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Electronics for Measuring Systems

Prerequisite for the second year course: Sensors and amplifiers in measuring instruments

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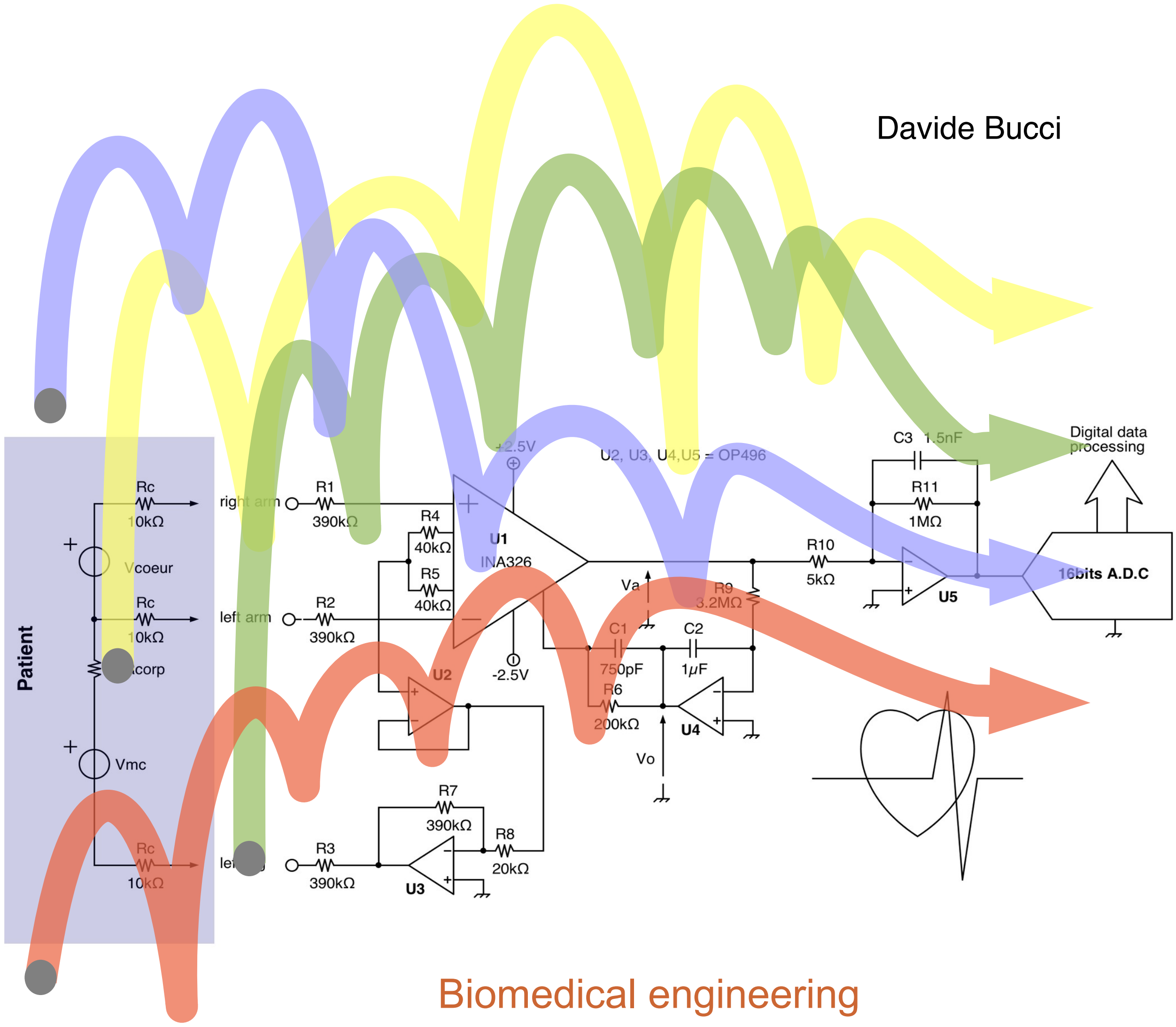

Biomedical engineering

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Cover freely inspired from Giovanni Pintori's advertisements for the Olivetti Lettera 22 typewriter. The circuit shown is a simplified system for acquiring electrocardiogram signals. Most circuit diagrams in this document were drawn with FidoCadJ, an open source and multiplatform editor, and then included in the pdfLaTeXdocument by means of the PGF/TikZ package.

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CHAPTER 1

Fundamentals of sensing and signal conditioning

1.1 Introduction

The first element in a classic electronic measurement chain is the sensor. Its role is to translate the physical quantity to be measured (called *measurand*) in an electrical quantity of some kind. Clearly, the goal is to obtain knowledge about the physical quantity: this translation should be done in a known and reliable way. Sensors are based on a wide range of principles and (our point of view being from the electronics side) we follow the classification proposed in [Asch, 2003]: we will categorise them depending on the electrical quantity which is of interest at their output: voltage, current, charge, resistance, reactance. This categorisation is not the only one applicable, but it allows to treat the signal conditioning at the same time as sensors.

1.2 Voltage generators

1.2.1 General description

Several physical phenomena entail in a specific equipment the presence of a voltage somewhere, related to a particular physical variable. They can thus be exploited to build sensors which can be seen as voltage sources, whose voltage depend on the measurand m. Very often, the electrical representation of the sensor might be a Thévenin-type equivalent circuit including a series impedance, as shown in figure 1.1.

The open-circuit voltage given by the sensor is $e(m)$ and its relation with m, the measurand, must be known and must not change appreciably with the time. The internal impedance of the sensor is represented by Z_c and determines the voltage drop between $e(m)$ and V_s when a load is attached, if the current I_s is different from zero.

We thus proceed by detailing some examples of such sensors.

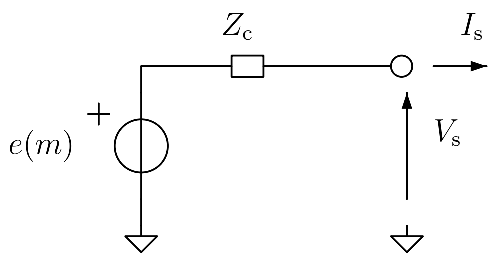


Figure 1.1: Thévenin representation of a voltage generator sensor.

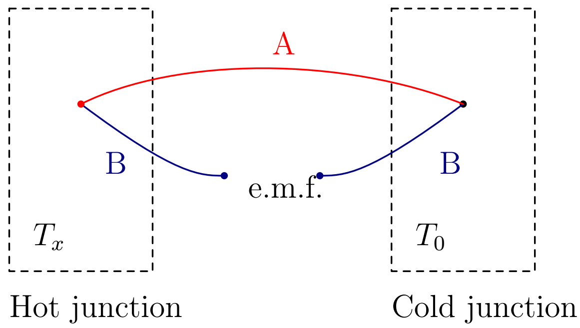


Figure 1.2: The Seebeck effect in two junctions of different conductors (A and B).

1.2.2 Examples

1.2.2.1 Thermocouples

A thermocouple is a temperature sensor based on the Seebeck effect. The principle, shown in figure 1.2 is that when two junctions between different metallic conductors are kept at a different temperature, a voltage difference can be measured. This voltage is approximatively proportional to the difference of temperature between the two junctions. This effect is a consequence of heat transport in conductors and the Seebeck coefficient is a volume property of each one. The Seebeck voltage, thus, is not generated in the junctions themselves, but on the whole length of the conductor: it is always present, even in a homogeneous circuit, but it becomes observable only with different conductors spliced together.

In the case of a thermocouple, we call conventionally "hot junction" and "cold junction" the connections between the two conductors and the temperature measurement is intrinsically differential. If one needs an absolute measurement, the cold junction should be kept at a constant and controlled temperature (for example employing water/ice slurry for $0\,\mathrm{C}$), or a *cold junction compensation circuit* may be used.

From a practical point of view, buckets containing ice and water have long ago been replaced by compensation circuits, more compact, much easier to be kept running and

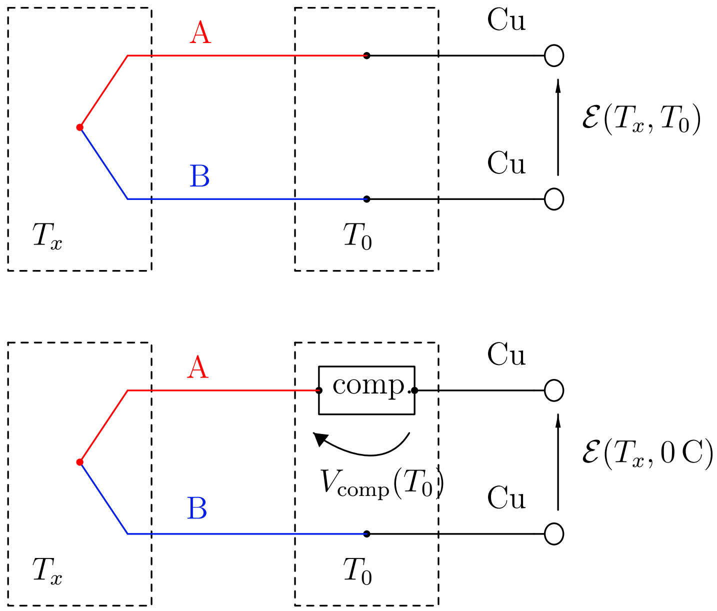


Figure 1.3: A cold junction compensation of a thermocouple measurement system. Temperature T_0 of the cold joint is measured and translated into a voltage $V_{\mathrm{com}}(T_0)$, substracted from the thermocouple output.

less expensive. The idea is to measure the temperature of the cold junction with a physical principle different from the Seebeck effect and subtract it to the signal delivered by the thermocouple. The advantage is that the cold junction is much less subjected to extreme temperatures or harsh conditions so the measurement is easy. Among the available strategies, a common solution is to make this subtraction directly to voltage delivered by the thermocouple. Figure 1.3 shows how it can be done by means of a simple analog circuit.

The need of a cold junction compensation circuit entails an increased complexity of the measurement system. Moreover, the generated voltages are quite small (the sensitivity is around 41 μV/ C for a K-type thermocouple) and an amplifier is always necessary. However, thermocouples are used in industry very often since they are extremely rugged. They can also work reliably in a huge range of temperatures (from cryogenic temperatures to beyond 1700 C).

For this reason, compact integrated solutions exist and are sold by microelectronic industries. For example, we cite the AD8494-7 family, whose a fraction of the data-sheet is visible in figure 1.4. Those integrated circuits are able to amplify thermocouple signals while doing an internal compensation of the cold junction. The chip is able to measure its own temperature, in order to do the compensation. Of course, this works only if the real cold junction is located very close to the integrated circuit.

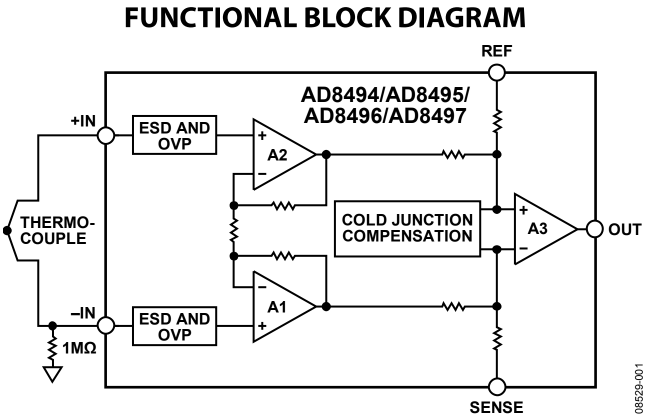


Figure 1.4: An extract of the data-sheet of Analog Devices AD8494-7 family. This device amplifies the thermocouple signal, compensating the cold junction temperature at the same time. ESD and OVP are the electrostatic and over voltage protections for input pins.

1.2.2.2 Thermocouple families

Various kinds of thermocouples are available on the market and are identified by a one letter code. The most frequently used families of thermocouples are:

E chromel (nickel-chrome)/constantan (copper-nickel)

K chromel (nickel-chrome)/alumel (aluminum-nickel)

J iron/constantan

S platinum-rhodium (10%)/platinum

R platinum-rhodium (13%)/platinum

B platinum-rhodium (30%)/platinum-rhodium (6%)

N nicrosil (nickel-chrome-silicon)/nisil (nickel-silicon)

T copper/constantan

In reality, more than a certain chemical composition, this letter indicates a well specified relation between temperature and voltage. The relation should be conformal to the ideal one within a known tolerance. The definition of the models and the tolerances is usually done by institutes of standards.

Typical thermocouple behaviours are shown in figure 1.5. Some thermocouples families (such as E or J) delivers a higher output signal, but they are affected by a stronger nonlinearity or they can be used only in a smaller range of temperature. To take into

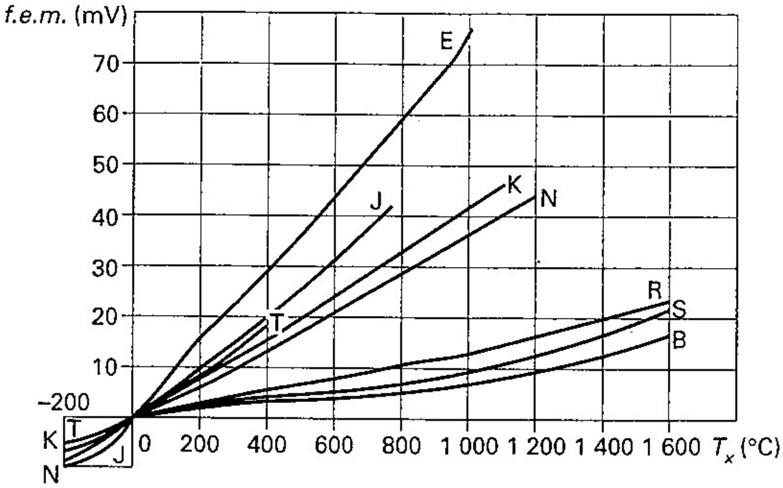


Figure 1.5: Relation between temperature and output voltage for some common thermocouple types. The cold junction is kept at 0 C. *Reproduced from [Asch, 2003].*

account the nonlinearities in a wide temperature range, the law $\mathcal{E}(T_x, 0\,\mathrm{C})$ is very often written as a polynomial equation, to which some exponential terms are added:

$$\mathcal{E}(T_x, 0\,\mathrm{C}) = \sum_{i=0}^{N} c_i T_x^i + a_0 \mathrm{e}^{a_1 (T_x - a_2)^2} \tag{1.1}$$

The National Institute of Standards and Technology (NIST) in USA publishes a catalogue of tables of coefficients to be adopted for the thermocouples cited above [NIST, ITS-90]. For instance, table 1.1 shows the coefficients useful for the K-type thermocouple in several temperature ranges.

Very often, for small temperature ranges (or when high accuracy is not sought), only the first linear term is kept into account in the calculations. This term is for short called sensitivity. For a K-type thermocouple, dealing with temperatures between 0 C and 100 C, the value $S \approx 41\,\mu\mathrm{V/\,C}$ may be considered. In this range, that simple strategy allows to obtain an accuracy of a few degrees.

1.2.2.3 pH measurement

Determining the pH of a solution is one of the most frequent characterisations employed when dealing with chemicals. It consists in the measurement of the acidity or basicity and it can be done using litmus paper, which dipped in a solution changes its colour following the pH. Another technique is based on the use of glass electrodes specifically built for this function. They contain a buffer solution which can exchange H^+ ions with the solution under test via a semi-permeable glass membrane. A voltage is obtained, proportional to the pH to be measured. The proportionality constant depends on the

	range: $-270\,\mathrm{C}$ to $0\,\mathrm{C}$	range: $0\,\mathrm{C}$ to $1372\,\mathrm{C}$
c_0	$0.000000000000 \times 10^{+0}$	$-0.176004136860 \times 10^{-1}$
c_1	$0.394501280250 \times 10^{-1}$	$0.389212049750 \times 10^{-1}$
c_2	$0.236223735980 \times 10^{-4}$	$0.185587700320 \times 10^{-4}$
c_3	$-0.328589067840 \times 10^{-6}$	$-0.994575928740 \times 10^{-7}$
c_4	$-0.499048287770 \times 10^{-8}$	$0.318409457190 \times 10^{-9}$
c_5	$-0.675090591730 \times 10^{-10}$	$-0.560728448890 \times 10^{-12}$
c_6	$-0.574103274280 \times 10^{-12}$	$0.560750590590 \times 10^{-15}$
c_7	$-0.310888728940 \times 10^{-14}$	$-0.320207200030 \times 10^{-18}$
c_8	$-0.104516093650 \times 10^{-16}$	$0.971511471520 \times 10^{-22}$
c_9	$-0.198892668780 \times 10^{-19}$	$-0.121047212750 \times 10^{-25}$
c_{10}	$-0.163226974860 \times 10^{-22}$	
	Exponential	
a_0	$0.118597600000 \times 10^{+0}$	
a_1	$-0.118343200000 \times 10^{-3}$	
a_2	$0.126968600000 \times 10^{+3}$	

Table 1.1: Table of coefficients to be used in equation (1.1) to calculate the output voltage (in millivolts) of a K-type thermocouple. Data published by [NIST, ITS-90].

exact configuration of the electrode, but a typical sensitivity is around $-60\,\mathrm{mV}$ for pH unit.

A difficulty of the pH measurement is that the series impedance of the glass electrode (called Z_c in the equivalent circuit of figure 1.1) is often very high, of the order of a few megaohms. For this reason, such probes may be equipped with an onboard amplifier very close to the measurement electrodes. Figure 1.6 shows part of the data-sheet of the CSIM-11 probe built by Campbell Scientific. The purpose of the amplifier is to buffer the signals at the probe output in order to reduce the perturbations on the cables.

A second difficulty is the strong temperature effect on the proportionality between the pH and the output voltage. Often, an automatic compensation must be done by measuring the temperature of the solution at the same time as the pH measurement. For example, concerning the CSIM-11 probe cited above, the sensitivity changes by $-0.2\,\mathrm{mV/pH/C}$. This means that at $20\,\mathrm{C}$ the slope is around $-58\,\mathrm{mV/pH}$, while it is equal to $-59\,\mathrm{mV/pH}$ at $25\,\mathrm{C}$ then reaching $-60\,\mathrm{mV/pH}$ at $30\,\mathrm{C}$. A calibration is hence required at regular intervals, using precision buffered solutions.

Body Material	ABS (Standard version) PPS / Ryton® (High Temperature version)
pH Range	0 to 14 pH
Output	±413 mV (±59 mV per unit pH @ 25ºC; 0 mV output at 7 pH)
Temperature Range	0°C to 80°C (Standard version) 0°C to +110°C (High Temperature version)
Pressure Range	30 psig (submerged). Mounting to pressurized pipes or tanks requires a non-refillable variation of the sensor. Speak with CSI for details.
Accuracy	±0.1% over full range
Sodium Error	Less than 0.05 pH in 0.1 Molar Na^+ ion @ 12.8 pH
Impedance	Less than 1 Mohm @ 25°C
Reference Cell	KC1/AgC1
Zero Potential	7.0 pH ±0.2 pH
Wetted Materials	ABS or PPS (Ryton®), Teflon®, Viton®, Glass
Response Time	95% of reading in 10 seconds
Drift	Less than 2 mV per week
Mounting	Threading at either end of probe is 3/4" NPT male
Power	Internally powered by two 3-volt lithium batteries

Figure 1.6: Extract from the data-sheet of the pH probe CSIM-11 from Campbell Scientific. This probe is powered by a battery, since there is an onboard amplifier.

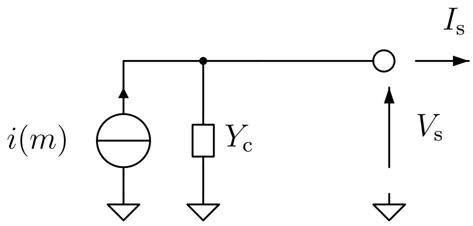


Figure 1.7: Norton equivalent circuit of a current generator sensor.

1.3 Current generator sensors

1.3.1 General description

When a physical action induce the generation of charge carriers in a material, this phenomenon may result in a variation of a current flowing in the device. The electrical output of a sensor is thus a current. Some examples can be seen:

- radiation induced ionisation effects
- carriers generation by photo-electrical effect.

As the information on the measurand is carried by a current, it is natural to adopt an equivalent circuit representation which is a Norton equivalent, as represented in figure 1.7. The measurand m is translated to a certain current $i(m)$ which flows through the terminals of the generator. The internal admittance of the sensor is given by Y_c. Of course, the model shown in figure 1.7 is greatly simplified and might not be able to represent the real behaviour of the sensor when V_s exceeds particular limits: nonlinearity often lurks around the corner. We discuss some examples of sensors of this kind.

1.3.2 Examples

1.3.2.1 Photomultipliers

When a photon impinges on a conductor, if the energy carried is high enough, it can extract an electron which becomes a free carrier. This effect is called external photo-electrical effect and it has been discovered in XIX century and studied by A. Einstein in one of his seminal papers [Einstein, 1905] published in his *annus mirabilis* 1905. By exploiting this principle, a light sensor can be built by applying an electrical field to move the carriers, while monitoring the current circulating in the system. The current will be in fact proportional to the flux of photons in the unit of time, related to the

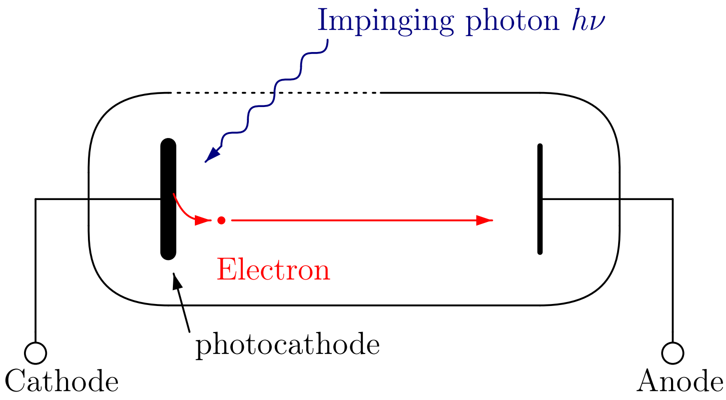


Figure 1.8: Working principle of a sensor based on the photo-electrical effect.

light intensity. An important condition is however that the energy of the photons is high enough to enable the photo-electrical effect.

Figure 1.8 shows a schematic view of a device which can be used to measure light via the photo-electric effect. An important value quantifying the overall quality of the device is called the *quantum efficiency* η, i.e. the ratio between the number of generated electrons and the impinging photons. A second parameter is the *sensitivity* R, the ratio between the generated current I_{ph} and the impinging light power P. The measurement unit of this parameter is thus A/W.

In practice, generated currents are often quite small and handling them might become tricky. The presence of a current amplification internal to the sensor itself might simplify the task of detecting very low optical intensities. This is done in photomultipliers by adding a number of intermediate electrodes (dynodes) to multiply the number of electrons by exploiting the secondary emission of electrons. A single photon thus results in a significant number of electrons thanks to the amplification process. This principle is represented in figure 1.9. Dynodes are often biased by a resistive network from the photomultiplier power supply rails.

An extract of the data-sheet of a multiplier tube can be found in figure 1.10. It detects light in the visible range of wavelengths and it is thus sensitive between 300 nm and 850 nm. The photocathode is followed by ten gain stages, and the electron gain depends exponentially on the bias voltage, adjusting the sensitivity of the sensor.

Photomultipliers are able to generate an event for just one photon impinging, even if not all photons will trigger it. They are relatively bulky, fragile and they need high voltages, yet photomultipliers are still in use today especially when a very high sensitivity and low noise are not an option.

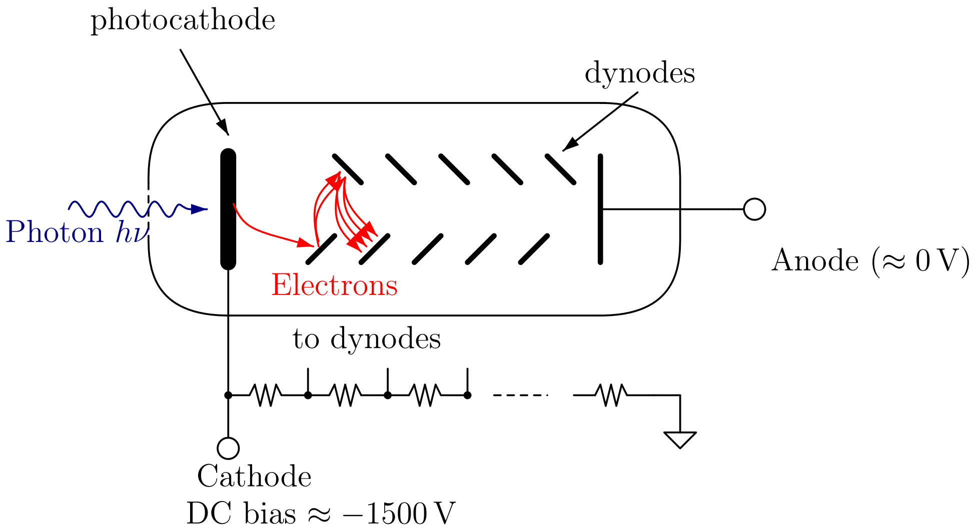


Figure 1.9: Working principle of a photomultiplier.

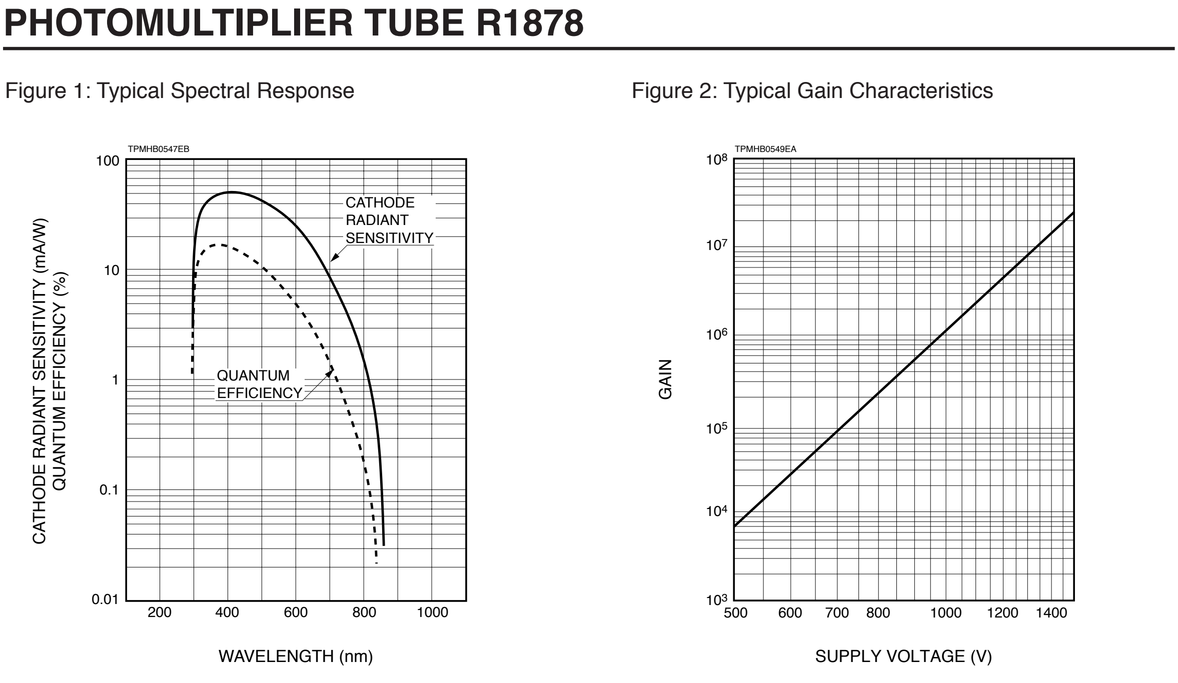


Figure 1.10: Extract of the data-sheet of a R1878 photomultiplier tube produced by Hamamatsu. On the left, the quantum efficiency as well as the responsivity of the photocathode versus the wavelength. On the right, the gain versus the bias voltage. Note the vertical logarithmic scale.

1.3.2.2 Photodiodes

A semiconductor is characterised by the presence of energy bands where carriers move reacting to an electric field. Electrons in valence band might be brought in the conduction band following the absorption of a photon having enough energy $E = h\nu$. This only happens if $E > E_g$, where E_g is the energy gap of the semiconductor. This is called *internal photoelectric effect*, in a similar way as we have already described in paragraph 1.3.2.1. An absorbed photon entails the generation of a pair of carriers, an electron in the conduction band and a hole in the valence band.

In a PN junction, the internal field allows to separate the electron/hole pair and this results in a certain current. This current is proportional to the number of generated carriers per unit of time. That is at its turn proportional to the number of absorbed photon flux in the junction, hence the absorbed optical power. The proportionality constant R is called responsivity or sensitivity and it is defined exactly as in photomultipliers:

$$R = \frac{I_{\mathrm{ph}}}{P}, \tag{1.2}$$

where I_{ph} is the photocurrent and P the optical power impinging on the device. The electrical symbol of a photodiode is shown in figure 1.11. The figure shows also how the current/voltage characteristics of the diode is vertically translated when the junction receives a certain flux of photons and a photocurrent appears.

Hugely different applications exist, ranging from light power measurements to colorimetry, detection of telecom signals in optical fibers and so on.

Figure 1.12 shows an extract of the data-sheet of a silicon photodiode. The silicon band gap is $E_g = 1.12\,\mathrm{eV}$ and this means that photons having wavelength longer than 1100 nm are practically not absorbed since they do not have enough energy to generate a carrier pair. The sensitivity is thus dramatically reduced when the impinging light has a longer wavelength. Nonlinear effects still exist above that limit, involving simultaneous absorption of two or three photons but play a role only for very intense light.

In the equivalent circuit shown in figure 1.11 there are two current generators. The first one I_{ph} is associated to the photocurrent obtained by the internal photo-electric effect. The second generator I_s represents the dark current, due to the minority carrier drift across the junction when it is reversely biased. We then have the internal resistance (ranging from several mega ohms to a few giga ohms) as well as the junction capacitance. In some applications, the capacitance C_d might become a limiting factor for the detection speed or the stability of the conditioning circuit.

A particular class of devices are the avalanche photodiodes. In this case, the carriers generated in the junction by a photon absorption are multiplied thanks to an avalanche effect in a strongly reversely biased device. Like photomultipliers, there is thus an internal gain of the photodiode which is precious for the detection of very low light intensities.

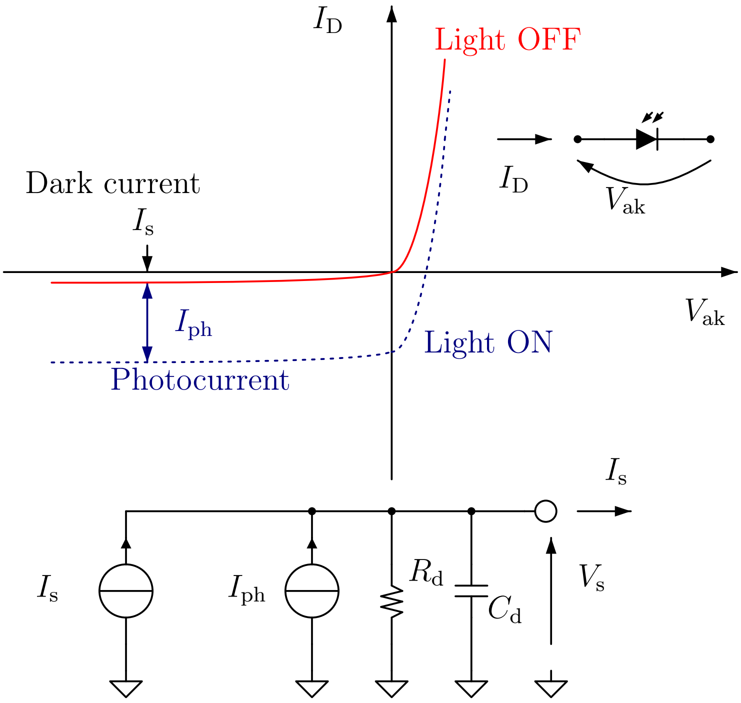


Figure 1.11: Electrical symbol of a photodiode, its typical I/V characteristics in obscurity and with light impinging, as well as its equivalent circuit.

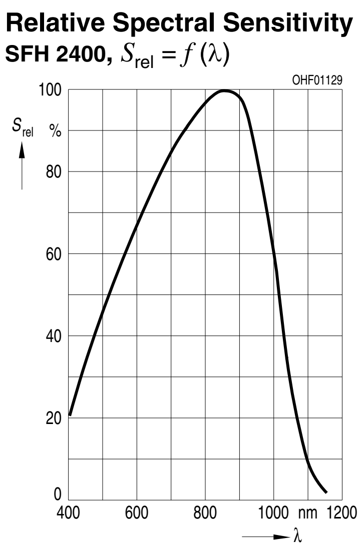


Figure 1.12: Sensitivity versus the wavelength of the SFH2400 silicon photodiode produced by Osram.

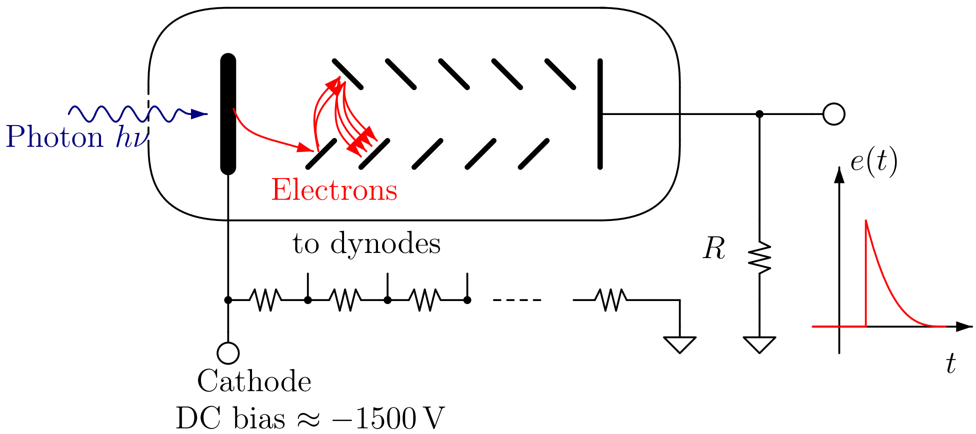


Figure 1.13: Conditioning of the current signal coming from a photomultiplier via the resistance R.

1.3.3 Conditioning circuits

A conditioning circuit converts the electrical signal at the output of the sensor into a voltage. Therefore, in this paragraph we deal with current to voltage converters. Several solutions exist, here we discuss just the simplest and most current ones.

1.3.3.1 One resistance

Sometimes, just a resistance in a strategic place of the circuit can do the job. This solution is often used with photomultipliers and an example can be seen in figure 1.13, where the voltage signal $e(t)$ is obtained via the resistance R.

A problem which might appear is when the voltage at the output of the circuit affects the biasing of the sensor, thus producing some nonlinear effects. In fact, this solution is also used sometimes with photodiodes, as shown in figure 1.14: a nonlinearity appears when the voltage $e(t)$ at the circuit's output is close to the threshold voltage of the diode. In this condition, the photodiode becomes directly biased and the voltage at its terminals is not anymore in a linear relation with the power of light impinging on the device. However, the method is quite simple and useful in radio frequency applications, where $50\,\Omega$ is used, eventually by superposing a reverse bias in order to reduce nonlinearity with strong light intensities.

1.3.3.2 Transresistance amplifier

In paragraph 1.3.3.1, we saw that one of the problems associated to the signal conditioning done with a simple resistance is that the sensor itself is subjected to the same voltage obtained at the output of the conditioning circuit. In some cases, the biasing of

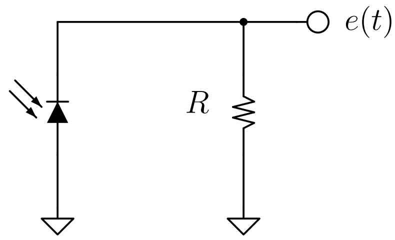


Figure 1.14: Conditioning of the current signal coming from a photodiode via the resistance R.

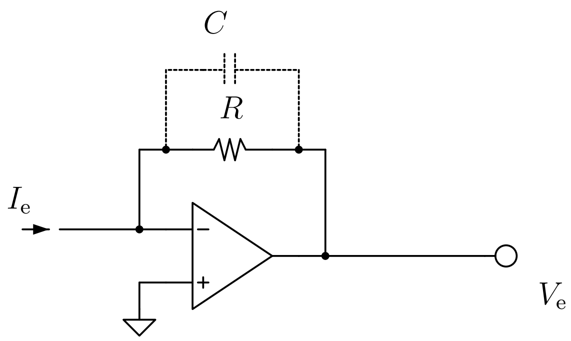


Figure 1.15: Classical circuit of a trans-resistance amplifier built around an operational amplifier.

the sensor might be perturbed, giving rise to non-linear effects. It would be better to make sort that V_s is always close to zero, and the current I_s gives rise to a proportional voltage *elsewhere* in the circuit, on nodes different from the terminals of the sensor.

A classical solution is the trans-resistance circuit shown in figure 1.15, often called "current to voltage converter". An advantage of the circuit is that, thanks to the properties of the operational amplifier, the voltage across the sensor is very close to zero, thus reducing the influence of the Y_c admittance in the equivalent circuit of fig. 1.7. In fact, if the operational amplifier is ideal, the feedback is provided by the R resistance and the output voltage of the circuit is:

$$V_e = -RI_e \tag{1.3}$$

If the goal is to obtain a very high trans-resistance, for example to detect very small currents, R should be very high. Values of resistance up to several gigaohms can be found in catalogues, but they tend to be very expensive and have to be handled with care. The circuit shown in figure 1.16 represents a classical workaround, avoiding extreme values of a single resistance. Circuit analysis leads to:

$$V_e = -\frac{R_1R_2 + R_2R_3 + R_1R_3}{R_3} I_e \tag{1.4}$$

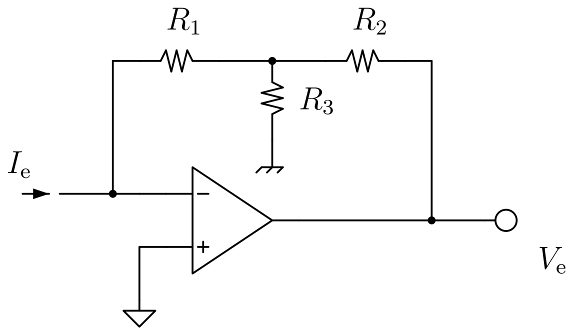


Figure 1.16: Use of a T-bridge feedback circuit in the current to voltage converter.

The proportionality term might thus be seen as equivalent to a very high value resistance. There are however some problems associated to this circuit, as the increased noise if compared to one resistance, as well as an increase of the offsets.

In the circuits shown in figures 1.15 and 1.16, some key points should be taken into account:

- Stability: sensors may possess a relatively large output capacitance, as we have seen for photodiodes (the C_d capacitor in the equivalent circuit). In this case, real world operational amplifiers might experience instability in the circuits proposed above. This is very well known and it is discussed in application notes [TI SBOA055A] or books [Franco, 2015] on the subject. A common solution consists in reducing the bandwidth of the circuit by introducing a dominant pole via the C capacitor, in parallel with the resistance R of figure 1.15.

- If small currents should be detected, the bias currents of operational amplifiers have to be very small too. MOS-based devices are available on the market with astonishingly low bias currents. For example, the LMC6001 declares in the data sheets a bias current of 25 fA at 25 C. As a second example, we cite the LMP7721, with the guaranteed maximum of 20 fA at 25 C, with a typical value of 3 fA. This is rather breathtaking if we observe that 1 fA represents something like 6000 electrons per second. Application note [TI SBOA061], from Burr-Brown/Texas Instruments, discusses some specific caveats of low current measurements with their OPA128.

- In a circuit, when currents of less of a nanoamper are to be treated, moist and dirt on the PCB play a role which is no longer negligible. Specific techniques must be adopted in the most delicate parts of the circuits (guard rings, shielding, teflon sockets or insulators. . .). Prepare yourself to wear gloves to handle the most delicate components and work as clean as possible.

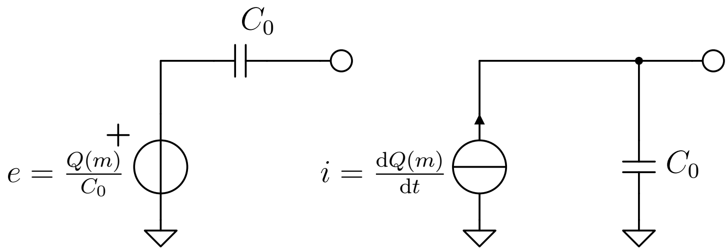


Figure 1.17: Equivalent circuits for modelling charge generating sensors.

1.4 Charge generator sensors

1.4.1 General description

Some sensors are based on the change of dielectric polarisation on a dielectric, in response to external stimuli. Examples can be cited:

- In the piezo-electric materials such as quartz crystals or specific ceramics and polymers, a mechanical deformation due to an applied force entails the appearance of an electric field. Therefore, if the sensor is inserted in a circuit, charges move in order to counterbalance the field.
- Small-scale rearrangements of the structure of some dielectrics occur when the temperature changes, giving rise to an electric field. This effect is called pyroelectricity. Examples include triglycine sulphate crystals, tourmaline...

In those situations, the measurand is translated into a certain charge unbalance from the equilibrium condition (so the charge is not *actually* generated inside the sensor). Figure 1.17 represents two different ways of modelling this family of sensors. The first circuit on the left is a Thévenin equivalent, which is completed sometimes by a resistance in parallel with the C_0 capacitor, to take into account internal sensor losses. The second equivalent circuit on the right is a Norton equivalence. The choice of the former or the latter equivalent circuit is often a matter of convenience, but both reflect the impossibility of performing DC measurements.

1.4.2 Examples

1.4.2.1 Pyroelectric sensor for infrared sensing

Probably the most widespread use of pyroelectric sensors is for thermal infrared detection. Motion detectors inside buildings are often based on the detection of sources of heat by a pyroelectric sensor. They are often used to switch on the light when a person enters a certain area, or to signal an intrusion. Figure 1.18 shows a typical sensor able to detect infrared radiation whose wavelengths are caught between 5 µm and 12 µm (in the thermal range).

It is interesting to remark the presence of a field effect transistor inside the sensor's package, to amplify signals coming from the sensor. A complete motion detector system usually couples the pyroelectric sensor with an amplifier/threshold circuit, a plastic Fresnel lens, and often a timer.

1.4.2.2 Piezoelectric sensors

The application of a force on a piezoelectric material entails the change of its polarisation. If the material is applied between two electrodes in an arrangement similar to a capacitor, a voltage appears between the two terminals of the sensor. This is due to a non equilibrium charge proportional to the applied mechanical constraint. This principle is reversible and a wide variety of sensors and actuators based on this phenomenon are of everyday use.

Some examples are: microphones, accelerometers, displacement sensors… We cite an interesting musical application: the percussion sensor in the pads of electronic drums is often piezoelectric.

1.4.3 Conditioning

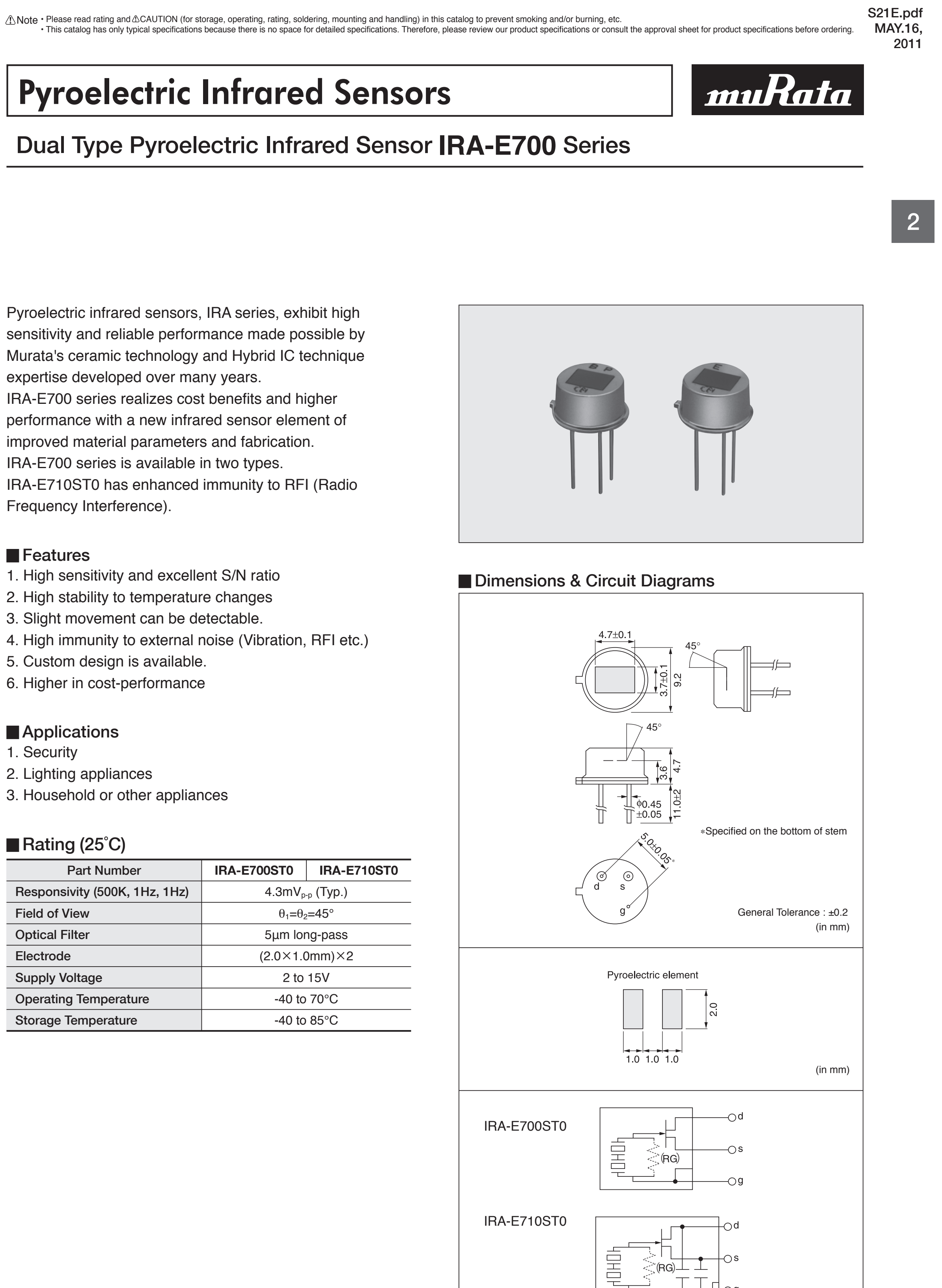
A variety of techniques exists for the conditioning of electrical signals coming from charge-based sensors (see for example [TI SLOA033A] for a discussion specific to piezoelectric sensors).

An idea which might be kept in mind is that if a capacitor is storing a certain charge, there is a resulting voltage, and vice-versa. This can be exploited at the sensor itself, by making it work in a open circuit, via a very high impedance amplifier. By using the Thévenin equivalent circuit shown in figure 1.17, it can be seen that the internal capacitor C_0 is charged by a certain voltage which can be measured if some care is taken not to discharge too much the capacitor during measurements.

The second possibility is to use the circuit shown in figure 1.19, where an operational amplifier is used as an integrator for the charge provided by the sensor.

Some circuit analysis done by assuming an ideal the operational amplifier leads to the observation that no current is circulating in capacitors C_0 (sensor capacitance) and C_c (cable/connection capacitance), since in this case, the sensor is connected to a virtual short circuit. In the same way, the resistance R_c representing the losses plays no role. All the provided charge is stored in the C capacitor, in the feedback loop around the operational amplifier. The voltage at the output of the circuit will thus be:

$$V_\text{s} = -\frac{Q(m)}{C} \tag{1.5}$$



⚠Note • Please read rating and ⚠CAUTION (for storage, operating, rating, soldering, mounting and handling) in this catalog to prevent smoking and/or burning, etc.
• This catalog has only typical specifications because there is no space for detailed specifications. Therefore, please review our product specifications or consult the approval sheet for product specifications before ordering.

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Pyroelectric Infrared Sensors

muRata

Dual Type Pyroelectric Infrared Sensor **IRA-E700** Series

2

Pyroelectric infrared sensors, IRA series, exhibit high sensitivity and reliable performance made possible by Murata's ceramic technology and Hybrid IC technique expertise developed over many years.
IRA-E700 series realizes cost benefits and higher performance with a new infrared sensor element of improved material parameters and fabrication.
IRA-E700 series is available in two types.
IRA-E710ST0 has enhanced immunity to RFI (Radio Frequency Interference).

■ Features

1. High sensitivity and excellent S/N ratio
2. High stability to temperature changes
3. Slight movement can be detectable.
4. High immunity to external noise (Vibration, RFI etc.)
5. Custom design is available.
6. Higher in cost-performance

■ Applications

1. Security
2. Lighting appliances
3. Household or other appliances

■ Rating (25°C)

Part Number	IRA-E700ST0	IRA-E710ST0
Responsivity (500K, 1Hz, 1Hz)	$4.3mV_{p\text{-}p}$ (Typ.)	
Field of View	$\theta_1=\theta_2=45°$	
Optical Filter	5µm long-pass	
Electrode	(2.0×1.0mm)×2	
Supply Voltage	2 to 15V	
Operating Temperature	-40 to 70°C	
Storage Temperature	-40 to 85°C	

■ Dimensions & Circuit Diagrams

Figure 1.18: Extract from the data-sheet of a IRA-E700 pyroelectric sensor built by Murata.

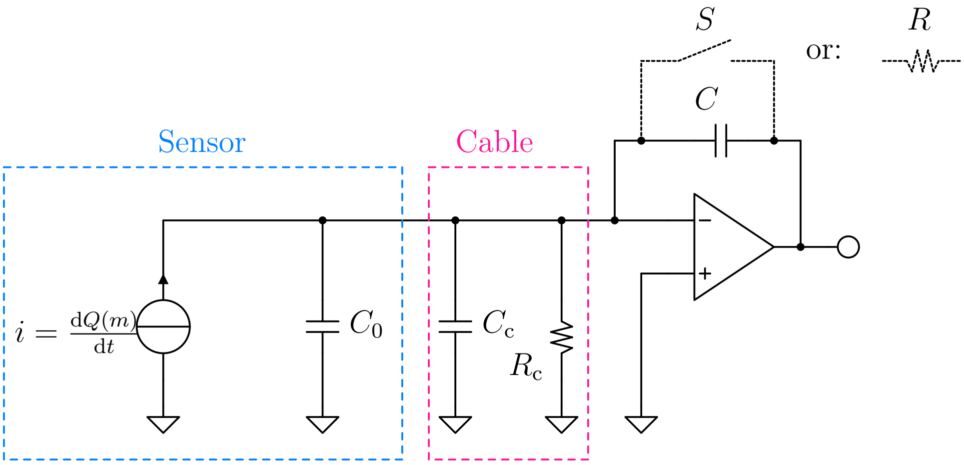


Figure 1.19: Conditioning circuit useful for charge-based sensors.

Something which must be taken into account in this circuit is that there is no DC feedback path (i.e. in the absence of S and R in the schematic of fig. 1.19). Once the circuit is switched on, the C capacitor will slowly charge itself with the bias currents of the operational amplifier. This means that after a while, V_s might become close to the power supply rails of the amplifier, thus giving rise to its inevitable saturation.

A first strategy to avoid this phenomenon is to place a switch (more realistically, a field effect transistor operating as a switch in response to a control signal) in parallel with the C capacitor. The purpose of the switch is to discharge it at the beginning of each measurement.

A second solution is to put a resistance R in parallel with C. A Laplace-domain analysis yields:

$$V_s(p) = -\frac{Q(p)}{C}\frac{pRC}{1+pRC}, \tag{1.6}$$

which is after all quite similar to equation (1.5), where we have a term which has a high-pass behaviour, with a cutoff frequency:

$$f_c = \frac{1}{2\pi RC}. \tag{1.7}$$

As long as R and C are chosen in such a way that the unattenuated band is compatible with the useful signal, this circuit will work reliably.

1.5 Resistive sensors

A lot of physical phenomena are closely related to the electrical resistance of a conductor or a semiconductor. In fact, every times something modifies the following characteristics,

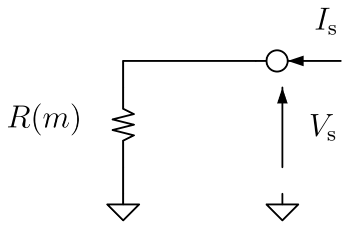


Figure 1.20: Equivalent circuit of a resistive sensor: a resistance whose value is dependent on measurand m.

there is a change in the resistivity:

- carrier mobility (temperature, constraint, magnetic field)
- carrier density (temperature, light absorption)
- geometrical dimensions (constraint, cursor displacement)

A wide class of sensors exploits those effects. Their equivalent circuit is shown in figure 1.20. We thus present some examples in the following paragraphs.

1.5.1 Examples

1.5.1.1 Light dependent resistors (LDR)

If we consider a generic semiconductor, resistivity is strongly dependent on the carrier concentration (electrons and holes) participating to the conduction. In an intrinsic semiconductor, carrier density is quite low, and the resistivity is comparatively high. If the semiconductor is subject to a photon flux, if their energy is sufficient to generate electron/hole pairs, a consequent reduction of the resistivity can be observed.

It is thus possible to obtain a device whose resistance is strongly dependent on the light intensity to which is subjected, also called a photo-resistance or Light Dependent Resistors (LDR). Devices based on cadmium sulphide (CdS cells) are widely used for a lot of applications involving visible light since they are simple to fabricate, low-cost and rugged. Figure 1.21 shows the resistance versus light intensity plot of a Silonex NORPS-12 LDR. It can be noticed that the device might be used on a quite large dynamic range, such as 4 decades, its resistance in complete dark being a few megaohms.[1]

[1]The light intensity is measured in the old foot-candles units; the correct unit in the SI standards is the lux. The conversion factor is 1 foot-candle = 10.764 lux. The multipliers are also nonstandard…

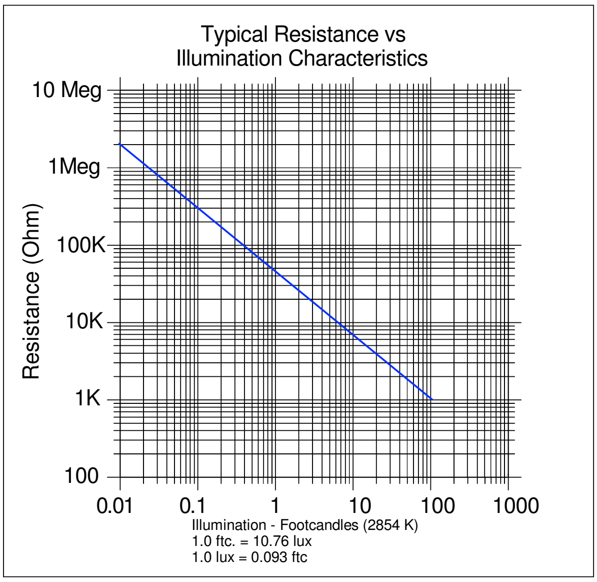


Figure 1.21: An extract of the data-sheet of the NORPS-12 LDR built by Silonex.

1.5.1.2 Platinum temperature probe "Pt100"

A "Pt100" probe is formed by a given length of platinum wire on a glass or ceramic insulating support. Its resistance R is $R_0 = 100\,\Omega$ at a temperature of $t_x = 0\,\mathrm{C}$ and the temperature dependance is well represented by the following equation:

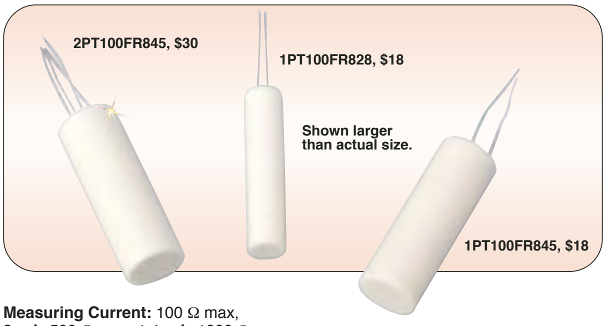
$$R(t_x) = R_0(1 + \alpha t_x) \tag{1.8}$$

where $\alpha = 3.85 \times 10^{-3}\,\mathrm{C}^{-1}$, the *average* nominal temperature coefficient in the range between 0 C to 100 C. The linear model of equation (1.8) allows to obtain an error of less than 0.5 C in that interval of temperatures. As we saw about thermocouples in paragraph 1.2.2.2, more complex models can be employed. The IEC 60751 norm (2008) specifies the relation between temperature and electrical resistance and defines two tolerance classes (A and B) depending on the desired precision.

A similar principle is used on variants such as the Pt500 or Pt1000, the only notable difference being that in equation (1.8) R_0 is respectively $500\,\Omega$ and $1000\,\Omega$.

Figure 1.22 shows the data-sheet common to several sensors Pt100, Pt500 and Pt1000. The tolerance class "B" is clearly indicated as well as the proportionality factor α (called TCR, ppm means "parts per million") to be used in equation (1.8).

Platinum Resistance Temperature Detector Sealed with Cement in a Cylindrical Ceramic Body



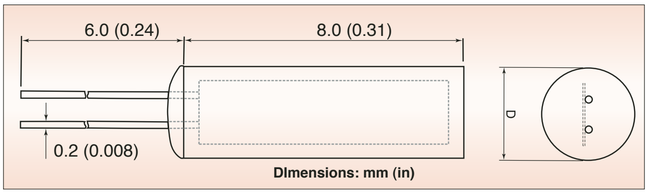
FR Series Starts at $18

FR Series elements are designed for applications where high vibration resistance as well as high temperature stability are vital. Typical industrial applications include analytical equipment, chemical plants, and mechanical equipment. Small tolerances on diameters allow problem-free installation in protective tubes.

Specifications

Tolerance: IEC751, Class B
Nominal Resistance: 100 Ω, 500 Ω and 1000 Ω at 0°C (32°F)
Temperature Range (Continuous Operation): -70 to 500°C (-95 to 930°F)
Temperature Coefficient: TCR = 3850 ppm/K
Leads: Platinum-clad nickel wire; 6.0 mm L (0.24")
Long-Term Stability: Max $R/_0$-drift 0.1% after 1000 h at 500°C (930°F)
Measuring Current: 100 Ω max, 3 mA; 500 Ω max, 1.4 mA; 1000 Ω max, 1 mA (self-heating has to be considered)
Vibration Resistance: At least 40 g acceleration at 10 to 2000 Hz; at least 100 g acceleration with 8 ms half sine wave
Environmental Conditions: Unhoused for dry environments only
Insulation Resistance: >10 MΩ at 20°C (70°F) >1 MΩ at 500°C (930°F)

Discount Schedule	
1 to 4 units	Net
5 to 10 units	4%
11 to 24 units	6%
25 to 49 units	10%
50 to 99 units	11%
100 and up	13%



MOST POPULAR MODEL HIGHLIGHTED!

Type Model No.	Nominal Resistance (Ω)	"D" Dimensions* mm (in)	Self Heating 0°C (K/mW)	Response Time in Seconds: Water V = 0.4 m/s 50%	Water V = 0.4 m/s 90%	Air V = 1 m/s 50%	Air V = 1 m/s 90%	Price
1PT100FR828	1 x 100	2.8 (0.11)	0.05	0.9	2.7	12.3	39.5	**$18**
1PT1000FR828	1 x 1000	2.8 (0.11)	0.05	0.9	2.7	12.3	39.5	**18**
2PT100FR828	2 x 100	2.8 (0.11)	0.16	0.9	2.7	12.3	39.5	**22**
2PT1000FR828	2 x 1000	2.8 (0.11)	0.16	0.9	2.7	12.3	39.5	**22**
1PT100FR845	1 x 100	4.5 (0.18)	0.04	1.5	4.6	24.8	78.8	**18**
1PT500FR845	1 x 500	4.5 (0.18)	0.04	1.5	4.6	24.8	78.8	**18**
1PT1000FR845	1 x 1000	4.5 (0.18)	0.04	1.5	4.6	24.8	78.8	**18**
2PT100FR845	2 x 100	4.5 (0.18)	0.08	1.5	4.6	24.8	78.8	**28**
2PT1000FR845	2 x 1000	4.5 (0.18)	0.08	1.5	4.6	24.8	78.8	**30**

The measuring point for the basic value is situated 8 mm from the end of the sensor body.
** **Note:** Dimensions are ±0.3 mm.*
Ordering Examples:** 12 **1PT100FR828** less 6% quantity discount = $18 x 12 x 0.94 = **$203.04.
2PT1000FR845,** 2 x 1000 Ω, 4.5 mm Dia RTD, **$30.

C-89

Figure 1.22: Data-sheet of a family of platinum temperature sensors, from Omega engineering. Note the self-heating data provided.

1.5.1.3 Strain gauges

Strain gauges (someone writes "strain gages"[2]) are based on the observation that a conductor subjected to a mechanical constraint tends to vary its electrical resistance. For example, we observe that a homogeneous conductor wire pulled with a certain force F will probably tend to increase its length l while reducing slightly its section S. This phenomenon is reversible, at least if the force F does not exceed the limits for the elastic deformation of the wire. The electrical resistance of the wire might be calculated via the well known equation:

$$R = \rho \frac{l}{S}, \tag{1.9}$$

where ρ is the resistivity of the material employed, which also depends from the strain applied (piezoresistive effect). The change of R is usually represented as follows:

$$\frac{\Delta R}{R} = K \frac{\Delta l}{l} \tag{1.10}$$

where the proportionality coefficient K is called gauge factor and might vary between 2 and 4 form most metals and 50 and 200 in module for semiconductors.

1.5.2 Caveats

When one needs to measure a resistance, several factors should be carefully considered:

- Resistance is always associated to heat generated by Joule effect. The electrical power employed for the measurement must therefore be controlled, to avoid self-heating effects which might induce an error on the measurements.
- The connection wires do not always have a negligible resistance.
- Sensors in an industrial environment might operate close to electrically noisy machines, giving rise to electromagnetic compatibility issues.
- Electrical insulation might be a delicate issue when dealing with harsh environments (high temperatures, aggressive chemicals, high voltages, radioactivity...)

1.5.3 Signal conditioning: measuring the total resistance $R(m)$

Two different situations must be considered:

- Information about the measurement is carried by the total resistance $R(m)$.

[2] *What's in a name? that which we call a rose*
By any other name would smell as sweet
W. Shakespeare, Romeo and Juliet

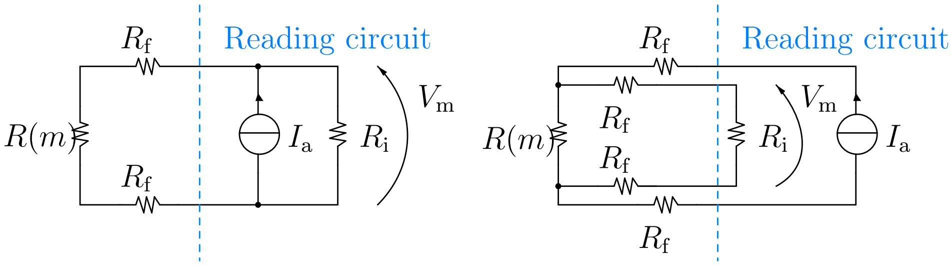


Figure 1.23: Conditioning strategies for a resistive sensor, measure of the total resistance $R(m)$. On the left: 2-wire measurement. On the right: 4-wire (Kelvin) measurement. R_f is due to cables and connections, R_i is the internal resistance of the voltmeter.

- Information is carried by a variation of the resistance around a certain mean value R_0, so that: $R = R_0 + \Delta R(m)$

In this paragraph, we discuss about the first case, where the information about the measurand m is translated by the sensor in a resistance value $R(m)$ to be determined. For example, it is the case of the platinum sensor described in paragraph 1.5.1.2. A simple solution might be to excite the sensor with a known current I_a: the voltage V_m obtained at the output is hence related to the excitation current by the Ohm's law:

$$V_\mathrm{a} = R(m) I_\mathrm{m} \tag{1.11}$$

In reality, the sensor might be placed quite far from the conditioning circuit and the connection resistances R_f might play a non negligible role in the measurement, as shown in figure 1.23 on the left. What it can be accessed in this configuration is just the resistance $R(m) + 2R_\mathrm{f}$. In fact, R_f might be difficult to control and dependent of a lot of factors (ageing, temperature, nature of the conductors, splices, connectors…). This kind of measurement is called “two wire measurement” to be distinguished from the more complex “four-wire measurement” which we are about to describe and that is depicted in figure 1.23 on the right. Most of the cases, the voltmeter resistance R_i might be sufficiently high to be considered infinite. With electronic voltmeters, it is not uncommon for R_i to range well above $10\,\mathrm{M\Omega}$.

The idea at the background of the 4-wire technique (also called “Kelvin contact”) is that the wires used to measure voltage carry almost no current, whereas the wires carrying the excitation current I_m are not involved in the voltage measurement. The measured voltage V_m becomes thus virtually independent from R_f, allowing to determine $R(m)$ very precisely.

1.5.4 Measuring a resistance variation: the Wheatstone bridge

When the goal is to measure a resistance variation around an equilibrium point (this is often the case when dealing with strain gauges discussed at the paragraph 1.5.1.3), a classic circuit is the Wheatstone bridge, shown in figure 1.24. It traces its origins in the first half of the nineteenth century and it was originally employed to measure an unknown resistance by carefully adjusting other (known) resistances. The principle was (and still is) to measure the voltage difference between nodes A and B in the bridge circuit. The voltage V_{AB} is equal to zero only for a perfectly balanced bridge, condition which corresponds to the following equation:

$$R_3 R_2 = R_1 R_4. \tag{1.12}$$

In order check wether or not this condition is satisfied, it was not needed to measure V_{AB} with precision. That was technically feasible even at the time when the bridge was invented; instead, at the time measuring a voltage with precision was a real challenge. In modern days, however, we have wonderfully linear instruments and it is far more convenient to measure V_{AB} and to explicit its relation with the values of the resistances. This is the path we will follow in the rest of this paragraph.

Two kinds of excitation are usually adopted:

- Constant voltage excitation, where the bridge is connected to a voltage generator E. The current flowing in the bridge varies during the measure because the total bridge resistance changes. The output voltage V_{AB} is:

$$V_{\mathrm{AB}} = \frac{R_2 R_3 - R_1 R_4}{(R_1 + R_2)(R_3 + R_4)} E. \tag{1.13}$$

- Constant current excitation, where the current flowing in the bridge is kept stable to a certain value I_{a}. The output voltage is as follows:

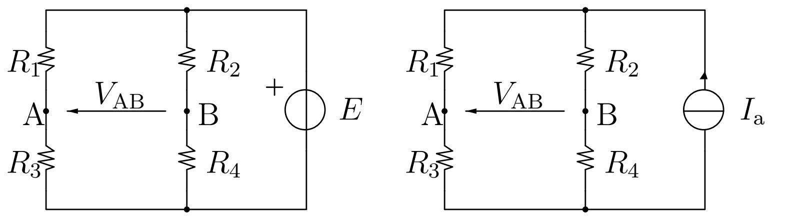
$$V_{\mathrm{AB}} = \frac{R_2 R_3 - R_1 R_4}{R_1 + R_2 + R_3 + R_4} I_{\mathrm{a}}. \tag{1.14}$$

Despite (or thanks to) its simplicity, the Wheatstone bridge, and its close relatives, is one of the most elegant solutions to a surprisingly wide range of problems [Williams, 1990]. In practical measurement situations, different variants are exploited, depending on the number of sensing elements in the bridge. In the following paragraphs, we will discuss the most useful configurations.

1.5.4.1 A single variable element

In the bridge shown in figure 1.24, just one of the four resistances is variable, for example R_2. We may write:

$$\begin{cases} R_2 = R_0 + \Delta R_2(m) \\ R_1 = R_3 = R_4 = R_0 \end{cases}. \tag{1.15}$$



Constant voltage excitation Constant current excitation

Figure 1.24: Wheatstone bridges, constant voltage and current excitations.

With a constant voltage excitation, by exploiting equation (1.13), we find:

$$V_{\mathrm{AB}} = \frac{\Delta R_2(m)}{R_0} \times \frac{1}{1 + \frac{\Delta R_2(m)}{2R_0}} \times \frac{E}{4}. \tag{1.16}$$

It can be noted that V_{AB} voltage is nonlinearly dependent from $\Delta R_2(m)$ and this might be considered a disadvantage in some situations. However, if $\Delta R_2(m)$ is small enough, the following simplification can be carried out:

$$V_{\mathrm{AB}} \approx \frac{\Delta R_2(m)}{R_0} \times \frac{E}{4}. \tag{1.17}$$

For a constant current excitation of the bridge, we obtain:

$$V_{\mathrm{AB}} = \Delta R_2(m) \times \frac{1}{1 + \frac{\Delta R_2(m)}{4R_0}} \times \frac{I_{\mathrm{a}}}{4} \approx \Delta R_2 \frac{I_{\mathrm{a}}}{4}, \tag{1.18}$$

which once again reflects a nonlinear relation between V_{AB} and ΔR_2 and the linearisation is allowable when the bridge unbalance is small. For the constant current excitation, however, the nonlinear term is twice smaller than for a constant voltage excitation.

1.5.4.2 Two variable elements

We insert two variable elements in the bridge, as follows:

$$\begin{cases} R_1 = R_0 + \Delta R_1(m) \\ R_2 = R_0 + \Delta R_2(m) \\ R_3 = R_4 = R_0 \end{cases} \tag{1.19}$$

The non null voltage V_{AB} can thus be calculated:

- constant voltage excitation:

$$V_{\mathrm{AB}} = \frac{\Delta R_2(m) - \Delta R_1(m)}{R_0} \times \frac{1}{1 + \frac{\Delta R_2(m) + \Delta R_1(m)}{2R_0}} \times \frac{E}{4} \tag{1.20}$$

- constant current excitation:

$$V_{\mathrm{AB}} = [\Delta R_2(m) - \Delta R_1(m)] \times \frac{1}{1 + \frac{\Delta R_2(m) + \Delta R_1(m)}{4R_0}} \times \frac{I_{\mathrm{a}}}{4} \tag{1.21}$$

We can see that both expressions obtained above might be simplified if a certain symmetry is respected. In fact, it $\Delta R_2 = -\Delta R_1 = \Delta R(m)$ (the so called "push-pull" configuration), we obtain:

- constant voltage:

$$V_{\mathrm{AB}} = \frac{\Delta R(m)}{R_0} \times \frac{E}{2} \tag{1.22}$$

- constant current:

$$V_{\mathrm{AB}} = \Delta R(m) \frac{I_{\mathrm{a}}}{2}. \tag{1.23}$$

An interesting point is that now the last expressions are perfectly linear and the sensitivity is twice as the case where one sensing element is present in the bridge. Very often, this configuration is adopted in cases where symmetrical resistance variations are due to a carefully tailored mechanical symmetrical arrangement.

1.5.4.3 Four variable elements

We consider a symmetric situation similar to the "push pull" strategy seen in the previous paragraph, but extending the same concept to four variable resistances:

$$\begin{cases} R_1 = R_0 - \Delta R(m) \\ R_2 = R_0 + \Delta R(m) \\ R_3 = R_0 + \Delta R(m) \\ R_4 = R_0 - \Delta R(m) \end{cases} . \tag{1.24}$$

The unbalance voltage V_{AB} is as follows:

- constant voltage excitation:

$$V_{\mathrm{AB}} = \frac{\Delta R}{R_0} E \tag{1.25}$$

- constant current excitation:

$$V_{\mathrm{AB}} = \Delta R I_{\mathrm{a}}. \tag{1.26}$$

Thanks to the symmetry, the expressions are linear.

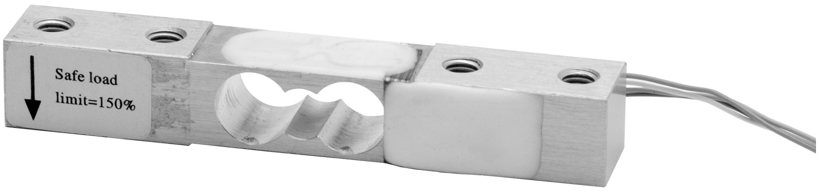


Figure 1.25: Photograph taken from the data-sheet of strain gauges DF2S-5 and DF2S-3 from HBM. The strain gauges are glued to a spring, whose deformation is translated into a differential voltage.

1.5.4.4 Example: strain gauges and Wheatstone bridge

The Wheatstone bridge configuration with four variable resistances is a widespread solution with strain gauges. In figures 1.25 and 1.26, we show part of the data-sheet of a sensor built with a strain gauge glued on a spring. Similar devices often constitute the heart of small weighing scales and it is very convenient to employ four gauges exploiting the compressive and tensile strains on the spring so that they work in a push-pull configuration as in eq. (1.24).

The parameter "sensitivity" shown in figure 1.26 indicates the ratio between the differential output voltage of the Wheatstone bridge contained in the device and the constant voltage excitation, when the full scale weight (max. capacity) is measured. For example, if one chooses to excite the sensor with 3.3 V, the full scale differential output V_{AB} is 6.6 mV for a model with a maximum capacity of 5 kg.

1.6 Reactive sensors

A vast family of reactive sensors exist. Those devices translate the measurand into a variation of capacity or an inductance. For example, a coil embedded in the floor may detect a car at the entrance of a parking via the change of its inductance, or loss factor. Conversely, there is a vast range of proximity or position sensors working on a capacitive principle.

We will not describe this class of sensors in much detail. We just present some points which may be considered:

- Measurement of a reactance involves employing AC signals, not always sinusoidal, but in any case always varying in time.

Specifications

Type		DF2S-5						DF2S–3					
Max. capacity (E_{max})	kg	1	3	5	10	15	20	1	3	5	10	15	20
Max. platform size	mm	150 x 150						150 x 150					
Sensitivity (C_n)	mV/V	2 ± 10 % (1 kg:1.8 ± 10 %)						2 ± 10 % (1 kg:1.8 ± 10 %)					
Zero balance		0 ± 0.5						0 ± 0.5					
Temperature effect on zero balance (TK_0)	% of C_n / 10 K	± 0.5000						± 0.2000					
Temperature effect on sensitivity (TK_C)		± 0.1000						± 0.0500					
Hysteresis error (d_{hy})	% of C_n	± 0.0500						± 0.0300					
Non-linearity (d_{lin})		± 0.0500						± 0.0300					
Creep (d_{cr}) over 5 min.		± 0.0500						± 0.0300					
Off center load error[1]	%	± 0.1000						± 0.0500					
Input resistance (R_{LC})	Ω	1000 ± 10						1000 ± 10					
Output resistance (R_0)		1000 ± 10						1000 ± 10					
Reference excitation voltage (U_{ref})	V	5						5					
Nominal range of excitation voltage (B_U)		5 ...15						5 ...15					
Insulation resistance (R_{is})	GΩ	> 2						> 2					
Nominal temperature range (B_T)	°C [°F]	–10... +40 [+14 ... +104]						–10 ... +40 [+14 ... +104]					
Service temperature range (B_{tu})		–20... +50 [–4 ... +122]						–20 ... +50 [–4 ... +122]					
Storage temperature range (B_{tl})		–30 ... +70 [–22 ... +158]						–30 ... +70 [–22 ... +158]					
Limit load (E_L)	% of E_{max}	150						150					
Lateral load limit (E_{lq}), static		300						300					
Breaking load (E_d)		150						150					
Deflection at E_{max} (s_{nom}), approx.	mm	< 0.4						< 0.4					
Weight (G), approx.	g	30						30					
Protection class according to EN60529 (IEC529)		IP54						IP54					
Material: Measuring element Coating		Aluminum Silicone rubber						Aluminum Silicone rubber					

[1] At loading with 30 % of max. capacity and 75 mm excentricity.

Figure 1.26: Technical specifications of the DF2S-5 and DF2S-3 from HBM.

- If the reactance is put in an oscillator, its changes may be detected through the output frequency variation, easily measurable with digital counters.
- Many variants of the Wheatstone bridge exist, adapted to reactive elements and working in AC (Maxwell bridge, De Sauty bridge, etc...)

1.7 Conclusion

In this chapter, several families of different electrical sensors have been described. Our description was oriented towards their representation in an electronics circuit. We have therefore focused rather on the electrical characteristics than on working principles. Thus, we have treated at the same time the conditioning circuits adopted for each kind of sensor. We have tried to follow a practical approach and several examples from datasheets have been discussed.

CHAPTER 2

Amplification and amplifiers

2.1 Introduction

In the previous chapter, we described a certain number of sensors, as well as the conditioning circuit used to obtain a voltage as an electrical representation of the measurand. Now, this voltage should be somehow treated: in most case it should be amplified, often filtered. This chapter is devoted to the amplification of low frequency signals and it is particularly focused on circuits based on operational amplifiers. We will therefore begin by briefly describing the working principles of operational amplifiers: in particular, we will focus on some parameters specified in the data-sheets quantifying their limits and defects. We will describe here the so called voltage-feedback operational amplifier (often just called operational amplifier), frequently adopted in low frequency circuits. A different element, the current-feedback operational amplifier bears some resemblance with it, but its use being more specific, it will not be described here. We will then give an overview of differential amplifiers, in particular the instrumentation amplifiers. The name of those circuits reflects their widespread use in instrumentation... The end of the chapter will be devoted to insulation amplifiers, precious when security or electromagnetic compatibility issues are of primary importance.

There are a lot of very good textbooks developing in detail the matter presented here. One of them is of course [Asch, 2003], also cited in the previous chapters. We recommend also the [Franco, 2015], which is very comprehensive and presents some advanced matters.

2.2 Introduction to operational amplifiers

2.2.1 The operational amplifier as a differential amplifier

First of all: an operational amplifier is an electronic circuit with two inputs and one output. It aims to be as close as possible to a *differential* amplifier with a very high gain.

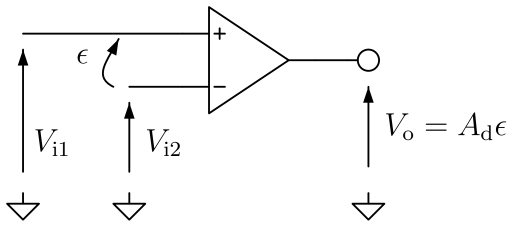


Figure 2.1: An operational amplifier as a differential amplifier.

Figure 2.1 shows an idealised circuit representation of what we expect from an operational amplifier: it takes the voltage difference ϵ measured between the non inverting "+" and the inverting "-" inputs, it amplifies it and the amplified output voltage V_o is now referred to the reference node:

$$V_o = A_d \epsilon \tag{2.1}$$

The input is differential and the output is single ended. In practice, the differential gain A_d is very high, but its exact value strongly depends on the frequency and a variety of other factors.

Usually, an operational amplifier is drawn as shown in Figure 2.1, where the two inputs and the output is represented[1].

There is of course important point which has been left out: the power supply rails V_{CC} and V_{EE}, as shown in Figure 2.2 on the left. The same picture depicts a more realistic model of the operational amplifier, by taking into account the saturation: no signal can (at least in ordinary cases) exceed the power supply rails in an operational amplifier circuit. Note that in the picture we represented the power supply rails symmetrically with respect to the reference potential. With A_d ranging usually between 10^5 and 10^6 (100 to 120 dB) at DC, if one wants to exploit the linear part of the characteristics, the

[1]While the situation shown in Figure 2.1 is quite common in textbooks, it is evidently a strongly idealised one, or in any case it can not be complete. In fact, *how can the operational refer its output to the reference node if it does not have any other connection to it?* This means that something must have been left out in the drawing and, for example, a lot of SPICE models for commercial operational amplifiers contain an artificial internal connection to the reference node. Be careful with SPICE simulations: believing that a result is accurate only because it comes from a computer is usually a good way to seek for catastrophes.

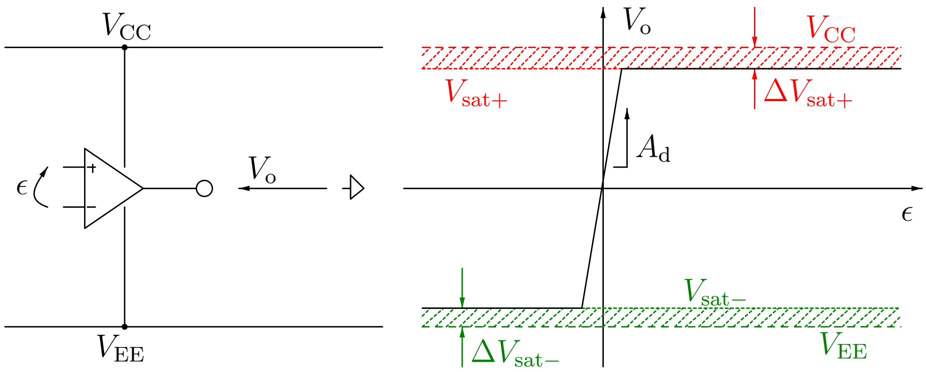


Figure 2.2: On the left: an operational amplifier with explicit representation of the power supply rails. On the right, a graphical representation of the (idealised) output characteristics.

only way is to use feedback. In other words, one must provide a link between the output and the input so that the output can be adjusted by the amplifier so that the differential voltage ϵ at the inputs is always in the linear range of the amplifier.

2.2.2 Modeling ideal operational amplifiers

Since the differential gain A_d is so high and moreover practically uncontrolled, we can say that *an operational amplifier is a circuit (often quite complex) optimised to be used as a linear block, by means of a feedback network.* A very useful model allows to simplify the calculations on the circuits by following two rules:

R1: The amplifier is able to measure a differential voltage without perturbing the circuit: no current flows in the inputs.

R2: The differential gain A_d is considered infinite. Therefore the only admissible situation where V_o is limited is $\epsilon = 0$. We may formulate that by saying that the amplifier does whatever it can with its output, to make sort that its inputs remain at the same voltage.

We notice that this way of seeing things is based on the presence of a feedback: otherwise, the output can not affect the inputs. We also remark that this simple model does not make a difference between the inverting and the non-inverting inputs, while we know that they can not be exchanged in a real circuit: saturation inevitably occurs if a confusion is done.

The rules associated to ideal operational amplifiers allow to understand the basic behaviours of a circuit. However, many key characteristics may only be deduced by taking into account more realistic models. In the following paragraphs, we will therefore briefly describe some of the most relevant defects of real operational amplifiers. Knowing their influence in a circuit allows to understand the data-sheets and select the best product in the huge catalogues proposed by the semiconductor companies.

2.3 Limits of real operational amplifiers

2.3.1 Saturation and rail to rail operational amplifiers

In figure 2.2, we notice that the output voltage of the amplifier is bounded inside a range defined by $V_{\text{sat}+}$ and $V_{\text{sat}-}$. One significant figure of merit of the amplifier is the difference between the saturation voltages and the power supply rails ($\Delta V_{\text{sat}+}$ and $\Delta V_{\text{sat}-}$), reflecting the capability of the circuit of working close to V_{CC} and V_{EE}. In classic integrated operational amplifiers $\Delta V_{\text{sat}+}$ and $\Delta V_{\text{sat}-}$ were between 1 and 2 volts. Today, we assist to the widespread use of wireless devices containing batteries, as well to a general trend coming from the digital circuits of reducing power supply voltage. Such a margin would represent a huge reduction of the available dynamic range. In fact, it is not uncommon to seek for high performance analog circuits with a single $V_{\text{CC}} - V_{\text{EE}} = 3.3\,\text{V}$ supply or even less. For this reason, a class of operational amplifiers (called rail to rail) has been optimised to make sort that $\Delta V_{\text{sat}\pm}$ do not exceed one hundred millivolts in the operating conditions.

What we described above is referred in particular to the output section of the operational amplifier, but something similar also happens at the inputs: most amplifiers do not work well if the voltages at their inputs are too close to V_{CC} and V_{EE}. Some of the modern rail to rail amplifiers, however, tolerate both inputs to be slightly above V_{CC} or below V_{EE}, adding flexibility in single supply configurations.

2.3.2 Input offset

For several reasons (mainly some small asymmetries in the fabrication process), when the voltage applied to the two inputs of a real operational amplifier is equal, the output voltage is not zero as predicted by equation (2.1). In fact, the very high value of the differential DC gain A_{d} will probably make sort that the asymmetry is such exasperated that output is saturated, either at $V_{\text{sat}+}$ or $V_{\text{sat}-}$. A small DC voltage, called the *offset voltage*, should be applied between the inputs in order that the output is no longer in this condition. The offset voltage can range between some microvolts in precision operational amplifier, to several millivolts. This effect being static, it affects only those circuits whose bandwidth include DC.

When feedback is present, the presence of offset changes Rule 2 in such a way that the difference between input voltages is no longer null, but equal to the value of the offset voltage. External nulling can often be performed via an external adjustable resistive network.

2.3.3 Common mode rejection ratio

In an ideal differential amplifier, the output voltage depends only by the voltage difference between the two inputs, that we called ϵ in fig. 2.1. In a real device, this is not completely true and the average of the two inputs voltages (V_{i1} and V_{i2}) plays a small role. In other words, by supposing that only this defect is present, equation (2.1) should be corrected as follows:

$$V_{\mathrm{o}} = A_{\mathrm{d}}\epsilon + A_{\mathrm{cm}}V_{\mathrm{cm}} \tag{2.2}$$

where V_{cm} is the so-called common mode, i.e. the arithmetic average of the voltages at the two inputs of the amplifier, each one referred to the reference node. To summarise:

$$\begin{cases} \epsilon = V_{\mathrm{i1}} - V_{\mathrm{i2}} \\ V_{\mathrm{cm}} = \frac{V_{\mathrm{i1}}+V_{\mathrm{i2}}}{2} \end{cases} \tag{2.3}$$

A new term of gain, namely A_{cm}, the common mode gain appears. A good differential amplifier (and thus a good operational amplifier) should make sort that A_{d} is much greater than A_{cm}. To quantify this characteristics, the data-sheets report the *common mode rejection ratio* in decibel, defined as follows:

$$C_{\mathrm{mrr}} = 20 \log \frac{A_{\mathrm{d}}}{A_{\mathrm{cm}}} \tag{2.4}$$

where the logarithm is base 10. Typical figures range between 80 to 120 dB.

2.3.4 Bias currents

In paragraph 2.2.2, the First rule states that no current flows in the inputs of an ideal operational amplifier. In real circuits, things are different: *some* current must flow to in the inputs, since a voltage measurement must be done. It is desirable to keep it as small as possible and operational amplifiers have been vastly optimised in this regard: currents as low as several picoampers are not uncommon in modern devices. We have already anticipated this issue while discussing conditioning circuits in paragraph 1.3.3.

2.3.5 Stability and frequency response

In our context, we call a circuit stable when the output to a bounded input is bounded. Other different definitions of stability exist. This definition is usually called with the acronym BIBO, from Bounded In Bounded Out. It is often highly desirable that a circuit remains stable. In low power applications, like operational amplifier circuits, the lack of stability will show up with non linearity, saturations, parasitic oscillations and head-scratching problems. In high power applications, lack of stability may yield expensive repairs, safety hazards, fires, explosions, nuclear meltdowns... If a circuit is unstable, most of the times it is practically useless.[2]

In modern voltage-feedback operational amplifiers there is an internal compensation network which tries to sort out a trade-off between overall speed and stability. The need of stability makes sort that the small signal bandwidth of the operational amplifier is often limited by introducing a low frequency dominant pole in the differential gain A_d. A consequence of that is that when a feedback network is present to obtain a circuit having a controlled gain G, the product Gf_p is approximatively constant, where f_p is the bandwidth obtained by the circuit. In other words, increasing the gain by acting on the feedback around the same operational amplifier entails a reduction of the frequency band treatable by the circuit. There is however a notable case in which the designer needs to take special care: most of operational amplifiers do not appreciate capacitive loads at their output (capacitors, long cables etc.).

If the limitation of the bandwidth is an effect associated to the small signal behaviour of the circuits, but a large signal (nonlinear) effect is also evident: there is a limitation to the slope of the variation of the output voltage versus the time (the so called *slew rate*).

2.3.6 Examples

Figure 2.3 shows an extract of the data-sheet of an operational amplifier optimised for low bias current. We notice the typical bias current value of 3 fA at a temperature of 25 C, which tends to rise with the temperature (5 pA at 125 C is a good achievement). Note also how the output saturation voltages are clearly specified: this is a rail to rail operational amplifier and this stuff matters! Its performances in terms of noise, gain-bandwidth product and slew-rate are also quite sound.

Figure 2.3 does not show the input offset voltage, specified elsewhere to be typically ±50 µV at 25 C and less than ±480 µV over an extended temperature range. This is a very decent offset performance, but it is clear that the device is not optimised towards this direction. As an exercise, compare this device with those in the following list (search for the data-sheet by yourself):

- The venerable general purpose bipolar µA741, designed in 1968 but still produced today. Compare it with the 1967 vintage LM101, today almost forgotten. Why µA741 was so successful?

[2]With the notable exception of oscillators, where a certain degree of instability is sought and kept under close control in order to initiate and sustain the oscillation.

2.5V Electrical Characteristics (continued)

Unless otherwise specified, all limits are specified for T_A = 25°C, V^+ = 2.5V, V^- = 0V, V_{CM} = (V^+ + V^-)/2. **Boldface** limits apply at the temperature extremes.

Symbol	Parameter	Conditions		Min (1)	Typ (2)	Max (1)	Units
I_{BIAS}	Input Bias Current	V_{CM} = 1V (4) (5)	25°C		±3	±20	fA
			−40°C to 85°C			±900	
			−40°C to 125°C			**±5**	pA
I_{OS}	Input Offset Current	V_{CM} = 1V (5)			6	40	fA
CMRR	Common Mode Rejection Ratio	0V ≤ V_{CM} ≤ 1.4V		83 **80**	100		dB
PSRR	Power Supply Rejection Ratio	1.8V ≤ V^+ ≤ 5.5V V^- = 0V, V_{CM} = 0		84 **80**	92		dB
CMVR	Input Common-Mode Voltage Range	CMRR ≥ 80 dB CMRR ≥ 78 dB		−0.3 **−0.3**		1.5 **1.5**	V
A_{VOL}	Large Signal Voltage Gain	V_O = 0.15V to 2.2V R_L = 2 kΩ to V^+/2		88 **82**	107		dB
		V_O = 0.15V to 2.2V R_L = 10 kΩ to V^+/2		92 **88**	120		
V_O	Output Swing High	R_L = 2 kΩ to V^+/2		70 **77**	25		mV from V^+
		R_L = 10 kΩ to V^+/2		60 **66**	20		
	Output Swing Low	R_L = 2 kΩ to V^+/2			30	70 **73**	mV
		R_L = 10 kΩ to V^+/2			15	60 **62**	
I_O	Output Short Circuit Current	Sourcing to V^- V_{IN} = 200 mV (6)		36 **30**	46		mA
		Sinking to V^+ V_{IN} = −200 mV (6)		7.5 **5.0**	15		
I_S	Supply Current				1.1	1.5 **1.75**	mA
SR	Slew Rate	A_V = +1, Rising (10% to 90%)			9.3		V/μs
		A_V = +1, Falling (90% to 10%)			10.8		
GBW	Gain Bandwidth Product				15		MHz
e_n	Input-Referred Voltage Noise	f = 400 Hz			8		nV/$\sqrt{Hz}$
		f = 1 kHz			7		
i_n	Input-Referred Current Noise	f = 1 kHz			0.01		pA/$\sqrt{Hz}$
THD+N	Total Harmonic Distortion + Noise	f = 1 kHz, A_V = 2, R_L = 100 kΩ V_O = 0.9 V_{PP}			0.003		%
		f = 1 kHz, A_V = 2, R_L = 600Ω V_O = 0.9 V_{PP}			0.003		

(4) Positive current corresponds to current flowing into the device.
(5) This parameter is specified by design and/or characterization and is not tested in production.
(6) The short circuit test is a momentary open loop test.

Figure 2.3: Some of the characteristics of the LMP7721 operational amplifier, from Texas Instruments.

- The JFET-input TL081.
- The first precision bipolar operational amplifier OP07.

Do not forget to search those amplifiers in the online catalogue of your favourite electronics dealer. Compare their costs.

2.4 Instrumentation amplifiers

2.4.1 Introduction

In chapter 1, we saw how the output of a sensor conditioning circuit is a voltage, which most of the times needs to be amplified. Moreover, in some situations, the voltage signal carrying information is not single ended (i.e. referred to the reference node) but differential. A classical example is the Wheatstone bridge with resistive sensors: the output signal is a voltage difference between two nodes, as shown in Figure 1.24.

We thus need a circuit able to extract the differential voltage signal without perturbing it, thanks to a high input impedance. We would also like to be able to easily adjust the differential gain of the circuit in a reasonable range, by changing only one component of the circuit. The ability of extracting the differential voltage regardlessly of the common mode voltage is quantified by the common mode rejection ratio parameter which is desirably very high. In the following paragraphs, we will describe a selection of classical circuits to achieve this goal, all based on operational amplifiers.

2.4.2 Differential amplifier with one operational amplifier

Figure 2.4 shows the classical differential amplifier made with one operational amplifier. The idea is to apply some amount of feedback in order to tame the the differential gain of the operational amplifier (as you remember from paragraph 2.2.1: it is very high, but it is variable, as affected by the frequency, power supply voltage, operating temperature etc.).

A little bit of circuit analysis, applying the rules given in paragraph 2.2.2, allows to write down the relation between voltages at the inputs V_{i1} and V_{i2} and the output V_{o}:

$$V_{\mathrm{o}} = \frac{R_1 + R_2}{R_1} \times \frac{R_4}{R_3 + R_4} V_{\mathrm{i1}} - \frac{R_2}{R_1} V_{\mathrm{i2}}. \tag{2.5}$$

To understand how this circuit can be exploited as a differential amplifier, we rewrite this expression by representing the electrical state of the inputs using the differential and common mode voltages. Thus, we apply the following relations, which can be seen as

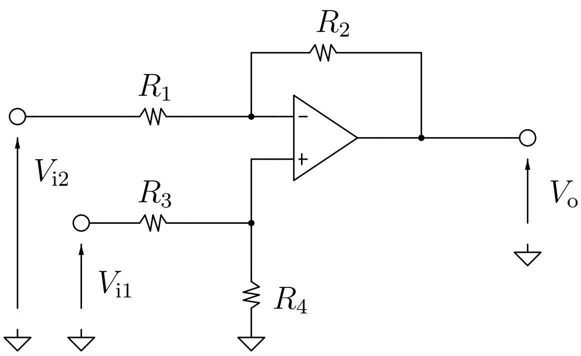


Figure 2.4: Differential amplifier built using one operational amplifier.

some sort of a coordinates change:

$$\begin{cases} V_d = V_{i1} - V_{i2} \\ V_{cm} = \frac{V_{i1}+V_{i2}}{2} \end{cases} \tag{2.6}$$

V_d is the differential mode (the signal carrying the information to be extracted) and V_{cm} is the common mode. By inverting the relations, we obtain:

$$\begin{cases} V_{i1} = V_{cm} + \frac{V_d}{2} \\ V_{i2} = V_{cm} - \frac{V_d}{2} \end{cases} . \tag{2.7}$$

This yields expressions of V_{i1} and V_{i2} to be injected in equation (2.5) to obtain equation (2.8). It relates the output voltage (single ended) to the input differential and common modes of the voltages:

$$V_o = \frac{R_1R_4 - R_2R_3}{R_1(R_3+R_4)}V_{cm} + \frac{R_1+R_2}{2R_1}\left(\frac{R_4}{R_3+R_4} + \frac{R_2}{R_1+R_2}\right)V_d. \tag{2.8}$$

In the expression (2.8) we recognise the contribution of the differential gain as well as the common mode gain:

$$A_{cm} = \frac{R_1R_4 - R_2R_3}{R_1(R_3+R_4)}, \tag{2.9}$$

$$A_d = \frac{R_1+R_2}{2R_1}\left(\frac{R_4}{R_3+R_4} + \frac{R_2}{R_1+R_2}\right). \tag{2.10}$$

If a perfect differential amplifier has to be built, resistances $R_1 \ldots R_3$ should be chosen in such a way that the common mode gain is equal to zero. This can be achieved by nulling the numerator of the expression (2.9), thus giving:

$$\frac{R_4}{R_3} = \frac{R_2}{R_1} \tag{2.11}$$

leading to a simple expression for the differential gain:

$$A_{\mathrm{d}} = \frac{R_2}{R_1}. \tag{2.12}$$

In practice, very often $R_1 = R_3$ and $R_2 = R_4$, yet perfectly achieving this condition is not possible, because of the inevitable tolerance of the resistances (this takes into account the effect of ageing and thermal drift). In practice, we know that every resistance has a certain relative shift from its nominal value. We suppose that the following conditions are verified (worst case scenario):

- The relative shift of the value of each resistance is equal to the tolerance r.
- The shifts are distributed in such a way that the common mode gain A_{cm} is maximised:

$$\begin{cases} R_1 = R_{1\mathrm{n}}(1+r) \\ R_3 = R_{1\mathrm{n}}(1-r) \\ R_2 = R_{2\mathrm{n}}(1-r) \\ R_4 = R_{2\mathrm{n}}(1+r) \end{cases} \tag{2.13}$$

  where $R_{1\mathrm{n}}$ and $R_{2\mathrm{n}}$ are the nominal values matching condition equation (2.11).

We obtain that the common mode gain is not zero, and it is proportional to the tolerance r:

$$A_{\mathrm{cm}} = \frac{4rR_{2\mathrm{n}}}{R_{1\mathrm{n}} + R_{2\mathrm{n}}}. \tag{2.14}$$

To calculate the common mode rejection ratio, we suppose that the differential gain has not changed very much if r is small, yielding:

$$C_{\mathrm{mrr}} = 20\log_{10}\frac{A_{\mathrm{d}}}{A_{\mathrm{mc}}} \approx 20\log_{10}\frac{R_{1\mathrm{n}} + R_{2\mathrm{n}}}{4rR_{1\mathrm{n}}}. \tag{2.15}$$

In the worst case scenario, this means that by adopting $r = 0.1\%$ tolerance for the resistances, by choosing a gain $A_{\mathrm{d}} = 100$, we might expect that the common mode rejection ratio is about 88 dB. This circuit has some defects:

- The input impedances are proportional to the values of the resistances. Very high resistance values are however associated to noise and the resulting low currents might be sensitive to stray capacitances, couplings…
- The gain can be modified, but the relation (2.11) should be respected. At least two matched resistances should be varied at the same time to vary the differential gain without disrupting the differential behaviour.

To solve the first problem, a second operational amplifier can be added, as discussed in the next paragraph.

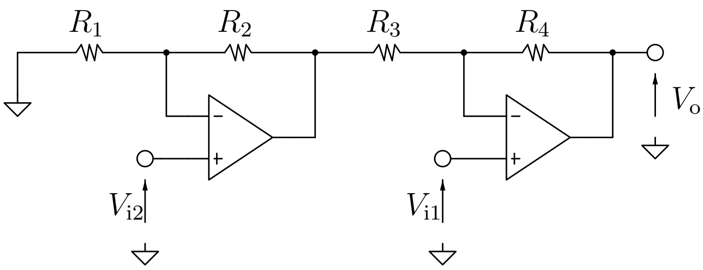


Figure 2.5: Differential amplifier built using two operational amplifiers.

2.4.3 Differential amplifier with two operational amplifiers

A useful way to vastly increase input impedances is to exploit the excellent input characteristics of operational amplifiers. The circuit shown in figure 2.5 solves the first issue seen at the end of paragraph 2.4.2, namely the low input impedances. In the circuit, the two inputs are directly connected to the inputs of the operational amplifiers. For this reason, once the correct biasing of the operational amplifiers is assured, inputs are extremely high impedance.

Analysing the circuit in the way described in paragraph 2.4.2, we calculate the differential gain, as well as the common mode gain:

$$\begin{cases} A_{\mathrm{d}} = \frac{1}{2}\left[1 + \frac{R_4}{R_3}\left(2 + \frac{R_2}{R_1}\right)\right] \\ A_{\mathrm{cm}} = \left[\frac{R_4+R_3}{R_3} - \frac{R_4}{R_3}\left(1 + \frac{R_2}{R_1}\right)\right]. \end{cases} \tag{2.16}$$

Nulling the latter, we obtain the balance condition of resistances $R_1 \ldots R_4$ to be respected:

$$\frac{R_1}{R_2} = \frac{R_4}{R_3} \tag{2.17}$$

thus yielding a simplified expression for the differential gain when the amplifier is purely differential:

$$A_{\mathrm{d}} = 1 + \frac{R_1}{R_2} \tag{2.18}$$

At a first sight, it might seem that this circuit is unable to solve the second problem described in the previous paragraph, i.e. the fact that it might not be easy to change the gain by modifying two matched resistances at the same time. In reality, a solution exists connecting a fifth resistor R_{g}, adjustable, which allows to trim the gain without bothering with two matched devices, as shown in figure 2.6. In this case, the ratio described by equation (2.17) must be respected, but the differential gain can be written as follows:

$$A_{\mathrm{d}} = 1 + 2\frac{R_1}{R_{\mathrm{g}}} + \frac{R_1}{R_2} \tag{2.19}$$

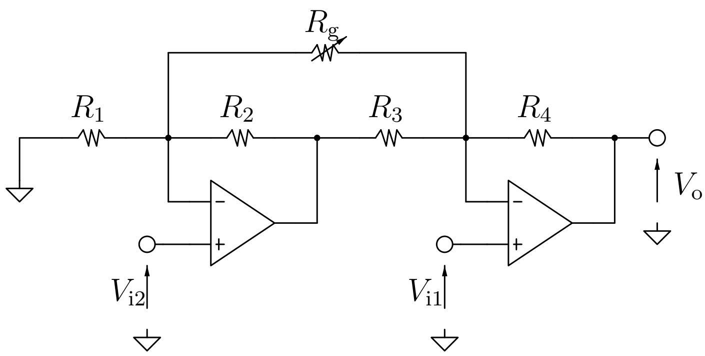


Figure 2.6: Differential amplifier with variable gain.

If the two issues of the differential amplifier discussed in paragraph 2.4.2 have been successfully addressed, this circuit still has a more subtile flaw. In fact, the signal paths are not symmetrical: the signal entering from V_{i1} passes through two operational amplifiers, whereas the signal entering from V_{i1} just have to pass through one. When the limitations of the operational amplifiers begin to play an important role (for example, when the frequency is relatively high), the asymmetry decreases performances and in particular the common mode rejection ratio of the circuit.

2.4.4 Differential amplifier with three operational amplifiers

The circuit shown in figure 2.7 is a more complex differential amplifier. It is quite commonly used in instrumentation chains and for this reason, when people say "instrumentation amplifier", they are often referring to this particular circuit. To understand its behaviour, we split it in two sub-circuits:

- An input stage, which has a differential input and a differential output, meant to boost the differential mode, while leaving untouched the common mode.
- A differential amplifier, to provide a single ended output related to the input differential mode.

The second stage is in fact the circuit discussed in paragraph 2.4.2, so we analyse now the input stage as shown in figure 2.8, which is perfectly symmetrical if $R_1 = R_1'$. If we suppose that the operational amplifiers are ideal, rule 2 seen in paragraph 2.2.2 states that the voltages at the nodes A and B are equal respectively to V_{i2} and V_{i1}. By supposing for a moment that V_{o1} and V_{o2} are known, we apply the Millman theorem:

$$\begin{cases} \text{node A: } \dfrac{\frac{V_{o2}}{R_1}+\frac{V_{i1}}{R_g}}{\frac{1}{R_1}+\frac{1}{R_g}} = V_{i2} \\ \text{node B: } \dfrac{\frac{V_{i2}}{R_g}+\frac{V_{o1}}{R_1'}}{\frac{1}{R_g}+\frac{1}{R_1'}} = V_{i1} \end{cases} \tag{2.20}$$

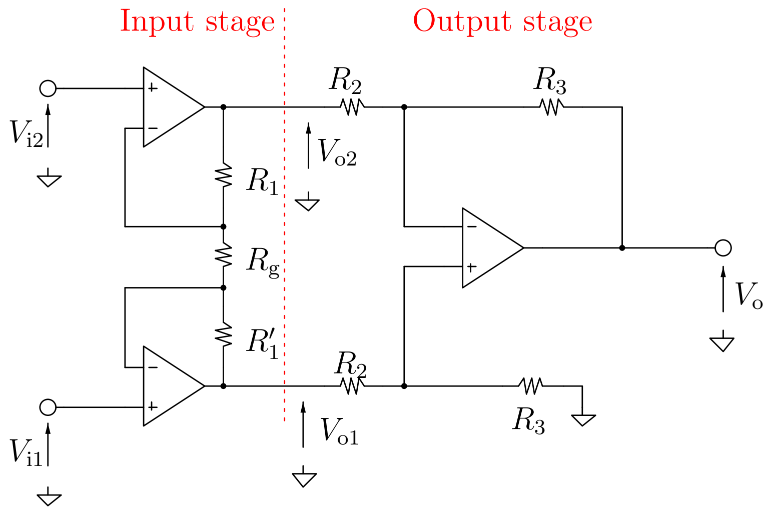


Figure 2.7: Differential amplifier with three operational amplifier: the instrumentation amplifier by antonomasia.

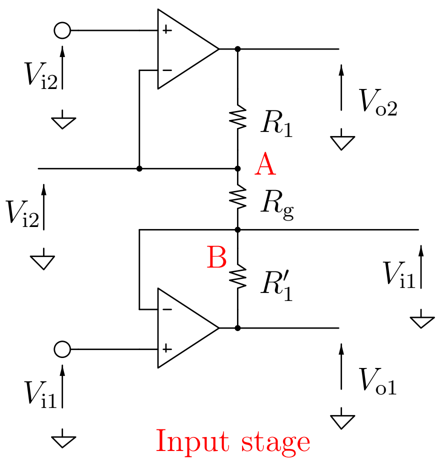


Figure 2.8: The symmetrical input stage of the instrumentation amplifier shown in figure 2.7.

By rearranging terms, we get:

$$\begin{cases} V_{\text{o1}} = \frac{R_{\text{g}}+R_1'}{R_{\text{g}}} V_{\text{i1}} - \frac{R_1'}{R_{\text{g}}} V_{\text{i2}} \\ V_{\text{o2}} = \frac{R_{\text{g}}+R_1}{R_{\text{g}}} V_{\text{i2}} - \frac{R_1}{R_{\text{g}}} V_{\text{i1}}. \end{cases} \tag{2.21}$$

Similarly to what done for the input signals with equations (2.6) and (2.7), we define differential and common mode voltages for the two outputs V_{o1} and V_{o1}:

$$\begin{cases} V_{\text{d}}' = V_{\text{o1}} - V_{\text{o2}} \\ V_{\text{cm}}' = \frac{V_{\text{o1}}+V_{\text{o2}}}{2}. \end{cases} \tag{2.22}$$

However, inevitable tolerances make sort that R_1 and R_1' are not identical. We consider the worst case scenario, as follows:

$$\begin{cases} R_1 = R_{\text{1n}}(1+r) \\ R_1' = R_{\text{1n}}(1-r) \end{cases} \tag{2.23}$$

where r is the tolerance of the resistances, whose nominal value is R_{1n}. In those conditions, we relate the output common mode with the input common and differential modes. After some algebra, we obtain:

$$V_{\text{cm}}' = V_{\text{cm}} + \frac{R_{\text{1n}}}{R_{\text{g}}} r V_{\text{d}} \tag{2.24}$$

as well as the output differential mode:

$$V_{\text{d}}' = \left(1 + \frac{2R_{\text{1n}}}{R_{\text{g}}}\right) V_{\text{d}}. \tag{2.25}$$

We notice that:

- If R_{g} is much smaller than R_{1n}, the input differential mode is greatly amplified.
- The input common mode is mostly not amplified nor attenuated by the first part of the circuit. This contribution is very often the most relevant one to the output common mode voltage.
- The output common mode is also affected by the input differential voltage, in a factor which is dependent on the $R_{\text{1n}}/R_{\text{g}}$ ratio (the same affecting the differential gain) as well as the tolerance r of the matching between R_1 and R_1'.

To summarise, the first circuit has a differential and common mode gains A_{d}' and A_{cm}' as follows:

$$\begin{cases} A_{\text{d}}' = 1 + \frac{2R_{\text{1n}}}{R_{\text{g}}} \\ A_{\text{cm}}' \approx 1 \end{cases} . \tag{2.26}$$

Paragraph 2.4.2 presented the analysis of the second half of the circuit of figure 2.7:

$$\begin{cases} A''_{\mathrm{mc}} = \frac{4rR_{3\mathrm{n}}}{R_{2\mathrm{n}}+R_{3\mathrm{n}}} \\ A''_{\mathrm{d}} = \frac{R_{3\mathrm{n}}}{R_{2\mathrm{n}}} \end{cases} , \tag{2.27}$$

where as usual r represents the tolerance of the resistances and the "n" subscript indicates their nominal values.

Putting together all these equations (and neglecting some cross terms) yields the differential and common mode gain of the complete amplifier:

$$\begin{cases} A_{\mathrm{mc}} = A'_{\mathrm{mc}}A''_{\mathrm{mc}} \approx \frac{4rR_{3\mathrm{n}}}{R_{2\mathrm{n}}+R_{3\mathrm{n}}} \\ A_{\mathrm{d}} = A'_{\mathrm{d}}A''_{\mathrm{d}} = \frac{R_{3\mathrm{n}}}{R_{2\mathrm{n}}}\left(1+\frac{2R_{1\mathrm{n}}}{R_{\mathrm{g}}}\right) \end{cases} . \tag{2.28}$$

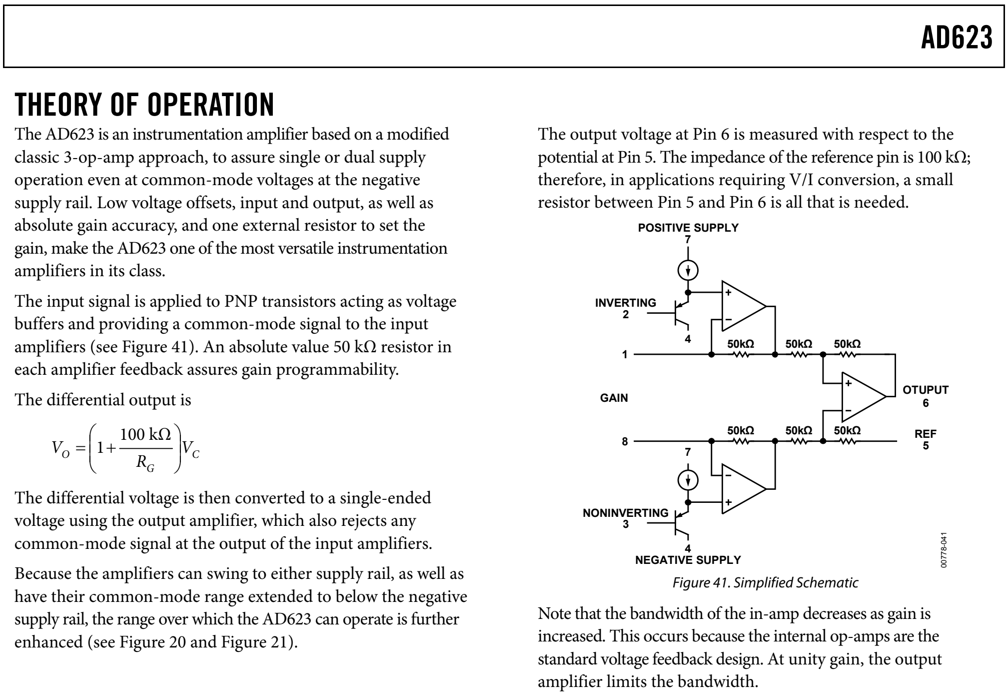
Those equations might be furthermore simplified when $R_{3\mathrm{n}} = R_{2\mathrm{n}}$, which is a frequent choice:

$$\begin{cases} A_{\mathrm{mc}} \approx 2r \\ A_{\mathrm{d}} = A'_{\mathrm{d}}A''_{\mathrm{d}} = 1+\frac{2R_{1\mathrm{n}}}{R_{\mathrm{g}}} \end{cases} . \tag{2.29}$$

In fact, the instrumentation amplifier built around 3 operational amplifiers is both flexible and very convenient to be integrated (for example the INA101, the AD623 and the INA333 and many others). In fact, in microelectronics it is difficult to control precisely the absolute value of a passive device, but symmetries such as those required in this circuit can be achieved quite conveniently. In fact, the end user just needs to choose the gain via the resistance R_{g} which is normally to be connected outside of the integrated circuit. This both provides outstanding performances, ease to use as well as flexibility. For example, have a look at figure 2.9, where the AD623 is described. Compare the expression given for the gain with equation (2.29).

2.5 Isolation amplifiers

Isolation amplifiers are able to decouple effectively two parts of a circuit which must exchange a signal, without having a direct galvanic connection between them. Figure 2.10 shows a schematic view of the way they are done: the signal e_{i} is transferred through an insulation barrier. Usually the transfer is done by an optical link (optocouplers), magnetically (transformers) or capacitively. Isolation amplifiers can be required for safety reasons and protection of equipments, for example when high voltages are involved. In this case, they are effective to eliminate very high common mode voltages V_{m} between the decoupled sections. A second important reason is to avoid ground loops, yielding severe electromagnetic compatibility issues. This is made explicit in figure 2.10 by the choice of two different symbols for the reference nodes, once the isolation barrier is crossed.



AD623

THEORY OF OPERATION

The AD623 is an instrumentation amplifier based on a modified classic 3-op-amp approach, to assure single or dual supply operation even at common-mode voltages at the negative supply rail. Low voltage offsets, input and output, as well as absolute gain accuracy, and one external resistor to set the gain, make the AD623 one of the most versatile instrumentation amplifiers in its class.

The input signal is applied to PNP transistors acting as voltage buffers and providing a common-mode signal to the input amplifiers (see Figure 41). An absolute value 50 kΩ resistor in each amplifier feedback assures gain programmability.

The differential output is

$$V_O = \left(1 + \frac{100\ \text{k}\Omega}{R_G}\right) V_C$$

The differential voltage is then converted to a single-ended voltage using the output amplifier, which also rejects any common-mode signal at the output of the input amplifiers.

Because the amplifiers can swing to either supply rail, as well as have their common-mode range extended to below the negative supply rail, the range over which the AD623 can operate is further enhanced (see Figure 20 and Figure 21).

The output voltage at Pin 6 is measured with respect to the potential at Pin 5. The impedance of the reference pin is 100 kΩ; therefore, in applications requiring V/I conversion, a small resistor between Pin 5 and Pin 6 is all that is needed.

Figure 41. Simplified Schematic

Note that the bandwidth of the in-amp decreases as gain is increased. This occurs because the internal op-amps are the standard voltage feedback design. At unity gain, the output amplifier limits the bandwidth.

Figure 2.9: A paragraph extracted from the data-sheet of AD623. Analog Devices describes it as an integrated version of the classic 3 operational amplifier instrumentation amplifier.

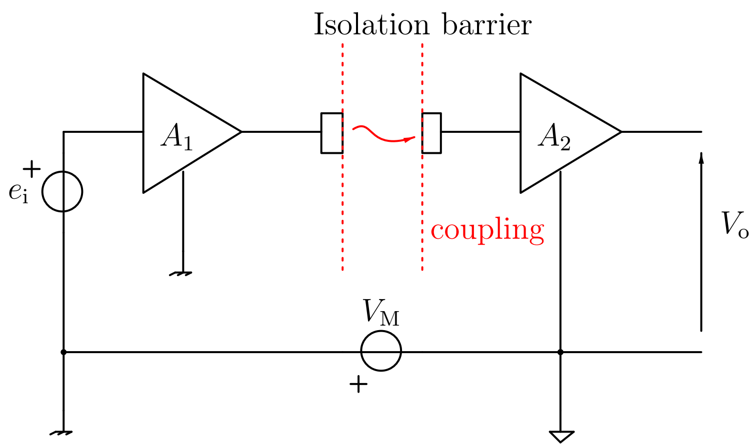


Figure 2.10: A schematic view of the principles of an isolation amplifier. The presence of an isolation barrier makes sort that the two reference nodes can be subjected to a voltage $\mathrm{V_M}$ without any current flowing and no risk for the signal integrity as long as $\mathrm{V_M}$ remains below a certain limit, specified in the datasheet.

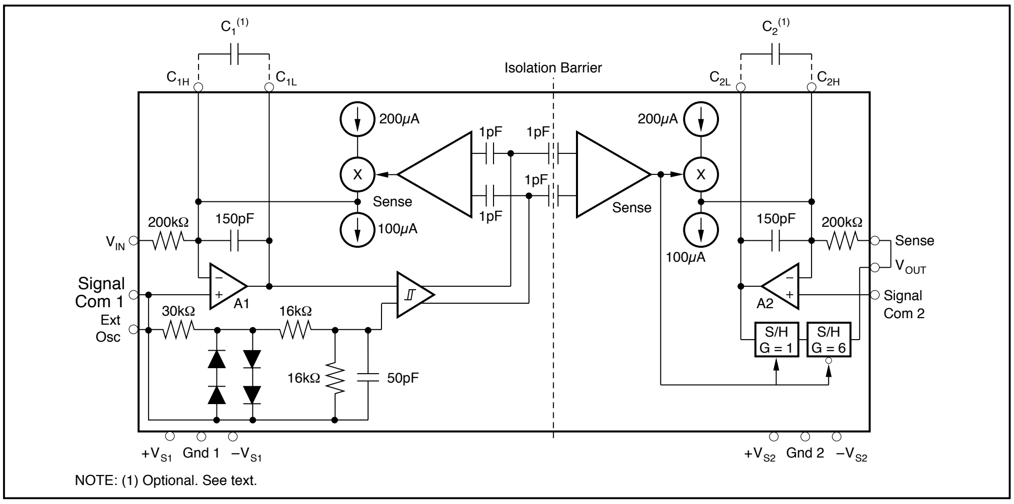


Figure 2.11: Block diagram of the internal structure of ISO120, a classic isolation amplifier from Texas Instruments.

Isolation amplifiers might be rated to guarantee several thousands volts of isolation. Examples include the classic ISO120 integrated isolation amplifier, whose internal structure is shown in figure 2.11. The isolation barrier is capacitive, so the input signal is used to modulate a carrier around 400 kHz. Note how a feedback loop is used on one side of the isolation barrier to achieve good linearity: this trick is effective when it is possible to obtain almost identical circuits on the two sides of the isolation barrier. Figure 2.12 contains the description of the working principle, the effect of sampling is visible in the oscillograms.

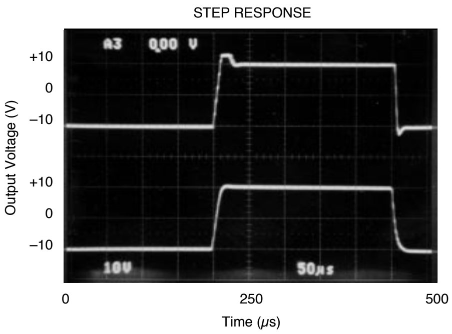
An example of a low-cost modern device proposed by Texas Instruments is the AMC1100, with once again a capacitive coupling. On the other side, the Analog Devices AD202 features a complete amplifier module, transformer-coupled, with an onboard isolated power supply converter and a cost which is aligned with the performances.

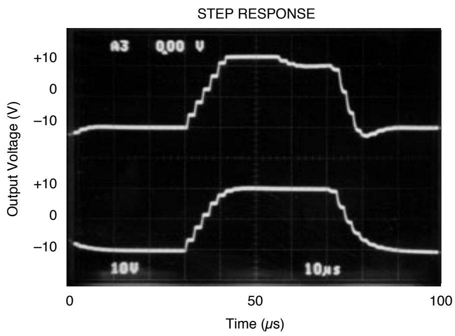
2.6 Conclusion

In this chapter, we saw rapidly the main characteristics of operational amplifiers. Then, we gave the desirable characteristics of instrumentation amplifiers and described three variants, often employed in practical situations. We finished our discussion by presenting isolation amplifiers.

TYPICAL PERFORMANCE CURVES (CONT)

T_A = +25°C; V_{S1} = V_{S2} = ±15V; and R_L = 2kΩ, unless otherwise noted.





THEORY OF OPERATION

The ISO120 and ISO121 isolation amplifiers comprise input and output sections galvanically isolated by matched 1pF capacitors built into the ceramic barrier. The input is duty-cycle modulated and transmitted digitally across the barrier. The output section receives the modulated signal, converts it back to an analog voltage and removes the ripple component inherent in the demodulation. The input and output sections are laser-trimmed for exceptional matching of circuitry common to both input and output sections.

FREE-RUNNING MODE

An input amplifier (A1, Figure1) integrates the difference between the input current (V_{IN}/200kΩ) and a switched ±100µA current source. This current source is implemented by a switchable 200µA source and a fixed 100µA current sink. To understand the basic operation of the input section, assume that V_{IN} = 0. The integrator will ramp in one direction until the comparator threshold is exceeded. The comparator and sense amp will force the current source to switch; the resultant signal is a triangular waveform with a 50% duty cycle. If V_{IN} changes, the duty cycle of the integrator will change to keep the average DC value at the output of A1 near zero volts. This action converts the input voltage to a duty-cycle modulated triangular waveform at the output of A1 near zero volts. This action converts the input voltage to a duty-cycle modulated triangular waveform at the output of A1 with a frequency determined by the internal 150pF capacitor. The comparator generates a fast rise time square wave that is simultaneously fed back to keep A1 in charge balance and also across the barrier to a differential sense amplifier with high common-mode rejection characteristics. The sense amplifier drives a switched current source surrounding A2. The output stage balances the duty-cycle modulated current against the feedback current through the 200kΩ feedback resistor, resulting in an average value at the Sense pin equal to V_{IN}. The sample and hold amplifiers in the output feedback loop serve to remove undesired ripple voltages inherent in the demodulation process.

SYNCHRONIZED MODE

A unique feature of the ISO120 and ISO121 is the ability to synchronize the modulator to an external signal source. This capability is useful in eliminating trouble-some beat frequencies in multi-channel systems and in rejecting AC signals and their harmonics. To use this feature, external capacitors are connected at C_1 and C_2 (Figure 1) to change the free-running carrier frequency. An external signal is applied to the Ext Osc pin. This signal forces the current source to switch at the frequency of the external signal. If V_{IN} is zero, and the external source has a 50% duty cycle, operation proceeds as described above, except that the switching frequency is that of the external source. If the external signal has a duty cycle other than 50%, its average value is not zero. At start-up, the current source does not switch until the integrator establishes an output equal to the average DC value of the external signal. At this point, the external signal is able to trigger the current source, producing a triangular waveform, symmetrical about the new DC value, at the output of A1. For V_{IN} = 0, this waveform has a 50% duty cycle. As V_{IN} varies, the waveform retains its DC offset, but varies in duty cycle to maintain charge balance around A1. Operation of the demodulator is the same as outlined above.

Synchronizing to a Sine or Triangle Wave External Clock

The ideal external clock signal for the ISO120/121 is a ±4V sine wave or ±4V, 50% duty-cycle triangle wave. The *ext osc* pin of the ISO120/121 can be driven directly with a ±3V to ±5V sine or 25% to 75% duty-cycle triangle wave and the ISO amp's internal modulator/demodulator circuitry will synchronize to the signal.

Synchronizing to signals below 400kHz requires the addition of two external capacitors to the ISO120/121. Connect one capacitor in parallel with the internal modulator capacitor and connect the other capacitor in parallel with the internal demodulator capacitor as shown in Figure 1.

Figure 2.12: Another extract of ISO120 data-sheet. Here is Ti's description of how the device works.

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