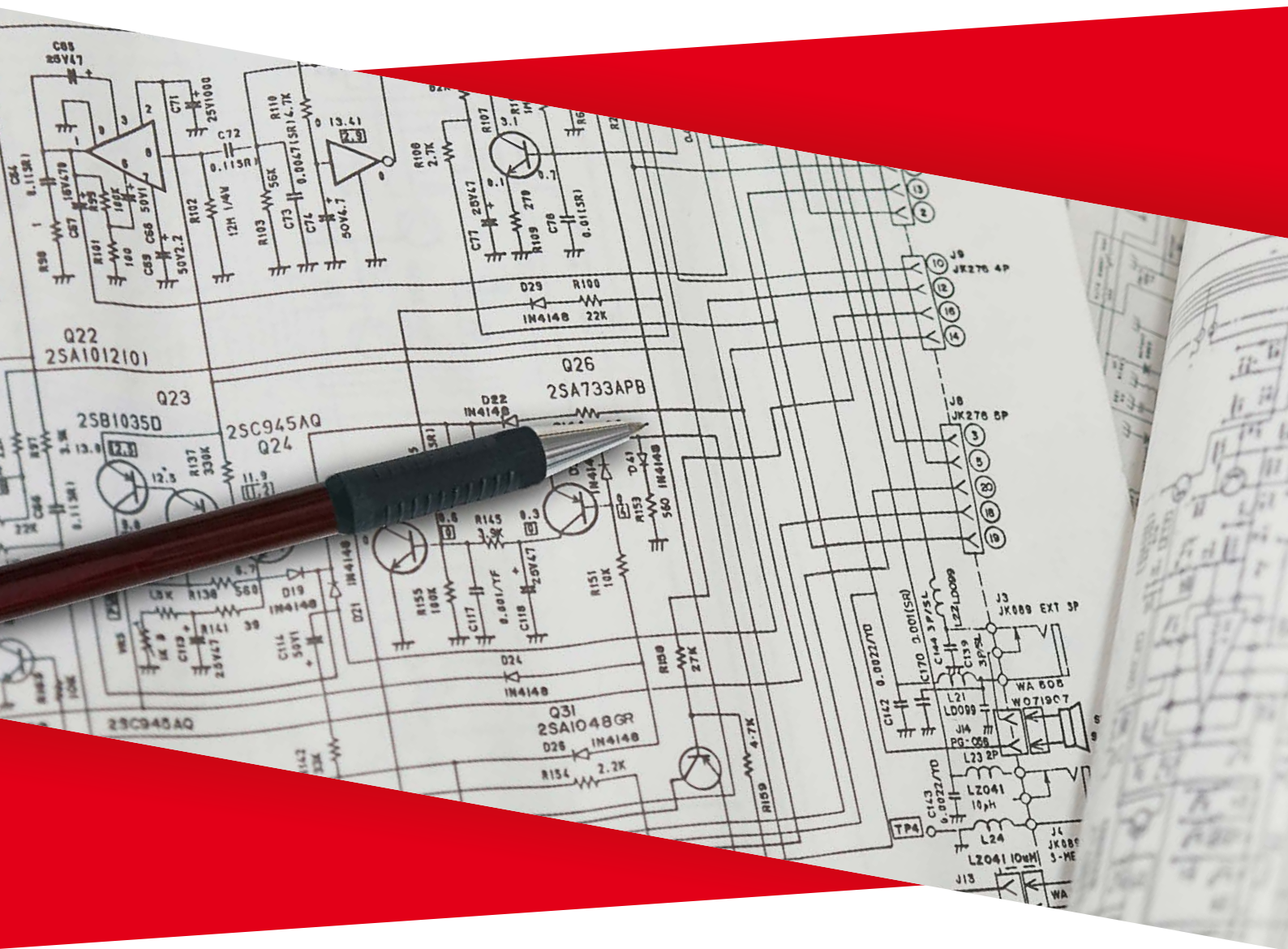


# Circuit Suggestions

Test and application circuits



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## Circuit Suggestions

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Since the signal processing and evaluation of pyroelectric detectors is complex, it is impossible to give a universal method. Therefore, we will give a few basic considerations below. We will consider the analogue signal processing and the design of the components based on simple methods of a functional test. As outlined in the chapter "Integrated Amplification", we distinguish between detectors in voltage mode and such in current mode here, too.

A final, more extensive circuit for measuring gas concentrations should demonstrate which other components are necessary for realising such a complete system.

### 5.1 Functional Test

**We would like to point out once again that pyroelectric detectors are sensitive to electrostatic discharges. Therefore, even for the functional test it is essential to follow the handling precautions which can be found at the end of the product catalogue and are also attached to each shipment.**

For a simple incoming goods inspection or to determine if a detector is functional or not, the parameters of the detector specified in the data sheet can be checked.

#### 5.1.1 Offset Voltage

Since almost all InfraTec detectors have an integrated amplifier (JFET or OpAmp), the easiest test is to measure the offset voltage. Here, the detector is connected to the supply voltage specified in the data sheet as shown in Figure 31 – Figure 33 and the offset voltage of the detector is measured at room temperature in a steady thermal state. This value should be within the limits specified in the data sheet.

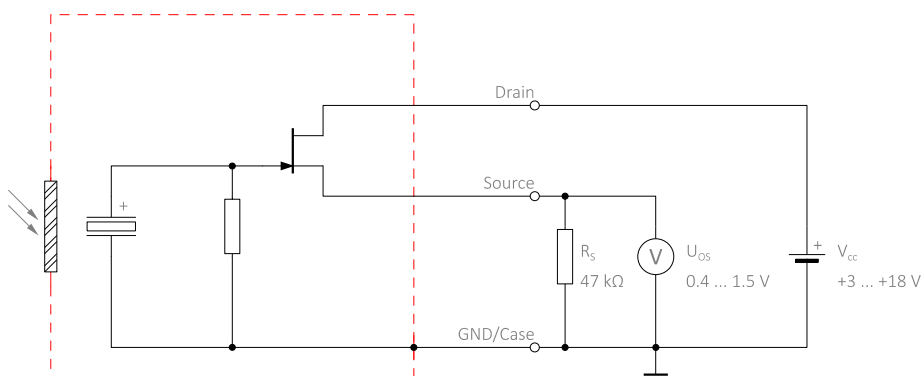


Figure 31: Measuring the offset voltage of detectors in voltage mode using the example of an LME-302

## Circuit Suggestions

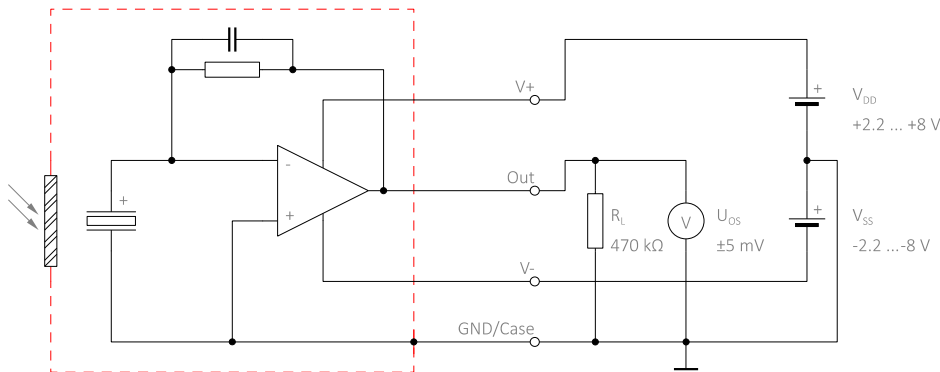


Figure 32: Measuring the offset voltage of split-supply detectors in current mode using the example of an LME-335

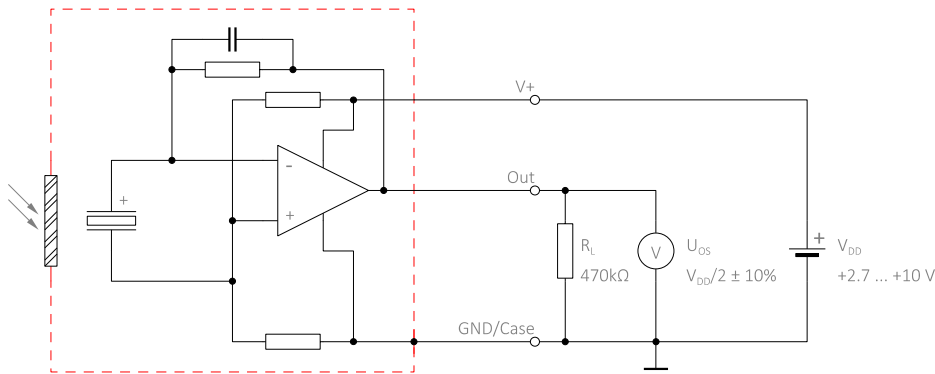


Figure 33: Measuring the offset voltage of single-supply detectors in current mode using the example of an LME-336

### 5.1.2 Noise

To measure the voltage noise density at the output of a detector accurately, please ensure the following:

- Cover the aperture of the detector so that it is insensitive to radiation from its environment
- Decouple the detector mechanically from its environment to minimize influences from vibrations
- Start the measurement only after the detector has reached a steady thermal state

The power supply of the detector must be stable and of low noise to make sure that detector noise is not influenced by the noise of the supply voltage. These requirements can be met easily by using batteries. As an alternative, a linear regulator which is decoupled from other parts of the circuit by an LC filter, for example, can be used. If the detector is operated in such a way, its voltage noise density can be resolved relating to frequency.

For this purpose, the output signal of a detector is logged. To adjust the detector signal to the input voltage range of the measurement device, an amplifier may be used. In this case, the bandwidth of the amplifier should be greater than the frequency band being analysed. In addition, the noise voltage generated by the amplifier must be lower than the noise voltage of the detector by at least a factor of three. The voltage noise density as a function of the frequency is then determined by Fourier transformation of this logged time signal into the frequency domain and division by the square root of the noise bandwidth. InfraTec uses this method for all noise density measurements by default.

To obtain repeatable results, we recommend a noise bandwidth of 1 Hz, an overlap of the individual measuring frames of  $> 70\%$ , the use of a Blackman-Harris window and an averaging of the measurements over a longer period of time.

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## Circuit Suggestions

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### 5.1.3 Responsivity

In addition to the electrical measurements described so far, it may also be necessary to check the responsivity to diagnose a fault. In this case, a measurement setup should be used as shown in Figure 34. To obtain results that are as repeatable as possible, we recommend the following:

- Use a source of radiation which is spectrally stable over time in the wavelength range of interest
- Modulate the IR radiation with a mechanical chopper at the frequency of interest
- As explained for measuring the noise voltage of the detector, use a power supply that is stable and of low noise
- Connect the detector to a high thermal mass during the measurement and allow the offset voltage to settle to a steady state (necessary particularly with thermally uncompensated detectors in voltage mode)
- Measure the effective value of the fundamental detector signal, e.g. using a lock-in amplifier.

Since the irradiated power is constant in this case, the measurable signal voltage is proportional to the responsivity of the respective detector.

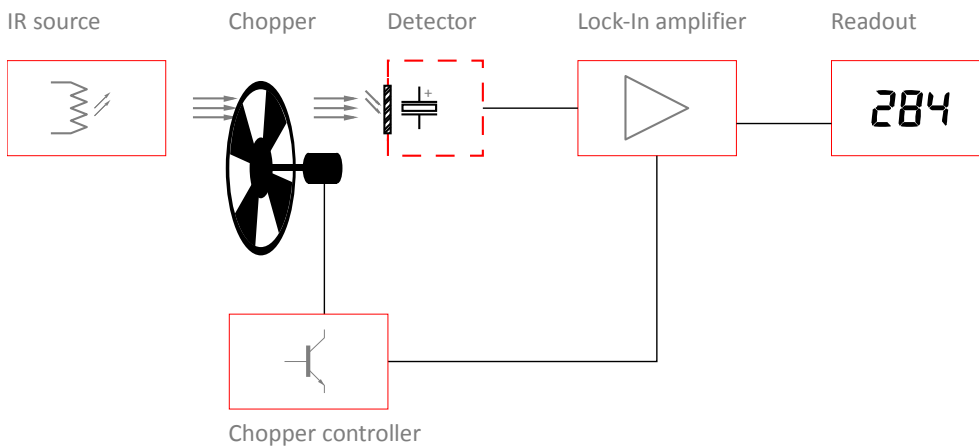


Figure 34: Schematic of a measuring setup for the responsivity measurement

# Circuit Suggestions

## 5.2 Analogue Signal Processing

The analogue signal processing adjusts the detector signal to the subsequent additional signal processing or digitization stage. As described in section 2.3, the signal voltages and noise densities of detectors in voltage mode and current mode differ in orders of magnitude. That is why, different requirements are placed on this part of the circuit depending on the particular operating mode.

In many applications, the pyroelectric detector is only excited with a single frequency<sup>2</sup> or narrow frequency band<sup>3</sup>. Parts of the signal with frequencies outside this band occur because of harmonics of the excitation, coupled interferences, the noise of the detector itself or low frequency changes in the ambient temperature. To reduce these undesirable parts of the signal, we recommend to filter the signal.

An active filter is normally used for this purpose to achieve amplification and filtering at the same time. An additional amplifier stage may be necessary to adjust the signal in terms of dynamic range and offset to the requirements of the AD converter.

### 5.2.1 Voltage Mode

The typical AC part of the signal of a voltage mode detector is in the range of a few hundred microvolts with effective noise voltages of a few microvolts. The offset voltage of the JFET (0.4 ... 1.5 V) is still superimposed on the AC part of the signal as constant component. An AD converter, however, has a typical input voltage range of 0 ... 3.3 V or 0 ... 5 V which is why an amplifier stage is used for adjustment. The simultaneous filtering of the signal by a bandpass filter results, in the simplest case, in the configuration of a non-inverting amplifier with a bandpass characteristic as shown in Figure 35.

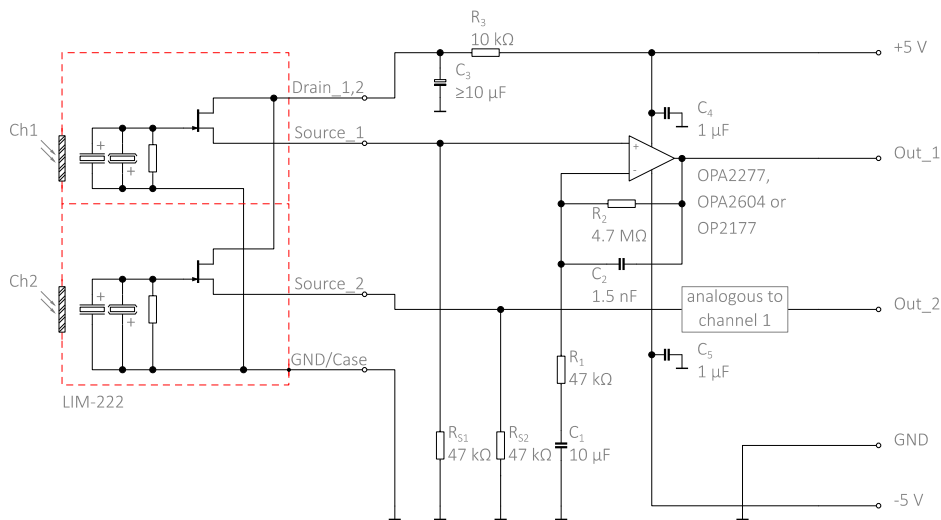


Figure 35: Non-inverting amplifier with bandpass characteristic for conditioning the output signals of a dual channel detector in voltage mode

<sup>2</sup> When using the detector for gas analysis

<sup>3</sup> When using the detector for flame detection

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## Circuit Suggestions

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To adjust the detector signal to the input voltage range of a subsequent AD converter, it has to be amplified. The maximum gain shall be calculated for the example in Figure 35. First of all, we consider the minimum offset voltage which is in the range of 0.4 ... 1.5 V in a steady thermal state at room temperature. As discussed in section 3.3.1, the offset voltage increases with an increasing temperature and decreases with a decreasing temperatures.

The smallest voltage possible at the output of the amplifier needs to be larger than the reference potential of the AD converter. To ensure this in practice, the lowest output voltage of the amplifier is selected typically at 0.1 V. Assuming that the output signal of the amplifier is symmetrical to the offset voltage, the maximum output swing can be calculated using

$$\Delta u_{\text{out,max}} = 2 \cdot (u_{\text{OS}} - 0,1 \text{ V}) \quad (44)$$

which results in 0.6 V for the example in Figure 35.

If the fundamental output signal of the detector has a maximum amplitude of 100  $\mu\text{V}$ , the peak-to-peak value is approximately four times larger due to the curves (compare Figure 7 and Figure 9). With these assumptions the maximum gain can be calculated using

$$g_{\text{max}} = \frac{\Delta u_{\text{out,max}}}{\Delta u_{\text{in,max}}} \quad (45)$$

In the example the maximum gain is 65 dB.

Following the amplification of the voltage to a maximum, it then should be digitised. With a reference voltage of  $u_{\text{ref}} = 5 \text{ V}$ , only  $\frac{0,4 \text{ V}}{5 \text{ V}} = 8 \%$  of the input voltage range of the AD converter can be used. A typical 12 bit AD converter integrated into a microcontroller is not suitable because the output voltage of the amplifier is quantised with only 320 digits. Assuming that the two least significant bits are not usable due to quantisation errors and other deviations of the AD converter from its ideal performance, only seven voltage levels can be distinguished after digitisation even with maximum gain. To measure a gas concentration, this is clearly insufficient.

Therefore, an AD converter with higher bit depth is chosen in practice so that a lower gain can be selected which then allows for a single stage realisation with standard components.

In the above circuit the signal is just amplified by a factor of 100 (40 dB) as shown in the frequency response Figure 36. To digitise the lower output signal with sufficient resolution a 16 Bit ADC is used. As a result, more than 100 different voltage levels can be distinguished.

The corner frequencies of the circuit were selected with  $f_{\text{HP}} = 0,3 \text{ Hz}$  and  $f_{\text{LP}} = 22 \text{ Hz}$  so that constant gain is resultant in the range between 1 ... 10 Hz, as the frequency response in Figure 36 shows.

## Circuit Suggestions

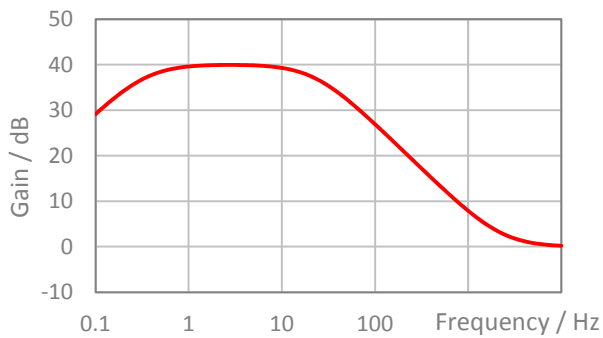


Figure 36: Frequency response of the gain of the active bandpass from Figure 35

A bandpass filter of a lower order is normally sufficient to eliminate low frequency signal and noise components as well as high-frequency disturbances prior to an analogue-to-digital conversion. Subsequently, the bandwidth is restricted even further by digital filters before the resulting signal is evaluated.

By determining the corner frequencies and gain, the value of the components can be quantified with

$$f_{HP} = \frac{1}{2\pi f R_1 C_1}, f_{LP} = \frac{1}{2\pi R_2 C_2}, |A_V| = \left| 1 + \frac{Z_2}{Z_1} \right| \quad (46)$$

using

$$Z_1 = R_1 + \frac{1}{j\omega C_1} = \frac{j\omega R_1 C_1 + 1}{j\omega C_1} \quad (47)$$

$$Z_2 = R_2 || C_2 = \frac{R_2}{j\omega R_2 C_2}.$$

Unlike with thermopiles, simple low-cost OpAmps are suitable for use with pyroelectric detectors. However, to achieve the optimum signal-to-noise ratio, we recommend using special low-noise amplifiers.

# Circuit Suggestions

## Noise

As already mentioned, the components of the circuit should be selected in such a way that their noise contribution to the detector noise can be ignored. If a noise source is at least three times smaller than the dominant noise source of the overall noise, then it can be ignored due to the accumulation of independent noise sources (see (28)). As shown in the noise model in Figure 37, there are six noise sources in addition to the detector noise for the example circuit given in Figure 35. The capacitors and resistors have been merged to complex impedances for simple representation.

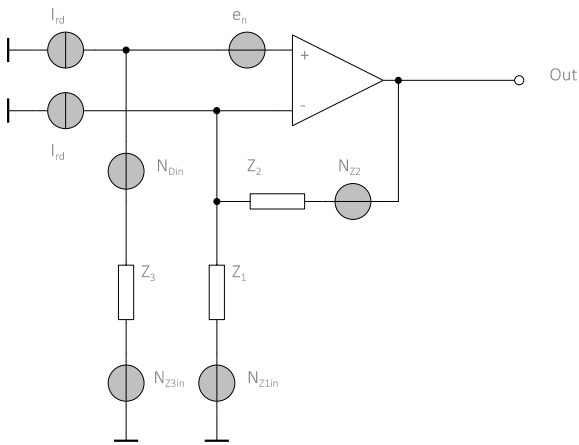


Figure 37: Noise model of the amplifier circuit of Figure 35

The impedance  $Z_3$ , which is located between the non-inverting input and ground, consists of the parallel connection of the source resistor  $R_{S1}$  and the dynamic output resistance of the JFET.

The noise sources result from the thermal noise of the resistors as well as from the input current and input voltage noise of the OpAmp. Their effect on the output of the circuit is given in Table 7.

Noise source	Noise density
Input voltage noise of the OpAmp	$N_{uOP} = e_n \cdot \left(1 + \frac{Z_2}{Z_1}\right)$
Input current noise of the OpAmp (non-inverting input)	$N_{iOP+} = I_{rd} Z_3 \cdot \left(1 + \frac{Z_2}{Z_1}\right)$
Input current noise of the OpAmp (inverting input)	$N_{iOP-} = I_{rd} Z_2$
Noise voltage at the non-inverting input	$N_{Zin+} = \sqrt{4kTZ_3} \cdot \left(1 + \frac{Z_2}{Z_1}\right)$
Noise voltage of the impedance between inverting input and mass	$N_{Zin-} = \sqrt{4kT \operatorname{Re}(Z_1)} \frac{Z_2}{Z_1}$
Noise voltage of the feedback impedance	$N_{Zfb} = \sqrt{4kT \operatorname{Re}(Z_2)}$
Noise voltage of the detector	$N_D = N_{Din} \cdot \left(1 + \frac{Z_2}{Z_1}\right)$

Table 8: Effect of the noise sources on the output noise of the amplifier shown in Figure 35

Figure 38 illustrates the frequency responses of the noise densities at the output of the OpAmp for the circuit presented in Figure 35. It can be seen that the detector is the dominant noise source up to a frequency of  $f \approx 10$  Hz.



## Circuit Suggestions

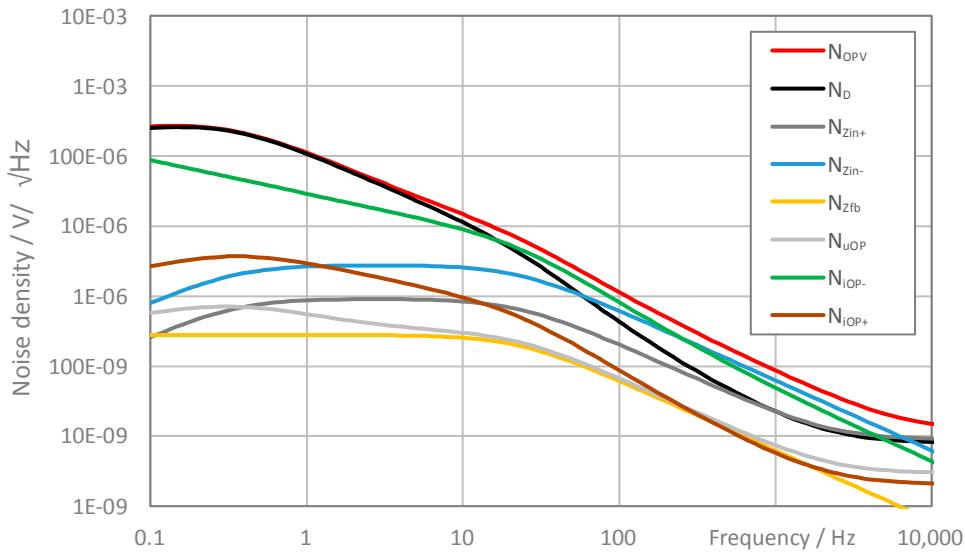


Figure 38: Frequency response of the noise density and shares of the various noise sources of the circuit shown in Figure 35

In the case of higher frequencies, the effect of the input current noise at the inverting input of the OpAmp is dominant. For this reason, when choosing a suitable OpAmp, attention was paid to low voltage noise and current noise at the excitation frequency. Each noise source of the amplifier should contribute a maximum of one third of the noise density of the detector to ensure that the noise at the output of the amplifier is dominated by the detector noise. In the example the current noise density is the dominant noise source of the amplifier in the frequency range 0.1 ... 100 Hz and its maximum value can be calculated using

$$I_{rd,max} = N_{Din} \cdot \left( \frac{Z_1 + Z_2}{3 \cdot Z_1 \cdot Z_2} \right) \quad (48)$$

We can assume maximum gain in the passband range of the filter so that equation (48) can be simplified to

$$I_{rd,max} = \frac{N_{Din} \cdot A_{V,max}}{3 \cdot Z_2} \quad (49)$$

In the example the requirement resulting from equation (49) can be fulfilled

$$I_{rd,max}(1 \text{ Hz}) = \frac{N_{Din}}{3 \cdot Z_2} = \frac{1000 \frac{\text{nV}}{\sqrt{\text{Hz}}} \cdot 100}{3 \cdot 4,7 \text{ M}\Omega} = 7 \frac{\text{pA}}{\sqrt{\text{Hz}}} > 6 \frac{\text{pA}}{\sqrt{\text{Hz}}}$$

As illustrated in Figure 38, the current noise density of the OpAmp already influences the noise at the output at 10 Hz. The maximum permissible current noise density is derived from equation (49)

$$I_{rd,max}(10 \text{ Hz}) = \frac{N_{Din}}{3 \cdot Z_2} = \frac{150 \frac{\text{nV}}{\sqrt{\text{Hz}}} \cdot 100}{3 \cdot 4,7 \text{ M}\Omega} = 1 \frac{\text{pA}}{\sqrt{\text{Hz}}}$$

According to the data sheet, however, the current noise density of OpAmp OPA227 is  $2 \frac{\text{pA}}{\sqrt{\text{Hz}}}$  and therefore greater.

## Circuit Suggestions

Depending on the circuit, there are additional criteria for selecting an OpAmp for filtering and amplifying the signals of voltage mode detectors:

- The supply voltage range
- The input offset voltage
- The current and voltage noise of the OpAmp in the modulation frequency range
- The input bias current
- The drift and temperature behaviour

### 5.2.2 Current Mode

The AC parts of the signal and noise densities of detectors in current mode are typically higher than those of voltage mode detectors by a factor of 100. Therefore, the signals generated by current mode detectors can often be processed using a suitable AD converter (e.g. LME-336). Due to the fact that the output signal of detectors with split supply (e.g. LME-335) is symmetrical to the reference potential as well as for reasons of noise suppression and signal amplification, an analogue signal processing can make sense for detectors in current mode, too. In this case, however, the demands placed on the OpAmps are lower than in voltage mode because the gain is significantly lower and the noise densities of current mode detectors are high.

Figure 39 shows the example of a circuit for a current mode detector LME-336 which uses an active first-order band-pass filter with an amplification of 20 dB and a passband between 0.7 Hz and 22 Hz.

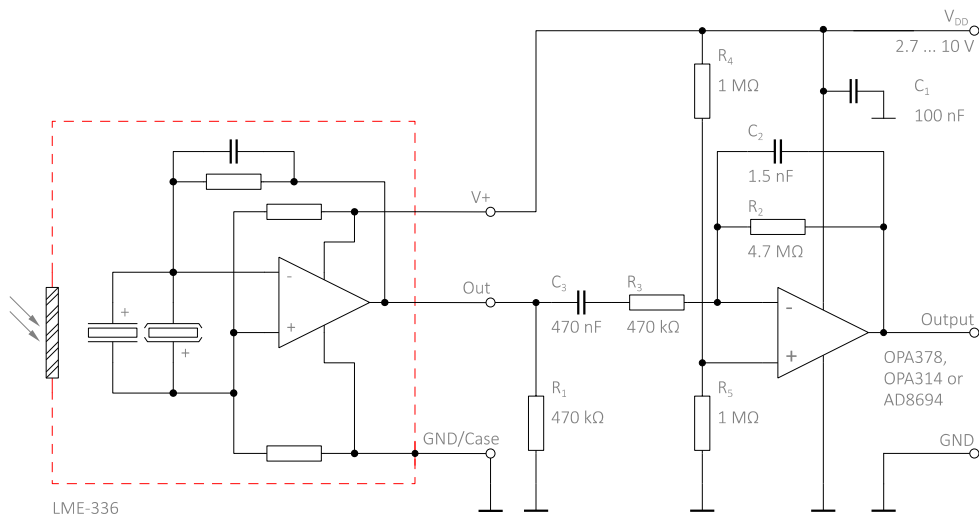


Figure 39: Circuit for an LME-336 with an active bandpass filter for signal processing

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## Circuit Suggestions

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The associated frequency response of this filter is shown in Figure 40. The considerations on how to design this circuit are similar for current mode and voltage mode. Due to the high-pass characteristics of the filter, the offset voltage of the detector signal is suppressed in current mode as well. Therefore, a DC voltage half of the supply voltage is superimposed on the amplified output signal across the voltage divider formed by  $R_4$  and  $R_5$ .

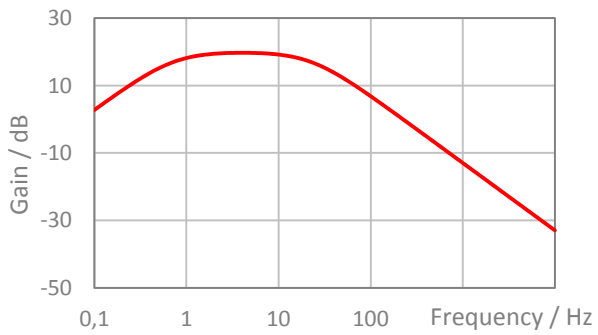


Figure 40: Frequency response of the gain of the filter in Figure 39

## Circuit Suggestions

### 5.3 Digital Signal Processing

To derive a proportional quantity of the incident radiant power from the AC part of the signal of the detector, the amplitude of the excitation frequency is often evaluated according to the definition of the responsivity (see section 1.3.1). After digitisation of the detector signal, this is possible in a number of ways:

- By means of Fourier transformation of the signal and evaluation of the spectral line of the excitation frequency
- By means of a lock-in process in which the excitation frequency is used as reference
- By means of a narrowband digital filter and subsequent RMS value measurement

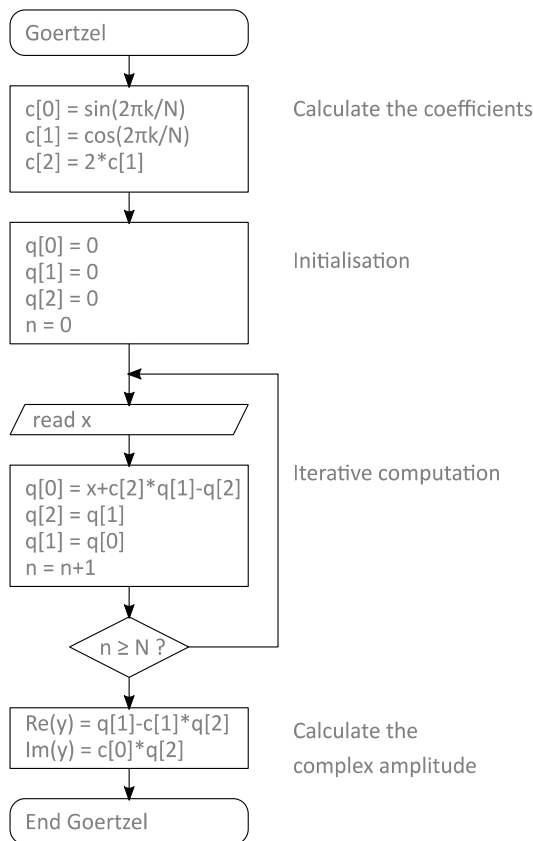


Figure 41: Exemplary implementation of the Goertzel algorithm

Since pyroelectric detectors are excited by a constant frequency in many applications, it is often sufficient to determine only the amplitude of this single excitation frequency in the output signal. A fairly simple method is the Goertzel algorithm which will be discussed briefly below. For a detailed derivation of the algorithm please refer to the relevant literature.

This algorithm is a special type of discrete Fourier transformation for the calculation of just one single spectral line.

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## Circuit Suggestions

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To determine the amplitude of the excitation frequency  $f_{\text{mod}}$  in the output signal of a detector, it is digitised with the sampling rate  $f_s$ . The amplitude is then determined from  $N$  sequential sampling values by the Goertzel algorithm whose exemplary implementation is shown in Figure 41. To reduce time and effort, the coefficients  $c[0] \dots c[2]$  can already be calculated and saved as constants before executing the program. Here,  $k$  is the number of the spectral line that is assigned to the modulation frequency. It is calculated using

$$k = \frac{N \cdot f_{\text{mod}}}{f_s} . \quad (50)$$

By choosing a greater number  $N$  of the sampling values to be processed the bandwidth of this digital filter decreases which, on the one hand leads to better frequency selectivity but, on the other hand reduces the tolerance to fluctuations of the excitation frequency.

Depending on the intended use of the detector, the amplitude and phase information obtained from the digital signal processing is further processed to show the desired value.

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## Circuit Suggestions

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### 5.4 Example of Application

Please note that this chapter discusses the basic structure and function of a gas measurement system with the sole purpose of demonstrating the application of our detectors. Of course, additional development is needed to perfect a true measurement system. It is also not possible here to go into the details of NDIR gas analysis. For reasons of clarity and comprehensibility, quite simple assumptions had to be made. Thus, for example, deviations from the Lambert-Beer law are not discussed. If you are interested, please refer to the corresponding literature on NDIR gas analysis.

Due to the excitation of molecular vibrations, many gases absorb infrared radiation in characteristic wavelength ranges. This property is used for measuring the concentration of gases. If the space between an IR source and the detector, which have the distance  $d$ , is filled with gas, the radiation intensity  $I$  decreases compared to the intensity  $I_0$  without the presence of the gas according to the Lambert-Beer law

$$I = I_0 \cdot e^{-\alpha c d} \quad (51)$$

with increasing concentration  $c$  of the absorbing gas. Here,  $\alpha$  is the wavelength-dependent absorption coefficient of the gas. To obtain the best result, the spectrally sensitive range of a detector is restricted by a bandpass filter which is selected to coincide with a particularly intensive absorption band of the gas being measured. The detector signal is then calculated using the integral of the following variables over the wavelength:

- Radiant exitance of the source
- Product of the transmission of the optical windows in the light path (e.g. of a gas cell)
- Transmission of the gas  $T = I/I_0$  (Lambert-Beer law)
- Transmission of the bandpass filter
- Responsivity of the detector

The ratio of the detector signals with and without gas represents a measure for the concentration of the gas in the optical path. In practice, the "signal without gas" is often measured in a second, so-called reference channel. The bandpass filter of the reference channel is transparent in a range where the gases of the mixture being measured do not have absorption bands.

Figure 42 shows the schematic of an entire system with a measurement and reference channel. The dual channel pyroelectric detector LIM-272-ZH converts the modulated radiation emitted by an IR source into an electrical signal. The signal is then filtered and amplified before being converted into a digital signal and evaluated by a micro-controller.

## Circuit Suggestions

Example of a circuit to measure CO<sub>2</sub> concentrations

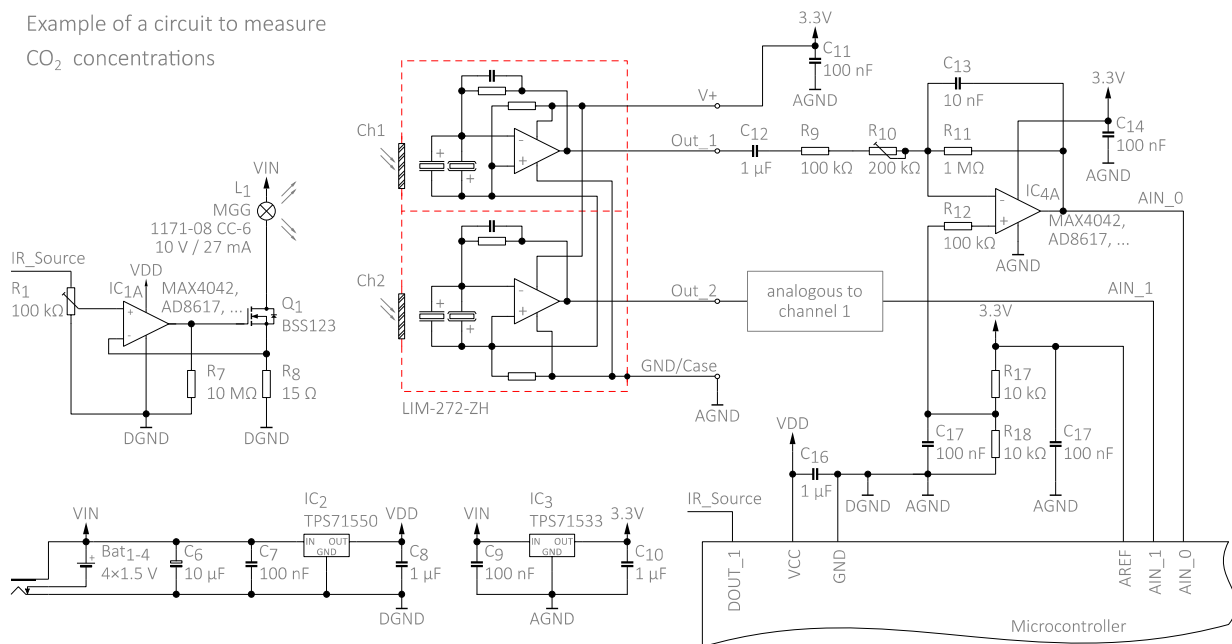


Figure 42: Circuit for the measurement of CO<sub>2</sub> concentrations using a single-supply dual channel current mode detector LIM-272-ZH

### 5.4.1.1 Power Supply

The power supply of the circuit is provided by batteries or by a DC power supply. The voltages  $V_{DD}$  for operating the digital parts of the circuit and 3.3 V for operating the detector and analogue signal processing are generated from a supply voltage  $V_{IN}$  of nominally 6 V by linear regulators.

To ensure a low-noise power supply, we recommend using ceramic or tantalum capacitors with a low equivalent series resistance (ESR) for all power supplies.

### 5.4.1.2 IR Source

When choosing an IR source, please make sure that it emits sufficient radiation in the spectral range that is of interest. A filament lamp whose bulb typically lets radiation pass of up to 4.5  $\mu\text{m}$  can be used for measuring CO<sub>2</sub> using a Z filter<sup>4</sup> (at a wavelength of 4.27  $\mu\text{m}$ ).

The setpoint for the maximum current of the filament lamp is set using the potentiometer  $R_1$ . A voltage of 1 V at the non-inverting input of the OpAmp IC<sub>1A</sub> results in a current of 66 mA.

<sup>4</sup> InfraTec standard NBP filter, see product catalogue section 2.4

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## Circuit Suggestions

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### 5.4.1.3 Analogue Signal Processing

The AC part of the output signal of the single supply detector LIM-272-ZH which is shifted by the offset voltage is amplified by 10 ... 20 dB using an active bandpass filter. The AC coupling formed by  $C_{12}$  eliminates the offset voltage of the detector. The amplified signal is then shifted by half of the supply voltage. Thus, interferences above 20 Hz and low frequency signal fluctuations below 2 Hz are suppressed to utilize the input voltage range of the subsequent AD converter to a maximum. Since the signal amplitudes of individual detectors may vary due to tolerances of components and materials used in the production of detectors as well as slightly changing installation conditions, adjustable gain was implemented at the potentiometer  $R_{10}$ . The frequency response of the bandpass filter used at minimum and maximum gain is shown in Figure 43.

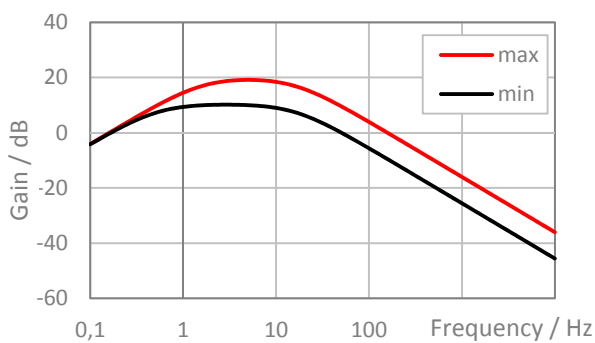


Figure 43: Minimum and maximum frequency response of the gain of the filter shown in Figure 42

### 5.4.1.4 Digital Signal Processing

The analogue signal is sampled at a frequency of  $f_s = 500$  Hz. The amplitude of the excitation frequency  $f_{\text{mod}} = 4$  Hz is then determined over 500 points each ( $N = 500; k = 4$ ) for both channels using a Goertzel algorithm. The ratio  $\frac{\text{Ch2}}{\text{Ch1}}$  of the amplitudes is formed from the  $\text{CO}_2$  and reference channel. Based on these results, the concentration  $c$  can be calculated using equation (52). Here,  $k$  is a device-specific factor.

$$c = \frac{1}{\alpha d} \ln\left(\frac{I_0}{I}\right) = \frac{1}{\alpha d k} \ln\left(\frac{\text{Ch2}(c)}{\text{Ch1}(c)}\right) \quad (52)$$

We would like to point out again that the above application example is greatly simplified to illustrate the principles of NDIR gas analysis. In practice, many additional details need to be considered to design an instrument which measures the concentration of gases at the required accuracy.

The circuit discussed here can only give a small insight, of course, into circuit technology and the signal processing for the measurement of gas concentrations with pyroelectric detectors. For this reason, the example chosen was simple and leaves scope for further developments and improvements.



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