltage will then be eliminated in the digital processing.

A two pole passive resistive-capacitive differential filter fitted before the multiplexer helps to reduce high frequency noise on the sensor signal. Depending on the frequency response required, a filter with its -3dB point at about 10Hz may be appropriate.

The differential instrumentation amplifier requires low bias currents and high Common Mode Rejection Ratio (CMRR).

An active Bessel three pole anti-aliasing filter (AAF), with its -3dB point at about a few kHz, is placed before the ADC. The Bessel filter is chosen for its fast settling time. An allowance of about 1ms is necessary before ADC readings commence to ensure that the voltage output from this filter has settled to within 0.01% of its final value.

Whichever type of ADC is employed, the most important parameter in a ratiometric application is non-linearity. So an ADC, with more bits than necessary for resolution purposes, may need to be used to achieve low non-linearity error. This is because improved linearity usually goes with a higher bit output ADC.

The amount of voltage headroom allowed for in the interface circuitry is important. Insufficient headroom could result in the sensor signal plus mains noise being peak clipped, due to exceeding voltage input or output ratings of devices. Then the apparent output signal, after averaging to remove mains noise, can be in error. Allowing voltage headroom equal to or greater than the sensor signal range may be appropriate. For example in the case of a 30mV full scale sensor signal and where long sensor cables of 30m are employed.

The sensor signal sampling period is chosen to be a multiple of the mains cycle time. By averaging the ADC readings over this sample period, the mains borne noise on the sensor signal is effectively cancelled out.

5. Practical Realisation of a Ratiometric Interface Circuit for a DC Current Driven Sensor

A Platinum Resistance Thermometer (PRT) sensor is an example of a sensor that can be driven by a DC current.

In figure 3, a schematic diagram for a practical realisation of ratiometric sensor interface circuitry is shown for such a sensor.

A sensor drive voltage of 1mA is typical. The reference high and reference low resistors are in series with the sensor, which may be a 100Ω PRT (at 0°C). So if a sensor range from 0°C to 400°C were required then reference low resistor would be 100Ω and reference high would be 247.09Ω.

The greatest design difficulty with this circuit topology is common mode voltage. Shown in Figure 3 is a feedback circuit which helps to minimise the problem. As the PRT sensor changes resistance with temperature, the operational amplifier adjusts the voltage on the resistor chain to maintain zero volts at the junction between the PRT and the reference resistors. Therefore the common mode voltage on the reference resistors is unchanged due to the sensor resistance changing with temperature.

The other sensor interface circuitry, i.e. the passive signal filter, the active anti-aliasing filter, the instrumentation amplifier, the multiplexer and the ADC are similar to those described for the voltage excited sensor.

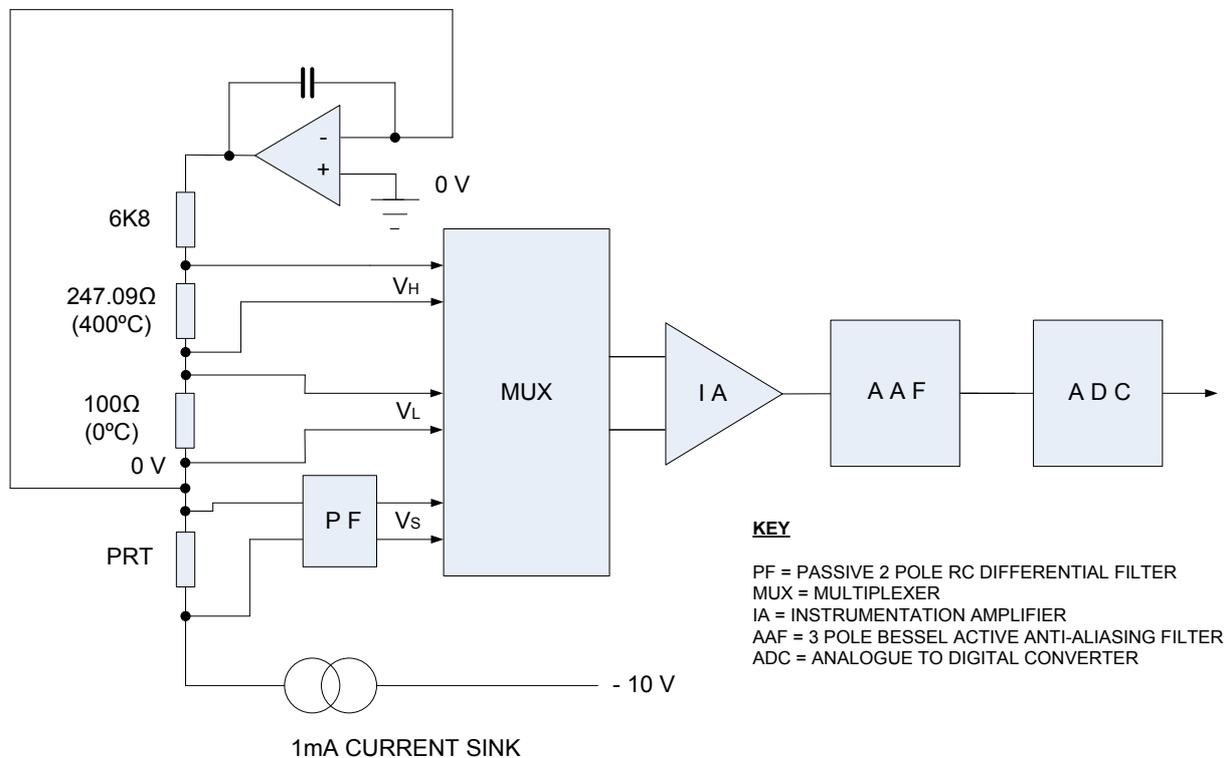


Figure 3: DC Current Driven Sensor Interface Schematic

A practical reason for choosing current drive is that contact resistances in the sensor drive circuitry have little effect on the sensor resistance measurement. Often there may be junction boxes in the sensor cabling which are situated in a hot environment where there is significant nuclear radiation. The hot environment can lead to corrosion of joints within the junction boxes, resulting in increased contact resistance.

By the use of a current drive the sensor interface circuitry is made tolerant to increased contact resistances. This means that the radiation dose of maintenance personnel can be reduced, since they will not be required to correct the high contact resistance.

6. Practical Realisation of a Ratiometric Interface Circuit for a Sensor which is Not Driven

A thermocouple sensor is an example of a sensor that is not driven.

In Figure 4, a schematic diagram for a practical realisation of ratiometric sensor interface circuitry is shown for such a sensor.

In this case a reference voltage with a potential divider is employed instead of a reference resistor. The potential divider values are chosen so that the voltage output is the same as full scale voltage output from the thermocouple.

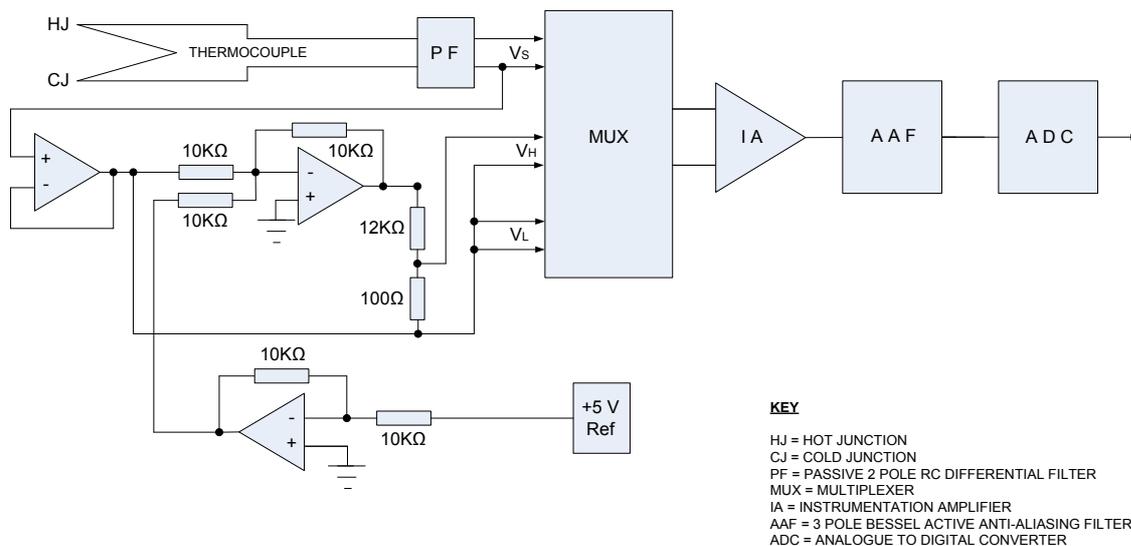


Figure 4: Non-Driven Sensor Interface Schematic

A thermocouple is usually nominally floating (no connection to 0V or earth), meaning that common mode voltage is not an issue. However provision needs to be made to allow the thermocouple to carry on working in the case of a fault that connects the thermocouple to earth and gives rise to a common mode voltage.

Such a situation can occur due to a leak in the hermetic sealing of a stainless steel sheathed, magnesium oxide insulated thermocouple. Due to ingress of water vapour there is a lower than normal resistance formed to the earthed sheath. Further a common mode voltage of up to about one volt can be generated due to a chemical reaction.

One method of dealing with this common mode voltage is to measure the common mode on the thermocouple. This voltage is then buffered and used as the zero volts reference V_L . Also the output of a voltage reference is added, in an operational amplifier circuit, to this buffered common mode voltage and used as the reference voltage V_H . See figure 4 for details of the circuitry.

The result of this is that, as in the case of the voltage driven sensor, residual common mode voltage offset errors are eliminated in the digital processing. By this means extremely high common mode rejection can be achieved and the circuit accuracy is not significantly degraded when there is this type of fault in the thermocouple.

Another aspect of the sensor interface circuitry design for thermocouples to be aware of is that due to long lengths of cable, the thermocouples may have significant output resistance of up to about 1000Ω .

There is an error due to a potential divider being formed between the thermocouple and the input bias current resistors fitted between zero volts and the inputs to the instrumentation amplifier. This error can be reduced by choosing the bias resistors to be in the ratio of the resistivities of the thermocouple metals (e.g. chromel/alumel).

Note that no mention has yet been made of either fixing or measuring the temperature of the thermocouple cold junction. One method that can be used is to place the cold junction in a copper bar along with a PRT sensor. The PRT sensor can then be current driven and measured ratiometrically as described previously.

The other sensor interface circuitry, i.e. the passive signal filter, the active anti-aliasing filter, the instrumentation amplifier, the multiplexer and the ADC are similar to those described for the voltage excited sensor.

7. Practical Realisation of a Ratiometric Interface Circuit for an AC Current Driven Sensor

An inductive rod position sensor is an example of a sensor that can be driven by an AC current.

In figure 5, a schematic diagram for a practical realisation of ratiometric sensor interface circuitry is shown for such a sensor.

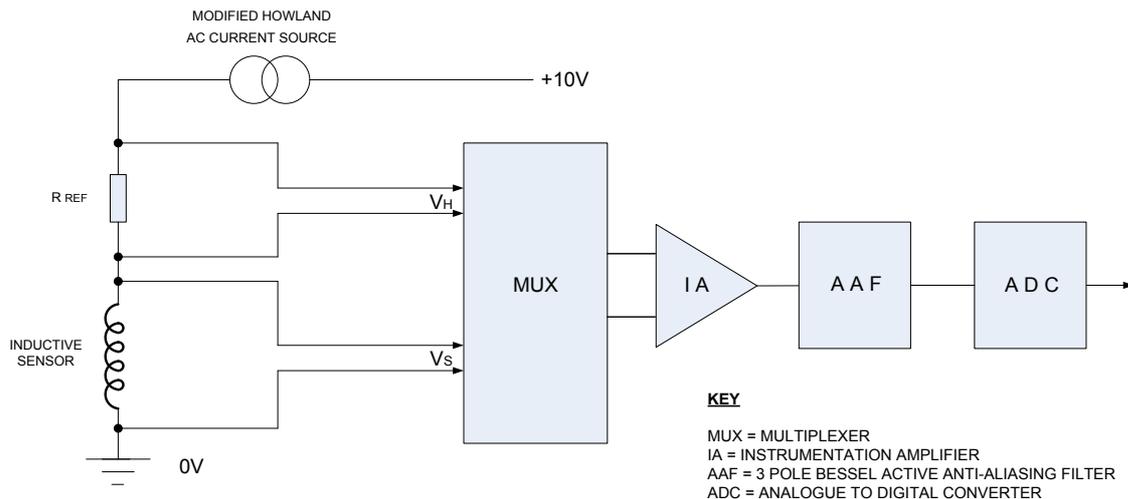


Figure 5: AC Current Driven Sensor Interface Schematic

The current sensing resistor RREF is in series with the inductive sensor being measured.

The phase shift and gain due to amplification and filtering does not need to be accurate, only stable over about 200ms due to ratiometric measurement principles being employed.

The sensor interface circuitry, i.e. the active anti-aliasing filter, the instrumentation amplifier, the multiplexer and the ADC are similar to those described for the voltage excited sensor.

The sensor signal sampling period is chosen to be a multiple of both the mains cycle time and the sensor cycle time. By this means mains borne noise and noise due to even harmonics on the sensor signal are eliminated. Second harmonic noise on the sensor signal often arises from magnetic hysteresis in the sensor ferromagnetic core material.

Designing the modified Howland current source to be stable can present difficulties. Firstly the inductive load with the sensor cable capacitance forms a resonant circuit at a particular frequency. Secondly, due to multiple feedback loops in the modified Howland current source, stability cannot be evaluated with the usual Bode plot analysis. So care must be taken in the design to ensure stability.

8. Benefits of Using Ratiometric Design

1. Allows the achievement of channel accuracies of about 0.2% of full scale, over 0°C to 60°C, 25 years and rigorous full worst case accuracy analysis. This means that sensor interface cards do not need calibration for the equipment lifetime (e.g. 25 years). The result of this is reduced costs and improved safety due to the removal of the possibility of a common mode fault arising in the periodic calibration. Further this low error of about 0.2% of full scale enables maximum power to be extracted from the reactor.

Accuracy is a key aspect of ratiometric design which is linked to both safety and performance. Therefore there is a more detailed discussion of accuracy analysis later in this paper.

2. By addition of an extra reference resistor, the channel calibration can be checked every measurement cycle to an accuracy of about 0.1% of full scale. This means component faults which give rise to errors above about 0.3% of full scale can be detected on-line. This is a very powerful diagnostic tool which yields a high degree of fault coverage and fault reporting in a few minutes or less.

When part of a multichannel protection system with voting (usually 2oo4), the high degree of fault coverage and rapid fault reporting mean that both safety and availability of the plant are enhanced.

3. Precision resistors can usually be used as reference devices. Provided these resistors are run at lower power dissipation, the drift with time will be small. Generally precision reference resistors run at low power will drift significantly less than a precision voltage reference. Precision resistors are also significantly less expensive than precision voltage references.
4. The nature of ratiometric design means that very few component characteristics, apart from those of the reference resistors or reference voltage, are important with regard to accuracy.

A list of these component characteristics is detailed below:

- a) ADC non-linearity
- b) Multiplexer leakage currents
- c) Multiplexer 'on' resistance
- d) Instrumentation amplifier input bias currents
- e) Instrumentation amplifier Common Mode Rejection Ratio (CMRR) - only important in some cases
- f) Intrinsic noise in the instrumentation amplifier.

5. No initial calibration or trimming of channel gain and offset is required.
6. Because only relatively simple mathematical operations are required by the radiometric measurement system, the digital processing is simple to implement and therefore safer. The equations suggest that addition, subtraction, division, multiplication and square rooting are required. However by utilising the approach given in reference (Bisdounis et al, 1993), the only digital processing required is addition, subtraction and bit shifting.

Note that sometimes a mathematical operation can be avoided. For example, consider the case where a trip point is at a particular value of a parameter. However a square root operation would be required to derive that parameter. Instead work out the square of the trip point and use this as the trip criterion, thus avoiding the need to perform a square root calculation on the sensor signal.

7. For voltage excited sensors the drive voltage does not need to be highly accurate, only stable over one measurement cycle of about 200ms. Similarly for a current excited sensor, the drive current only needs to be stable over the short term. These benefits apply to both AC and DC excited sensors.

9. Benefits of Using Phase Quadrant Demodulation

1. Measurements of inductance are relative to a reference resistor and a crystal clock (frequency). Both of these are physically much smaller and cost less than any feasible inductive reference with similar accuracy.
2. The measurement of resistance of the inductive sensor depends only on the accuracy of the reference resistor, neglecting circuit errors such as nonlinearity.
3. Phase shift in both the drive amplifier and the signal amplifier and filter are of no consequence.
4. Only relatively simple mathematical operations need to be performed on the measured signals to extract the sensor inductance, resistance and phase shift. This means the digital processing required is simpler which equates to safer.

10. Accuracy Analysis

It has previously been stated that ratiometric design can achieve channel accuracies of about 0.2% of full scale, over 0°C to 60°C, 25 years and rigorous full worst case accuracy analysis.

The details of the rigorous full worst case accuracy analysis are given below.

In the accuracy analysis the following error sources are considered:

1. Component initial tolerance
2. Component temperature coefficient
3. Component drift with time
4. Effect of soldering
5. Common mode voltage
6. Power supply noise (Effect of this on accuracy measured not calculated)
7. Multiplexer leakage and 'on' resistance
8. Instrumentation amplifier input bias current
9. ADC non-linearity
10. Intrinsic noise in semiconductors
11. Mains frequency noise on sensor signal due to pickup by sensor cable
12. Digital processing accuracy

In the case of component drift with time, the drift is assumed to vary:

1. As the square root of time
2. Linearly with power but limited to the shelf life drift at zero power
3. Linearly with temperature with respect to 0°C

The claim of 0.2% of full scale accuracy may not appear very impressive when judged against a digital voltmeter (DVM) that claims 0.005% accuracy. However on close examination of the DVM specifications it will be found, for example, that the quoted accuracy is valid over 23°C±5°C, for one year and that some statistical analysis has been carried out when combining errors. Further the rigor with which the assessment of errors is not known, i.e. have all sources of error been considered.

11. Concluding Remarks

It has been argued that ratiometric design and phase quadrant demodulation can bring significant benefits to sensor interface circuitry design for nuclear applications. These include increased channel accuracy, no calibration for the life of the equipment, increased on-line fault coverage (diagnostics) and only simple digital signal processing required.

Advice on the practical realisation of these design principles for a range of sensor types has been detailed. These include pressure, temperature PRT, thermocouple and inductive rod position sensors.

Typical accuracy of the sensor interface circuitry that can be achieved using ratiometric design is quoted for the full ambient temperature range and the lifetime of the equipment.

12. References

1. Bisdounis et al (1993). VLSI Implementation of Digit-Serial Arithmetic Modules. "Microprocessing and Microprogramming Journal", Vol 39 Dec 1993.
2. Sheingold, D.H. (1964). Impedance & admittance transformations using operational amplifiers. "The Lightning Empiricist", Vol 12, No.1, January 1st 1964 http://www.philbrickarchive.org/1964-1_v12_no1_the_lightning_empiricist.htm.