Mid-Infrared Microspectrometer Systems

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Gerrit DE GRAAF
HBO Ingenieur in de Elektrotechniek
tegen Delft
Dit proefschrift is goedgekeurd door de promotor:

Prof. dr. ir. J. H. Huijsing

Toegevoegd promotor: Dr. ir. R.F. Wolffenbuttel

Samenstelling promotiecommissie:

Rector Magnificus, voorzitter
Prof. dr. ir. J. H. Huijsing, Technische Universiteit Delft, promotor
Dr. ir. R.F. Wolffenbuttel toegevoegd promotor
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Prof. dr. ir. G.C.M. Meijer, Technische Universiteit Delft
Prof. dr.ir. P.P.L. Regtien Universiteit Twente

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Introduction

You have to have an idea of what you are going to do, but it should be a vague idea.

Pablo Picasso

1.1 Motivation and Objectives

Spectrometers are systems that separate light into its basic spectral components. The most common example of light separation is a rainbow. Raindrops separate (disperse) sunlight into a spectral pattern that is observed by the human eye as a band of different colours in the sky. Optical spectrometers are used in many fields of science and in many industrial areas.

Development trends

A typical laboratory spectrometer is shown in figure 1-1a. These spectrometer systems are large and expensive and need trained personnel (spectroscopists) for operating and sample handling. There is a great demand for making these spectrometers portable, cheaper and more user-friendly. In many cases this can be achieved by making spectrometers dedicated to a specific application. A recently developed spectrometer, shown in figure 1-1b, illustrates the trend towards portability and miniaturisation. The next and future step in spectrometer development is a further miniaturisation resulting in the microscale spectrometer as illustrated by the research prototype [1.1] as shown in figure 1-1c.
Introduction

Although there is a trend towards miniaturisation, most spectrometers are still large and complex systems and would benefit greatly from miniaturisation.

Fig. 1-1. Developments in recent spectrometer systems:
  a). A modern laboratory instrument.
  b). A recent type of a portable mini spectrometer.
  c). A micro-scale spectrometer prototype [1.1].

There are two main benefits of microspectrometers:
• Miniaturisation, scaling down of spectrometers has the following benefits:
  • Portability: reduced size and weight
  • Small samples can be measured
  • Fast response
Motivation and Objectives

- System functional integration
  - logistics
  - ease of use.

Applications

Applications of spectrometers are abundant and a discussion of many types of spectrometers and their applications is beyond the scope of a thesis. Therefore, only some important applications of infrared spectrometers are considered in this thesis. Mid-infrared spectrometers are applied for analysis in agriculture, the food industry, soil biology, remote sensing and the chemical industry.

Examples of the use of mid-infrared spectrometers are in the measurement of gases during chemical or biological processes, e.g. monitoring greenhouse emissions or combustion processes. Typical applications in the food industry are in the detection of oils or fat, the analysis of dairy products such as butter and cheese or measurement of the ripening or freshness of food by the monitoring of water content. These applications will be discussed in more detail in chapter 3. In many of these systems one can accept a lower performance compared to a laboratory setup.

Motivation

Even larger opportunities of microspectrometers are in measurements where only very small sample sizes are available or the systems need to very be small or lightweight. Key applications for small and lightweight microspectrometers in the mid-infrared are in the fields of:

- Forensic analysis, e.g. the analysis of small samples of textile fibres, hair, drugs, toxic materials, explosives, paint, toners, inks etc.
- Analysis of pharmaceutical products and biomedical materials, e.g. tumor detection in tissue samples or the detection of acetone for diabetes by breath analysis.
- Infrared measurements aboard spacecraft or satellites.

The impressive amount of progress that has been made in the development of technology for integrated-circuits (IC’s) and, more recently, Micro Electro Mechanical Systems (MEMS) technology in the last decade and has resulted in the
**Introduction**

fabrication of many new miniaturized optical and electro-optical components, a field generally referred to as Micro-Optical Micro Electro Mechanical Systems or MOMEMS [1,4]. IC technology enables the batch fabrication of micro optical components like lenses, mirrors, gratings and filters. In addition MEMS technology can make these devices moveable, so they can be used for actuation. Probably even more important is that large numbers of identical devices, e.g. arrays of actuators or detectors, can be fabricated. This introduces many new opportunities for the miniaturisation of complex optical systems. Modern CMOS integrated circuit technology allows integration of analog and digital signal processing. Fast processing of data from detectors and the control of actuators is possible at low cost.

Nowadays many sensors and actuators using IC-compatible MEMS technology are developed. The scientific challenges are in developing new and optimization of existing components. Also research in the integration of these components into systems, thereby developing miniaturized systems combining optics and electronics, for many different applications is very challenging task.

**1.2 Organisation of this Thesis**

The previous paragraphs have introduced the objectives and motivation of the work presented in this thesis. An overview of the contents of this thesis is given below.

**Chapter 2 Introduction to Spectrometer Systems**

In Chapter 2 an overview of the operating principles and realisations of the various types of spectrometers is given. Their basic properties, such as resolution, will be discussed and, based on this, some conclusions with respect to the down-scaling of these instruments to chip size level will be given.

**Chapter 3 Mid-infrared Spectroscopy and its Applications**

Chapter 3 starts with a brief introduction of the interaction of infrared light with matter and the basic theory of vibrational spectroscopy is discussed. The interpretation of infrared spectra and some practical issues such as sample interaction and handling are discussed. Applications of infrared spectroscopy are abundant and some examples in the mid-infrared are given at the end of this chapter.
Chapter 4 IC-compatible Infrared Microspectrometers
This chapter starts with an overview of the optical properties in the mid-infrared of the materials commonly used in IC technology. Next an overview, of the state-of-the-art in miniaturized spectrometers in the mid-infrared will be given. In order to have a broader view on the field, spectrometers that operate in the Near-infrared or visible part of the spectrum are also discussed, if there are no fundamental limits for their operation in the mid-infrared.

Chapter 5 Fabrication of Thermally Isolated MEMS Structures
High thermal isolation is required for thermal infrared detectors and sources. An overview of the state-of-art technologies for the fabrication of thermopiles, bolometers and thermal infrared emitters is given in this chapter. Based on this overview, the selected process for the fabrication of thin siliconnitride membranes and cantilevers developed at DIMES will be described. Moreover, some new developments in the process are discussed.

Chapter 6 Integrated Thermal Infrared Sources
The light source is generally an integral part of the spectrometer system, therefore they are discussed in this chapter. The analysis, design, fabrication and testing of infrared sources in MEMS technology for microspectrometer applications is described also.

Chapter 7 Detectors for Mid-infrared Microspectrometers.
Chapter 7 starts with a general overview of the state-of-art of infrared detectors. After a discussion on compatibility with IC processes, a selection is made of mid-infrared detectors suitable for integration in a micro-system.

Chapter 8 Optimization of Infrared Thermal Detectors
The design and optimization of thermopiles and bolometers in the thermal, electrical and mechanical domain is treated in this chapter. Thermopiles and bolometers are most suitable devices in terms of compatibility with modern MEMS and IC technology.

Chapter 9 Signal Processing for Spectrometer Systems
After a short introduction into coherent detection and chopping, chapter 9 presents low offset and noise analog signal processing for the readout of the thermopile arrays. A new offset and drift reduction technique for thermal sensor readout is presented in this chapter.

Chapter 10 Low-Noise Readout Amplifiers for Thermal Detectors
Introduction

This chapter presents a design and circuit implementations of a low-noise, low-offset amplifier for the readout of thermopile arrays.

Chapter 11 Microspectrometers based on Attenuated Total Reflection
A new mid-infrared microspectrometer based on Attenuated Total Reflection is presented in this chapter.

Chapter 12 Microspectrometers based on Multilayer Interference Filters
This chapter discusses the theory, the design and fabrication of thin-film mirrors and Fabry-Pérot filters for use in Mid-Infrared microspectrometers.

Chapter 13 Conclusions
The conclusions of the work presented in this thesis and future research will be discussed in this chapter.

1.3 Summary: Original contributions

This thesis contains discussions on spectrometer systems in general and many different topics from the fields of sensor physics, optics, IC technology and electronics all related to spectrometers.

The main original contributions in this thesis can be found in:

- The design and fabrication of the first Mid-infrared microspectrometer based on Attenuated Total Reflection in bulk silicon as presented in chapter 11 [1.2].
- A new offset and drift reduction technique for thermal sensor readout as presented in chapter 9.
- The design optimization of the sensitivity and the signal-to-noise ratio of thin-film thermal sensors as presented in chapter 8.
- The IC-compatible multilayer Fabry-Pérot filters as presented in chapter 12 [1.3].
- The thermal analysis of infrared emitting thin-film beams in chapter 6.

In addition, some interesting overviews and analyses have been compiled within the context of this thesis:
Summary: Original contributions

- The study of the literature on infrared systems and microspectrometers has resulted in the overview of state-of-art infrared microspectrometers given in chapter 4.
- An overview of uncooled infrared detectors and their properties is given in chapter 7.
- The systematic noise analysis of folded cascode amplifiers as described in chapter 10.
- The noise analysis of current sources as described in chapter 10.

1.4 References


2.1 Introduction

As discussed in the introductory chapter, spectrometers are systems that separate light into its basic spectral components and/or can individually measure certain spectral components of electromagnetic radiation in general. Figure 2-3 gives the spectral range in which spectrometers can operate. Also there are many different types of spectrometers developed for different applications.

Spectrometers operating in the visible, UV and infrared can be classified according to a few basic optical principles. The following paragraphs introduce the basic operating principles of spectrometers. Typically, the application results in specifications that determine the type of spectrometer to be used. For this reason and to enable a comparison between the different types of spectrometers, some important specifications of spectrometers will be discussed.
Introduction to Spectrometer systems

2.2 General Specifications

An important specification of a spectrometer is its ability to resolve different spectral components, indicated by the term spectral, or chromatic, resolving power $R$ [2.5]. As an example figure 2-3 shows the output plot of a spectrometer measuring the spectral power distribution of light from a metal-halide discharge lamp with a resolution of 100 nm. Much more detail is provided by the curve measured with a wavelength resolution of 5nm.

The Rayleigh criterion is the generally accepted criterion for the minimum resolvable detail in the projected image of an optical system. Diffraction in an optical system results in an image of two nearby point sources as shown in figure 2-3b, called Airy patterns [2.4].
The Rayleigh criterion is an empirical estimate of the ability of the human eye to discern the overlap of these patterns in an image.

The minimum resolvable distance or resolution according to Rayleigh is said to be reached at a distance $\Delta x$ when the first diffraction minimum of the

**Fig. 2-2. Spectral plots of a lamp measured with a resolution of 5 and 100nm**

**Fig. 2-3. Resolving power definitions:**

a). Spectral.

b). Spatial.

The minimum resolvable distance or resolution according to Rayleigh is said to be reached at a distance $\Delta x$ when the first diffraction minimum of the
image of one source point coincides with the maximum of another [2.4], as shown in figure 2-3b. This distance $\Delta x$ is called the resolution of the imaging system.

In a spectrometer a monochromatic light source also produces Airy patterns on the detector due to diffraction, as will be discussed in section 2.3.2.1. Two spectrally closely spaced monochromatic light sources, e.g. from a mercury or sodium lamp, produce Airy patterns as shown in figure 2-3b. In this case the Rayleigh criterion indicates the ability of a spectrometer to resolve details in a spectral plot, as shown in figure 2-3a. Equivalent to imaging systems the resolution of a spectrometer is given by $\Delta \lambda$ and the chromatic resolving power $R$ of a spectrometer is defined by:

$$R = \frac{\lambda}{\Delta \lambda}$$

(2-1)

The resolving power determines the amount of detail in the spectrogram as required by different applications [2.12].

Another specification is the FWHM (Full Width Half Maximum) value, defined by the wavelength difference between two half power points, as shown in figure 2-3 also.

Another important issue is the ability of a spectrometer system to collect light at its input (“entendu”) [2.6]. Macro-scale spectrometers use lenses or convex mirrors to collect and transport as much light as possible from the input through the optical path. In micro systems the structures are generally planar and curved devices such as lenses and focusing mirrors are much more difficult to fabricate, as will be shown in this thesis.

### 2.3 Spectrometer Classification based on Optical Principle.

Spectrometers can be classified into three categories based on the three main operating principles:

- Refraction or prism types
- Diffraction or grating types
Spectrometer Classification based on Optical Principle.

- Interference types.

The next paragraphs briefly discuss the common types of spectrometers and their properties and limitations.

2.3.1 Refraction

Refraction of light occurs at the boundary between two different transparent media as shown in figure 2-5.

\[ \sin \theta_1 = \frac{n_2}{n_1} \sin \theta_2 \]  

Fig. 2-4. Light refraction and reflection at an interface.

The angle of refraction is given by Snell’s equation:

If the speed of light in a medium varies with the wavelength of the light, or in other words, if the refractive index of that medium is wavelength dependent, the angle of refraction will also change with the wavelength. This is illustrated in figure 2-5a. The different spectral components will be spatially deflected in the medium, according to Fermat’s principle, ending up at different locations in a plane after the prism. Glass prisms were the first devices used to break up, or disperse, visible light into their basic colour components.
Introduction to Spectrometer systems

The deflection angle is proportional to the change of speed in the medium, i.e. proportional to the dispersion or change in the index of refraction \( n \) of the medium. Because generally the index of refraction varies by only a few percent across the spectrum, different wavelengths are separated by small angles. It can be shown that if the ray traverses the prism parallel with the base the deviation angle is minimum and this results in the lowest losses [2.2].

In a practical system a beam, or wavefront, rather than a single ray is incident on the prism, and the dispersed wavefronts will overlap. An optical system imaging system, as shown in figure 2-5b, using a slit and lenses to form an image of that slit on the detector is therefore used in spectrometers.

The resolving power of a properly aligned prism is given by [2.2]:

\[
R = \frac{\delta n_z}{\delta \lambda} (b_2 - b_1) = D(\lambda) (b_2 - b_1)
\]  

(2-3)

The term \( D(\lambda) \) is the dispersion coefficient of the prism material and the path length differences \( b_2 \) and \( b_1 \) are shown in figure 2-5b. The size and material properties of the prism thus mainly determines the resolving power. The dispersion of most materials is non-linear, causing a varying resolving power when using prism spectrometers over a wide range. For highest resolution of the prism the optical system should completely fill the prism with the input beam. It must be noted that the presence of a very narrow slit can limit the resolution of the system by diffraction also. Diffraction will be discussed in section 2.3.2.

Fig. 2-5. Ray paths in a prism

a). Single ray.
b). Collimated beam
2.3.1.1 Prism-based Spectrometers

Figure 2-7 shows a basic prism spectrometer system. Input light enters via the entrance slit and is collimated by a concave mirror to the prism. To improve the resolution, spectrometers using prisms generally use two prisms or guide the beam twice through a single prism. In figure 2-7 the light is dispersed twice in the prism by the flat mirror and the spherical mirror focuses the beam again on the exit slit. In this system a simple single element detector has been used after the exit slit and the spectrum is scanned by rotating the spherical mirror. The typical resolution that can be obtained is $R = 200-500$.

![Fig. 2-6. Basic Prism Spectrometer.](image)

A more advanced prism spectrometer is shown in figure 2-7. The figure shows the light path in a dual beam transmission spectrometer system. Light rays from the source are split into two beams, one is lead through a reference cell and the other is directed to the sample. Time multiplexing of the reference and the sample signal has been implemented by a reflecting/transmitting mechanical chopper wheel. The spectrum is formed on a detector array by a concave mirror.
Introduction to Spectrometer systems

Introduced around the year 1954, this instrument was the first widely used infrared spectrophotometer [2.3]. Nowadays these infrared spectrometers have been superseded by grating or Fourier Transform Infrared Spectrometers (FTIR) systems in most applications.

Concluding it can stated that systems based on refraction are used only when low spectral resolving power is required. Since the size of the prism determines the resolving power, scaling down of the device will result in an even lower resolution and therefore prisms are not well-suited for spectrometers.

2.3.2 Diffraction

Diffraction manifests itself as the apparent bending of waves around small obstacles or the spreading out of waves past small openings. Diffraction gratings are formed by closely spaced patterns that reshape the wavefronts of incoming light into new wavefronts. These new wavefronts form spatial interference patterns depending on the wavelength. Diffraction in gratings can be observed in transmission or reflectance. Transmission gratings are generally formed by small openings or slits on a flat surface. Patterns of grooves on a flat or curved surface can form both reflective and transmission gratings. Generally the these gratings are covered with a thin reflecting layer for higher reflectance. The regular pattern of grooves in a compact-disc is a simple example of a reflective grating.
2.3.2.1 Transmission Gratings

If monochromatic (=single wavelength) bundled light is directed toward a transmission grating, the wavefronts pass through and spread out at the slits as secondary waves, as shown in figure 2-8.

![Fig. 2-8. Fraunhofer diffraction pattern of a single slit.](image)

At some exit angles, the secondary waves from adjacent slits of the grating are delayed by exactly one wavelength and constructive interference takes place. As a result light can be observed in directions where the spacing between the adjacent radiators is delayed by one wavelength. Constructive interference is also observed for delays of integral numbers of wavelengths. If the distance of the image plane is much larger than the width of the slit $W$ the intensity pattern of a single slit is given by the Fraunhofer approximation [2.4],

$$I = I_0 \left( \frac{\sin \left( \frac{k W}{2 \sin \theta} \right)}{k W \sin \theta} \right)^2 \approx I_0 \left( \frac{W \frac{\pi}{\lambda} \sin \theta}{W \frac{\pi}{\lambda} \sin \theta} \right)^2,$$

(2-4)

where $I_0$ is the intensity on the screen at $Y=0$ ($\theta=0$) on the detector. The right part of the figure shows the calculated intensity pattern of the slit. The minimum intensities occur at $\sin \theta = \pm m \frac{\lambda}{D}, m=1, 2, 3\ldots$ etc. The resolution of a transmission grating can be improved by a multi-slit configuration as shown in...
The figure shows a grating of \( N \) slits equally spaced at a distance \( a = 2W \). The intensity pattern of a multi-slit grating [2.4] is given by the following equation:

\[
I = I_o \left( \frac{\sin \left( \frac{W \pi \sin \theta}{\lambda} \right)}{W \frac{\pi \sin \theta}{\lambda}} \right)^2 \left( \frac{\sin \left( \frac{N a \pi \sin \theta}{\lambda} \right)}{a \frac{\pi \sin \theta}{\lambda}} \right)^2
\]  

(2-5)

Basically this equation represents the pattern of a single slit multiplied with envelopes caused by the interference of the other slits. The right figure shows the calculated density plots for \( N = 1,2,4,6,8 \) slits each wide a distance \( W \) and equally spaced at a distance \( a = 2W \). A larger number of slits results in narrower peaks, a smaller grating constant yields a larger separation between the spectral lines and a smaller slit width produces a wider envelope. Also it can be seen that the power (intensity) in the higher order patterns increases with \( N \). If the input light source consists of several different spectral components, these wavelengths will be diffracted in slightly different patterns and due to the interference of waves they can be observed and recorded separately.

Because of the relatively long wavelength of Mid-infrared radiation, multi-slit transmission gratings can be readily realized within the capabilities of the lithography typically used state-of-the-art IC processes.
2.3.2.2 Reflective Gratings

Reflective gratings are more commonly used than transmission gratings. A reflective grating is shown in figure 2-10.

![Fig. 2-10. Ray path of a reflection grating.](image)

Reflecting gratings can be described in the same manner as transmission gratings. The regular patterns on a reflective substrate produce new wavelets and from the figure can be seen that:

\[
\begin{align*}
\phi_1 &= a \sin \theta \\
\phi_2 &= a \sin \varphi \\
\end{align*}
\]

The path difference is then:

\[
d_2 - d_1 = a (\sin \varphi - \sin \theta)
\]

Constructive interference occurs at wavelengths \( \lambda \) that are multiples of the path difference. This leads to the well-known grating equation:

\[
m \lambda = a (\sin \varphi - \sin \theta)
\]

With: \( a \) is the spacing between the slits (the grating period), \( m \) is the order of diffraction (\( m = 0, \pm 1, \pm 2, \ldots \)) and \( \lambda \) is the wavelength. A lens or concave mirror can be used to produce images of the spectral lines. Gratings are usually blazed (engraved), as shown in figure 2-11, so that each order reflects specularly in the same direction, leading to a much higher output [2.4].
Introduction to Spectrometer systems

Mid-Infrared Microspectrometer Systems

![Fig. 2-11. Ray path of a blazed grating.](image)

The reflection condition of the blazed grating equals:

\[ \theta_i - \theta_b = \theta + \theta_b \]  

(2-8)

There are several ways to use a blazed grating. If \( \theta_i = 0 \) this leads to the grating equation:

\[ m \lambda = a \sin (2\theta_b) \]  

(2-9)

If \( \theta_i = -\theta_i \) the grating is in the auto-collimation, or Littrow mount, condition and the grating equation becomes:

\[ m \lambda = 2a \sin (\theta_b) \]  

(2-10)

The largest values of the resolving power \( R \) of a grating is obtained when it is used in auto-collimating and \( R \) is then given by:

\[ R_{\text{auto}} = \frac{2aN \sin (\theta_b)}{\lambda} \]  

(2-11)

where \( N \) is the number and \( a \) is the pitch of the grooves. The resolving power of a grating therefore typically increases with its width, \( W = Na \) and decreases with its central wavelength. Transmission gratings and non-blazed reflective gratings are much easier to fabricate in planar IC technology compared to blazed reflective gratings.
2.3.2.3 Grating-based Spectrometers

The most common grating spectrometers are the Littrow, Ebert and Czerny-Turner configurations. All of these instruments are built from the same basic components, an entrance slit, a dispersive element, a focusing element and an image sensor as shown in figure 2-12.

![Fig. 2-12. A grating spectrometer.](image)

As the grating in a spectrometer is rotated about an axis parallel to the slit axis, the spectral lines of a sample are transmitted successively through the instrument. A detector placed behind the slit can then be used to measure the intensity of light at each wavelength in the spectrum.

2.3.2.4 Resolving Power of a Grating Spectrometer

The way in which the diffraction angle $\theta$ behaves when light composed of different wavelengths is directed at a grating is important when considering the separation of light into its spectral components. If the incident angle $\phi$ is regarded constant, differentiating both sides of equation (2-7) with respect to $\theta$ yields:

$$\frac{d\theta}{d\lambda} = \frac{m}{a \cos \theta},$$  \hspace{1cm} (2-12)

where $d\theta/d\lambda$ is generally referred to as the angular dispersion of the spectrometer. Multiplying both sides of (2-12) by the focal distance $f$ of the...
Introduction to Spectrometer systems

optical system and substitution of \( dx = d\theta \) \( f \) yields the following relation for the linear dispersion:

\[
\frac{dx}{d\lambda} = f\frac{m}{a\cos\theta}
\]  

(2-13)

Here, \( dx / d\lambda \) is called the linear dispersion and represents the difference in wavelength per unit length on the surface of the exit slit in the spectrometer. Division by the slit width \( x_w \) gives an indication of the resolving power \( R \):

\[
R \equiv \frac{dx}{d\lambda} \frac{1}{x_w} \equiv f\frac{a\cos\theta}{x_w m}
\]  

(2-14)

From above equation it can be seen that the resolving power of a grating spectrometer depends on the pitch \( a \) of the grooves in the grating, the focal distance of the optical system and the width of the exit slit. Since the first two parameters generally have fixed values, the resolving power or resolution of grating spectrometers is normally controlled by varying the width of the exit slit. Increasing the width of the exit slit decreases the spectral resolving power but increases the total amount of light on the detector.

2.3.3 Interference

A third class of devices for isolating frequencies or wavelengths in spectra are known as interferometers. These instruments divide the light in semi transparent surfaces, producing two or more beams that travel different paths and then recombine.

In spectroscopy, the principal interferometers are those developed by the American physicist A.A. Michelson (1881) and by two French physicists, Charles Fabry and Alfred Pérot (1896) [2.8]. Figure 2-13 shows the two basic configurations. In the Michelson type of spectrometer two beams interfere, while multiple beams interfere in the Fabry-Pérot type of filters, as will explained in the next section.
Spectrometer Classification based on Optical Principle.

2.3.3.1 Dual Beam Interferometers

In the Michelson interferometer, illustrated in figure 2-14, an incident wavefront strikes an angled semi transparent mirror (called a beam splitter) and is divided into a reflected and transmitted wave.

These waves continue to their respective mirrors, are reflected, and return to
Introduction to Spectrometer systems

the semi transparent mirror. A common implementation of the Michelson interferometer has one movable mirror mounted on a slider so that length of the light path in that path can be varied If the total number of oscillations of the two waves during their separate paths adds up to be an integral number, the light from the two beams will add constructively. Destructive interference, with a difference in path length \( \Delta L \), gives minima of light intensity on the detector according to:

\[
I_o = I \left( 1 + \cos \left( \frac{2\pi \Delta L}{\lambda} \right) \right) \quad (2-15)
\]

For light with a spectral distribution \( S(\lambda) \) the intensity as a function of the path length can be written as:

\[
I(\Delta L) = \left( 1 + \cos \left( \frac{2\pi \Delta L}{\lambda} \right) \right) \int_0^\infty S(\lambda) d\lambda
\]

or:

\[
I(\Delta L, \lambda) = \int_0^\infty S(\lambda) \cos \left( \frac{2\pi \Delta L}{\lambda} \right) d\lambda
\]

This interference pattern \( I(\Delta L) \) is the Fourier cosine integral of the spectrum of the optical input signal. The interferograms are recorded while scanning \( \Delta L \) by means of the moveable mirror. The original input spectrum can be obtained again by calculating the inverse cosine Fourier transform of the measured interferogram:

\[
S(\lambda) = \frac{1}{2\pi} \int_{-\infty}^{\infty} I(\Delta L) \cos \left( \frac{2\pi \Delta L}{\lambda} \right) d\lambda \quad (2-17)
\]

The accuracy of the system is directly related to the ability to accurately measure the optical pathlength \( \Delta L \) [2,12]. For accurately measuring the displacement of the moving mirror, most Fourier spectrometers use a stable monochromatic reference light source, generally a He-Ne laser (\( \lambda = 633 \text{ nm or } \nu = 15800 \text{ cm}^{-1} \)), following the light path also. By counting the minima in the fringe patterns of this reference wavelength source a large series of accurate reference points are generated when the mirror moves.

Basically these types of spectrometers convert the power spectrum of an optical input signal into spatial information (fringe patterns) by a Fourier
transform. The minimum resolvable power is therefore limited by the number of samples taken for the FFT operation and the Nyquist criterion [2.12]. Nowadays fast digital processing such as FFT and spectral averaging using large blocks of data from the detector can be readily done using digital signal processing (DSP) circuits.

There are several types of Fourier Transform (FTIR) infrared spectrometers all based on the Michelson principle, some configurations are illustrated in figure 2-15.

![Different types of interference-based spectrometer systems](image)

**Fig. 2-15. Different types of interference-based spectrometer systems.**

a). Michelson.
b). Sagnac.
c). Mach-Zehnder.

**Advantages of Fourier Transform Spectrometers**

The first advantage of Fourier spectrometers is known as the Fellgett advantage. The Fellgett, or multiplex advantage arises because all of the spectral elements are measured simultaneously. Thus, a spectrum can be obtained very quickly. This leads to a better signal to noise ratio (SNR) for a given measurement time, as compared to a dispersive instrument. Alternatively, it makes the Fourier
transform systems a quicker technique to achieve the same quality spectrum. Fourier systems are better than dispersive systems for kinetic work, and in spectra where the signal-to-noise ratio is poor.

The second advantage of Fourier transform spectrometers is known as the Jacquinot advantage. The Jacquinot, or throughput advantage, arises because unlike dispersive spectrometers, Fourier spectrometers have no slits which attenuate the input light. This means the spectrometer has greater optical throughput, and again, a higher S/N ratio than dispersive infrared spectrometers.

The third advantage of Fourier systems is known as the Connes advantage. The Connes advantage arises because the frequency scale of the spectrum is known very accurately. This enables good control over the calibration in the wavelength domain and means additional operations such as spectral averaging are possible, allowing the collection of many spectra and their co-addition. The addition of many spectra increases signal to noise ratio in FTIR spectroscopy. An additional advantage of the accurate wavelength calibration is that since the wavelength is known, techniques such as spectral subtraction can be performed easily.

**Disadvantages of Fourier Spectrometer Systems**

Actually Fourier spectrometers do not measure spectra; they measure interferograms. Interferograms are light intensity patterns that are difficult to interpret without first performing the inverse Fourier transform to produce a spectrum. This used to be a major problem, but since the dramatic increase in computing power, performing a Fourier transform is now a quick and easy process. However, the way the interferogram is transformed can affect the results, and care must be taken.

In systems that are source noise-limited, the Fellgett disadvantage applies. This arises because all regions of the spectrum are observed simultaneously. Therefore, noise will be spread throughout the spectrum. In a dispersive system, the noise would be seen only in the region of the spectrum in which it arose. Fortunately, in mid-infrared systems, as most commonly used, the system is detector-noise limited. Therefore the Fellgett advantage applies, but not the Fellgett disadvantage.

Most importantly, FTIR instruments have a single beam, whereas dispersive
Spectrometer Classification based on Optical Principle.

Instruments usually have a double beam, as can be seen in the instrument of figure 2-7. For highly sensitive work or long experiments, changes in infrared absorbing gas concentrations can severely affect the results in single beam systems. Therefore, in these cases, when using an FTIR instrument, it is necessary to purge the instrument of CO₂ and water vapour using infrared transparent gases such as nitrogen or helium.

The resolution of Fourier Transform spectrometers cannot be calculated easily since, besides the number of samples in the Fourier Transform calculations, it depends on many other parameters and instruments settings [2.7].

2.3.3.2 Fabry-Pérot Spectrometers

If a thin transparent spacer is placed between two highly reflective coatings, multiple reflections occur between the two mirrors as shown in the Fabry-Pérot etalon of figure 2-16.

![Fabry-Perot etalon](image)

**Fig. 2-16. Fabry-Perot etalon.**

Constructive interference occurs if the transmitted beams are in phase, and this corresponds to a high-transmission peak of the etalon. If the transmitted beams are out-of-phase, destructive interference occurs and this corresponds to a transmission minimum. The transmission of a Fabry-Pérot device is given by [2.4]:

\[
T = \frac{1}{1 + F \sin^2 \left( \frac{\delta}{2} \right)}
\]

(2-18)

With: \( F = \frac{4R}{(1+R)^2} \) and: \( \delta = \left( \frac{4\pi n}{\lambda} \right) L \cos \theta \)
Where $R$ is the reflectivity of the mirror material and $F$ is called the coefficient of finesse. Narrow peaks are formed in transmission if $R$ approaches 1 and the sharpness of the peaks increases with reflectivity. The resolving power of a Fabry-Pérot device can be approximated by [2.13]:

$$R_p = \frac{\lambda}{\Delta \lambda} = m F = \frac{m \pi R}{(1 - R)} \quad (2-19)$$

A Fabry-Pérot interferometer differs from a Fabry-Pérot etalon in the fact that the distance between the plates can be tuned in order to change the wavelengths at which transmission peaks occur in the interferometer.

### 2.3.3.3 Non-dispersive Spectrometers

Non-dispersive spectrometers use the Fabry-Pérot filters, as described in the previous section. They are described here since they are the most basic types of spectrometers and are commonly used for gas detection in the mid-infrared [2.16] [2.17]. They have a simple and straightforward optical path, enabling low-cost and are typically used for dedicated systems in automotive applications [2.18] or in portable instruments [2.19] [2.20]. The term Non-Dispersive Infrared (NDIR) spectrometer refers to the fact that in these types of spectrometers all light passes through the sample and optical filtering takes place directly before the detector, as shown in figure 2-17.

![Fig. 2-17. A non-dispersive infrared spectrometer.](image)

The optical filter is tuned at the absorption peak of the sample that needs to be analysed. For instance the gas-cell setup of figure 2-17 can be used for the detection of the concentration of gases using a Fabry-Pérot interference filter with a narrow peak transmission at a wavelength $\lambda = 3.4 \, \mu m$ for hydrocarbons or for $CO_2$ at $\lambda = 4.26 \, \mu m$. IC-compatible interference filters for NDIR spectrometry
Conclusions

will be discussed in chapter 12.

2.4 Conclusions

In this chapter the common types of macro-size spectrometers and their basic properties have been discussed. A good comparison of the different principles has not been found in literature and has been included therefore in this chapter. Although it is difficult to compare the different types of spectrometers, some general features of the different types have been summarized in Table 1.

TABLE 1. Comparison of different spectrometer systems.

<table>
<thead>
<tr>
<th>Spectrometer type</th>
<th>Resolution</th>
<th>Light through put</th>
<th>Size</th>
<th>Cost</th>
<th>Microsystem compatible</th>
</tr>
</thead>
<tbody>
<tr>
<td>Prism</td>
<td>Low</td>
<td>Low</td>
<td>Very large</td>
<td>Moderate</td>
<td>--</td>
</tr>
<tr>
<td>Grating</td>
<td>High (adjustable)</td>
<td>Low</td>
<td>Large</td>
<td>high</td>
<td>+</td>
</tr>
<tr>
<td>Fabry-Perot</td>
<td>Very high (fixed)</td>
<td>High</td>
<td>Small</td>
<td>low</td>
<td>++</td>
</tr>
<tr>
<td>Fourier Transform</td>
<td>Very high</td>
<td>High</td>
<td>Very large</td>
<td>Very high</td>
<td>-</td>
</tr>
</tbody>
</table>

The effect of down scaling on the performance of the different types of spectrometers is a very important issue for microspectrometers in general. In section 2.3.1 has been shown that for a prism type of spectrometers the resolving power is mainly determined by the size of the prism and scaling down of the optical system will therefore result in a lower resolution. In section 2.3.2 it has been shown that for the grating types of the spectrometers the resolving power depends on the exit slit width, on the properties (size) of the grating and on the focal length of the system. In interference type of spectrometers a very short optical path is in principle possible. The optical path can be as short as a few wavelengths without affecting the interference principle. Alignment problems have a larger influence when using a very short optical path and can cause performance degradation.
An important conclusion from the discussions in this chapter is that basically the interference type of spectrometers are more suitable for scaling down than the diffractive or dispersive systems. In grating or prism type of spectrometers the light beams are deflected a certain angle. So a high resolution requires a large deflection angle or a long optical path, which generally complicates miniaturisation.

The ability of the various types of spectrometers to collect light at their input (also called: entendu) and its internal light efficiency is not considered here, since it is depends largely on the application how light is coupled and transported into the spectrometer.

In conventional spectrometers lenses, blazed gratings and focusing mirrors are used to collect and transport as much light as possible through the system to the detector. Fortunately, much effort is put nowadays into the development and production of microscale optical components [2.15]. Generally, it can be stated that scaling down microspectrometers requires a scaling down of the complete system, the input coupling, the optical path and the detector and in many applications also the sample, the sample holder and the light source.

### TABLE 2. Basic performance when down-scaling spectrometers.

<table>
<thead>
<tr>
<th>Spectrometer type</th>
<th>Resolution when downscaled</th>
<th>IC compatability</th>
<th>Ease of assembly</th>
</tr>
</thead>
<tbody>
<tr>
<td>Prism</td>
<td>Decreases</td>
<td>--</td>
<td>--</td>
</tr>
<tr>
<td>Grating</td>
<td>Decreases</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Fabry-Perot</td>
<td>independent</td>
<td>++</td>
<td>++</td>
</tr>
<tr>
<td>Fourier Transform</td>
<td>independent</td>
<td>-</td>
<td>--</td>
</tr>
</tbody>
</table>

The ability of the various types of spectrometers to collect light at their input (also called: entendu) and its internal light efficiency is not considered here, since it is depends largely on the application how light is coupled and transported into the spectrometer.

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### 2.5 References


Conclusions


Introduction to Spectrometer systems


3.1 Introduction

This chapter starts with a short introduction on the different ways in which light interacts with matter. The main application of mid-infrared spectrometers is in absorption and emission spectroscopy, and these techniques will be discussed subsequently. Finally an overview of the main applications of mid-infrared spectroscopy are given, and based on this, some potential applications of microspectrometers in the mid-infrared will be given.

3.2 Interaction of Light with Matter

Spectroscopy is the use of the absorption, emission, or scattering of electromagnetic radiation by materials to study atoms or molecules, or to observe physical processes. The interaction of radiation with matter can cause redirection of the radiated field and/or transitions between the energy levels of the atoms or molecules.
A transition from a higher level to a lower level is called emission if energy is transferred to the radiating field. The term non-radiative decay is used if no radiation is emitted.

A transition from a lower level to a higher level with transfer of energy from the radiation field to the atom or molecule is called absorption.

Redirection of light due to its interaction with matter is called scattering, and may or may not occur with transfer of energy, i.e., the scattered radiation has a slightly different or the same wavelength.

### 3.2.1 Emission

Atoms or molecules that are excited to high energy levels can decay to lower levels by emitting radiation (emission or luminescence). For atoms excited by a high-temperature energy source this light emission is commonly called atomic or optical emission, and for atoms excited with light it is called atomic fluorescence or molecular fluorescence. For molecules it is called fluorescence if the transition is between states of the same spin and phosphorescence if the transition occurs between states of different spin. The measurement of line spectra emitted by heated gases (sparks, flames) or plasma emission mainly occur in the UV or visible range, but emission spectroscopy also has many applications in the mid-infrared and some examples will be given in section 3.4.1.

### 3.2.2 Absorption

When atoms or molecules absorb light, the incoming energy excites quantized energy level structures to higher energy levels. The type of excitation depends on the energy required and hence on the wavelength of the light. Electrons are promoted to higher orbitals by ultraviolet or visible light, molecular vibrations are excited by infrared light, and molecular rotations are excited by microwaves. Infrared absorption spectroscopy or vibrational spectroscopy is the most important application of spectrometers in the mid-infrared and this technique will be discussed in this paragraph.

Infrared or vibrational spectroscopy exploits the fact that most basic resonant modes of molecules occur at wavelengths in the mid-infrared [3.6]. When two or more atoms are bonded, forming a molecule, many modes of mechanical vibrations are possible. A diatomic molecule can be roughly treated
as two particles bonded together by a spring, neglecting the effect of the electrons. Diatomic molecules only have a stretching mode of vibration. The frequency of the vibration is associated with the masses of the molecules and the strength and type of their bonds, leading to infrared absorptions at characteristic wavelengths. More complex molecules have many bonds and vibrations can be conjugated, leading to absorption peaks that can be related to groups of materials. Figure 3-1 shows the six different modes of vibration of a triatomic molecule.

![Fig. 3-1. Molecular vibration modes.](image)

The Newtonian mechanical model is a very simple approximation. Better and more accurate models are based on quantum theory [3.6]. Infrared absorption spectroscopy is the most commonly used method to identify various types of bonds and groups of materials in poly-atomic molecules.

### 3.2.3 Scattering

When electromagnetic radiation passes through matter, most of the radiation continues in its original direction but a small fraction is scattered in other directions. Light that is scattered at the same wavelength as the incoming light is called Rayleigh scattering. Light that is scattered due to vibrations in molecules or optical phonons in solids is called Raman scattering. Raman spectroscopy is also widely used in the infrared and will be briefly discussed in this section.

When light is scattered from a molecule, most photons are elastically scattered. The scattered photons have the same energy (frequency) and, therefore, wavelength, as the incident photons. A small fraction of light (approximately 1 in
$10^7$ photons) is scattered at optical frequencies different from the frequency of the incident photons. This inelastic scatter is the Raman effect. The difference in energy between the incident photon and the Raman scattered photon is equal to the energy of a vibration of the scattering molecule. A plot of intensity of scattered light versus energy difference gives the Raman spectrum of a material. A Raman spectrometer therefore, requires a very high power laser source to generate sufficient inelastic scattering. Applications of Raman spectroscopy in the infrared are still hampered by the cost and size of high power mid-infrared laser systems. The fabrication of micro-scale size Raman spectrometers is not feasible at present and Raman systems are not discussed in more detail here.

### 3.2.4 Interpretation of Infrared Spectra

In nearly all applications the interpretation infrared spectra is preferred in terms of wavenumbers and not in wavelengths. Unfortunately, two different definitions of the wavenumber exist in physics. In theoretical physics it is common to use the definition of wavenumber or wave vector as:

$$\nu = \frac{2\pi}{\lambda} \quad (0-1)$$

In spectroscopy the wavenumber $\nu$ is expressed in cm$^{-1}$ and is related to the wavelength $\lambda$ by

$$\nu = \frac{10^4}{\lambda [\mu m]} \quad [c m^{-1}] \quad (0-2)$$

Figure 3-2 shows the infrared absorption of ethanol as an example of a typical spectral plot.
Pronounced absorption peaks for this material can be observed at wavenumbers:

- O-H stretch: a broad band at $\nu = 3391$ cm$^{-1}$ (typical for alcohols)
- C-H stretch: at $\nu = 2951$ cm$^{-1}$ and bending at $\nu = 1450$ cm$^{-1}$.
- C-O stretch: at $\nu = 1102$ cm$^{-1}$ and $\nu = 1055$ cm$^{-1}$.

Not all molecular vibrations modes lead to observable infrared absorption peaks, e.g. for symmetric molecules some of the symmetric modes of vibration may be absent in the infrared spectrum. In general, a vibration must cause a change in the charge distribution within a molecule to absorb infrared light. The larger the change in charge distribution, the stronger the absorption.

### 3.2.5 The Fingerprint Region

The absorption spectra of region of 1-propanol and 2-propanol are shown in figure 3-3. Both compounds contain exactly the same bonds. Both compounds have very similar troughs in the left hand side of the figure, showing the presence
of an oxygen-hydrogen bond. The region to the right-hand side of the diagram (from about $\nu = 1500-500 \text{ cm}^{-1}$ or $\lambda = 5-20 \mu \text{m}$) usually contains a very complicated series of absorptions. These are due to all modes of bending vibrations within the molecule. This is called the **fingerprint region**.

![Infrared spectra of two types of propanol](image1)

**Fig. 3-3. Spectra of two types of propanol (from [3.8]).**

It is much more difficult to pick out the individual bonds in this region than at the higher wavenumbers. The importance of the fingerprint region is that each different compound produces a different pattern of troughs in this part of the spectrum. The pattern of compound materials in the fingerprint region is completely different and could therefore be used to uniquely identify various compounds. Spectra of many materials or groups of materials are both free and commercially available in databases on the Internet [3.7] [3.8] [3.9] [3.10].
3.3 Sample Interaction

There are several methods for interaction of samples with infrared light. Measuring the transmittance and/or reflectance spectrum of samples are the most common techniques in spectroscopy. Specular reflectance of a surface is the direct mirror-like reflection while diffuse reflection includes the light reflected in all directions. For both transmission and reflection we can distinguish between *specular* and *diffuse* measurement techniques. Diffuse reflectance or transmittance requires the collection of the scattered light in all possible directions. This is done by placing the sample at the entrance of (for diffuse transmittance) or inside an *integrating sphere* (for reflectance) as shown in figure 3-4.

Two other commonly used infrared spectroscopic techniques are photo-acoustic and total reflection techniques. In *photo-acoustic spectroscopy* [3.1] the absorbed energy is transformed into kinetic energy, resulting in local heating and thus a pressure wave or sound. By measuring the sound pressure at different wavelengths an absorption spectrum of a sample can be recorded.

*Attenuated Total Reflection spectrometry* [3.13] is based on the interaction of light with a sample when reflection occurs at the boundary of a medium. Reflection at a boundary surface produces an electric field, called the evanescent wave. In case of an absorbing sample, the intensity of the reflected light decreases. Scanning of the spectrum provides information about the type and amount of absorption in the sample as a function of the wavelength. An important property of Attenuated Total Reflection spectroscopy is, since it is a surface measurement technique, the ability to measure highly absorbing samples. The measurement of absorption spectra by Attenuated Total Reflection will be

![Diffuse reflectance and transmittance setups.](image)
discussed in more detail in chapter 11.

### 3.3.1 Qualitative Analysis

The Beer-Lambert law is a relationship that relates the absorption of light to the properties of the material through which the light is travelling. The transmission $T(\lambda)$ through a material with length $L$ and an incident light intensity $I_o$ is given by [3.1]:

$$T(\lambda) = \frac{I(\lambda)}{I_o(\lambda)} = 10^{-\alpha L c}$$

with $\alpha(\lambda)$ is the absorption coefficient or molar absorptivity:

$$\alpha(\lambda) = \frac{4\pi k(\lambda)}{\lambda}$$

and $k(\lambda)$ is the complex refractive index or extinction coefficient.

The units of absorber concentration $c$ and absorption coefficient $\alpha$ depend on the material. For a solid the extinction coefficient $k$ is generally used. If the material is a liquid, the absorber concentration is usually expressed as a mole fraction. If the concentration is expressed in moles per unit volume then $\alpha$ is a molar absorptivity (usually indicated as $\varepsilon$) in units of mol$^{-1}$cm$^2$. In the case of a gas, the concentration may be expressed as a number density (cm$^{-3}$), in which case $\alpha$ is the absorption cross-section (cm$^2$).

### 3.3.2 Sample Preparation

Gaseous samples require little preparation beyond purification and a sample cell with a long pathlength (typically 5-10 cm) is normally needed as gases show relatively weak absorbances [3.11]. Extra gases that have low absorption in the infrared such as nitrogen of helium can be added.

Liquid samples can be placed in a special cuvette (quartz) or sandwiched between two plates. The sample holder should be infrared transparent and should not introduce extra any lines onto the spectra. Common materials for these plates or cuvettes are potassium bromide (KBr) or calcium fluoride (CaF$_2$) in the experiments in this thesis silicon and ZnSe have been used as infrared transparent materials.
Applications of Infrared Spectrometer Systems

Solid samples are prepared either by grinding and mixing with an anhydrous (without water) solvent forming a paste on the sample plate or by crushing and mixing the sample with a specially purified salt (usually potassium bromide: KBr) finely (to remove scattering effects from large crystals). A translucent pellet through which the beam of the spectrometer can pass is formed from the powder mixture in a mechanical die press. It is important to note that spectra obtained from different sample preparation methods will look slightly different due to differences in the physical state of the samples [3.12].

3.4 Applications of Infrared Spectrometer Systems

Important applications of infrared technology nowadays are in infrared vision systems [3.14] and spectrometers [3.1] [3.2] [3.3] [3.4]. Infrared camera systems for vision at night, in fog or in smoke [3.14] have many applications, e.g. in the military, the automotive industry [3.16][3.17], for fire-fighting [3.18] and personal safety [3.19]. Thermographic camera systems, where the temperature profile of objects is accurately measured by measurement of the spectral distribution of the emitted infrared radiation of that object, can be considered as a combination of vision and spectrometer system [3.15]. There are many applications for mid-infrared vision systems in medical analytical systems. [3.20] [3.21] [3.22] [3.23]. Most camera systems operate in the near infrared, using special CCD camera detectors [3.5]. The limited use of camera systems in the mid- or long-infrared range at present is still governed by the price of the detectors.

Low cost mid-infrared vision systems definitely would enable new applications and open new markets. The reason for the high cost of infrared systems is partly in the cost of the materials used for infrared optics, but even more in the price of the detectors. Many vision applications would benefit greatly from the availability of small, low cost detector arrays in the mid-infrared. Chapter 6 in this thesis is therefore dedicated to the design and optimization of thermal infrared detectors. The next sections give a short overview of the emerging applications of mid-infrared spectrometer systems.
3.4.1 Infrared Emission Spectroscopy Applications

In emission spectroscopy the spectral distribution of an infrared light is analysed. Mid-infrared telescopes with emission spectrometers are used in astrophysics to identify objects in space or analyze the emission spectra of stars, galaxies etc. The generally narrow spectral lines from heated gases in flames, sparks or plasmas can give information on the type and composition of gases. Generally these peaks are in the ultraviolet and visible spectral range, however some materials also give spectral lines in the infrared. Other applications are gas sensing during Plasma Etching (HCL, HBr) in semiconductor processing [3.24][3.25].

3.4.2 Infrared Absorption Spectroscopy Applications

Mid-infrared spectroscopy is the most common method to identify various types of bonds and groups of materials in poly-atomic molecules. This is the reason for the popularity of infrared spectroscopy in both organic and inorganic chemistry. It is a well-established technique and many books and journals are devoted to the theory, developments and applications of infrared spectroscopy in chemistry, material research and physics.

Traditional infrared spectrometers instruments in analytical laboratories are universal instruments that need trained personnel for operation. These systems are hard to compete with in terms of reproducibility and accuracy. The advances in sensors, signal processing and optics for spectrometers allows scaling down the conventional spectrometer systems, making them portable. As an example figure 3-3 shows a commercial miniature spectrometer with a fibre optic input that has a typical size of less than 10 cm.

![Image of a spectrometer](image.png)

Fig. 3-5. The USB2000 spectrometer (Courtesy of Ocean Optics).
Applications of Infrared Spectrometer Systems

The system shown is modular and can be configured by different gratings for operation in the UV to the Near-infrared, for different applications. There is an increasing demand for low cost, small and dedicated systems for many different applications. The competition of small portable microspectrometers with the large bulky, expensive, laboratory instruments is a great challenge.

Some examples of the use of mid-infrared spectrometers are:

- The measurement of water content in samples, by measuring the absorption peak of water at $\lambda = 2.94 \, \mu m$, e.g. in agriculture [3.26] or soil biology [3.27].
- The detection of frost, snow and ice in remote sensing [3.28][3.32].
- The measurement of combustion and greenhouse emissions by measuring the very high absorption peak of CO$_2$ at $\lambda = 4.23 \, \mu m$ [3.33]
- Typical gases for the monitoring of combustion or for chemical or biological processes are O$_2$, H$_2$O, CO, CO$_2$, NO, NO$_2$, NH$_3$, H$_2$S and CH$_4$. Most of these gases have pronounced absorption peaks in the mid-infrared [3.6].
- Many different materials need to be analysed in the food and agricultural industry. Typical applications in the food industry are in the detection of oils or fat [3.44], the analysis of diary products such as butter and cheese [3.45] or the ripening [3.29] or freshness of food [3.30] by measurement of water content. Other applications are e.g. in the measurement of fluorescence spectra to monitor the photosynthetic efficiency of plants [3.31].

Even larger opportunities of microspectrometers are in systems where only very small sample volumes are available or the systems need to very be small. Volume and weight are decisive issues for spectrometers in a satellite.

Besides replacement of the conventional spectroscopic instruments as mentioned in the applications above, key applications for microspectrometers in the mid-infrared are therefore also in the fields of:

- Forensic analysis [3.37]: For the analysis of small samples of textile fibres, hair, drugs, toxic materials, explosives, paint, toners, inks etc.
- Analysis of pharmaceutical products [3.38]
3.5 Conclusions

In this chapter infrared vibrational spectroscopy has been introduced as an important application of mid-infrared spectrometers. In absorption spectroscopy a well-defined infrared source is needed in the instrument and infrared light sources are therefore considered an integral part of the spectrometer. Microspectrometer infrared light sources will be discussed in Chapter 6. Interaction with the sample and some typical infrared absorption spectra have been discussed. Finally a brief overview of the applications of infrared spectroscopy has been given and some potential opportunities for mid-infrared spectrometers have been identified.

3.6 References

Conclusions


[3.9] http://webbook.nist.gov/chemistry, National Institute of Standards and Technology, NIST, 100 Bureau Drive, Gaithersburg, MD 20899-1070, USA.


Conclusions


Mid-infrared Spectroscopy and its Applications


IC-compatible Infrared Microspectrometers

4

Best wide-angle lens?
Two steps backward.
Ernst Haas, Photographer

4.1 Introduction

This chapter presents the state-of-the-art in the field of Mid-infrared microspectrometers, with an emphasis on those systems that can be realized using integrated circuit technology. Spectrometers, for the near-infrared or visible part of the spectrum and can be extended for operation in the mid-infrared, will be discussed also. Since the research field of optical MEMS is rapidly expanding in many directions the overview presented here is intentionally restricted to the basic options.

The previous chapters have already discussed the major types and principles of conventional spectrometers, some commercial realizations, and their major applications. Much on-going research is performed on the individual components for miniaturized spectrometers, e.g. in the field of micro-optics [4.1] [4.2], integrated light sources (solid state lasers) and detectors. Not many articles describe the design, fabrication of complete systems in the infrared and based on integrated circuit technology.

This chapter starts with a short overview of the optical properties of IC-compatible materials in the infrared. The optical properties of IC-compatible
IC-compatible Infrared Microspectrometers

Materials such as silicon, polysilicon, metals and isolating layers such as SiN and the various types of oxides play a crucial role in the design of the optical path of a spectrometer system. Fabrication of the optical components in a microspectrometer system in IC-compatible technology has many advantages, since it enables the fabrication of fully integrated systems. Advantages of these systems are e.g. an accurately defined optical path, small size and low cost. Chapter 11 and 12 describe two realisations of such fully integrated microspectrometers.

Subsequently an overview of recently published infrared microspectrometers, based on their basic operating principle (as described in Chapter 2) is given in this chapter.

4.2 Optical properties of IC-compatible Materials

In IC technology the bulk material is silicon and thin layers of metals (Aluminium), polysilicon and many glassy dielectric materials are available. Silicon has a negligible absorption in the mid-infrared, however as in most optical systems, behaviour at the interface, where the infrared beam meets another medium are of crucial importance.

4.2.1 Infrared Optical Properties of Silicon.

Silicon has a very high refractive index \(n=3.42\) compared to the common optical materials used in the visible part of the spectrum [4.3]. In infrared optics it is, together with germanium \(n=4.0\), one of the materials with the highest refractive index and commonly used because of its hardness and chemical resistance. Figure 4-3 shows that the refractive index of crystalline silicon is almost constant in the Mid-infrared region. The dispersion in silicon is very low and therefore silicon is not a suitable material for dispersive devices such as prisms.

The reflection loss at the interface between two isotropic transparent media depends on the refractive index, the incident angle and polarization. The large step in refractive index of silicon with respect to air or free space, causes high
reflection losses at the interface.

Typically the reflection losses for a silicon/air interface at normal incidence are given by:

\[ |R| = \left| \frac{n_{\text{Si}} - n_{\text{Air}}}{n_{\text{Si}} + n_{\text{Air}}} \right| = \frac{3.41 - 1}{3.41 + 1} = 0.55 \]  \hspace{1cm} (4-1)

Germanium and silicon lenses or windows are therefore generally coated with thin anti-reflective layers, to increase the transmission through the interface in a certain wavelength interval. Figure 4-3 shows the imaginary part of the refractive index \( k \), or also called extinction coefficient [4.5]. The absorption \( \alpha \) in a dielectric material is related to the extinction coefficient \( k \) by

\[ \alpha = \frac{4\pi k}{\lambda_{\text{Si}}} , \]  \hspace{1cm} (4-2)

where \( \lambda_{\text{Si}} \) is the wavelength in the medium. The figure shows a very low absorption of silicon exceeding wavelength of 1.1 \( \mu \)m and silicon can be considered a highly transparent material in the infrared. Hence, the bulk silicon of a wafer can be used to in the optical path in the infrared spectral range. Poly-silicon has a slightly higher refractive index than crystalline silicon and a much higher absorption coefficient in the infrared than crystalline silicon [4.5] [4.11].

Fig. 4-1. Refractive index of silicon as a function of wavelength..
4.2.2 Infrared Optical Properties of Metal Interconnect.

Practical IC-compatible materials, such as silver, copper, gold and aluminium, show a high reflectivity over a broad spectral range in the visible and infrared, as shown in figure 4-3. Aluminium is the standard material for the metallisation process both in integrated circuits (IC’s) and in Micro Electro-Mechanical Systems (MEMS). The high reflectivity of aluminium for visible and infrared wavelengths enables the realisation of high quality mirrors compatible with standard IC processing.

The reflectance of aluminium at infrared wavelengths rapidly increase at film thicknesses above 5 nm [4.6]. Consequently, aluminium is a very suitable material in terms of fabrication compatibility and simplicity, while it shows high reflectance at the thickness used in standard metallisation processes. The reflectance is even higher than that of silver or gold in the near-infrared spectral region [4.6]. Hence aluminium can be used as a highly reflecting material for light shielding and reflection.
4.2.3 Optical properties of Siliconoxides and Siliconnitrides.

Dielectric layers such as siliconoxides and siliconnitride always have a refractive index lower than crystalline silicon. Optical data of these materials in the visible range and near-infrared can be found in literature and in some databases for thin-film calculations. Available data on the optical properties of these material in the mid-infrared is scarce. Bulk crystalline SiO₂ (quartz) has a reasonable transparency up to wavelengths of 20 μm [4.3] while other IC fabricated glasses such as TEOS and (B)PSG generally can be used as low loss materials up to 1.6 μm [4.7] i.e. in the visible or near-infrared. This limits the use of these materials for near-infrared waveguides.

Optical interference filters as, described in chapter 12, use very thin layers, typically in the range of hundreds of nanometers and absorption plays a much lesser role in these layers. Literature on silicon oxides thin-films [4.7][4.8] shows a large absorption peak at λ = 11 μm, due the Si=O2 double bond and a smaller peak at λ = 7 μm due the Si-O bonds. Silicon nitride films have an broad absorption peak around λ = 10 μm [4.9][4.10] and low loss up to λ = 8 μm [4.4]. As a result both materials are considered to be useable for thin-films for wavelengths up to λ = 8 μm [4.4].

Fig. 4-3. Spectral reflectance of silver, gold and aluminium.
4.3 Infrared Microspectrometers

The next paragraphs give an overview of published articles on miniaturized spectrometers. There are only a few overview articles on integrated spectrometers [4.12][4.13][4.14]. Most articles discuss spectrometers or parts of spectrometers in the visible range. Most systems are restricted by the use of glasses as the optical material to the near-infrared. Only a few articles discuss mid-infrared spectrometers.

4.3.1 Grating Spectrometers

A spectrometer based on a transmission grating and multiple reflections on bulk-micromachined wafer for the visible range has been published by Kwa [4.15]. Miniature VIS spectrometers based on different types of micromachined gratings on top of CCD camera have been published by Yee [4.16].

A mid-infrared microspectrometer based on a transmission grating shown in figure 4-4 has been described by S.H.Kong et al. [4.18] [4.19].

A resolution of 200 nm in the range 2-5 μm has been obtained with this device. A much higher resolution has been obtained by the grating spectrometer in the visible range of Grabarnik et al. [4.20] shown in figure 4-4.
Planar gratings in IC technology that can be used in integrated spectrometers have been reported by Goldman [4.27], Mohr [4.24] and Muller [4.26]. Hocker et al. developed a MEMS programmable diffraction grating [4.17] consisting of an array of electrostatically actuated long (1cm) and thin (10 μm), reflective beams. Figure 4-6 shows a MEMS tuneable binary grating published by Manzardo/de Rooij [4.30].

---

**Fig. 4-5. Double grating Microspectrometer (Grabarnik [4.20]).**

**Fig. 4-6. Tuneable binary grating (Manzardo [4.18])**

\[ a \). Principle: Wavefront shaping by a binary grating. 
\[ b \). Detail of the movable binary grating. \]
IC-compatible Infrared Microspectrometers

The depth of the binary grating can be varied mechanically by an electrostatic comb drive actuator. The devices have been fabricated using DRIE etching of SOI wafers and a spectral resolution of 6 nm at a wavelength of 633 nm has been reported. Recently the design and simulations on an infrared spectrometer based on an etched Bragg grating, operating in the range 1-10 μm, has been presented by Butler et al. [4.19].

A Near-infrared (λ = 1.0-1.9 μm) microspectrometer based on a reflective grating placed on a micromachined comb-drive has been developed and published by Fraunhofer IPMS [4.21]. The detector is a single InGaAs infrared photodetector and the spectrum is scanned by electrostatic actuation of the grating. This spectrometer is commercially available [4.22] and has a resolution of 10 nm and a size of 10x8x8 cm³.

A Near-infrared spectrometer based on a blazed grating and bulk micromachining has been reported by Zhou [4.23]. A grating with Vgrooves has been fabricated by anisotropic etching of silicon and gold-polysilicon thermopiles have been used for the detectors.

4.3.2 Fourier Transform Devices

Miniaturized Fourier transform infrared spectrometers based on silicon technology and the Michelson principle have been reported by Manzardo using SOI MEMS [4.28] [4.29], by Solgaard [4.31] and by Mohr using the LIGA technique [4.33] [4.34]. An example of a near-infrared Fourier spectrometer is shown in figure 4-7.
4.3.3 Fabry-Pérot Microspectrometers

The largest number of microspectrometers, as have been reported in literature, are the Fabry-Pérot type of interference devices. Many different technologies have been used for the fabrication of different types of devices.

Arrays of 16 Fabry-Pérot etalons in a CMOS compatible process, as shown in figure 4-8, have been published by J.H. Correia et al. [4.35] [4.36] [4.37]. Although this system is designed for the visible wavelength region, the system could be realised in the infrared also, by replacing the photodiode array with an array of infrared detectors and a redesign of the Fabry-Pérot etalons. Other systems base on fixed Fabry-Pérot resonant devices have been reported also by [4.46] and [4.47].
Tuneable Fabry-Pérot devices can be realized by a cavity formed by a moveable and a fixed mirror. A typical device is shown in figure 4-9. Fabrication of these devices by bulk micromachining and wafer bonding has been published by Jerman [4.38], Raley [4.39], Correia [4.41], Kung [4.45] [4.46] and some others. Infrared Fabry-Pérot devices with detectors have been fabricated by Rossberg [4.44] and Musca [4.42] [4.43].

Fabry-Pérot spectrometers based on surface micromachining techniques have been published by Wolffenbuttel [4.49] and Tran [4.48]. A thermally actuated Fabry-Pérot device using porous silicon has been published by Renaud [4.50]. IC-compatible Fabry-Pérot etalons based on dielectric mirrors for the
Mid-infrared have been published and fabricated by Emadi et al. [4.56]. Chapter 12 of this thesis discusses these IC-compatible thin-film optical filters.

Linear Variable Filters, or Wedge Filters are basically Fabry-Pérot etalons with a linearly tapered thickness of the resonance cavity, as illustrated in Figure 4-11.

The transmission peaks vary therefore also linearly at different positions along the device. Macro scale LVF optical filters are expensive and commercially available in the visible range. Recently they are used in combination with CCD camera detectors in some portable visible range spectrometer systems. An infrared gas sensor based on an integrated LVF device has been published by Hara et al. [4.57].

4.3.4 Waveguide Devices

Other interesting miniature infrared spectrometer systems are based on waveguides. An extensive overview of interferometers based on waveguides is given by Lambeck [4.51]. Figure 4-11 shows the typical setup of a waveguide spectrometer [4.27].
IC-compatible Infrared Microspectrometers

A commercially available microspectrometer [4.25] system in the visible range is shown in figure 4-12. Light is coupled in by a fiber to a silicon waveguide based on the LIGA technology, as has been published by Mohr et al. [4.34].

A curved self-focussing grating has been fabricated in the waveguide resulting in a higher efficiency of the grating. The advantage of this system is the high efficiency of the grating.

A miniature mid-infrared gas sensor based on a waveguide has been described by [4.52]. The gas-cell is formed by a hollow waveguide made of Ag/
Ag-halide. The infrared interacts with the gas sample by the evanescent field as described in chapter 11 of this thesis. The device is not IC-compatible but the principle could be applied using silicon micromachining also. A spectrometer based on a silicon waveguide for measuring liquids in the visible has been presented by [4.53]. In this thesis a fully integrated spectrometer, using a bulk silicon wafer as a waveguide and based on evanescent wave detection is presented in chapter 11 [4.55].

4.4 Conclusions

An overview has been given on both the infrared optical properties of the main materials used in IC processing and current research on micro-scale spectrometers. An important conclusion is that silicon has very low loss above wavelengths \( \lambda > 1.2 \mu m \) and can be used as a high quality material for guiding mid-infrared light.

Since most spectrometers are designed with an application in mind, the different papers on microspectrometers rarely give figures of merit or a comparison of performance with existing devices. Often it is possible to identify a few characteristic properties of the design that limits its performance on any of the following: Throughput, resolution, spectral range, switching time, multiplex advantage, contrast ratio, and simplicity of operation. Some of these figures of merits would be easy to compare in a table, but others are very complex and difficult to define in order to make a strict and just comparison. Similarly, many micro-scale spectrometer systems are not intended to be a competitive alternative to laboratory Fourier transform instruments, but may be tailored to solve specific problems, where one can accept a lower throughput, resolution or free spectral range.

4.5 References

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Mid-Infrared Microspectrometer Systems
5.1 Introduction

Arrays of high-sensitivity thermal sensors and low power thermal actuators require MEMS structures with high thermal isolation. The design of these devices is mainly a trade-off between thermal and mechanical properties of the final device. More specifically, the isolation can be improved by removing as much as possible of the surrounding material of the sensor or actuator reducing the mechanical rigidity of the devices. The next section discusses the different geometries of thermally isolated MEMS structures. Subsequently, some commercially available MEMS processes will be discussed and the following sections are devoted to the DIMES processing technology.

5.2 Thermal Isolation

Thermal isolation is a common issue for all thermal infrared detectors and improving thermal isolation can dramatically improve the performance of these devices. MEMS technology can improve thermal isolation by removing material in critical paths. The basic methods are placing the sensor on micromachined thin
Fabrication of Thermally Isolated MEMS Structures

membranes and removing the bulk silicon (Bulk micromachining) underneath or by the fabrication of thin structures built on sacrificial layers above a silicon substrate (Surface micromachining).

5.3 Available MEMS Processes

Basically two different MEMS techniques can be distinguished, i.e. Bulk Micromachining and Surface Micromachining [5.1]. There are several foundries that offer multiproject prototyping in surface micromachining technology, e.g. the XM-SP process of XFab [5.4]. Very large volume commercial applications of these devices are in accelerometers [5.4], adaptive optics [5.5] and pressure sensors [5.6]. Surface micromachining allows the fabrication of small and thin structures above a wafer. Sacrificial layers define the thin (generally a few μm) spacing between these structures and the substrate. This implies that the thermal loss due to conduction via the underlying air layer will be dominant. Nowadays there is a strong trend towards integration of CMOS and surface micromachined MEMS processes, available for commercial but also for Multi-Project series by eg. XFab, CSMC and MEMSCAP [5.4][5.18][5.10].

High thermal isolation of these structures is only possible if the thermal conduction to the bulk is minimized by vacuum encapsulation. There are several commercial foundries that offer vacuum encapsulation both on the die, and on the wafer level [5.3][5.10]. Another approach is to combine surface and bulk micromachining by removing the bulk underneath the surface micromachined structures as provided by e.g. the MetalMUMPS or SOIMUMPS processes of MEMSCAP Inc.[5.7][5.8] or the MultiMEMS process from Infineon SensoNor [5.9]. Focal plane arrays, i.e. arrays of resistive thermal sensors such as bolometers, can be fabricated as infrared camera arrays, using surface micromachining on top of CMOS technology. Surface micromachined infrared focal plane arrays using Schottky-barrier photodetectors have been reported in literature [5.2]. Fabrication of thermopiles however requires isolated junctions of two materials with different Seebeck coefficients. At present this is not possible in the multi-project processes available. Another approach is the fabrication of thin isolated membranes by bulk micromachining only, eg. the XM-BT process of XFab [5.4]. Commercial processes using these MEMS technologies generally produce closed membranes for pressure sensors applications and do not allow extra process steps for the cutting of membranes into cantilevers or bridges,
Available MEMS Processes

which is required for highest isolation or to avoid crosstalk between neighbouring elements in an array. As a result of all above issues the bulk micromachining technology for the fabrication of thin silicon nitride membranes [5.14] developed by the DIMES institute was choosen.

Figure 5-1 show some possible structures for thermal isolation of the sensor or actuator from the environment using bulk micromachining

![Fig. 5-1. Some thermally isolated MEMS structures](image)

- b). Rectangular closed membrane
- c). Bridge type ( dual-clamped beam)
- d). Hinged cantilever beam.
- e). Cantilever beam.
- f). Corner hinged structure
- g). Crableg hinged structure

For sensors that require only a few wires, such as bolometers, highest thermal isolation can obtained using hinged structures. Thermopiles require many wires to generate a sufficient voltage level and therefore bridges or cantilevers are always used for thermopile sensors.
5.4 Fabrication Process

The first series of thermopiles and infrared emitters (batch EI913) have been fabricated in an available process at the DIMES institute for the fabrication of integrated thermoelectric coolers based on polycrystalline silicon germanium [5.11]. Some large heater wires and thermopiles using N- and P-type PolySiGe strips on thin SiN bridges and cantilevers have been successfully fabricated in the process described in [5.12] and [5.13]. Processing of the next series of devices, especially designed for the microspectrometer system (batches EI-1305 and EI-1448) has been described in [5.14]. From the run EI1305 it has been found, that the final RIE etch step is the most critical in terms of mechanical stresses on the structures. In the run of EI1448 two different approaches have been used for the final MEMS steps. The preprocessing steps are described in the next paragraph.

5.4.1 Basic Processing Steps

The fabrication process flow chart is based on the DIMES process described in [8] and shown in figure 5-2.

1. Deposition of 800 nm low-stress silicon nitride film by low pressure chemical vapor deposition (LPCVD) on P-type 2-5 ohm-cm double polished <100> wafers without annealing and the stress is measured.
2. Deposition of a 300 nm layer of polysilicon by LPCVD is applied and subsequently the stress level is measured.
3. The front side of the wafer is then processed further by ion implantation using the PP mask with Boron with 40 keV, and 5x10^{15}/cm^2 for the P-type PolySi. After a cleaning procedure the NP mask is applied and the N-type PolySi is implanted with Phosphor with 40 keV, and 7.5x10^{15}/cm^2.
4. Next the PolySi layers are patterned using the PE mask and stripped from the front and backside of the wafer and the stress is measured.
5. Activation of the PolySi layers by annealing for 35 mins at 1000°C.
6. Next a 300nm layer of LPCVD TEOS oxide or 300 nm low stress SiN is deposited by LPCVD to make the isolation between metal contact and the PolySi.
7. Subsequently a 350 nm SiN layer for the bulk etch is applied on the back side of the wafer and dry etched, using the KOH mask.
8. The CO mask is applied for making contacts between the PolySi and the metallization applied in next steps and for thinning the SiN for RIE etching of trenches in the membranes later.

**Fig. 5-2. The flow chart of the preprocessing.**
9. Subsequently aluminium interconnect (Al/1%Si) is deposited and patterned and etched using the IC mask.

10. Final step of the postprocessing is the application of the RIE etch mask, defining the windows where trenches in the membranes should be formed by reactive ion etching.

Two different techniques have been applied for the MEMS postprocessing on run EI-1305. Anticipating on the fragility of the membranes during the final steps the batch of wafers have been split in an A and a B series and are differently processed as described in the next two sections.

### 5.4.2 Postprocessing MEMS steps (option A)

In the postprocessing sequence of series-A is as follows:

- Trenches in the membranes are first dry etched from the frontside through the SiN layers.
- Next the bulk is removed from the backside by a wet etch.

These steps are shown in figure 5-4. Postprocessing start with a RIE using the windows as defined already in the preprocessing steps:

1. RIE etching on the front side. Figure 5-3 shows a SEM photograph of a resulting trenches for isolation of the thermopile arrays and the heater wires. After this the front side of the wafer is protected by a KOH resistant photoresist.

![Fig. 5-3. SEM photograph of the RIE etch.](image-url)
2. Next the bulk is wet etched from the back side using 33 wt % KOH solution at 85 °C. The wafer is mounted in a special holder in order to prevent the damage to the front side structures during most of the etching time.

3. The KOH stops on the nitride and finally the photoresist is removed and the wafer is cleaned.

5.4.3 Postprocessing MEMS steps (option B)

In the postprocessing sequence of series-B wafers the bulk is removed first by a KOH etch stopping on the nitride membrane or with a timed etch stop leaving a few μm of silicon better mechanical strength. The final MEMS step is a RIE etch from the frontside forming trenches in the membranes as shown in

*Fig. 5-4. The flow chart of the A series postprocessing.*
Postprocessing starts with bulk etching on the wafer backside using a 33wt % KOH solution at 85 °C in the windows as defined already in the preprossing steps.

2. After this the structures in the membranes are released by a reactive ion etch using the windows as defined already in the preprossing steps.

Each wafer holds a pattern of 4 x 4 identical structures of 15x15 mm as shown in the photo of a final wafer of series EI1448 in figure 5-6.

Figure 5-7 shows a linear array of 23 thermopile pixels with a total length
Measurement results

of 1060 μm and a spacing of 10 μm on a wafer of run EI1448.

5.5 Measurement results

In run EI1448 layer stresses have been analysed by wafer bow measurements. Table 1 summarizes the results of these measurements giving the both the total stress and the stress for each layer.

**TABLE 1. Stress measurement results of the EI1448 layers**

<table>
<thead>
<tr>
<th>Wafer batch: EI 1448 Layer</th>
<th>Total Stress [MPa]</th>
<th>Layer Stress [MPa]</th>
</tr>
</thead>
<tbody>
<tr>
<td>SiN (Both sides)</td>
<td>σ_{2SiN}</td>
<td>7.9</td>
</tr>
<tr>
<td>SiN (1 side)</td>
<td>σ_{SiN}</td>
<td>158.2</td>
</tr>
<tr>
<td>PolySi on SiN 2 sides</td>
<td>σ_{2PSiN}</td>
<td>365</td>
</tr>
<tr>
<td>PolySi on SiN 1 sides</td>
<td>σ_{PSiN}</td>
<td>433</td>
</tr>
<tr>
<td>After anneal</td>
<td>σ_{PSiNA}</td>
<td>711</td>
</tr>
<tr>
<td>SiN (Both sides) second</td>
<td>σ_{2SiN2}</td>
<td>2201</td>
</tr>
<tr>
<td>SiN (1 side) second</td>
<td>σ_{SiN2}</td>
<td>2388</td>
</tr>
</tbody>
</table>

*Fig. 5-7. Photograph of a thermopile detector array.*
A negative value indicates a compressive stress. The table shows a large tensile stress around 2.3 GPa is present after the deposition of the second nitride layer. The fracture stress of thin SiN layers as reported in [5.19] [5.20] [5.21] is typically a 6-12 GPa.

Many teststructures for the measurement of electrical, thermal and mechanical parameters have been included in the design. Sheet resistance measurements on the N- and P-type PolySi layers have been performed by measuring VanDerPauw structures. Contact resistance of N- and P-type PolySi to the Al interconnect have been measured by 4-wire resistance measurements on structures with different contact sizes [5.17]. Differently sized structures, as described in [5.17], [5.15] and [5.16], for measuring thermal conductivity and thermal diffusivity of the various layers have been included. The photo in figure 5-5 shows a test structure on run EI-1448 for measuring the thermal conductivity of the various layers.

![Test structure for thermal conductivity measurement (EI-1448).](image)

Structures for measuring the Seebeck coefficients of N- and P-type PolySi, as described by [5.11] [5.17] have been included also. Table 2 gives an overview of the measured parameters that are relevant for the calculations in this thesis.

**TABLE 2. Data and measurement results of the fabricated structures**

<table>
<thead>
<tr>
<th>Thickness of SiN membrane</th>
<th>d₁</th>
<th>0.8 [μm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thickness of PolySi</td>
<td>d₂</td>
<td>0.3 [μm]</td>
</tr>
<tr>
<td>Thickness of SiO (or second SiN)</td>
<td>d₃</td>
<td>0.5 -(0.3) [μm]</td>
</tr>
<tr>
<td>Thickness of Al</td>
<td>d₄</td>
<td>0.6 [μm]</td>
</tr>
</tbody>
</table>
5.6 Conclusions

The fabrication of micromachined thermopiles and bolometers on thin isolating membranes has been described in literature for more than 20 years now. The fabrication of large bridges and cantilevers is much more critical and a commercial fabrication process that could fulfill the requirements for the intended spectrometer applications is not available at present.

Therefore an in-house technology developed by the DIMES institute at the Delft University of Technology for the fabrication of thin silicon nitride
membranes has been used. The process has served many research projects by prototypes and a few small commercial series of devices also, during the past 10 years. From the difficulties we have encountered during several runs we may conclude that the fabrication of cantilevers or bridges is much more critical than the fabrication of closed membranes. Large tensile stresses can easily cause failures of bridge and cantilever structures during the final etch step where the structures become freestanding. A tensile stress of around 2.3 GPa is present after the deposition of the second nitride layer and a very low yield has been obtained for most of the bridge structures after the final etch step. The reason for this is still unknown, since the fracture stress of thin SiN layers is typically 6-12 GPa. The wafers fabricated by process option A were lost by a large underetched of the protective resist by the KOH etching. The improvements in quality and reproducability of the recent devices indicate better perspectives for high sensitivity thermopile arrays using this technology in the future.

5.7 References

[5.5] http://www.dlp.com/, Texas Instruments DLP Products, Plano, TX, USA.
Conclusions


Integrated Thermal Infrared Sources

If you can’t make them see the light, make them feel the heat.
Unknown.

6.1 Introduction

Spectrometers are used to measure the spectral energy distribution of light. In most infrared spectrometer applications however, the absorption, reflection or transmission of a material needs to be measured and a well-defined high power infrared source is also needed. In conventional infrared spectrometers these sources are almost exclusively thermal sources:

- Tungsten incandescent lamps can be used as black body sources for the Near-Infrared only.
- Nichrome (or rhodium) wires are heated by resistance to incandescence and a black oxide layer forms on the surface. Temperature 1100°C. Requires little maintenance and no cooling. Emits in the Mid-infrared but less power than other sources.
- Globar: Globars are rods of silicon carbide. They are self starting and operate at 1300 °C. Their output power is between the nichrome wire and the Nernst glower. Globars must be water cooled.
- Nernst Glowers (rare earth oxides) are constructed from a mixture of fused oxides of Zirconium, Thallium and Cesium and have a more intense emitted radiation. Non-conducting at ambient temperatures but at
temperatures >800 °C the material becomes electrically conducting, and can maintain a high temperature by resistive heating. The energy output of Nernst Glowers is typically twice the intensity of a nichrome wire or Globar.

- Thin-film heaters and wire-wound resistors are commercially available devices, made from metals [6.10] or amorphous carbon [6.7] typically operating at 700-800 °C. The output power is low, but they can be made small and are low cost. Small thin-film heaters have a low thermal mass and can be used for pulsed operation. The peak output power for pulsed operation can be higher than in continuous mode.

Figure 6-1 shows two commercially available small mid-infrared sources: a robust 4 watt wire-wound resistor and a 1 Watt thin-film device suitable for pulsed operation.

![A wire-wound and a thin-film infrared source](image)

Mid-infrared LED’s are expensive and have a peaked response. Moreover, they generally emit less power than thermal sources, because of their low efficiency (<1%) [6.23]. Small lasers are not available yet in the mid-infrared.

Small infrared sources that can be modulated are needed in the readout system using coherent detection, as described in chapter 7. IC-compatible thermal sources on thin membranes have been designed and fabricated as infrared radiating elements for microspectrometer applications. In the following sections the thermal, electrical and mechanical properties of these devices will be discussed.
6.2 Integrated Thermal Infrared Sources

Polysilicon resistors, thermally isolated on thin silicon nitride bridges, have been fabricated in compatible MEMS technology as sources in the infrared [6.1][6.2] and as microlamps in the visible range [6.5][6.6][6.12]. Many papers on MEMS hotplates describe the properties and fabrication of thermally isolated membranes operating at high temperatures [6.4][6.3]. An interesting new development is the co-integration of thin-film infrared sources with thin-film optical filters enabling the fabrication of narrow-band infrared sources [6.8][6.9].

After a short description of the fabrication and device geometry, a static thermal analysis will be given in the next sections. Subsequently, the dynamic properties of the thin-film heater wires will be given. The final sections discuss some measurement results of the fabricated infrared sources.

6.3 Geometries of the Infrared Emitting Heater Strips

The geometry of the infrared emitting element is governed by the requirement of the optical system, however also high thermal isolation to the substrate of the chip is required to reach a temperatures high enough for Mid-infrared emission with a reasonable amount of electrical input power. Details on the fabrication of the bridges have been given in chapter 5. The polysilicon heaters are sandwiched in between a low-stress nitride and an oxide layer, as shown in.

![Fig. 6-2. Cross section view of an IR emitting beam.](image)

The typical stack composition used is 300 nm Si₃N₄, 600 nm polySi and 300 nm SiO₂. The typical length $L_b$ of these dual-clamped beams is between 1000-4000 μm and the width $W$ varies between 20-100 μm, as shown in figure 6-3.
Table 2 gives the data of the microwire strips of IC runs EI 913 and EI 1305 that will be used for the analysis in this chapter. Long and thin wire elements with high thermal isolation have been fabricated and will be analyzed in the next sections. Some different structures of thin beams have been fabricated to relieve the stress of the device at the clamping point to the substrate, as shown in figure 6-7.

Fig. 6-3. Back and front view photos of some fabricated beams.

Fig. 6-4. Details of the clamping structures of some fabricated beams.

a). Dogbone clamped wires \( (L = 2000\mu m \ W = 20\mu m) \)
b). Serpentine clamped wires \( (L = 2000\mu m \ W = 20\mu m) \)
6.4 Thermal analysis of Infrared Emitting Heater Strips

The thermal properties of heater strips on thin dielectric bridges are analysed in the next sections. It will be shown that the thermal analysis can be simplified by uncoupling the conduction and radiation problem in the structure.

Blackbody radiation is assumed from the thermal infrared source and therefore the required average temperature of the infrared emitter directly follows from the peak in the infrared wavelength region (1 to 5μm) by Wien’s Displacement Law. Typically a peak in the emission spectrum at 3μm requires a blackbody radiator operating at

\[
T_\text{str} = \frac{c_v}{\lambda} = \frac{2.897 \times 10^{-3}}{3 \times 10^{-6}} = 966 \text{K or } 693 \degree \text{C}
\]  

(6-1)

This equation shows a direct relation between the temperature (power) of the source and the maximum emitted power at a certain wavelength, i.e. the operating wavelength of the spectrometer. Increasing the power of the thermal source results in emission beyond the wavelength region of the spectrometer and is therefore rather useless.

6.4.1 Thermal Model

The fabricated bridges are thermally symmetrical and their analysis can be simplified by considering only one half of the structure (L = L/2) resulting in a cantilever structure of one half of the bridge length as shown in figure 6-7.

Conservation of energy in an infinitesimal element of the beam results in following sets of equations:

\[
W_e = W_{st} + W_{loss}
\]

\[
W_e = J^2 \rho_e
\]

\[
W_{st} = c_v \rho_v \frac{\delta T}{\delta t} \delta y \delta x \delta z
\]

\[
W_{loss} = -\kappa \nabla T
\]

(6-2)
\[ W_g = \text{energy generated in the volume by power dissipation caused by a current density } J \text{ through the conductor with an electrical resistivity } \rho_e. \]

\[ W_s = \text{stored energy in the volume with a specific heat } c_p \text{ and a density } \rho_v. \]

\[ W_l = \text{loss of energy in the volume by heat conduction. The heat flux by conduction is given by the heat diffusion or Fourier equation:} \]

\[ q_{\text{loss}} = -\kappa \nabla T = -\left( \frac{\delta}{\delta x} \left( \kappa \frac{\delta T}{\delta x} \right) + \frac{\delta}{\delta y} \left( \kappa \frac{\delta T}{\delta y} \right) + \frac{\delta}{\delta z} \left( \kappa \frac{\delta T}{\delta z} \right) \right). \]  

where \( \kappa \) is the thermal conductivity of the beam and the minus sign indicates that the flux is in the direction of decreasing temperature. Temperature gradients by conduction within the beam are only significant along the beam in the \( x \) direction, since the beam length is much higher than its width (\( W \)) or thickness (\( d \)). Therefore we can define a thermal conductance \( k_x \) per unit of length \( x \) along the beam:

\[ k_x = \frac{k}{W d} \]  

Generally the beam is fabricated from layers of different materials and the total conductance of the beam equals the conductances of the individual layers in parallel, thus:

\[ G_{\text{c}} = G_{\text{poly}} + G_{\text{SiN}} + G_{\text{SiO}} \]

\[ G_{\text{c}} = \frac{k_x A_x}{L} = \frac{k_{\text{poly}} A_{\text{poly}}}{L} + \frac{k_{\text{SiN}} A_{\text{SiN}}}{L} + \frac{k_{\text{SiO}} A_{\text{SiO}}}{L} \]  

Fig. 6-5. Thermal model of the beam.
The equivalent conductivity of the composite beam $\kappa_c$ can be, therefore, expressed as a combination of the three different layers as:

$$\kappa_c = \frac{1}{\frac{W}{W_x} d_p \kappa_p + \frac{d_{SiN}}{W} \kappa_{SiN} + \frac{d_{SiO}}{W} \kappa_{SiO}}$$

(6-6)

In this equation $\kappa_c$ is the composed conductivity of the different layers in the beam.

Convection and radiation are considered surface effects that are only relevant at the boundaries of the medium. Since the beams are uniform and generally very thin ($y >> z$) we can neglect the heat flow in the $y$ direction and the heat loss by conduction through a quiet gas from the top and bottom surface $W$ in the $z$ direction can be written as:

$$q_{loss-Z} = -2 \kappa_{air} \frac{\delta T}{\delta z} \int dy = -2 W \kappa_{air} \frac{\delta T}{\delta z}$$

(6-7)

It will be shown that the thermal conductivity $\kappa_{air}$ of the air surrounding the beam causes high power loss, that has to be compensated by a high current through the heater. Also fast ageing of the polysilicon layer occurs if the device operates in air. For these reasons we assume that the device operates in vacuum and we can neglect convection and conduction through the surrounding air. Heat transport can now take place by conduction through the beam to the clamping points and by radiation only.

### 6.4.2 Radiation

The heat exchange by radiation from the beam surface to the environment can be approximated by the general equation of grey-body radiation exchange [6.17]:

$$q_R = AF_{e-amb} \varepsilon \sigma (T(x)^4 - T_{amb}^4)$$

(6-8)

where $\sigma$ is the Stefan-Boltzmann constant, $\varepsilon$ is the emission coefficient, $A$ is the emission area and $F_{e-amb}$ the View Factor of the emission area with respect to the environment. If we assume a small object in a large closed cavity we can
assume \( F_{\text{e,amb}} = I \) [6.11].

\[
\left( T(x)^4 - T_{\text{amb}}^4 \right) = \left( T(x)^2 + T_{\text{amb}}^2 \right) \left( T(x) - T_{\text{amb}} \right) \equiv T(I) \frac{\delta T(x)}{\delta x}
\]

Substitution in equation (6-8) while considering the fact that the average temperature of the emitter \( T_e \) is much higher than \( T_{\text{amb}} \), above equation can be approximated by

\[
q_{\theta} \equiv A F_{\text{e,amb}} c \sigma T_e^3 \frac{\delta T(x)}{\delta x} \equiv 2WL\varepsilon \sigma T_e^3 \frac{\delta T(x)}{\delta x}.
\]

where \( W \) and \( L \) are the width and the length of the beam respectively. In the next section the heat equation will extended with the radiation effect.

### 6.5 Static thermal analysis of a Heater beam

The static equilibrium for the beam \( (W_{st} = 0) \) follows from equation (6-2) and equation (6-10) yielding:

\[
J^2 \rho_e (1 + \alpha_e) \frac{\delta T(x)}{\delta x} - \left( 2WL\varepsilon \sigma T_e^3 \right) \frac{\delta T(x)}{\delta x} - \kappa_e \frac{\delta^2 T(x)}{\delta x^2} = 0
\]

In a beam with a uniform cross-sectional area the current density \( J \) is constant throughout the heater. Both the thermal and electrical conductivity \( \kappa \) and \( \rho_e \) are temperature dependent, and therefore depend on the position \( x \) along the beam. The temperature coefficient of doped poly-silicon layers is positive and relatively constant \( \alpha_e \approx 10^{-3} \text{K}^{-1} \) [6.15]. The temperature coefficient of the thermal conductivity of several thin poly-silicon films has been published up to temperatures of 500 °C by [6.14] and [6.22] and shows no significant change \( (< 10\% \) at temperatures above \( T = 100 \text{K} \). The first term represents the heat dissipation by an electrical current \( I \) and neglecting the temperature coefficient of electrical resistance \( R \) the heat generation rate per unit of length \( x \) is constant and
can be written also as:

\[ q_x = J^2 \rho_x = \frac{I^2 \rho_x}{W_{poly} d_{poly}} = \frac{I^2 R}{L} = \frac{P_e}{L} \quad (6-12) \]

Equation (6-11) can now be simplified and written as:

\[ \frac{P_e}{L} - (2WL \varepsilon \sigma T(x)^4) \frac{\delta T(x)}{\delta x} - \kappa_x \frac{\delta^2 T(x)}{\delta x^2} = 0 \quad (6-13) \]

This is a second-order differential equation with a non-linear term caused by radiation. Generally these equations cannot be solved for non-linear terms higher then power 2 \[6.21][6.14\]. From literature is found that the emitted radiation loss from the polysilicon heater is rather low with respect to the conductive heat loss. In general the efficiency of almost all types of thermal sources is only a few percent \[6.20\]. Furthermore it must be noted that accurate calculations of the radiative quantities are generally difficult since many parameters like e.g. the emission coefficient are either unknown or show a very large spread at high temperatures.

The emission coefficients of thin nitride films on silicon have been investigated by \[6.22\]. Values for the emission coefficient \( \varepsilon = 0.65-0.95 \) have been found depending on the layer thickness and the angle of incidence. A value \( \varepsilon = 0.7 \) will be used in the calculations in this chapter. In the following paragraphs an analysis and estimation of the emitted power of the heater strip will follow using a two-step approach. In the next section we will calculate the temperature profile neglecting the radiation term in equation (6-13) and in the following paragraph we will use this temperature profile to calculate the optical power using a grey-body emission approximation.

### 6.5.1 Temperature distribution along the beam

From the previous section it can be concluded that amount of energy transported by conduction in the thin MEMS beam is much higher than the heat exchange by radiation. For calculation of the temperature distribution in the emitter, only the thermal conductivity then needs to be taken in account. The temperature profile in the beam can be analysed by considering the thermal conductivity to the clamping of the bridge only. The temperature distribution along the beam now becomes a one-dimensional problem.
situation, the electrically generated heat flow in the resistor must equal the conductive heat loss through the beam. The temperature distribution can now be derived from equation (6-13) as:

$$\kappa_x \frac{\delta^2 T(x)}{\delta x^2} = \frac{I^2 \rho_o}{W_{Poly} d_{Poly}} = \frac{P_e}{L}$$  \hspace{1cm} (6-14)

Equation (6-14) is a linear second-order differential equation and can be solved by integration and setting two boundary conditions. Assuming a constant electrical input power dissipation $P_e = I^2 R$, a Neumann boundary condition (no heat flow $dT/dx = 0$) in the middle of the beam ($x = L$) and a Dirichlet boundary condition ($T = \text{ambient}$) at the clamped end of the beam the heat equation can be solved as [6.11]:

$$\left[ T(x) \right]_{x=0}^{x=L} = T_{\text{amb}}$$

$$\left[ \frac{\delta^2 T(x)}{\delta x^2} \right]_{x=L} = 0$$

$$T(x) = T_{\text{amb}} + K_w \left( Lx - \frac{x^2}{2} \right)$$  \hspace{1cm} (6-15)

With:

$$K_w = \frac{I^2 \rho_o}{\kappa_x W_{Poly} d_{Poly}} = \frac{P_e}{\kappa_x L}$$

Equation (6-15) results in a parabolic temperature distribution across the beam as can be seen from the plot of $T(x)$ in figure 6-6.

**Fig. 6-6. Temperature profile along the beam.**
Static thermal analysis of a Heater beam

The maximum temperature of the beam is, as expected, at half the length of the bridge

$$\frac{\delta}{\delta x} \left( Lx - \frac{x^2}{2} \right) = 0 \Rightarrow x = L.$$ \hspace{1cm} (6-16)

The maximum $T_m$ and average temperature $T_{av}$ of the beam are given by:

$$T_m = T_{amb} + \frac{P_c}{\kappa_4} \left( Lx - \frac{x^2}{2} \right)^{\infty} = T_{amb} + \frac{P_c}{\kappa_4} \left( L \right)^{\infty}$$

$$T_{av} = T_{amb} + \frac{P_c}{\kappa_4} \frac{1}{L} \int_{-L}^{+L} \left( Lx - \frac{x^2}{2} \right) \left( Lx - \frac{x^2}{2} \right) \frac{Lx^2 - x^3}{6} \left( \frac{Lx^2 - x^3}{6} \right)^{\infty}$$

In the next section the temperature profile of equation (6-15) is used to calculate the emitted radiation of the heated wire.

6.5.2 Emitted Power of the Polysilicon Element

The heater wire can be approximated by a line of point sources emitting radially into ambient space. By applying Planck’s equations for “grey-body” radiation both the emitted power and its spectral distribution can be evaluated from the parabolic temperature profile. The emitted power can be approximated by combining the Stefan-Boltzmann law for radiation equation (6-8) and the temperature profile of equation (6-15), yielding:

$$P(x) = K \cdot \left[ T_{amb} + I^T \cdot \left( Lx - \frac{x^2}{2} \right) - T_{amb}^g \right]$$

With: $K = \frac{\rho_c}{\kappa_4 \cdot W_{poly} \cdot d_{poly}}$ and $K = 2WLE\sigma$

Using some typical values given from Table 2 the radiation profile along the beam has been calculated and plotted in shown in figure 6-7 for a few values of the electrical input power. From figure 6-7 can be seen that most of the power is
emitted from the part of the beam where \( x > L/2 \). By integration over this usable part of the beam we can find the total emitted power of the beam as:

\[
P_i = \frac{K_x}{L} \int_{x=L/2}^{x=L} \left( T_{amb} + I_x K \left( 2x^2 \right) \right)^4 \left( T_{amb} \right)^4 \, dx
\]

(6-19)

Using the given material parameters we find that total amount of optical power from the middle part of a bridge heated by 2.5 mW electrical power, reaching an average temperature of 730 °C, is about 41 μW. This may appear as a low value. However, comparison with an expensive Mid-infrared LED operating at 200mA giving a total output power of only 8 μW [6.23] in a narrow band (\( \lambda = 2.8-3.0 \mu m \)) shows that the performance of these thermal sources is comparable and in many applications superior. A more useful specification for micro-optical systems is the infrared power intensity emitted directly above the heater wire with length \( L \) and width \( W \):

\[
I_{rad} = \frac{P_i}{WL} = \frac{41 \times 10^{-6}}{2 \times 10^{-7} \times 44 \times 10^{-3}} = 465 \, W/m^2
\]

(6-20)

The efficiency of the Polysilicon infrared heater of around 1% at 700 °C increases dramatically with temperature, due to the fourth power relation with the electrical input power, but it has been found that operating temperatures beyond 900°C are not recommended due to the strongly reduced lifetime of the devices.
6.6 Dynamic analysis of the Heater Beam

Since the fabricated device are small and have a low mass, their response time can be fast and high chopper frequencies are feasible.

6.6.1 Analysis using lumped elements

A rough estimation of the thermal time constant of the heater wires can be given by the lumped thermal capacitor and resistor model shown in figure 6-7.

![Lumped element thermal model of the beam.](image)

The thermal conductance $G$ equals:

$$G = \frac{1}{R_h} = \frac{\kappa_c W d}{L}$$  \hspace{1cm} (6-21)

where $\kappa_c$ is the combined thermal conductivity of the beam as given by equation (6-6), $W$ the width and $d$ the total thickness. The beam temperature response $T(t)$ when cooling down can be therefore written as:

$$T(t) = P_e(0) R_h \left[1 - \exp\left(\frac{t}{\tau}\right)\right] = \frac{P_e L}{\kappa_c W d} \left[1 - \exp\left(\frac{t}{\tau}\right)\right]$$  \hspace{1cm} (6-22)

The time constant is then given by $\tau = C/G$, where $C_c$ is the heat capacity of...
the beam and \( G \) is the effective thermal conductance:

\[
G = \frac{\kappa_c W d}{L} = 0.27 \times 10^{-6} \text{ [W/K]}
\]

\[
C = \rho_c c_c (W L d) = 2.6 \times 10^{-7} \text{ [K/W]}
\]

(6-23)

(6-24)

(6-25)

(6-26)

The time constant \( \tau \) follows from the RC product or

\[
\tau = C \frac{G}{\kappa} = \frac{L^2}{D_c} \text{ [s]}
\]

In this equation \( D_c \) is the combined thermal diffusivity of the three-layer SiN-PolySi-SiO stack in the beam:

\[
D_c = \frac{\kappa_c}{\rho_c c_c} \text{ [m}^2/\text{s]}
\]

(6-25)

where \( \rho_c \) is the combined density, \( \kappa_c \) is the combined thermal conductivity and \( c_c \) the combined specific heat capacity. The heat capacity \( C_c \) is the combined heat capacity of the layers in the beams and follows from the product of the combined specific density and combined specific heat of the Polysilicon, silicon oxide and silicon nitride in the beam:

\[
c_c = \frac{1}{d_p + d_{SN} + d_{SO}} \left( \frac{W_p}{W} d_p c_p + d_{SN} c_{SN} + d_{SO} c_{SO} \right)
\]

and:

\[
\rho_c = \frac{1}{d_p + d_{SN} + d_{SO}} \left( \frac{W_p}{W} d_p \rho_p + d_{SN} \rho_{SN} + d_{SO} \rho_{SO} \right)
\]

(6-26)

Table 1 shows the calculated values for the combined specific heat, diffusivity, conductivity and density based on the data and beam dimensions given in Table 2 on page 98. The resulting thermal time constant equation (6-26) for a typical beam is \( \tau = 10 \text{ ms} \).
6.6.2 Analysis using Differential Equations

More accurate results can be found by solving the dynamic heat equation of the beam. Adding the energy storage term $W_s$ to the simplified static equation of the beam (eqn 6-14) from the previous paragraph, results in:

$$J(x,t)^2 \rho_c(T) - \kappa_c \frac{\partial^2 T(x,t)}{\partial x^2} - \frac{1}{D_c} \frac{\partial T(x,t)}{\partial t} = 0$$

(6-27)

This equation can be solved analytically by the method of separation of variables. The general solution of this type of partial differential equation (PDE) results in an infinite series of damped sine and cosine functions [6.13]:

$$T(x,t) = e^{-\alpha^2 t} \sum_{n=1}^{\infty} \left[ C_n \cos\left(\frac{x}{\lambda}\right) + D_n \cos\left(\frac{x}{\lambda}\right) \right]$$

(6-28)

The variables $C_n$ and $D_n$ are Fourier coefficients that must meet the boundary conditions. For a bridge of length $2L$ substitution of $T(x,t) = 0$ and $T(L,t) = 0$ results in a set of Fourier coefficients:

$$D_n = 0 \quad \text{and} \quad C_n = \frac{1}{L} \int_{0}^{L} \sin\left(\frac{n\pi x}{2L}\right) T(x,0) \, dx$$

(6-29)

and therefore:

$$T(x,t) = \sum_{n=1}^{\infty} \exp\left(-D\left(\frac{n\pi}{2L}\right)^2 t\right) C_n \sin\left(\frac{n\pi x}{2L}\right) T(x,0)$$
Here $T(x,0)$ is the temperature profile at $t=0$ and since this solution should also include the static solution, we can write for $T(x,0)$ the static heat equation (6-15)) from the previous section.

$$T(x) = T_{amb} + \frac{I^2 \rho_o}{\kappa L W_{poly} d_{poly}} \left( Lx - \frac{x^2}{2} \right)$$ \hspace{1cm} (6-30)

The maximum temperature $T_m$ is in the middle of the beam for $x=L$. The temperature response when cooling down from the static equilibrium at $T_m = 700^\circ C$ results in a temperature response as shown in figure 6-9. If $n$ is large the more higher order terms are chosen and the better $T(x,t)$ approaches the real solution. However since the initial temperature distribution is rather smooth the higher order terms $n \geq 3$ do not contribute significantly to the solution anymore. As time continues each coefficient decays to zero, but at different rates. The large $C_n$ coefficients decrease very rapidly and the information about the high order components in the initial solution disappear first, resulting in a smoothing of the temperature distribution. The equivalent thermal time constant follows from the
Electrical properties

Equation (6-31) gives an expression for the time constant $\tau$ for a homogeneous bridge or beam when switched off. The \textbf{thermal time constant of the lumped model of a bridge or beam, as given by equation (6-24), is around 33\% higher than the exact solution using differential equations.} This should be taken in account when these devices are used in pulsed mode.

The dynamic response of the heater in terms of radiation can be calculated by substitution of the radiation pattern $P(x)$ of equation (6-26) in equation (6-29) resulting in the dynamic optical response when switched off, as shown in figure 6-10.

6.7 Electrical properties

Theoretically, the temperature of the heater increases with the fourth power of the electrical power dissipation and, as a result with the sixth power of the
current $I$ which can be critical when current steering of the heater is used. The polysilicon layers used have a positive temperature coefficient $\alpha_e = 10^{-6}$, resulting in an increased power dissipation when heater rises in temperature, causing thermal runaway above certain temperatures, according to:

\[
\begin{align*}
\text{Current switching: } P &= I^2 R (1 + \alpha_e \Delta T) \\
\text{Voltage switching: } P &= \frac{U^2}{R (1 + \alpha_e \Delta T)}
\end{align*}
\]

(6-32)

Therefore, to prevent thermal runaway, voltage switching is preferred because then the electrical input power decreases if the heater temperature increases.

### 6.8 Measurement results

Measurement methods and results of the electrical and thermal properties of wafer series EI-913 have been given in [6.24] and of wafers EI 1305 in [6.25].

<table>
<thead>
<tr>
<th>Thin-film parameter</th>
<th>EI913 NpSiGe</th>
<th>EI913 PpSiGe</th>
<th>EI1305</th>
</tr>
</thead>
<tbody>
<tr>
<td>Beam length (L) [μm]</td>
<td>2000</td>
<td>2000</td>
<td>---</td>
</tr>
<tr>
<td>Beam width (W) [μm]</td>
<td>44</td>
<td>22</td>
<td>---</td>
</tr>
<tr>
<td>Polysilicon width $W_{\text{poly}}$ [μm]</td>
<td>40</td>
<td>18</td>
<td>---</td>
</tr>
</tbody>
</table>

**TABLE 2. Typical sizes and data of the fabricated infrared emitters.**
Measurement results

### TABLE 2. Typical sizes and data of the fabricated infrared emitters.

Some values have been given in Table 2 also. Measurement methods and results of the electrical and thermal properties of wafer series EI-913 have been given in [6.24] and of wafers EI-1305 in [6.25].

For measuring the optical power of the devices on EI 913 a controllable vacuum chamber has been used. A calibrated thermopile (Newport model 71937) has been placed above the heater. Figure 6-12 shows the device. The emitted power without a window cover on the housing has been measured with an estimated angle $FOV = 0.6$ in the range $\lambda = 1$-10 $\mu$m. Figure 6-3 shows a camera.

<table>
<thead>
<tr>
<th>Thin-film parameter</th>
<th>EI913 NpSiGe</th>
<th>EI913 PpSiGe</th>
<th>EI1305</th>
</tr>
</thead>
<tbody>
<tr>
<td>PolySi conductivity $\kappa_{SiP}$ $[\text{Wm}^{-1}\text{K}^{-1}]$</td>
<td>4.8</td>
<td>4.8</td>
<td>24</td>
</tr>
<tr>
<td>PolySi Specific Heat $c_{SiP}$ $[\text{J g}^{-1}\text{K}^{-1}]$</td>
<td>0.77</td>
<td>0.77</td>
<td>0.77</td>
</tr>
<tr>
<td>PolySi Specific density $\rho_{SiP}$ $[\text{kg m}^{-3}]$</td>
<td>2860</td>
<td>2860</td>
<td>2860</td>
</tr>
<tr>
<td>PolySi Diffusivity $\alpha_{SiP}$ $[\text{cm}^2\text{s}^{-1}]$</td>
<td>0.65</td>
<td>0.65</td>
<td>0.65</td>
</tr>
<tr>
<td>PolySi thickness $d_{poly}$ $[\mu$m]</td>
<td>0.6</td>
<td>0.6</td>
<td>0.3</td>
</tr>
<tr>
<td>SiN thickness $d_{SiN}$ $[\mu$m]</td>
<td>0.3</td>
<td>0.3</td>
<td>0.8</td>
</tr>
<tr>
<td>SiN conductivity $\kappa_{SiP}$ $[\text{Wm}^{-1}\text{K}^{-1}]$</td>
<td>1.55</td>
<td>1.55</td>
<td>1.5</td>
</tr>
<tr>
<td>SiN Specific Heat $c_{SiN}$ $[\text{J g}^{-1}\text{K}^{-1}]$</td>
<td>1.5</td>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>SiN Specific density $\rho_{SiN}$ $[\text{kg m}^{-3}]$</td>
<td>2500</td>
<td>2500</td>
<td>2500</td>
</tr>
<tr>
<td>SiN Diffusivity $\alpha_{SiP}$ $[\text{cm}^2\text{s}^{-1}]$</td>
<td>0.012</td>
<td>0.012</td>
<td></td>
</tr>
<tr>
<td>SiO thickness $d_{SiO}$ $[\mu$m]</td>
<td>0.5</td>
<td>0.5</td>
<td>0.3</td>
</tr>
<tr>
<td>SiO conductivity $\kappa_{SiO}$ $[\text{Wm}^{-1}\text{K}^{-1}]$</td>
<td>1.1</td>
<td>1.1</td>
<td>1.17</td>
</tr>
<tr>
<td>SiO Specific Heat $c_{SiO}$ $[\text{J g}^{-1}\text{K}^{-1}]$</td>
<td>1.0</td>
<td>1.0</td>
<td>1.0</td>
</tr>
<tr>
<td>SiO Specific density $\rho_{SiO}$ $[\text{kg m}^{-3}]$</td>
<td>2200</td>
<td>2200</td>
<td>2200</td>
</tr>
<tr>
<td>Polysilicon resistivity $[\text{ohm/sq.}]$</td>
<td>47</td>
<td>48</td>
<td>58</td>
</tr>
<tr>
<td>Beam Total resistance $[\text{ohm}]$</td>
<td>2300</td>
<td>5100</td>
<td>---</td>
</tr>
</tbody>
</table>
Integrated Thermal Infrared Sources

picture of a heater emitting visible light. The heater devices on the spectrometer chips of EI-1305 have been mounted in a metal can housing covered with an infrared transparent window of ZnSe or silicon. A tube on the back side of the housing has been connected to the vacuum chamber to pump out the air as shown in figure 6-13.

The measured emitted total optical power versus electrical input power is plotted in figure 6-16. Continuous operation of several weeks at $P_e=15\text{mW}$, emitting about $250\mu\text{W}$ optical power has been measured. Figure 6-15 show a typical step response of an infrared heater.

The spectral distribution of the emitted infrared radiation could not be measured since there was no suitable instrument available to perform infrared emission measurements on such small devices.

Fig. 6-12. CCD camera capture of an operating IR emitter.

Fig. 6-13. Photo of the IR emitter EI-1305 connected to a vacuum pump.
Conclusions

6.9 Conclusions

An analysis of thermal Mid-infrared emitters using polysilicon wires on thin membranes has been presented. The temperature profile and the resulting emitted power have been calculated. The thermal analysis and its results are valid for cantilevers and bridges. It has been shown that the exact solution for calculation of the thermal time constant differs 33% from the lumped element model. This indicates that a lumped element approach for a uniform beam can lead to rather inaccurate results.

Fig. 6-14. Emitted power versus dissipated power of the heater (EI-913).

Fig. 6-15. Measured step response of a typical heater wire (EI-913).
A general remark on the analysis is that accurate calculations of the radiative quantities are generally difficult, since many parameters like e.g. the emission coefficient are either unknown or show a very large spread at high temperatures. Thermal radiation is a highly non-linear phenomenon and therefore including radiation terms in the thermal analysis generally results in equations that cannot be solved analytically and often also not numerically.

The required temperature of the infrared emitter is governed by the wavelength region required by the application. Mid-infrared operation requires an operating temperature of around 700 °C. Increasing the input power of the thermal source to obtain more emitted optical power mainly results in more emission at shorter wavelengths and is therefore not very effective. The device should operate in vacuum to obtain sufficient low thermal isolation to reach the high average temperature of the beam. The calculated temperature profiles show that half of the length of the beam (in the middle) emits about 80% of the total optical power. The Thompson effect [6.20] can cause asymmetry in the temperature profile of incandescent heaters at high temperatures, as described by [6.18]. This effect has not been observed in our devices.

Fast ageing and failure of the polysilicon layer has been observed if the device operates in air above 400 °C. As in all types of thermal emitters it is found that the efficiency is only a few percent. The total optical power is higher than the output of Mid-infrared LED’s fabricated by InAs heterostructures. The use of narrowband silicon infrared filters as described in chapter 12 as windows for the
housing of these devices also enables their use in narrowband applications.

6.10 References


Integrated Thermal Infrared Sources


7 Detectors for Mid-Infrared Microspectrometers

7.1 Introduction

Traditionally the principal detection methods used in infrared spectrometers are photoconductive (semiconductor), photoemissive (photomultipliers), and photographic (e.g., film). Prior to about 1940, most spectra were recorded with photographic plates or film, in which the photographic material is placed at the image point of a grating or prism spectrometer. An advantage of this technique is that the entire spectrum of interest can be obtained simultaneously, and low-intensity spectra can be taken with sensitive film using long exposure times to the optical signal. Advances in material science and semiconductor technology has enabled the fabrication of many more and different types of infrared detectors for spectrometers. Nowadays high-performance, wide-range (UV to IR) spectrometers use photo multiplier tubes for the UV and visible range and thermoelectric cooled InGaAs detectors or MCT (cooled Mercury Cadmium Telluride) detectors for the mid-IR measurement range [7.1].

The next paragraphs will give a general overview of different principles and the basic properties of infrared detectors. This is followed by an overview of the basic specifications of photon and thermal detectors, such as spectral response, sensitivity and detectivity. Subsequently the IC compatibility of infrared detectors...
will be discussed and finally a selection will be made of the different types of infrared detectors that are suitable for the fabrication of single detectors or arrays of sensors for the intended microspectrometer applications.

7.2 Infrared Detectors

Almost any physical interaction phenomenon in the range of about 0.1-1 eV can be utilized for infrared detection [7.2]. The most common effects applied are:

- thermoelectric power (thermocouples).
- changes in electrical conductivity (bolometers).
- gas expansion (Golay cells).
- pyroelectricity (pyroelectric detectors).
- Josephson effect (Josephson junctions),
- internal emission (PtSi Schottky barriers),
- fundamental absorption (intrinsic photodetectors),
- impurity absorption (extrinsic photodetectors).
- low dimensional solids (quantum well detectors).
- Thermo-mechanical effects (bimetallic devices.)
- Photoemissive effects (Photomultipliers).

In general, infrared detectors are classified by means of their physical principle into photon-based detectors and thermal detectors.

7.2.1 Photon-based Detectors

In photon detectors, the radiation is absorbed within the material by interaction with electrons either bound to lattice atoms or to impurity atoms or with free electrons. The observed electrical output signal results from the change in electron energy distribution. The photon detectors show a selective wavelength dependence of response per unit incident radiation power. These exhibit both good signal to noise performance and a very fast response. The detection of Mid-IR radiation using quantum based detectors is hampered by the low energy of the photons, making non-cooled devices very noisy. Almost all types of quantum
detectors operating above about 3 μm wavelength need to be cooled, which is expensive and also limits speed. Photon detectors in silicon cannot be used in the mid-infrared region since the energy of the photons is much lower then the bandgap energy of silicon [7.5]. Cooling requirements are the main obstacle to the more widespread use of IR systems based on semiconductor photo detectors making them bulky, heavy, expensive and inconvenient to use.

Photoemissive detectors quickly replaced photographic plates in most spectrometer instrumentation. When a photon with sufficient energy strikes a surface, it can cause the ejection of an electron from the surface into a vacuum. A photoemissive diode [7.6] consists of a surface (photocathode) appropriately treated to permit the ejection of electrons by low-energy photons and a separate electrode (the anode) on which electrons are collected, both sealed within an evacuated glass envelope. A photomultiplier tube [7.7] has a cathode, a series of electrodes (dynodes), and an anode sealed within a common evacuated housing. Appropriate voltages applied to the cathode, dynodes, and anode cause electrons ejected from the cathode to collide with the dynodes in succession. Each electron collision produces several more electrons; after a dozen or more dynodes, a single electron ejected by one photon can be converted into a fast pulse of as many as 10^7 electrons at the anode. In this way, individual photons can be counted with good resolution. Arrays of photoemissive detectors include imaging tubes (e.g., television cameras [7.8]), which can measure a spatial variation of the light across the surface of the photocathode [7.9], and microchannel plates [7.10], which combine the spatial resolution of an imaging tube with the light sensitivity of a photomultiplier.

Solid-state detectors, such as semiconductor photodiodes [7.5], detect photons by exciting electrons from immobile, bound states of the semiconductor (the valence band) to a state where the electrons are mobile (the conduction band). In a photodiode the mobile electrons in the conduction band and the vacancies in the valence band can be moved through the solid with externally applied electric fields, collected onto a metal electrode, and sensed as a photo-induced current. In a photoconductor these free carriers will eventually recombine, but until this happens the conductivity is increased. Microfabrication techniques developed for the integrated-circuit semiconductor industry can be used to construct large arrays of individual photodiodes closely spaced together. The charges that are collected by the individual diodes, are read out separately in charge-coupled devices (CCD) or in CMOS camera arrays and can be displayed as an image.
Detectors for Mid-Infrared Microspectrometers

Photon based detectors using low energy bandgap materials, such as PbSe, PbS, InAs and InSb [7.3] are commonly used in infrared instrumentation for the detection of Mid-IR radiation. These detectors are expensive and compatibility issues excludes the application of these types of detectors in integrated circuit technology.

7.2.2 Thermal Infrared Detectors

Another group of infrared detectors are thermal detectors. These device are tandem transducers. In thermal infrared detectors the optical energy is transferred via the thermal domain into the electrical domain in two steps:

1. Incident optical radiation is absorbed causing a temperature rise in an absorbing layer.
2. This temperature gradient results in a change of some physical property is used to generate an electrical output.

Three main concepts have found the most widespread application in infrared technology, namely, bolometers, pyroelectric and thermoelectric effects. In pyroelectric detectors a change in the internal electrical polarisation is measured, whereas in the case of bolometers a change in the electrical resistance is measured. In contrast to photon detectors, thermal detectors typically operate at room temperature. The relatively slow process of heating and cooling of the detector element results in a slow response and low electrical bandwidth. The

Fig. 7-1. Basic photo-voltaic and photoconductive detectors.
conversion of optical radiation in thermal effects is generally wavelength independent [7.3] and as a result the optical bandwidth of thermal infrared detectors can be very wide. The sensitivity of infrared thermal detectors is generally low, but high thermal isolation from the environment can improve this.

In some thermal detectors, such as piezoelectric sensors, the heat flow is typically in the direction of the incident radiation, as shown in figure 7-2a, while in most types of bolometers and thermopile sensors is that in piezoelectric sensors the heat flow is typically in the lateral direction figure 7-2b. This gives an extra degree of freedom to the designer as will be used in next chapter.

Fig. 7-2. Two types of thermal detectors:
  a). Parallel flow.
  b). Lateral flow.

### 7.3 Specifications of Detectors

In order to compare the different types of sensors their specifications have to be compared. The main specifications of detectors are defined in this next section.

#### 7.3.1 Detector Sensitivity

The output signal of a photovoltaic detector is generally a current \( I_{ph} \) and its sensitivity, generally called responsivity, and is therefore expressed as:

\[
R_{\phi}(\lambda) = \frac{I_{ph}}{E_{opt}(\lambda) A} \quad [\text{A/W}].
\]  

(7-1)
where $E_{opt}(\lambda)$ is the optical irradiance in W/m² and $A$ the detector area [7.3]. In terms of quantum efficiency $\eta$ above equation can be written as:

$$R_i(\lambda) = \frac{q\eta\lambda}{hc} = \frac{\eta\lambda}{1.24}$$, with $\lambda$ in \(\mu\text{m}\). \hspace{1cm} (7-2)

The output signal of photoconductive detector is a change in resistance and a constant voltage $V$ across the detector results in a current response [7.3]

$$R_i(\lambda) = \left( \frac{q\eta\lambda}{hc} \right) \left( \frac{\mu\tau}{L} \right) V,$$ \hspace{1cm} (7-3)

where $L$ is the length of the detector, $\mu$ the majority carrier mobility and $\tau$ the carrier life time. Equation (7-3) shows that the responsivity of the photoconductor depends on its bias voltage. The first term corresponds to the responsivity of the photo-voltaic detector and we can compare the sensitivity of a photoconductive and a photoelectric detector by comparing equation (7-2) and equation (7-3).

Because only photons with sufficient energy can excite charge carriers across the material’s bandgap, the relation between the bandgap $E_g$ and the long wavelength cut-off frequency of a photon detector $\lambda_c$ can be written as [7.3]:

$$\lambda_c = \frac{1.24}{E_g}, \text{ with } \lambda \text{ in } \mu\text{m}. \hspace{1cm} (7-4)$$

### 7.3.2 Detector Noise

The signal-to-noise ratio of a detector is given by the ratio of the signal power and the noise power as:

$$\text{SNR} = \frac{\mathcal{S}(\lambda)P_s}{\mathcal{S}_n}$$ \hspace{1cm} (7-5)

where $\mathcal{S}_n$ is the noise power spectral density of the detector, $\mathcal{S}(\lambda)$ the spectral sensitivity and $P_s$ the optical signal power. The noise equivalent power NEP of a detector is defined as the optical input power for a signal-to-noise ratio $\text{SNR} = 1$.
The NEP of a detector is thus given by:

\[ NEP = \sqrt{\frac{S_n}{S(\lambda)}} \]  \hspace{1cm} (7-6)

The NEP can thus be written as the noise divided by the responsivity \( S(\lambda) \). Generally the output signal of a detector is a current or a voltage. For a current output detector the noise \( S_n \) should be expressed in terms of an output current noise \( \tilde{i}_n \) [A] and the responsivity \( R(\lambda) \) expressed in Ampere-per-Watt [A/W] yielding for the NEP:

\[ NEP = \frac{\tilde{i}_n}{R(\lambda)} \]  \hspace{1cm} (7-7)

The responsivity of a detector depends on its size, and its noise depends on the bandwidth, therefore the definition \( D^* \) is more often used as a figure of merit, \( D^* \) is the NEP of a detector per area unit and per bandwidth unit:

\[ D^* = \frac{\sqrt{A_D \Delta f}}{NEP} \]  \hspace{1cm} (7-8)

Figure 7-3 shows the detectivity \( D^* \) of some common types of detectors. The noise of a detector also depends on its mode of operation and, as a consequence several expressions for the NEP for different operation conditions can be distinguished.

### 7.3.3 Noise limits of Infrared Detectors

A current through the detector causes shot-noise and the noise current can be expressed by the well-known shot-noise equation [7.2]:

\[ \tilde{i}_n^2 = 2qI_s \Delta f \]  \hspace{1cm} (7-9)

where \( I_s \) is the signal current. This current can be the result of an optical signal or by background radiation. The signal-limited NEP is therefore given
from equation (7-5) and equation (7-9) as:

\[
NEP_{\text{SL}} = \frac{2qI_{\text{avg}} \Delta f}{R(\Delta f)}
\]  

(7-10)

For a photon detector this yields:

\[
NEP_{\text{SL}} = \frac{2qI_{\text{avg}} \Delta f}{I_s^2} = \frac{hc}{\lambda} \frac{2q \Delta f \eta q}{\eta \Delta f} = \frac{2hc}{\eta \Delta f}
\]  

(7-11)

For photon detectors this type of noise is only significant for very wideband detectors (\( \Delta f > 1 \, \text{GHz} \)) with a low quantum efficiency. In (uncooled) infrared detectors this shot-noise can become dominant if a large signal current is produced by background radiation.

Even if a detector is not illuminated an output signal is present due to intrinsic thermal effects.

- For photon detectors this is caused by thermal excitation of charge carriers, a process that competes with photon excitation. If this is the dominant noise source the photodetector is said to be dark-current limited.
- A thermal detector is in interaction with the environment by random processes of conduction, convection and radiation. When this thermally generated output signal of photodetector dominates the detector is said to be background limited or in BLIP (Background Limited Infrared Photodetector) mode. The highest detector temperature at which this happens is called the BLIP temperature. Background here means the (not cooled) surroundings or the scene within the detector field of view.
- For both types of detectors the electrical resistance \( R \) of the detector causes thermal- or Johnson noise. The noise current can be expressed as:

\[
\overline{i_s^2} = \frac{4kT \Delta f}{R}
\]  

(7-12)

7.3.4 Noise limits by Background Radiation

7.3.4.1 Photon Detectors

In a photon detector thermal noise is generated by the random process of
generation and recombination of charge carriers by radiation exchange with the background. Normally the noise in an infrared detector is background limited (BLIP), i.e. the noise caused by the background light level is much higher than the signal level. In this case the detectivity depends on the so-called generation-recombination noise caused by the background light level $E_{bg}$[7.4]. For a photovoltaic detector the detectivity can be derived [7.3] as:

$$D^* = \left( \frac{\lambda}{hc} \right) \sqrt{\frac{\eta}{2E_{tr}}}$$  \hspace{1cm} (7-13)

In a photoconductor the random fluctuations of the carrier lifetime are another source of noise, and addition of both noise sources results in an increase in an intrinsic noise level of a factor $\sqrt{2}$ higher than a photodiode detector. For a photoconductive detector the detectivity due to background radiation can be derived [7.3] as:

$$D^* = \left( \frac{\lambda}{hc} \right) \sqrt{\frac{\eta}{E_{sh}}}$$  \hspace{1cm} (7-14)

### 7.3.4.2 Thermal Detectors

The fundamental detection limit of a thermal detector is given by the laws of thermodynamics. The temperature of an object is always in interaction with the surroundings by random processes of conduction, convection and radiation. Even in equilibrium the temperature of a body will fluctuate random around a mean value $\Delta T^2$. This thermal noise can be derived by applying statistical thermodynamics on small bodies and the following expression for the total noise is can be derived [7.4] as

$$\Delta T^2 = \frac{kT^2}{C_{th}}$$  \hspace{1cm} (7-15)

where $C_{th}$ is the heat capacity [J/K] of the system. The noise level decreases with increasing thermal capacitance, however the detector bandwidth decreases by the same amount. The noise spectrum of a thermal detector can be written as a
lowpass filter:

\[
\Delta T^2(\lambda) = \Delta T_0^2 \left| \frac{1}{1 + 2\pi jf\tau} \right|^2 = \Delta T_0^2 \frac{1}{1 + (2\pi f\tau)^2}
\]  \hspace{1cm} (7-16)

with \(\Delta T_0^2\) the noise at low frequencies, i.e. below the thermal bandwidth of the device. The relation between the total noise and the low frequency noise is given by integration of the noise over the whole signal band:

\[
\Delta T^2 = \Delta T_0^2 \int_0^\infty \frac{1}{1 + (2\pi f\tau)^2} df = \Delta T_0^2 \frac{1}{4\tau}
\]  \hspace{1cm} (7-17)

And substitution of equation (7-3) yields:

\[
\frac{\Delta T_0^2}{\Delta f} = \frac{4kT^2 \tau}{C_{th}}
\]  \hspace{1cm} (7-18)

The combination of heat conduction, convection or radiation is represented by an equivalent thermal heat conductance \(G_{th}\). The thermal time constant is formed by the heat capacitance \(C_{th}\) and an equivalent thermal heat conductance \(G_{th}\) to the environment, hence \(\tau = C_{th}/G_{th}\). The noise spectral density in the signal band can be written therefore as:

\[
\frac{\Delta T_0^2}{\Delta f} = \frac{4kT^2}{G_{th}}
\]  \hspace{1cm} (7-19)

The thermal noise spectral density \(\Delta T_0^2\) of a detector, in a signal band smaller than the thermal bandwidth, is thus inversely proportional to its thermal conductance. A high thermal conductance results in low noise but also in a reduced sensitivity. The NEP in the low frequency part can be found by equating the temperatures of the signal \(\Delta T^2\) and noise \(\Delta T_0^2\) and the corresponding signal power giving:

\[
\Delta T^2 = \Delta T_0^2 \quad \Rightarrow \quad \text{NEP} = \sqrt{\frac{4kT^2 \Delta f}{G_{th}}}
\]  \hspace{1cm} (7-20)
and the detectivity $D^*$ is given as:

$$D^* = \sqrt{\frac{A_d G_{th}}{4kT^2}}$$  \hfill (7-21)

A low thermal conductivity of the detector thus results in a low detectivity $D^*$. In most sensors the thermal conductance of the connecting wires or conduction through air to the environment is dominant. However for a perfectly isolated thermal detector, operating in vacuum, there is always heat exchange by radiation with the environment. The heat flow is then given by [7.11]

$$q = \varepsilon \sigma A_d (T_d^4 - T_{amb}^4)$$  \hfill (7-22)

The equivalent conduction $G_{rad}$ can be expressed by differentiating above equation:

$$G_{rad} = \frac{\delta q}{\delta T} = 4\varepsilon \sigma A_d T_d^3$$  \hfill (7-23)

Combining equation (7-23) and equation (7-21) and assuming $\varepsilon = 1$ yields for the $NEP$:

$$NEP = G_{th} \sqrt{\frac{2kT^2\Delta f}{\pi}} = \sqrt{4\sigma A_d T_d^3} \sqrt{\frac{2kT^2\Delta f}{\pi}} = \sqrt{\frac{8\sigma A_d kT^4 \Delta f}{\pi}}$$  \hfill (7-24)

and for the detectivity $D^*$:

$$D^* = \frac{\sqrt{A_d \Delta f}}{NEP} = \frac{1}{\sqrt{\frac{4\varepsilon \sigma kT^5}{\pi}}}$$  \hfill (7-25)

At room temperature ($T=300K$) equation (7-25) gives the maximum theoretical detectivity $D^*$ for any thermal detector at room temperature:

$$D^* = 1.8 \times 10^3 \text{ W cm Hz}^{-1}$$  \hfill (7-26)

It must be noted that the values of $D^*$ of practical detectors are typically a factor 10-50 lower than this theoretical minimum. Comparison of equation (7-26) and equation (7-21) for a certain type of detector gives a figure of merit for that
Detectors for Mid-Infrared Microspectrometers

Device and $D^*$ also allows a comparison between thermal and photon detectors. The detectivity $D^*$ of a thermal detector is generally much smaller than the detectivity of a typical photon detector.

7.3.5 Spectral Response

The spectral responses of the most common types of uncooled infrared detectors are shown in figure 7-6. This figure gives a clear overview of available uncooled infrared detectors. For the mid-infrared the lead-salt detectors and thermopiles are the best uncooled types of detectors.

![Spectral response of common infrared detectors](image)

**Fig. 7-3.** Spectral response of common infrared detectors [7.3].

7.3.6 Conclusions

Different types of detectors have been analysed in the previous paragraphs. A global overview of the differences in performance of photodetectors and
therm detectors is given in Table 2.1.

**TABLE 1. Comparison of mid-Infrared Thermal and Photon Detectors.**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Photon Detector</th>
<th>Thermal detector</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sensitivity</td>
<td>High</td>
<td>Low</td>
</tr>
<tr>
<td>Detectivity</td>
<td>Low (uncooled)</td>
<td>Moderate</td>
</tr>
<tr>
<td>Spectral response</td>
<td>Selective</td>
<td>Wide and flat</td>
</tr>
<tr>
<td>Response time</td>
<td>Fast</td>
<td>Slow</td>
</tr>
<tr>
<td>Cost</td>
<td>High</td>
<td>Low</td>
</tr>
</tbody>
</table>

For accurate infrared detection uncooled thermal detectors are limited to a few basic types, i.e. photon detectors using low energy bandgap materials or thermal detectors. High thermal isolation is required for a high detectivity thermal detectors and can be realized by placing the sensor on micromachined thin membranes, bridges of cantilevers. The next sections discuss the integrated circuit compatibility of mid-infrared detectors.

### 7.4 IC-compatible Infrared Detectors

#### 7.4.1 IC-compatible Infrared Photon Detectors

Schottky barrier types of photodetectors can be fabricated compatible in standard IC processes, contrary to the quantum well detectors such InGaAs or the low bandgap materials PbS, PbSe, etc. Schottky barrier infrared photodetectors consist of a metal film, generally deposited on a P-type silicon substrate, as depicted in figure 7-4. The device is back-illuminated through the silicon die. Internal photoemission is produced across the metal/Si barrier by photon generated holes. Typical thickness of the metal layer is 2-10 nm. The silicon die forms an optical filter that transmits only wavelengths beyond the bandgap at 1.1 μm. The efficiency can be increased by an optical resonance cavity formed by an oxide layer placed on top of the metal layer and an aluminium mirror. The quantum efficiency is rather low ( η = 0.1-1% ). Ni/Si Schottky diodes have an optical bandwidth to about λ_c = 2.2 μm. Pd/Si Schottky diodes have an optical...
bandwidth up to $\lambda_c = 5.4 \, \mu m$.

Schottky barrier photodetectors can be easily fabricated in IC technology and large (e.g. 512x512) detector arrays, with the pixels having a non-uniformity of less than 1% can be fabricated [7.3]. The major problem of Schottky barrier detectors is that they need to be cooled, Ni/Si devices operate typically at 200K while Pt/Si detectors should be cooled down to 77K.

### 7.4.2 IC-compatible Infrared Thermal Detectors

Based on different physical principles, there are basically four types of thermal detectors that can meet to some extent the material and fabrication compatibility issues of standard IC processing and can be fabricated as arrays of sensors:

- Seebeck sensors
- Bolometers
- Pyro-electric sensors
- Opto-mechanical sensors.

#### 7.4.2.1 Thermopiles or Seebeck Sensors

Thermocouples are temperature sensors based on the Seebeck effect and inherently measure temperature gradients. This is an important advantage for infrared radiation detection over other type of thermal sensors, since slow ambient temperature variations are inherently compensated. Thermopiles can be readily fabricated in most IC-compatible processes, as shown in
IC-compatible Infrared Detectors

The optical bandwidth of these devices can be very high; in principle from the visible into the far-infrared (0.3μ-50 μm) depending on the absorber material [7.39]. The separation of infrared absorption and temperature measurement is a benefit in terms of IC compatibility. Typical output voltages of macro-size thermopiles are $S = 60\text{V/W}$ and $D^* = 5 \times 10^8 \text{cmHz}^{-1}\text{W}^{-1}$. For integrated thermopiles values in the same order of magnitude can be obtained using standard materials [7.28] [7.34] [7.30]. Using bismuth-telluride and IC processing values of $S=2000\text{V/W}$ and $D^* = 1 \times 10^7 \text{cmHz}^{-1}\text{W}^{-1}$ have been reported [7.35].

7.4.2.2 Bolometers

Bolometers are temperature sensitive resistors fabricated using doped silicon or metal films on a thermally isolated membranes [7.21]. Bolometers can be made small and can be fabricated in surface micromachining technology. Arrays for Mid-IR cameras, the so-called FP (=Focal plane) arrays can be
Detectors for Mid-Infrared Microspectrometers

fabricated by repetition of the pixel element shown in figure 7-6 [7.12] [7.14].

![Surface micromachined bolometer pixel](image)

*Fig. 7-6. Surface micromachined bolometer pixel.*

Traditionally thin metal films have been used [7.13], such as platinum [7.23], nickel and titanium [7.26]. Vanadium oxide (VO or V$_2$O$_3$) [7.21] and Yttrium Barium Copper Oxide (YBaCuO) [7.24][7.25] are most commonly used for high sensitivity bolometers because of their very high temperature coefficients ($\alpha_e = 0.018$ K$^{-1}$) and ($\alpha_e = -0.03$ K$^{-1}$) respectively. A lot of effort is undertaken in the IC-compatible processing of these materials for fabrication of vanadium oxide bolometer arrays while maintaining compatibility [7.20][7.21]. Doped (poly)silicon resistors or diffused resistors have a temperature coefficient of about a factor 10-30 lower [7.40] and have been fabricated in standard processing [7.32][7.33]. Uncooled infrared cameras using bolometers with amorphous silicon or vanadium-oxide pixels elements are expensive and commercially available [7.22].

For accurate measurements bolometers are less suitable because their temperature coefficient is generally not very stable [7.40] and highly non-linear. This causes the sensitivity of a bolometer to be dependent also on the ambient temperature. Accurately measurement of temperature differences using bolometers requires compensation of the large offset, e.g. by using more devices in bridge configurations or using by temperature balancing techniques [7.27]. Typical output voltages of 5000 V/W and $D^* = 2 \times 10^8$ cmHz$^{-1}$W$^{-1}$ for vanadium oxide array element sensors can been obtained.

### 7.4.2.3 PN junctions as Temperature Sensors

Silicon junction diodes can be used as accurate temperature sensors in the
IC-compatible Infrared Detectors

milliKelvin range [7.15]. These small temperature differences can be achieved only by good thermal isolation of the junction from the bulk silicon. A disadvantage the relatively large offset of around 600 mV with a typical temperature sensitivity of 2 mV/K. Thermal detector arrays have been fabricated by bulk micromachining of CMOS wafers with PN junctions in the epilayer, as shown in figure 7-6 [7.16].

A sensitivity $S=4900 \text{ V/W}$ and a detectivity $D^*= 9.7 \times 10^8 \text{ cmHz}^{-1}\text{W}^{-1}$ has been reported.

7.4.2.4 Pyro-electric Sensors

Pyro-electric detectors are very sensitive for measuring temperature differences and, depending on the materials used, high output voltages can be obtained. Several papers discuss the use of piezoelectric polymer materials processed on silicon [7.36] [7.37].

---

**Fig. 7-7.** Bulk micromachined PN diode thermal detector.

**Fig. 7-8.** Pyro-electric detector pixel.
These materials are deposited as the last step in the processing of the device since they cannot withstand the regular processing steps in IC technology or they are not clean-room compatible. A disadvantage is that the heat flow in the piezoelectric layer in the same direction as the incident radiation and if the devices are very thin their thermal isolation is low. Nevertheless piezoelectric detectors are very promising as detector arrays in microsystems because of the recent developments of many new types of polymers in IC technology and also because of their small size, high speed and output voltage. For macro scale commercial devices typical output voltages of 4000 V/W and detectivities $D^* = 1.2 \times 10^9 \text{ cmHz}^{-1}\text{W}^{-1}$ have been obtained. For IC-compatible devices typical output voltages of 460 V/W and a detectivity $D^* = 2.6 \times 10^8 \text{ cmHz}^{-1}\text{W}^{-1}$ have been obtained [7.37].

7.4.2.5 Opto-mechanical Thermal Detectors

In opto-mechanical thermal detectors, yet another domain is added to the sensor transduction process. Infrared radiation absorbed by MEMS micromirrors causes small rotation angles by thermal expansion of their hinged beams. The small displacements of these mirrors can cause large deflections of a visible light source such as a laser diode that can be measured by e.g. an array of photodetectors [7.38]. In addition to the MEMS structure these sensor systems need a rather complicated long and aligned optical path. Although these optomechanical devices can be now readily fabricated using current MEMS processes, the extra optical path makes these devices rather unpractical.

7.5 Conclusions

Thermal isolation is a common issue for all thermal infrared detectors and improving thermal isolation can dramatically increase both the sensitivity and the detectivity of these devices. High thermal isolation can be realized in MEMS technology by placing the sensor on thin micromachined membranes, bridges or cantilevers, as has been discussed in chapter 5. Table 2 shows a quantitative
Conclusions

comparison between the discussed thermal sensors

**TABLE 2. Main properties of Thermal Infrared sensors.**

<table>
<thead>
<tr>
<th>Type</th>
<th>Compatibility</th>
<th>Thermal isolation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thermo-electric detectors (Seebeck effect)</td>
<td>++</td>
<td>+</td>
</tr>
<tr>
<td>Bolometers (High temp, coefficient resistors)</td>
<td>-</td>
<td>++</td>
</tr>
<tr>
<td>PN junctions as temperature sensors (diodes)</td>
<td>++</td>
<td>-</td>
</tr>
<tr>
<td>Pyro-electric detectors</td>
<td>--</td>
<td>+</td>
</tr>
<tr>
<td>Opto-mechanical thermal detectors</td>
<td>+</td>
<td>+</td>
</tr>
</tbody>
</table>

Although all of the above mentioned principles could be used for IC-compatible infrared detector arrays in infrared microspectrometers, thermopiles have been selected for use as detectors in the microspectrometer, because of the following features:

- Inherent sensitivity for temperature gradients
- Flat and wide spectral sensitivity
- Ease of fabrication.

The next chapter is therefore dedicated to the design and optimization of thermopile detectors on thin micromachined membranes.

7.6 References


Detectors for Mid-Infrared Microspectrometers


[7.18] T. Ichihara, Y. Watabe, Y. Honda and K. Aizawa, “A high performance amorphous Si-1xC_{x}:H thermistor bolometer based on micromachined structure”, Presented at International Conference on Solid-State Sensors
Conclusions

and Actuators (TRANSDUCERS'97), Chicago, USA, 16-19 June 1997, Page(s) 1253-1256.


Detectors for Mid-Infrared Microspectrometers


Conclusions

Detectors for Mid-Infrared Microspectrometers
Optimization of Thermal Infrared Detectors

An original idea? That can't be too hard.
The library must be full of them.
Stephen Fry.

8.1 Introduction

This chapter deals with the design and optimization of the sensitivity and signal-to-noise ratio of thermal infrared detector arrays based on thermopiles or bolometers. In literature several papers can be found that discuss the design and optimization of the sensitivity and noise of thermal detectors [8.9][8.10] [8.12][8.13]. Most papers compare the properties of bolometers and thermopiles in terms of different technologies, different materials, layer thickness, etc. used. Geometrical optimization of thermal sensors, given technological design constraints, is limited to remarks. Since this is an important issue for arrays of thermal detectors this is the main new topic in this chapter.

It is almost impossible to take into account all parameters that influence the trade-off between size, noise level, speed and sensitivity of a thermal detector and therefore some general assumptions have been made in section 8.2. The presented analysis in the following paragraphs result in guidelines and indicative values. Since many material properties and fabrication tolerances are generally not well known, this is a valid approach.
8.2 Design of Thermal Detectors

Figure 8-1 illustrates how design specifications and technology (design rules) interact in the different signal domains. The design specifications and constraints for these sensors must meet in the electrical, the thermal, optical and mechanical domain, resulting in an optimal choice of materials and design geometry.

![Design specifications](image)

*Fig. 8-1. Interactions in the design of MEMS thermal sensors.*

The possible geometries of thermally isolated MEMS structures have been discussed already in chapter 5. In this chapter it has been shown that in terms of thermal isolation bridges and cantilevers, as shown in figure 8-2, are the optimal shapes for thermopiles and hinged structures for bolometers. Some design parameters are contradictory in different domains. For example if the total sensor area is given, increasing the optical sensitivity by enlarging the area of the absorber decreases the space available for the thermal isolation and therefore the thermal sensitivity.
Figure 8-2a illustrates the optimization problem that will be discussed in section 8.3.4 and section 8.3.6. The sensitivity or the signal-to-noise ratio of the detector can be changed by the designer by moving the position of \( L_b \). In view of the fact that thermal sensors have a low sensitivity, design optimization is definitely worthwhile.

Sets of mathematical equations relating the design parameters in the different domains will be derived in the next sections leading towards scalable rules for an optimal geometry of the devices. As the thermopile output signals are small, the noise and drift of the readout-amplifier has been included in the analysis.

The design procedure of thermopile structures can be significantly simplified if some general basic assumptions are made:

1. The design area is limited. The membrane area of the device is limited either by optical requirements or by the required mechanical strength.
2. The temperature difference between the membrane and ambient is small. Therefore conductive and convective heat transfer in the beam dominate and heat transfer by radiation, other than from the infrared source, can be neglected.
3. The infrared absorber area of the detector can be considered an isothermal area. The absorber has a very high thermal conductance and a large heat capacitance with respect to the membrane and the thermopile strips.
4. A lumped element model can be used for the dynamic analysis (response time) of the detector. The heat capacitance is mainly determined by the absorbing layer and the thermal conductance by the supporting membrane.

5. The noise of the thermopile sensor itself should be less or in the same order of magnitude as the noise of a typical low-noise CMOS read-out amplifier, as will be discussed in chapter 10.

6. The number of pixels in the detector array is defined by the required optical resolution of the spectrometer system.

In addition for the targeted spectrometer applications, as will be discussed in chapters 12 and 13, the following statements apply also:

7. The device operates in vacuum. As a consequence we can assume no pixel crosstalk and no heat flow in the vertical direction. This reduces the thermal optimization to a two-dimensional problem.

8. The size of the absorber is defined by the size of the spectrometer exit slit.

9. Pixel crosstalk the detector array should be minimized.

Statement [5] defines a limit on either the thermal noise of the detector, i.e. the total resistance of the thermopile, if the noise of the amplifier is given, or defines the maximum noise of the readout amplifier if the detector noise is given. This sets a relation between the electrical domain i.e., the value (layout) of the parasitic resistance of the thermopile strips and the geometrical properties in the thermal domain.

Items [1] and [3] indicate that the absorber area is a highly thermal conducting surface. In practice this is true, since good infrared absorbers are generally relatively thick (metal) films with a low surface reflection coefficient [8.22] [8.23] [8.20]. Furthermore it is assumed that the absorber is the only area converting the incident optical energy into heat. This isolates and simplifies the thermal problem.

8.2.1 Thermal Sensors on Bridges and Cantilevers.

The above statements indicate that cantilever or bridge type of beams are the most efficient shapes for thermal isolation for Seebeck sensors. In these shapes the heat transport from the absorber to the environment takes place only via the thermopile strips. Hinged structures are best for bolometers, where only
two electrically conducting wires are needed for connections. Figure 8-3a and figure 8-3b show the cantilever and bridge type of thermopile sensor respectively.

![Fig. 8-3. Fabricated basic thermopile detectors:](image)

If we consider the cross-sectional area defined by the line A-A in the bridge detector of figure 8-2b it is easily seen that, because of symmetry, there is no heat flow across the boundary A-A. Therefore there is no difference in the thermal and electrical analysis between the cantilever and the symmetrical bridge structure. For the mechanical analysis this is not valid and for this reason the bridge and the cantilever detector are discussed separately in the mechanical analysis of sections [8.6.2] and [8.6.1] in this chapter. A detailed thermal analysis is given in the following paragraphs.

### 8.2.2 Thermal Design

Figure 8-4 shows the thermal model of a thermopile sensor. The thickness of the layers in the structure is generally prescribed by the fabrication process (doping level or deposition method). The main design parameters for optimization in the thermal domain are therefore the width and the length of the layers in the thermopile legs. The thermal conductivity of the resistive layers in IC technology is always much higher than that of dielectric layers, typical values from literature and our measurements are shown in the Appendix A Table-A3. The thermal conductivity of metals is even higher, and those metals which are the best electrical conductors are also the best thermal conductors. This relationship is based upon the fact that the heat and electrical transport both involve free electrons in the metal. In the following section the temperature profile of a thermopile on a cantilever beam is derived. The energy generated in the absorber
Optimization of Thermal Infrared Detectors

region \( (L_a < X < L) \) of the beam can be written as:

\[
P_{opt}(x) = A_{Abs} \eta(\lambda) I(\lambda) = (L - L_a) W \eta(\lambda) I(\lambda)
\]  \quad (8.1)

Here \( I(\lambda) \) is the uniform incident radiation intensity (W/m\(^2\)) and \( \eta(\lambda) \) the absorption coefficient of the absorber material.

![Fig. 8-4. Cantilever beam with thermopile and absorber area.](image)

In the steady state situation the heat flux by conduction is given by:

\[
q_{loss} = -\kappa \nabla T = -\left(\kappa_a \frac{\partial T^2}{\partial x^2} + \kappa_{air} \frac{\partial T^2}{\partial y^2} + \kappa_{air} \frac{\partial T^2}{\partial z^2}\right)
\]  \quad (8.2)

where \( \kappa \) is the thermal conductivity of the beam and the minus sign indicates that the flux is in the direction of decreasing temperature. In a first approach we will neglect the lateral flow in the \( y \) direction to the neighbouring pixels. The first term represents the heat transport in the \( x \) direction by conduction through the beam to the clamping points and equals:

\[
q_{loss-x} = -\kappa_{beam} \frac{\partial T^2}{\partial x^2}
\]  \quad (8.3)

The last term represents the heat loss by conduction from the surface of the beam and the absorber, via the gas, to the environment. The heat transport per unit of length in the \( z \) direction, by conduction through a quiet gas, equals:

\[
q_{loss-z} = -\kappa_{gas} W \left(\frac{1}{d_B} + \frac{1}{d_T}\right) \frac{\partial T}{\partial x} = -\left(\frac{W \kappa_{air}}{d^3}\right) \frac{\partial T}{\partial x}
\]  \quad (8.4)
Where \( d_B \) and \( d_T \) are the distances of the beam to the bottom and the top of the housing respectively. Combining equation (8.4) and equation (8.2) and energy conservation in the beam yields the static heat equation for the beam:

\[
\left( \frac{\kappa_{beam}}{Wt} \right) \frac{\delta T^2}{\delta x^2} + \left( \frac{W \kappa_{Air}}{d^3} \right) \frac{\delta T}{\delta x} = 0
\]

and the boundary equations:

\[
T(x = 0) = 0
\]

\[
\frac{\delta T(x = L_b)}{\delta x} = q_{opt}
\]

The resulting temperature profile along the length of the beam then can be found by solving above differential equation (Appendix B):

\[
T(x) = e^{\beta(L_b-x)} \left( e^{\beta x} - 1 \right) q_{opt}
\]

With:

\[
\beta = \left( \frac{W \kappa_{beam}}{d^3 \kappa_{beam}} \right)
\]

\[
q_{opt} = \frac{P_{opt}}{Wt} = \frac{(L-L_b)}{t} \eta(\lambda) I(\lambda)
\]

The temperature difference \( \Delta T \) across the cold \((x=0)\) and the hot junction \((x=L_b)\) equals:

\[
\Delta T = \int_{x=0}^{x=L_b} \frac{e^{\beta(L_b-x)} \left( e^{\beta x} - 1 \right) q_{opt}}{\beta} \, dx
\]

\[
\Delta T = \frac{q_{opt}}{\beta} \left( e^{\beta L_b} \left( 1 - e^{\beta L_b} \right) - 1 \right)
\]

Equation (8.7)) shows the temperature difference across the thermopile elements as a function of the optical power \( q_{opt} \), taking into account the thermal conductance of the beam and the heat loss through the surrounding gas. The next paragraph shows that an equivalent thermal conductance of the beam cross-sectional area can be defined and reveals the conditions when the heat loss through the gas can be neglected.
8.2.3 Beam Cross-section Thermal Analysis

In practice the beam consists of several materials and an equivalent thermal conductivity $\kappa_{\text{beam}}$ for the cross-sectional area can be defined. Depending on the materials and process used, the conductor layers can be stacked or placed side-by-side and we can distinguish between planar and stacked thermopile structures as shown in figure 8-5.

![Cross-section of a thermocouple element.](image)

**Fig. 8-5. Cross-section of a thermocouple element.**

- a). Planar structure.
- b). Stacked structure.

In IC technology metals or doped polysilicon layers are used for the fabrication of thermopiles. N- and P-doped polysilicon layers can be used in the planar structure of figure 8-5a. Stacked structures used differently doped polysilicon layers are unpractical, since they would require much more processing steps. A metal layer is always needed for interconnect and the fabrication of the planar and stacked thermopiles of figure 8-5 using a polysilicon and a metal interconnect layer requires less processing steps.

The thermopiles can be considered as $N$ single thermocouple unit cells thermally in parallel and electrically in series. Two possible types of unit cells are shown in figure 8-5a and b. It must be noted that, as can be seen from figure 8-5, using the minimum widths the stacked thermopile can hold two times more thermocouples on the same total area.

The conductors in a stacked structure are generally composed of a polysilicon and a metal layer. The equivalent thermal conductance $G_{\text{ucf}}$ of a unit cell for the planar structure with length $L_{\text{tp}}$ is given by a parallel combination of
the conductances of the different layers, and this can be calculated as:

\[
G_{wi} = \sum_{i=1}^{N} \frac{\kappa_i A_i}{L_{tp}} = \frac{2(W_c + W_s)}{L_{tp}} \left( \kappa_{SiO} t_{SiO} + \kappa_{SiN} t_{SiN} + 2 \kappa_{SiP} t_{SiP} \frac{W_c}{W_c + W_s} \right) \quad (8.8)
\]

\[
G_{wc} = \frac{2W_c \kappa_{SiP} t_{SiP}}{L_{tp}} \left( 1 + \frac{\kappa_{SiN} t_{SiN} + \kappa_{SiO} t_{SiO}}{\kappa_{SiP} t_{SiP}} \left( 1 + \frac{W_c}{W_c} \right) \right) \quad (8.9)
\]

where \(\kappa_i\) is the thermal conductivity and \(A_i\) is the cross-sectional area \((t_i W_i)\) of the layer \(i\). Equal widths and conductivity of the N- and the P-type polysilicon layer is assumed also. The minimum width of the polysilicon layer \(W_c\) and the spacing \(W_s\) between two conductors is given by process tolerances and mask alignment accuracy. The maximum size can be chosen by the designer. Also the equivalent thermal conductance \(G_{ucs}\) of a unit cell for the stacked structure with length \(L_{tp}\) can be calculated as:

\[
G_{ucs} = \frac{W_c \kappa_{SiP} t_{SiP}}{L_{tp}} \left( 1 + \frac{\kappa_{SiN} t_{SiN} + \kappa_{SiO} t_{SiO}}{\kappa_{SiP} t_{SiP}} \left( 1 + \frac{W_c}{W_c} \right) \right) \quad (8.10)
\]

If we assume that total width \(W\) of the cantilever is filled with \(N\) thermopile unit cells, the number of unit cells follows from:

\[
N_f = \frac{W}{2(W_c + W_s)} \quad \{ \text{planar structure} \} \\
N_s = \frac{W}{(W_c + W_s)} \quad \{ \text{stacked structure} \}
\]

The total thermal conductance \(G_f\) for the planar structure and \(G_s\) for the stacked structure can be derived from equation (8.8), equation (8.9) and equation 8.10 as:

\[
G_f = \frac{W_c W \kappa_{SiP} t_{SiP}}{(W_c + W_s) L_{tp}} \left( 1 + \frac{\kappa_{SiN} t_{SiN}}{\kappa_{SiP} t_{SiP}} \left( 1 + \frac{W_c}{W_c} \right) \right) \quad (8.11)
\]

or:

\[
G_f = G_{SiP} \left( 1 + \beta_{SiO} + \beta_{SiN} \right)
\]

or:

\[
G_f = G_{SiP} \left( 1 + \beta_{SiO} + \beta_{SiN} \right)
\]
In these equations the first term represents the thermal conductance of the polysilicon layer and the $\beta$ terms show the influence of the other layers. A high thermal conductance results in a lower temperature gradient and therefore the $\beta$ terms should be preferably $< 1$. The polysilicon layers also determines the electrical resistance and hence also the minimum noise level of the thermopile.

Generally, the thickness of the layers $t$ is prescribed by the fabrication process (doping level or deposition method). The thermal conductivities $\kappa$ of the layers are also governed by the fabrication process. The ratio of the width of the polysilicon layers $W_c$ and the spacing between these layers $W_s$ can be chosen within limits and can be approximated for the planar structure by:

$$2 \left( \frac{\kappa_{SiN}}{\kappa_{SiP}} \right) \left( \frac{W_c + W_s}{W_c} \right) \leq 1 \quad \text{\{planar structure\}} \quad (8.12)$$

The thermal conductivity $\kappa_M$ and the thickness $t_M$ of the metal layer are generally much larger than the polysilicon layer. The ratio of the width of the polysilicon layer $W_c$ and the metal layers $W_M$ can be approximated for the stacked structure by:

$$\left( \frac{\kappa_M}{\kappa_{SiP}} \right) \left( \frac{W_M}{W_c} \right) \leq 1 \quad \text{\{stacked structure\}} \quad (8.13)$$

The material and geometrical data for the process described in chapter 5 have been taken from Appendix A and Table 1, substitution of these values yields:

$$\left( \frac{W_c + W_s}{W_c} \right) = \frac{30 \times 0.3}{2 \times 1.3 \times 0.5} \equiv 7 \quad \text{\{planar structure\}} \quad (8.14)$$

$$\left( \frac{W_M}{W_c} \right) = \frac{30 \times 0.3}{156 \times 0.6} = 0.006 \quad \text{\{stacked structure\}}$$

Equation (8.8) shows that in order to reduce the effect of high thermal conductivity $\kappa_M$ of the metal layer its width should be very small compared to the polysilicon. Considering the minimal width design rule of the metal layer this is generally not possible. The unit cells for the stacked structure hence become very wide and number of elements $N$ reduces. In case of the planar structure the spacing between the conductors can be approximated to a factor 6.
larger than the width on the polysilicon conductors before the thermal conductivity of the nitride layer becomes dominant.

<table>
<thead>
<tr>
<th>TABLE 1. Typical geometries of the fabricated beams.</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Thickness SiN d_{SiN}</td>
<td>0.5 [μm]</td>
</tr>
<tr>
<td>Thickness PolySi d_{SiP}</td>
<td>0.3 [μm]</td>
</tr>
<tr>
<td>Thickness SiO d_{SiO}</td>
<td>0.3 [μm]</td>
</tr>
<tr>
<td>Thickness of metal (Al) d_M</td>
<td>0.6 [μm]</td>
</tr>
<tr>
<td>Width PolySi W_c</td>
<td>6.0 [μm]</td>
</tr>
<tr>
<td>Spacing of PolySi W_s</td>
<td>6.0 [μm]</td>
</tr>
<tr>
<td>Width Metal (Al) W_M</td>
<td>1.2 [μm]</td>
</tr>
<tr>
<td>Beam Width (Al) W</td>
<td>64 [μm]</td>
</tr>
<tr>
<td>Beam Length L</td>
<td>2000 [μm]</td>
</tr>
</tbody>
</table>

Using the dimensions of the process described in Chapter 5 we can calculate the sensitivity per unit of length for both types of thermopiles using data from Appendix A and Table 1. The thermal conductance of a beam with the planar structure (N/P-poly) of 1mm length has been found as $G_{f(1mm)} = 0.82 \times 10^{-6} \text{ W/K}$ and for the stacked structure (Metal/P-poly) $G_{f(1mm)} = 50 \times 10^{-6} \text{ W/K}$. This equivalent thermal conductance enables a comparison between the thermal conductance of the beam and the surrounding gas, which will be further evaluated in the next section.

### 8.2.4 Gas Thermal Conductance

The temperature difference $\Delta T$ across the thermopile, taking into account the heat loss through the air and the beam, has been given by equation (8.7):

$$\Delta T = \frac{q_{opt}}{\beta^2} \left( e^{\beta L_b} \left( 1 - e^{\beta L_b} \right) - 1 \right)$$

(8.15)

with:

$$\beta = \left( \frac{W \kappa_{air}}{d \kappa_{gas}} \right).$$
Substitution of some typical values of the fabricated beams yields for the ratio $\beta$ of both heat losses

$$\kappa_{\text{air}} = \kappa_{\text{beam}} = 0.024 \text{ Wm}^{-1}\text{K}^{-1}$$

$$d' = \text{wafer thickness} = 530 \mu\text{m}.$$  \hspace{1cm} (8.16)

$$\beta = \left(\frac{W \kappa_{\text{air}}}{d' \kappa_{\text{beam}}}\right) = 10^{-5} \text{ W[}\mu\text{m]}$$

where $W$ is the width of the beam in $\mu$m. For bulk-micromachined devices $W$ is in the same order of magnitude as $d'$ and $\beta$ will be small (typically $< 0.1$). Hence the thermal conductance of the beam is dominant in most cases for large air gaps and the temperature profile along the beam can be approximated by the linear equation:

$$T(x) = \lim_{\beta \to 0} \frac{e^{\beta(L_b-x)}(e^{\beta x} - 1)}{\beta} q_{\text{opt}} = x \ q_{\text{opt}}$$ \hspace{1cm} (8.17)

The thermal conductance $G(x)$ now depends on the beam conductance only and increases with the length given by:

$$G(x) = \frac{\kappa_{\text{beam}} W t_b}{x}.$$ \hspace{1cm} (8.18)

The total thermal conductance between the thermopile junctions $G(L_b)$ is hence given by:

$$G(L_b) = \frac{\kappa_{\text{beam}} W t_b}{L_b}.$$ \hspace{1cm} (8.19)

where $\kappa_{\text{beam}}$ is the equivalent thermal conductance of the beam as described in section 8.2.3.

### 8.2.5 Beam Temperature Profile

If we assume the thermal conductance in the bulk and the absorber region much higher than in the rest of the beam, we can write the temperature profile in
the beam region $0 < x < L_B$ as given by equation (8.18) and equation (8.6) as:

$$ T(x) = \frac{P_{in}(x)}{G(x)} = \frac{x(L-L_B)W}{\kappa_{beam}} I(\lambda) = \frac{x(L-L_B)}{\kappa_{beam}} I(\lambda) \quad \text{For: } (0 > x > L_B) \quad (8.20) $$

The temperature $T(L_B)$ at the absorber region $L_B < x < L$ is constant and follows from equation (8.20)) by substitution of $x = L_B$

$$ T(L_B) = \frac{\eta(\lambda)I(\lambda)}{\kappa_{beam}} L_B (L - L_B), \quad \text{For: } (L_a > x > L). \quad (8.21) $$

Figure 8-6 shows the temperature profile along the beam.

**Fig. 8-6. Temperature profile $T(x)$ along the beam.**

The temperature difference between the cold and the hot junction is therefore:

$$ \Delta T = \frac{L_B}{\kappa_{beam} t_B} (L - L_B) \eta(\lambda) I(\lambda) \quad (8.22) $$

A high $\Delta T$ can be obtained by a large absorber area $(L-L_B)$, long thermocouple wires $L_B$ and a thin beam $t_B$ with a low thermal conductivity $\kappa_{beam}$.

The temperature profile of equation (8.20) and equation (8.21) will be used for calculation and optimization of the sensitivity $S$ of both thermopile and bolometer sensors on bridges or cantilevers in the next sections.
8.3 Sensitivity

Another design approach is to maximize the thermopile sensitivity for obtaining the highest possible output signal. This is a better approach if the noise floor is not determined by the thermal noise of thermopile itself, but e.g. by the readout amplifier.

Material parameters that influence the sensitivity of a thermopile are the thermal conductivity \( \kappa_m \) and the electrical resistivity \( \rho_m \) of the conductors. The electrical resistance between the thermopile junctions generates thermal noise that adds directly to thermopile signal. A higher thermal conductivity of the membrane and the electrical connections results in a smaller temperature gradient and a lower output voltage. The thermal conductivity of metals and semiconductors is high, and those metals which are the best electrical conductors are also the best thermal conductors. In order to compare the different materials, a common definition for the figure of merit \( Z_m \) for thermo-electric materials is given as:

\[
Z_m = \frac{\alpha_m}{\kappa_m \rho_m^2}
\]

Since a thermocouple always needs two materials, the combination of the values of \( Z_m \) for both materials selected, should be as high as possible if the fabrication technology allows this. Highly doped N and P type polysilicon have the highest figure of merit \( Z \) of the IC-compatible materials. Table A1 in Appendix A gives the figure of merit \( Z_m \) for some typical IC-compatible materials. In the next section we will discuss more in detail the relations between the thermal and electrical domain and the consequences for the thermal sensor design.

8.3.1 Signal Level

The relation between the temperature gradient \( \Delta T \) and the output voltage \( V \) of an N element thermopile is given by:

\[
V_{TP} = (\alpha_1 - \alpha_2) N \Delta T
\]

where \( \alpha_1 \) and \( \alpha_2 \) are the Seebeck coefficients of the thermocouple materials.
For the highest electrical output signal, the difference in Seebeck coefficients of the thermocouple materials should be as high as possible and the number of elements $N$ should be large also. Appendix A Table-2 shows the Seebeck coefficients of some typical IC-compatible materials. From this table it can be seen that the Seebeck coefficient of N- and P-type polysilicon is relatively high with respect to the other materials. The Seebeck coefficient of the poly-silicon layers also depends on temperature [8.20] causing a non-linearity, however since the temperature gradients are in the milliKelvin range this effect can be neglected.

The temperature gradient $\Delta T$ between the thermopile elements has been given by equation (8.22) and substitution in equation (8.24) yields for the sensitivity $S$ of an $N$ element thermopile:

$$S = \frac{V_{tp}}{I(\lambda)} = \eta(\lambda) (\alpha_1 - \alpha_2) N \frac{(L - L_B)}{G_{beam}} = \frac{\eta(\lambda) (\alpha_1 - \alpha_2) N (L - L_B)L_B}{\kappa_{beam} W t_B}$$  \hspace{1cm} (8.25)

The thermal conductivity of a beam with the planar structure of 1mm length has been calculated already in section 8.2.3 and equals $G_f(1mm) = 0.82 \times 10^{-6}$ W/K and for the stacked structure $G_s(1mm) = 50 \times 10^{-6}$ W/K. The high thermal conductivity of the metal (Al-1%Si) results in a conductance of the stacked thermopile structure (Al-PolySi) of about a factor 60 higher than the N-and P-type polysilicon layers using some practical values.

The number of thermocouples of the planar structure that can be placed on a cantilever of width $W$ is given by $2N(W_c + W_s)$ where $W_s$ is the spacing between the wires and $W_c$ the width of the conductors. The number of thermocouples $N$ thus equals:

$$N = \frac{W}{2(W_s + W_c)}$$  \hspace{1cm} (8.26)

Substitution in equation (8.25) yields for the sensitivity $S$:

$$S = \frac{\eta(\lambda)(\alpha_1 - \alpha_2)}{\kappa_{beam} W t_B} \frac{W}{2(W_s + W_c)} (L - L_B)$$  \hspace{1cm} (8.27)

The sensitivity $S$ can be calculated from equation (8.27) using the data from Appendix A-Table A3 and Table 1. With $\eta(\lambda) = 1$ and $L = 1mm$ a value of
$S_f = 50 \, V/W$ (planar structure) and a value of $G_s = 10 \, V/W$ (stacked structure) have been found.

As a consequence the **planar structure is preferred due to its much higher thermal isolation**, while also the electrical signal will be higher since the Seebeck coefficient of the N-and P-type PolySi combination is higher. The stacked structure is easier to fabricate (one implantation mask less).

### 8.3.2 Noise Level

Because of the low signal levels of thermal detectors the signal-to-noise ratio (SNR) of a detector and its readout circuit is generally the most important design criterion. The Johnson noise level of the detector system is a combination of the noise of the thermopile resistance and the noise level of the amplifier.

#### 8.3.2.1 Amplifier Noise

Coherent detection, as will be described in chapter 9, is commonly used for the detector readout signal processing. Using coherent detection the relevant noise is in a narrow band around the chopper frequency. The noise of a CMOS amplifier, although generally not white but operating in the $1/f$ noise region, can be assumed white in this narrow band and the equivalent input voltage noise around the chopper frequency can be converted into an equivalent noisy resistor:

$$R_{aeq} = \frac{\nu^2_{n_{aeq}}}{4k_B T} \triangleq 7.8 \times 10^9 \nu^2_{n_{aeq}}$$

(8.28)

An equivalent input noise voltage the amplifier of e.g. $\nu^2_{neq} = 15 \, nV/\sqrt{\text{Hz}}$ at the chopper frequency equals the noise of a thermopile resistance $R_{TP} = 120 \, \Omega$. The equivalent noisy resistor, representing the amplifier and thermopile noise can now be written as the sum of the amplifier equivalent noise and the thermopile noise:

$$R_N^2 = R_{aeq}^2 + R_{TP_{eq}}^2$$

(8.29)
8.3.2.2 Sensor Noise

The resistance $R_{TPeq}$ of the total thermopile depends on the number of wires $N$ and also on the length $L_b$ of the thermopile wires. Each thermocouple consists of a strip of N and P-doped poly-silicon connected by a short metal interconnect wire. Neglecting contact resistances, this resistance can be written for the planar structure of figure 8-5 as:

$$R_{TPeq} = N \frac{L_b}{W} (R_{sqN} + R_{sqP}) = \frac{W}{2(W_e + W_f)} L_b (R_{sqN} + R_{sqP}),$$  \hspace{1cm} (8.30)

where $R_{sqN}$ and $R_{sqP}$ equal the electrical resistivity of the N- and P-type polysilicon.

The design of a thermal detector with a noise level far below the equivalent noise level of its readout amplifier is rather useless. Either the noise level of the readout amplifier at the chopper frequency, or the Johnson noise of the thermopile itself, can be a starting point for the design of the thermal detector. In general it is more practical to match the noise level of the amplifier to that of the detector.

8.3.3 Signal-to-Noise Ratio

The signal to noise ratio is by definition the ratio of the irradiant signal power and the noise power:

$$SNR(\lambda) = \frac{S_{TP}(\lambda)}{S_n^2},$$  \hspace{1cm} (8.31)

where $S_{TP}(\lambda)$ and $S_n^2(\lambda)$ are the power spectral densities of the output signal and the noise of the electrical output signal respectively.

Considering only the noise of the thermopile elements, the combination of equation (8.31) and equation (8.27) gives an expression for the signal-to-noise
ratio for a Seebeck sensor as in figure 8-6:

\[
SNR(\lambda) = \left[ \frac{\eta(\lambda)(\alpha_i - \alpha_s)}{\kappa_{Beam} \sqrt{2k_bT \left( R_{pp} + R_{np} \right)}} \right] \left[ \frac{1}{t_{Beam}} \sqrt{\frac{W}{(1 + \frac{W_s}{W_c})^2 L_b (L - L_b)}} \right] \quad (8.32)
\]

This expression gives the sensor signal-to-noise ratio as a function of the device geometry and some material parameters and some important conclusions for the design of thermopiles can be given, i.e.

- The \( SNR \) is proportional to the difference of the Seebeck coefficients and inversely proportional to the thermal conductivity \( \kappa_{Beam} \) and the thickness \( t_{Beam} \) of the beam or the dominant thermal conductor material.
- Increasing the spacing \( W_s \) of the conductors \( W_c \) with respect to their width increases the \( SNR \) of the detector. In general it will be practical to decrease the spacing of the conductor wires to the minimum design rule.

Assuming that the thickness \( t_{Beam} \) is set by the fabrication process or mechanical strength of the structure and a small ratio between the width and the spacing of the thermopile conductors (\( W_s/W_c \)) equation (8.32) can be simplified as:

\[
SNR = C_1 \sqrt{W} \sqrt{L_b (L - L_b)} \quad (8.33)
\]

where \( C_1 \) is a constant. Some conclusions, with respect to the device geometry, can be clarified from this expression, i.e.

- Increasing \( L \) increases the length of the absorber area and increases the \( SNR \).
- Increasing \( L_b \) increases the length of the thermocouple wires and decreases the thermal conductivity of the beam and increases the \( SNR \).
- Increasing the total width \( W \) of the cantilever or bridge increases the \( SNR \) (square root relation).
8.3.4 Thermopile Detector SNR Optimalisation

The total area of the cantilever of bridge is shared by the absorber and the thermopile area. Chip area and mechanical strength of the cantilever or bridge sets a limit to the total length \( L \) shared by the absorber and the actual thermopile. Therefore if we define the ratio of the thermopile and absorber area as \( \gamma \) and therefore \( L_b = \gamma L \) the signal to noise ratio can be rewritten as:

\[
SNR = C_2(L - \gamma L)\sqrt{\gamma L}
\]  

(8.34)

Taking the derivative of equation (8.34) with respect to \( \gamma \) and setting to zero gives the value for \( \gamma \) for the maximum SNR as:

\[
d\left(\frac{SNR}{\gamma}\right) = \left(-L\sqrt{\gamma L} + \frac{L(L - \gamma L)}{2\sqrt{\gamma L}}\right) = 0
\]

(8.35)

gives: \( \gamma = 1/3 \)

Equation (8.35) shows that for a maximum SNR \( \gamma = L / L_B = 1/3 \), i.e. the length of the thermopile legs should be 1/3 of the total length of the cantilever or 1/6 the total length of a bridge structure.

The next two paragraphs discuss the value of \( \gamma \) for maximum sensitivity and SNR for similar bolometer thermal detector arrays.

8.3.5 Thermopile Sensitivity Optimization

Defining the absorber / thermopile area ratio as \( \gamma \) and hence \( L = \gamma L_B \), equation (8.30) can be written as:

\[
S_{TP} = \frac{V_{TP}}{I} = \frac{(\alpha_1 - \alpha_2)N}{\kappa_{beam} I_{beam}} \gamma L (L - \gamma L) = c_1 \gamma L (L - \gamma L)
\]

(8.36)

The value of \( \gamma \) for a maximum sensitivity of the thermopile can be found by taking the derivative of equation (8.36) with respect to \( \gamma \) and setting to zero.
Optimization of Thermal Infrared Detectors

yielding:

\[
d\left(\frac{V_{ep}}{d\gamma}\right) = -\gamma L^2 + L (L - \gamma L) = 0
\]  

(8.37)

gives: \( \gamma = 1/2 \)

Equation (8.37) shows that for maximum sensitivity \( \gamma = L/L_B = 0.5 \), i.e. the length of the thermopile legs should be 1/2 of the total length of the cantilever or in case of a bridge 1/4 of total the length of that bridge.

8.3.6 Bolometer Sensitivity Optimization

The optimal sensitive of a bridge type detector array using resistive elements, as shown figure 8-2, can be different since a bolometer measures the average absolute temperature of the cantilever beam or bridge, while a thermopile measures the temperature difference at two points.

![Fig. 8-7. Single element of a bridge type bolometer array.](image)

Since the temperature differences in the sensor are small we can assume a constant temperature coefficient \( \alpha \) of the resistor material and the relative change in resistance being proportional to the average temperature rise in the membrane, therefore:

\[
\Delta R_{av} = \alpha R_{0} \Delta T_{av}
\]  

(8.38)

The average temperature of the membrane can be calculated from the
temperature profile in the beam of figure 8-6 as:

\[
\Delta T_{av} = \frac{1}{L} \int_{x=0}^{x=L} T(x) dx = \frac{1}{L} \left[ \int_{x=0}^{x=L} T(x) dx + \int_{x=L}^{x=L_B} T(L_B) dx \right]
\]  

(8.39)

Using equation (8.19) and equation (8.39) yields:

\[
\Delta T_{av} = \frac{P_{opt}}{\kappa_{Beam} \tau_{Beam}} \frac{(L - L_B)W}{L} \left[ \frac{1}{2} L_B^2 + L_B (L - L_B) \right]
\]  

(8.40)

Again if we define the ratio of the absorber to beam area as \( L_B = \gamma L \) it follows:

\[
\Delta R_{av} = \alpha \gamma R_0 \Delta T_{av} = c_1 (1 - \gamma) \left( \frac{1}{2} \gamma^2 L^2 + \gamma L (L - \gamma L) \right)
\]  

(8.41)

Or:

\[
S = \frac{\Delta R_{av}}{P_{opt}} = c_2 (1 - \gamma) \left( \frac{1}{2} \gamma^2 + \gamma (1 - \gamma) \right)
\]  

(8.42)

The maximum sensitivity \( S \) is for a value \( \gamma \):

\[
\frac{dS}{d\gamma} = \left( -2 + 2 \gamma + 3 \gamma^2 \right) = 0
\]

gives: \( \gamma = \sqrt{\frac{1}{3}} - \frac{1}{3} \equiv 0.55
\)

(8.43)

The calculated value of \( \gamma \) for maximum sensitivity of a bolometer is 0.55.

8.3.7 Bolometer Detector SNR Optimization

The signal to noise ratio of a bolometer with a resistance \( R_B \) follows from equation (8.33) and is given by:
**Optimization of Thermal Infrared Detectors**

\[
\text{SNR}(\lambda) = \frac{\Delta R B I}{\sqrt{4kT_B}} = \frac{\eta(\lambda)\alpha_R \Delta T_{av} I}{\sqrt{4kT_B}}
\]  
(8.44)

where \(\eta(\lambda)\) is the absorber efficiency and \(\Delta T_{av}\) is the average temperature rise of the device. Substitution of \(\Delta T_{av}\) from equation (8.39) yields:

\[
\text{SNR} = \frac{\alpha_R \sqrt{R_B I}}{\kappa_{\text{beam}}} \sqrt{\frac{W}{4kT}} \left( \frac{1}{2} L_B^2 (L - L_B) + L_B (L - L_B) \right)
\]
(8.45)

With \(L_B = \gamma L\) it follows:

\[
\text{SNR} = C_1 \frac{L^3}{\sqrt{L}} \left( \gamma (\gamma - 1)(\gamma - 2) \right)
\]
(8.46)

The maximum SNR can be found for a value \(\gamma\)

\[
\frac{d\text{SNR}}{d\gamma} = (\gamma (\gamma - 1)(\gamma - 2)) = 0
\]
(8.47)

gives: \(\gamma = 1 - \frac{\sqrt{3}}{3} \equiv 0.42\)

The calculated value of \(\gamma\) of a bolometer for an optimal SNR is 0.42.

### 8.3.8 Conclusions

Table 2 shows the calculated values for \(\gamma = \frac{L_B}{L}\) of Seebeck sensors and bolometers.
The calculated values of $\gamma$ for a maximum SNR of a bolometer and a thermopile detector differ around 20% and for a maximum sensitivity they differ only around 10%.

It should be noted that the described one-dimensional bolometer geometries are most likely used only in detector arrays with small widths. The proper use of meanders for the resistor and folded springs for the beam suspension can increase the sensitivity dramatically since this allows the fabrication of both high resistor values and high thermal isolation. Figure 8-2 illustrates a good and a bad design example of a bolometer detector on a hinged membrane.

<table>
<thead>
<tr>
<th>Table 2. Calculated values for the Absorber / total area ratio $\gamma$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\gamma = L / L_B$</td>
</tr>
<tr>
<td>Thermopile (maximum sensitivity) 0.5</td>
</tr>
<tr>
<td>Thermopile (maximum SNR) 0.33</td>
</tr>
<tr>
<td>Bolometer (maximum sensitivity) 0.55</td>
</tr>
<tr>
<td>Bolometer (maximum SNR) 0.42</td>
</tr>
</tbody>
</table>

A major problem of bolometers is that they are passive devices i.e. energy is needed to measure the change in resistance. A current or voltage has to be supplied causing both a voltage offset and a thermal offset due to self-heating.

Fig. 8-8. Bolometer design examples

a). $\Delta R$ is large (Good design).
b). $\Delta R$ is low (Bad Design).

A major problem of bolometers is that they are passive devices i.e. energy is needed to measure the change in resistance. A current or voltage has to be supplied causing both a voltage offset and a thermal offset due to self-heating.
8.4 Response Time of a Thermal Detector

In the introduction of this chapter has been assumed that the thermal conductivity of the absorbing layer is much higher than the thermal conductivity of the thermopile part of the membrane. A lumped element approach can be used for calculating the time constant of the detector, i.e. the thermal mass of the beam can be neglected and the heat capacity is formed by the absorber. The thermal conductance is from the thermopile strips only. The equivalent electrical circuit for the thermal analysis is shown in figure 8-9.

The transfer function of the detector can be written as:

\[ T_{Ab}(f) = \left( \frac{R_{tp}}{1 + 2\pi f (R_{tp} C_{tp})} \right) q(f) \]  

(8.48)

The time constant of this first order system equals:

\[ \tau = R_{tp} C_{tp} = \frac{C_{Abs}}{G_{Beam}} = \frac{L_B}{\kappa_w W t_B} \rho_{Ab} c_{Ab} W_{A} L_A \]  

(8.49)

In the above equation \( C_{Ab} \) is the heat capacity, \( \rho_{Ab} \) the density and \( t_{Ab} \) the thickness of the absorber. The thermal conductivity is the combination of thermal conductivities of the layers in the membrane and has been given in equation (8.8). If \( L \) is the total length of the beam we can write \( L_B = \gamma L \) and \( L_A = (1-\gamma)L \).
yielding:

\[ \tau = \frac{P_{AB} C_{AB} l_{AB}}{\kappa_B} \gamma (1 - \gamma) L^2 \] \hspace{1cm} (8.50)

The thermal time constant of a thermopile detector is thus proportional to the square of the length of the membrane and does not depend on the width of the beam.

8.5 Pixel Crosstalk in Detector Arrays

In the array configuration trenches are made between pixels to prevent heat leakage from one pixel to neighbouring one, as can be seen in figure 8-14. The heat leakage by air through the gap between neighbouring pixels is hard to predict using an analytical model because of the large fringe effects. The cross-talk between two neighboring pixels with a 4 μm gap has been simulated with finite element model software [8.25]. Heat was imposed on one pixel and the temperature differences between the hot and cold junction of the heated pixel and the neighbouring pixels have been simulated using COMSOL. The thermal conductivity of air has been assumed \( \kappa_{\text{Air}} = 0.03 \) W/mK. The temperature difference of the simulation indicates approximately 47% energy leakage from the heated pixel to the neighbouring pixels.

8.6 Mechanical Design

In order to compare mechanical properties of cantilevers and bridges with same active area and fabricated in the same process an analysis the strength of cantilever beams and bridges with a uniform load will be discussed in the next section.

8.6.1 Cantilever Stress Analysis

The deflection \( z \) of a cantilever beam with uniform load \( \sigma_l \) as shown in figure 8-10 can be found by solving the Euler-Bernoulli equation [8.15]:
Optimization of Thermal Infrared Detectors

\[ \frac{\delta^2}{\delta x^2} \left( EI \frac{\delta^2 z}{\delta x^2} \right) - \sigma_t = 0, \quad (8.51) \]

where is the $E$ is the Young’s modulus and $I$ the area moment of inertia or structural rigidity. The forces, shear stresses and bending moments for many types of structures can be found systematically by repetitive integration of this deflection equation and setting proper boundary conditions.

In Appendix B listings of a Mathematica notebook that symbolically solves this problem for homogeneous cantilever beams and a bridge type of structures. The bending moment in a cantilever beam with a uniform load $\sigma_p$ and length $L$ has been found as:

\[ M (x) = -\sigma_l \left( \frac{L_p^2}{2} + L_p x + \frac{x^2}{2} \right) \quad (8.52) \]

This bending moment is largest at the clamping point $x=0$ and this results in a stress at the clamping point:

\[ \sigma_{t,0} = \frac{M_{z,0} d_z}{I} = -\sigma \left( \frac{t_b L^2}{2I} \right) \quad (8.53) \]

where $t_p$ is the distance from the clamping point to the neutral layer of the beam and $t_b$ the thickness of the beam. The area moment of inertia $I$ of a beam cross section is given by:

\[ I = \int z^2 dA \quad (8.54) \]
where \( z \) is the distance of the cross section area \( a \) from the neutral area as shown in figure 8-10. For the homogeneous rectangular beam of figure 8-10 the neutral plane is the mid-plane, yielding for \( I \):

\[
I = \int_{-\frac{W}{2}}^{\frac{W}{2}} \int_{-\frac{t_b}{2}}^{\frac{t_b}{2}} z^2 \, dz \, dy = \frac{W \, t_b^3}{12} \tag{8.55}
\]

with \( W \) the width and \( t_b \) the thickness of the beam. With \( A_b = W t_b \) the beam area and combing equation (8.53) and equation (8.55) yields for the stress at the clamping point of a cantilever loaded with a constant uniform load \( \sigma_L \) in the beam can be written as:

\[
\sigma_M = \frac{6L_b^2}{2Wt_b^2} \sigma_L \quad \text{or} \quad \sigma_M = \frac{3L_b^3}{A_b t_b} \sigma_L \tag{8.56}
\]

Conversely equation (8.56) shows that for a given maximum load of the beam and a maximum allowable stress and thickness of the nitride layer we could globally calculate the maximum allowable slenderness (\( L/W \) ratio) of the cantilever beam. However in practice it is hard to predict what the mechanical load during handling of the wafer and the die will be. The maximum deflection (at the end point of the beam) resulting from a load pressure \( \sigma_P \) on the beam can be written as:

\[
\varepsilon_{\text{max}} = \frac{qL^4}{24EI} = \frac{L^4}{2EI_b t_b} \sigma_P \tag{8.57}
\]

### 8.6.2 Bridge Stress Analysis

Figure 8-11 shows a thin bridge structure that has fixed supports on both ends, resulting in two forces and a moment at each support.
A Mathematica notebook that solves this differential equation using the boundary equations for a bridge type of structure can be found in Appendix B. For a uniform load \( \sigma_l \) we find for the bending moment at the clamping points \( l = 0 \) and \( l = L \):

\[
M_{l=0} = \frac{L^2}{12} \sigma_l \tag{8.58}
\]

Substitution of the flexural rigidity \( I \) of a rectangular beam results in a maximum stress at the clamping points:

\[
\sigma_{l=0} = \frac{M_{l=0} t}{I} = -\frac{L^2}{2W_b t_b^2} \sigma_l \tag{8.59}
\]

The stress on a bridge type of beam with a constant total area of the beam \( A_b = W_b L \) can be derived from equation (8.60) as:

\[
\sigma_b = \frac{L}{2A_b t_b^2} \sigma_l \tag{8.60}
\]

In order to comply with the results of the thermal analysis the mechanical analysis of the cantilever structure should be compared with a half bridge structure. Substitution of \( L = 2L_b \) in equation 8.60) yields an expression for the stress at the clamping point for the half bridge equivalent to the cantilever
structure:

\[ \sigma_{HH} = \frac{4L_b^3}{A_b t_b^2} \sigma_t \quad (8.61) \]

From equation (8.61) and equation (8.56) two important conclusions can be drawn:

The maximum stress in single or dual clamped cantilevers with a fixed area loaded with a uniform mechanical load is proportional to the third power of the length of the beam. In addition, by comparing equation (8.56) and equation (8.61) it can be seen that the maximum stress for a uniform load in cantilever beams (fixed-free beams) is comparable with bridge type of beams (fixed-fixed beams).

The maximum deflection of a dual-clamped cantilever is in the middle of the beam and calculated in Appendix B as:

\[ z_{max} = \frac{L_b^4}{384EI} \sigma_t = \frac{L_b^4}{32EW_b t_b^3} \sigma_t \quad (8.62) \]

The deflection of the cantilever should be compared with a half bridge, so \( L = 2L_b \) yields for the deflection of an equivalent bridge beam:

\[ z_{max} = \frac{L^4}{2EW_{bb} t_b^3} \sigma_t \quad (8.63) \]

which is the same as the deflection of the single-clamped cantilever beam. It must be noted, however, that residual stresses can cause deflections (buckling) in dual clamped beams while in cantilevers this is not possible since one end is free. Bending of single clamped beams bending can only occur by bi-layer effects, i.e. differences in internal stress of the different layers in the beam [8.26].

The strength of the beams is an important issue in the fabrication process. Internal stresses in the deposited layers can cause breaking of the membranes during etching. Certain deposition methods and combinations of layers can result in high built-in stresses in the nitride and polysilicon layers [8.27]. The residual stress in the beams and cantilevers can be reduced using silicon-rich Low Pressure Chemical Vapour Deposition (LPCVD) nitride layers [8.27]. The beams
are formed either by etching the bulk silicon from the backside by a wet etch step (KOH) and a final RIE step or by RIE etching of the membrane on the front side first and a final KOH etch to remove the bulk silicon. The RIE etch removes membrane material causing a better thermal isolation, at the same time the stress in the remaining structure increases and fracture of the SiN beams takes place during the final etch step. Another cause of failure is the dicing and sawing process which can cause high mechanical stresses on the very fragile beams.

Typical values of the fracture stress $\sigma_{fr} = 3-6$ GPa in thin SiN MEMS cantilever beams have been reported by [8.31], [8.32] and [8.33]. It must be noticed that there is a large spread in fracture strength over different samples, because LPCVD silicon nitride is a brittle material and fractures are most likely initiated by small flaws at the clamping point where the nitride layer meets the silicon.

Some details to improve the mechanical strength of the beams are shown in figure 8-12. Figure 8-12 shows a bridge thermopile supported by thin beams, these extra beams should not contribute significantly to the thermal conductivity but can improve the strength of the beam by reducing the bending moment at the clamping point. Stress concentrations in sharp corners can avoided by a gradual column shaped suspension near the corners of the beam clamping points.
As a result from the analysis in the previous paragraphs it can be conclude that in general bridges have a better mechanical strength for surface loads than single-clamped cantilever beams of the same effective area. The following sections give an optimization for the signal-to-noise ratio of thermopiles and optimization of the sensitivity of bolometers and thermopiles.

**Fig. 8-12.** Hinged bridge Thermopile.

- a). Detail photo of the clamping point.
- b). Detail showing corner stress and a crack.

As a result from the analysis in the previous paragraphs it can be conclude that in general bridges have a better mechanical strength for surface loads than single-clamped cantilever beams of the same effective area. The following sections give an optimization for the signal-to-noise ratio of thermopiles and optimization of the sensitivity of bolometers and thermopiles.
8.7 Results

Several thermopiles and arrays of thermopiles have been fabricated in the processes described in chapter 5 and by [8.26]. Figure 8-13 and figure 8-14 show some devices.

Fig. 8-13. Micro photographs of cantilever thermopiles (run EI-914, EI-1305).

Fig. 8-14. Micro photograph of a closed membrane with arrays of thermopiles on run EI-918.

Table 3 summarizes the performance of the thermopiles designed on run EI1442 of the fabricated devices.
Conclusions

This chapter has discussed the optimization of thermal sensors mainly in terms of sensitivity and geometry, assuming a given maximum active area of the device.

Highest thermal isolation for thermopile arrays on thin SiN membranes is obtained using cantilever or bridge type structures. For small sensors (L<1mm) on thin membranes on 500 μm substrates the heat loss through the air in the package and the resulting decrease in sensitivity, is relatively small. In a first approximation the sensitivity of the detector does not change with the width of the detector. There is no real optimum value for the length \( L \) of the beams; increasing the length increases both the sensitivity and the signal-to-noise ratio of the detector. A good design approach, in terms of sensitivity, is therefore to maximize the length/width ratio \( L/W \) of the cantilever (or a bridge) if possible. This is however in contradiction with the mechanical analysis of the beams, which states that for higher strength of cantilever structures the ratio \( L/W \) should be small. Also the response time of the detector increases with the length of the beam.

For highest sensitivity the thermal conductance of the electrical conductors should be in the same order of magnitude as the supporting isolating membrane.
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materials. In practice this is generally not feasible because the thermal conductance of electrical conductors is much higher than that of insulators.

The total area of the cantilever of bridge is shared by the absorber and the thermopile. For highest sensitivity the absorber area should be 1/2 of the total area. For the highest signal-to-noise ratio $\text{SNR}$, i.e. including the thermal noise of the thermopile resistance, results in an optimum value for the absorber area of 2/3 the total area of the device. The mechanical strength in terms of surface loading of cantilevers and bridges are comparable.

Assuming an ideal absorber a sensitivity $S$ of more than 200 V/W and a detectivity $D^*$ of $1.5 \times 10^8 \text{cm} \sqrt{\text{HzW}^{-1}}$ can be obtained by the devices fabricated. Due to the absence of absorber material, a typical sensitivity of around $S = 28 \text{ V/W}$ has been measured for most of the devices.

8.9 References


Conclusions


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Don’t worry about your problems with mathematics, I can assure you mine are even greater.
Albert Einstein.

9

9.1 Introduction

This chapter discusses the accurate amplification of the low frequency and low level signals of thermal detectors. Modulation of the optical signal or pulsing of the infrared source is generally required to distinguish the infrared optical signal from optical or thermal signals such as background radiation and drift of the infrared source or the detector. These disturbing signals are generally large and can saturate the sensitive electronics circuits for the processing of the optical signal. Generally these disturbing signals are removed by AC coupling. The low bandwidth of thermal optical detectors requires large time constants of the RC filters that cannot be realized with on-chip components.

A new method for offset reduction of modulated signals will be introduced in this chapter. The method is based on a-priori information about the modulated signal, which is generally a square wave. The presented method can be regarded as a sampled system where the zero samples are taken by synchronizing the sample and hold clock with the input signal. The system can be described as an autozero system if the optical domain is included. The term autozero can be confusing since autozero is normally used for offset reduction in amplifiers [9.11] [9.24] [9.25]. We will refer to this method therefore as Synchronous Correlated
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Sampling (SCS).

The proposed system will be explained in the time domain in section 9.4.1 and further analysed in the frequency domain. Two circuit implementations using this technique will be described and measurement results will be given.

9.2 Coherent Detection and Choppers

The low energy level of infrared signals and the low sensitivity of thermal detectors results in low electrical signal levels. The sensor signals are typically in the μV range and hence require high gain and low-noise signal processing circuits. Another major problem of thermal sensors in infrared systems is that unwanted optical and thermal effects such as background illumination, stray light and thermal and electrical drift of the sensor can be a few orders of magnitude higher than then the signals that have to be measured [9.1].

Modulation is commonly used to distinguish the optical signal from these large DC or slow drifting signals in infrared systems. Lock-in amplifiers [9.2][9.16], dedicated electronic circuits for coherent detection [9.3] [9.4] or special amplifiers [9.5] [9.6] are commonly used for signal processing. A typical readout system using optical chopping and synchronous detection is shown in figure 9-1.

![Fig. 9-1. A typical optical system using a mechanical chopper.](image)

Infrared light from the spectrometer is mechanically modulated
Coherent Detection and Choppers

("chopped") by the chopper wheel, and the low level signal from the detector is amplified and demodulated and filtered by a low-pass filter.

Optical chopping by periodically blocking the light source by a chopper wheel is generally the simplest and most adequate solution to modulate the sensor signal in conventional systems [9.1]. Typical chopper frequencies depend on the type of sensor and are in the range of 10 to 80 Hz depending on the bandwidth of the thermal infrared detector. The detector signal is amplified by a high-gain AC amplifier, demodulated, amplified and lowpass filtered. Coupling capacitors at the input of the amplifier are required for removing the offset in the detector signal and in the amplifier itself.

Another method is to modulate the infrared source, as shown in figure 9-2. Electronic chopping is preferred since no moving parts are needed and the system can be significantly reduced in size.

![Fig. 9-2. Coherent detection systems using a pulsed infrared source.](image)

High performance signal choppers (or phase reversal switches) can be readily realized in CMOS technology using sets of 4 identical PMOS or NMOS switches. A basic CMOS chopper circuit is shown in figure 9-3.
Chopping techniques are very popular in high performance analog circuits due to the simplicity and high performance of CMOS switches. Circuits with fast switches can be implemented that can switch from a very high off to a low on resistance using very low amounts of energy.

Most thermal infrared sources are large and have a high thermal time constant and are not suitable for electrical modulation. The thin-film infrared emitters presented in chapter 6 have a bandwidth in the same order of magnitude as the detectors and can be easily electrically modulated by applying a pulsed voltage. The typical sensor signals are described in the next paragraph.

### 9.2.1 Thermal Sensor Signals

Generally the field of view of the infrared detector is not completely filled by the infrared source and the detector is also illuminated by background radiation from the heater environment. A large DC signal level results and is added to the periodic on/off chopped signal from the source. A typical detector signal is shown in figure 9-4.

---

**Fig. 9-3. CMOS chopper circuit with equivalent symbol.**
The shape of the detector signal depends on the transient time of the infrared source and the detector. Typically this signal is a low-pass filtered square wave, due to the thermal time constant of the sensor. The information of interest is in the amplitude of the square wave and for measuring sensor signals a bandwidth of a few Hz or less is generally required. In terms of the signal spectrum the signal of interest is in narrow bands $B_S$ (0.1-4 Hz) around the fundamental frequency and the harmonics of the chopper frequency, as shown by

*Fig. 9-4. Typical signal shape of a chopped thermal detector.*
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the shaded regions in figure 9-5.

![Typical spectral plot of a chopped detector signal.](image)

**Fig. 9-5. Typical spectral plot of a chopped detector signal.**

The DC offset signal can be easily one of two orders of magnitude higher than the chopped AC signal. This DC offset can saturate the high gain amplifier(s) or the mixer that follows. As a result the amplifier input should be AC coupled to remove drift and DC signals before the amplifier. The offset the amplifier itself should be low also.

The transfer of chopper signals as low as 20 Hz with a small phase shift requires a corner frequency of the highpass filter in the order of magnitude of a few Hz. Capacitive coupling of amplifier stages with a time-constant of the several hundreds of milliseconds requires external capacitors. The fabrication of very large resistors and large coupling capacitors is an essential problem in IC technology. Low frequency high-pass filters are needed to remove the DC component in the input signal and will be discussed in the next section.

**9.3 High-pass Filter Techniques**

High-pass filters can be realized by separation and subtraction of the DC component from the signal. Basically this function can be realized at the system level by periodic nulling of the input signal, sampling the offset and subtraction.
from the signal. There two sampling techniques commonly used in analog IC
signal processing, i.e. autozero (AZ) and correlated double sampling (CDS).

- **Autozero**: Autozero circuits in amplifiers [9.12] [9.24] periodically
  remove the input signal, sample the amplifier offset and compensate it
  [9.7]. Autozero can be used reduce the offset of an amplifier but not the
  DC component in its input signal.

- **CDS**: In CDS systems the input signal is not removed and previous values
  of the input signal are delayed and subtracted from the input signal [9.7].
  A sample and hold circuit at the output ensures that the output signal is
  also available during the zeroing phase.

Autozero (AZ) and Correlated Double Sampling (CDS) are closely related
techniques based on sample-and-hold (SH) circuits to sample, store and subtract
unwanted signals. Correlated Double Sampling techniques can be used to create
high-pass filters [9.19] and will be described in more detail in the next paragraph.

Correlated Double Sampling is a high-pass filtering technique that was
originally used for compensation of drift and offset effects in CCD cameras [9.14]
and nowadays it is still used in both CCD and CMOS camera sensor read-out
circuits [9.15] [9.17] and in switched capacitor circuits for high-pass filtering and
noise reduction [9.20] [9.22]. The basic operation of Correlated Double Sampling
is illustrated in figure 9-6.

![Fig. 9-6. Block diagram of a CDS system.](image)

The input signal is periodically sampled, delayed and subtracted from the
Correlated Double Sampling (CDS) systems behave like highpass filters at frequencies below the sample frequency [9.7] and can be used for noise and drift reduction, as will be shown. The Synchronous Correlated Sampling (SCS) offset reduction method described in the next sections is closely related to the CDS technique and both systems can be described by similar mathematical equations.
9.4 Offset reduction by Synchronous Sampling

The offset reduction technique described in the next sections can be considered as a Correlated Double Sampling system with a sampling clock related to the input signal. Alternatively it can be regarded as an Autozero system at the system level. At the system level the optical input signal is periodically nulled by the optical chopper, or by switching off the infrared source, and the system samples and is nulled at the moment when the optical signal is at zero level. A description of the offset cancellation by sampling will follow in the next section.

9.4.1 Time domain analysis

The block diagram of the synchronous sampling technique is shown in figure 9-8.

*Fig. 9-8. Block diagram of the synchronous sampling system*

Figure 9-9 shows the timing diagram.
The chopped detector signal $V_{in}$ of the sensor is shown in the top plot of figure 9-9. A reference (zero) level is inherently available at periodic intervals within the detector signal, because of the on/off chopping of the sensor signal. A short pulse $\phi_{SH}$ with a repetition rate equal to the period of the optical signal is derived from the chopper. This sample pulse is also delayed. The reference level for sampling can be selected by the delay of the sampling clock $\phi_{SH}$ anywhere in the time window of the input signal period, but has been choosen at levels where changes in input signal are smallest, i.e. at $V_{min}$ or at $V_{min}+V_{sig}$. The sample timing is less critical here and phase jitter in the sampling pulse will have less influence here. The voltage for zeroing has been choosen at $V_{min}$ (light blocked or IR source switched off). When the sample clock signal $\phi_{SH}$ is high, the sample-and-hold is
in the track mode, following the input signal. On the falling edge the momentary value of \( V_{\text{min}} \) will be transferred to a hold capacitor, until the next pulse. The hold value of \( V_{\text{min}} \) is then subtracted from the input signal and the offset is reduced, as can be seen from figure 9-9 also. The transfer function for clock synchronous input signals equals:

\[
h(t) = \frac{V_{i}(t)}{V_{i}(t)} = \frac{V_{\text{sig}}(t) - V_{AZ}}{V_{\text{sig}}(t) - V_{\text{min}}} = 1 \quad (9.1)
\]

This equation shows that the AC component of the signal, \( V_{\text{sig}} \), is not attenuated. The presented method can be regarded as a CDS system where the zero samples are taken by synchronizing the sample and hold clock with the input signal. The system can be described as an autozero system if the optical domain is included. The synchronous sampling method as has been introduced here, should not be confused with the demodulation technique called synchronous sampling \[9.8][9.9][9.10\], although both concepts use input synchronized interpolating clock signals.

Aliasing and noise-folding are a common problem in sampled systems. The offset reduction technique should not introduce extra noise in the overall system and for this reason a frequency domain analysis will be given in the following sections.

### 9.4.2 Frequency Domain Analysis

#### 9.4.2.1 Introduction

In the next section an analysis of synchronous correlated sampling in the frequency domain is given. Equations for the output power spectrum will be derived. The presented method for the reduction of DC signals has not been found in literature. A simplified straightforward analysis will presented here for the sake of clarity. In Appendix 1c a more in-depth analysis of the system in the frequency domain is presented, based on the work of Pimbley and Michon [9.21] and a paper of Enz and Temes [9.7]. The effects of wideband signal and noise aliasing, which is a major problem in sampled systems, will be discussed here in more detail.
The basic sampler is shown in figure 9-10 and the timing diagram is shown in figure 9-10.

\[ v_s(t) = \sum_{m=-\infty}^{\infty} h_z(t) v_i(t) + h_H(t) [v_i(t) - v_i(t - mT_S)] \]  \hspace{1cm} (9.2)

where \( h_z(t) \) and \( h_H(t) \) are the window functions of the sample and the hold phase respectively, defined by:

\[
h_z(t) = \begin{cases} 1, & 0 < t < T_{AZ} \\ 0, & \text{otherwise} \end{cases} \quad \text{and} \quad h_H(t) = \begin{cases} 1, & T_{AZ} < t < T_S \\ 0, & \text{otherwise} \end{cases} \]  \hspace{1cm} (9.3)

Using these window definitions the output voltage \( v_o(t) \) can be written as

\[ v_o(t) = v_i(t) - v_s(t) = \sum_{m=-\infty}^{\infty} h_H(t - mT_S) \left[ v_i(t) - v_i(t - mT_S) \right] \]  \hspace{1cm} (9.4)

Equation (9.4) shows that the output signal equals the input signal minus delayed samples of the input signal filtered by the time window of the hold circuit. The Fourier transform of the rectangular time window function \( h_H(t) \) has
been derived in Appendix 1a as

\[ H_H(f) = \frac{1 - e^{-j2\pi f T_H}}{j2\pi f T_H} = e^{-j\pi f T_H} \text{Sinc}[\pi f T_H] \]  

where \( \text{Sinc}[\pi f T_H] \) represents the magnitude transfer function and \( e^{-j\pi f T_H} \) is a phase term. In the frequency domain the hold circuit can be considered as a lowpass filter. If the sample time \( T_AZ \) is very small then \( T_H = T_S \). The magnitude transfer function of the sample- and hold is then given by:

\[ |H_s(f)| = |\text{Sinc}[\pi f T_S]| \]  

The filter has a zero transfer function for frequencies \( f = m T_S \) where \( m \) is an integer. In other words, signals near the sampling clock or multiples thereof will be highly attenuated. The transfer function of the system equals:

\[ \left| \frac{v_o(f)}{v_s(f)} \right| = 1 - |\text{Sinc}[\pi f T_S]| \]  

Figure 9-11 shows the block diagram in the frequency domain. The transfer function of the sample- and hold and the total system is shown in figure 9-12.

**Fig. 9-11.** Frequency domain block diagram.
The overall system transfer $H(f)$ has a high-pass characteristic, reducing drift and offset and a signal gain of approximately unity at non-zero multiples of the clock frequency. The shaded areas in figure 9-12 show the gain for signals in a band around the chopper frequency ($f_T = 1$) and its harmonics ($f_T = 2, 3, \ldots$). This indicates that periodic input signals with a fundamental frequency equal to the sampling frequency can pass almost unattenuated. The time domain analyses in the previous section has confirmed this statement for a square wave signal.

**9.4.2.2 Noise Spectrum of SCS Systems**

The bandwidth of the amplifier in a chopper system needs to be typically a factor 5 higher than the chopper frequency for the amplifier to follow the slopes of the input signal and to avoid large phase shifts in the signal path. The synchronous sampling system operates at a sampling clock with a frequency equal to the input signal frequency, and wideband signals such as noise will be undersampled. Periodic sampling of the input signal results in duplicates at multiples of the repetition rate in the frequency domain, as has been derived in

\[ a) \text{ the sampler: } H_s(f) = \text{Sinc}(\pi f T_s). \]
\[ b) \text{ the total system: } H(f) = 1 - H_s(f). \]
Appendix 1b. The sampling action therefore results in periodic repetition of the input signal spectrum at the output, given by:

\[ v_s^*(t) = v_i(t) \sum_{m=-\infty}^{\infty} \delta(t-mT_s) \quad \Leftrightarrow \quad v_s^*(f) = v_i(f) \odot \frac{1}{T_s} \sum_{n=-\infty}^{\infty} \delta(f-n/T_s) \quad (9-8) \]

The output of the sampler is the convolution of the input noise spectrum with a series of Dirac pulses in the frequency domain. Figure 9-13 shows the aliasing of the noise in the input spectrum \( v_i(f) \) due to the undersampling of the white noise input spectrum \( S_{nw}(f) \) with a noise bandwidth \( B_n=4f_s \). As a result the noise in the spectrum of the output of the sampler increases \( v_s^*(f) \).

**Fig. 9-13. Aliasing of white noise due to undersampling.**

In the SCS system the sampling frequency equals the chopper frequency and the noise spectrum is filtered by the transfer function of the hold circuit, as given by equation (9-6). The frequency diagram of figure 9-14 shows the aliased noise and the filter function of the hold circuit.
Figure 9-14 shows an important feature of the synchronous sampling method, the contribution of aliased wide band noise in the signal bands, represented by the shaded areas around the sampling frequency and its multiples, is reduced by the filter function of the hold circuit.

**9.4.3 Summary**

An important conclusion from the previous analysis is that the synchronous sampling technique can suppress the DC component in modulated signals. Drift and other signals below the sampling frequency will be suppressed also. A-priori information on the shape of the signal is needed for deriving the sampling pulses. If the sampling time is short compared to the period of the sample clock the chopper signal passes through unattenuated. The effect of the aliasing of wideband signals and noise in narrowband modulated systems is reduced due the filter function of the sample-and-hold circuit. If the sampler is placed at the output of an amplifier the offset of the amplifier can be compensated also. In the following paragraphs two basic circuit implementations of the method will be introduced.
9.5 Circuit Implementations

Two circuit implementations of the SCS technique described in the previous paragraph have been designed, tested and fabricated to prove the concept of operation, namely:

- A circuit using sampling at the input of the amplifier: This is a very straightforward implementation, but the DC signal is sampled at the input and the amplifier offset is not compensated.
- A circuit implementation using output sampling: The DC signal is measured and compensated at the output of the amplifier and the amplifier drift and offset can be compensated also.

The circuits are described in the next two sections.

9.5.1 Sampling at the Amplifier Input

A very straightforward realisation of a zero-order hold and subtraction circuit is shown in figure 9-15. A simple passive sample-and-hold switch is placed at the input of a low-noise high impedance non-inverting voltage amplifier.

During the short sampling phase switches $S1$ opens and $S2$ close, and the
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hold capacitor \( C_h \) tracks the input signal. On the falling edge of the sample clock \( S2 \) opens and \( S1 \) closes again. The momentary value of the sensor signal is then stored on the hold capacitor \( C_h \) and subtracted from the input signal during the rest of the period. The input voltage minus the sampled value, is read by a non-inverting voltage amplifier with a very high input impedance as not to load the sampling capacitor. A low-noise and low-offset CMOS amplifier has been designed and fabricated and will be described in detail in the next chapter.

Some limitations of this configuration should be discussed:

- The offset of the amplifier is not compensated in this scheme. This results in a gain limit of the amplifier and a large ripple after the second chopper. For low-noise and low-offset operation the amplifier needs to be internally chopper stabilized. For low noise the chopper frequency needs to be above the \( 1/f \) noise corner frequency of the amplifier [9.25]. Chopper switches operating, typically in the range 10-40 kHz, are therefore needed at the input. The switching action results in a net input current that discharges the hold capacitor causing droop of the hold capacitor and a compensation error.

- Switches \( S1 \) and \( S2 \) cause charge injection to the amplifier input. Since the sampling frequency is low, the charge injection will be small also. For low-noise design the input transistors of the amplifier should be large, resulting in a large input capacitance compared to the switch capacitance, but since the input signals are small charge injection can be a problem. The charge injection of switches \( S1 \) and \( S2 \) can be, partly, compensated by adding identical dummy switches at the inverting input of the amplifier. The output resistance of the sensor limits the sampling pulse width.

The charge time of the hold capacitor by the sensor resistance poses a limit on the sampling pulse width and also poses a bandwidth limit on the input signal. If the source resistance is well-defined this can be taken into account in the design. A limit on the signal bandwidth reduces the amount of undersampled noise, which will shown later.

An advantage of this circuit is that large offsets can be compensated, since the system uses only passive switching. Measurement results of this circuit are given in section 9.5.3.

A circuit implementation using sampling at the output of the amplifier to
compensate for the amplifier offset also has been designed, fabricated and tested and is described in the next sections.

9.5.2 Sampling at the Amplifier Output

The block diagram of the AC amplifier using synchronous sampling at the output is shown in figure 9-16.

![Block diagram of the SCS system using an amplifier with an auxiliary input.](image)

The output signal is sampled with delayed sample pulses synchronized with the optical chopper signal as has been described before in section 9.4.1 and is shown in figure 9-16. The value $V_{min}$ is sampled by S1 each period of the chopper signal by a very short pulse ($T_{AZ} = 10 \mu s$ to $100 \mu s$) to the hold capacitor and a DC feedback loop is formed by the auxiliary input of the amplifier $g_{m2}$ and the amplifier output stage.
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The input amplifier stage $g_{m1}$ has a high transconductance, and the output stage has a transconductance $1/R_p$. If S1 is open, the output voltage $V_o$ can be calculated as

$$V_o = \left(\frac{1}{g_{m1}R_b} + \beta\right) V_i - \left(\frac{g_{m2}}{g_{m1}}\right) \left(\frac{1}{g_{m1}R_b} + \beta\right) V_{oc} \quad (9-9)$$

If we assume that the AC signal will be suppressed by the low-pass filter $R_pC_h$ then $V_{oc} = 0$ and the AC voltage gain of the amplifier can be derived from equation (9-9) as:

$$A_{ac} = \frac{V_o}{V_i} = \left(\frac{1}{g_{m1}R_b} + \beta\right) = \left(\frac{1}{g_{m1}R_b} + \frac{R_o}{R_f + R_o}\right) = \frac{R_f + R_o}{R_f} \quad (9-10)$$

The DC gain can be calculated by neglecting the RC filter and by setting $V_{oc} = V_o$ (switch S1 is closed) in equation (9-9), yielding a reduction in DC gain $\gamma$. 

---

Fig. 9-17. Timing diagram of the sample switch S1.
Circuit Implementations

of:

\[
A_{\text{DC}} = \frac{V_o}{V_i} = \left( \frac{A_{\text{AC}}}{1 + A_{\text{AC}} \frac{g_{m2}}{g_{m1}}} \right) = \frac{A_{\text{AC}}}{\gamma} \quad (9-11)
\]

Equation (9-9) shows that the DC gain can be reduced by increasing the feedback in the integrator feedback loop \( g_{m2} \) with respect to the gain \( g_{m1} \) and the gain setting \( A_{\text{AC}} \) (by \( C \) and \( R_f \)). As an example; if the DC offset reduction should be a factor \( \gamma = 20 \) for an AC amplifier with a gain \( A_{\text{AC}} = 100 \) and \( g_{m1} = 2 \text{ mA/V} \) the transconductance in the feedback loop should be equal to:

\[
g_{m2} = \frac{g_{m1}}{A_{\text{AC}}} (\gamma - 1) = 0.38 \text{ mA/V} \quad (9-12)
\]

In order not to reduce the AC gain, the time constant in the feedback loop needs to be much larger than the period of the input signal, i.e. the chopped detector signal. A time constant in the order of a few hundreds of milliseconds is required if the chopper frequency is in the range 20-100 Hz. A typical maximum size of an on-chip capacitor is around 100 pF and a time constant of 200 ms would require an on-chip resistor of 2 GΩ, which is not feasible. However the sampling of the output offset with a small duty-cycle can effectively enlarge this time constant since the hold capacitor can only charge/discharge via \( R_h \) during the short sample pulse. The time-constant of the RC filter is therefore enlarged by the reciprocal value of the duty-cycle of the sampling clock, as given by:

\[
\tau = \frac{T_s}{T_{AZ}} R_h C_h \quad (9-13)
\]

Using a sampling capacitor \( C_h = 100 \text{ pF} \) and a resistor \( R_h = 10 \text{ MΩ} \) a duty cycle of the sample pulse \( \delta = T_{AZ}/T_s = 0.01 \) results in an effective time constant of \( \tau = 100 \text{ msec} \). The highest practical value of the time constant is limited by leakage currents, discharging the hold capacitor. The capacitor should be a low leakage on-chip (metal) capacitor and the switch should be a small MOS transistor for low leakage.
Double Sampling

The output signal of the amplifier is not symmetric around this reference level as can be seen in figure 9-17. A DC offset of half the amplitude of the chopper signal is inherently present. For optimal use of the dynamic range of the following circuits a symmetrical signal around a certain reference level is often preferred. To remove all DC components sampling on both the high- and the low level of the chopped square wave is implemented as shown in figure 9-18.

The large time constant $\tau$ of the RC filter causes averaging over many samples. The feedback loop resulting in a DC signal on the hold capacitor that compensates the average DC signal level at the output. In practice $V_{oc}$ can be referenced to ground, but also to a reference level for the amplifier or demodulator that follows.

9.5.3 Measurement Results

For both implementation CMOS amplifiers have been designed with low noise as the main parameter. For the system with input sampling a low-noise CMOS amplifier has been designed and tested. The amplifier with an auxiliary input can be chopped internally also, to modulate and reduce the effect of the 1/f noise. The amplifiers are described in detail in the next chapter and the
measurement results of the overall systems using these amplifiers are given in the next section.

Complete readout systems including choppers for modulation and demodulation, based on the block diagram of figure 9-2, have been realized in the AMIS 0.7u process. A chip photo of the fabricated circuits is shown in figure 9-19. The top right corner shows the sampling capacitors. The chopper amplifier is in lower right corner. The lower left corner shows the chopper switches for modulation of the infrared source.

Fig. 9-19. Chip photograph of the integrated coherent detection system for thermal sensors.

The measurement conditions have been choosen for a typical thermopile sensor application. The autozero and optical chopper frequency have been chosen at 44 Hz using a sampling pulse width of $T_{AZ} = 100 \mu s$. The output voltage of the thermopile was $100\mu V$ peak-peak with an offset of $2 \text{ mV}$. The measured performance of the circuit is summarized in Table 1.

The offset and drift reduction can be easily compared by applying or removing the clock. The effect of extra noise by the autozero circuit has been measured with the input shorted and synchronous detection at 44 Hz using a low-
pass filter with a 1 Hz bandwidth and a 18dB/oct rolloff.

TABLE 1. Measurement results of the AC readout system

<table>
<thead>
<tr>
<th>parameter</th>
<th>Input sampling</th>
<th>condition</th>
</tr>
</thead>
<tbody>
<tr>
<td>Effective chip size</td>
<td>2.2x2.1mm²</td>
<td>C_h = 100 pF</td>
</tr>
<tr>
<td>Offset reduction factor by SCS</td>
<td>30</td>
<td>Input offset = 1mV</td>
</tr>
<tr>
<td>Total output noise without SCS</td>
<td>16 μV</td>
<td>Amplifier gain: 100</td>
</tr>
<tr>
<td>System bandwidth 1 Hz</td>
<td></td>
<td>Amplifier BW: 70 KHz</td>
</tr>
<tr>
<td>Total output noise with SCS</td>
<td>29 μV</td>
<td>Amplifier gain: 100</td>
</tr>
<tr>
<td>System bandwidth 1 Hz</td>
<td></td>
<td>Amplifier BW: 70 KHz</td>
</tr>
<tr>
<td>Gain</td>
<td>100x</td>
<td>external resistors</td>
</tr>
</tbody>
</table>

The bandwidth of the amplifier (f_s=70 KHz) is much higher than the signal frequency at 44 Hz, for proper operation of the choppers. The output noise level of the amplifier is higher than expected because of this large bandwidth. The SCS switching spikes also add to the noise of the system. The design should be improved in this respect.

It must be noted that the circuit with the auxiliary input can compensate only small offset signal levels. The reason for this is that the auxiliary input stage is biased at a much lower current level than the main input stage in order to keep its noise contribution low, as will be discussed in chapter 10. As a result the auxiliary input stage can not fully compensate the input stage current. However if large DC offsets need to be compensated the auxiliary input stage must provide more current and its bias current and transconductance should be increased. For this the transistors in the auxiliary input stage need to be scaled up in size also, in order to keep their noise contribution low.
9.6 Conclusions

The signals of chopped thermal sensor readout systems have been analysed and an offset and drift reduction technique has been presented. A high-pass characteristic has been realized with a sampling capacitor, avoiding the use of large (external) coupling capacitors, by means of offset-sampling and offset-subtraction within the optical chopper loop. Drift and other signals below the sampling frequency will be suppressed also. The sampling frequency must be synchronized with the chopper signal and the zero samples are taken at the moments when the optical signal is at zero level by the optical chopper, or by periodically switching off the infrared source.

In this chapter the system has been explained in the time domain and further analysed in the frequency domain to study its transfer function in the chopper signal bands and the effects of noise folding by the sampling action into the chopper signal bands.

Two circuit implementations using this technique have been described and measurement results have been given. The concept allows integration of all critical components of the system in a CMOS process, avoiding large coupling capacitors. Two low-noise CMOS circuit implementations have been realized, i.e. an autozero circuit that removes DC signals at the input of CMOS amplifier using a zero-order hold circuit and a circuit that uses DC feedback with a sample-and-hold at the output of a low-noise CMOS amplifier, and an auxiliary input for offset compensation. The circuits will be applied for the readout of thermopile detectors in a miniaturized infrared spectrometer system.

The circuits presented prove the concept and can be further improved for a specific detector or by the implementation of other circuit topologies. Fully differential versions of the system, using instrumentation amplifiers can result in a higher CMRR and will have a potential better performance.

9.7 References

Signal processing for Thermal detectors


Conclusions


Signal processing for Thermal detectors
10

Low-Noise Amplifiers for Thermal Detectors

Art has to move you and a design does not, unless it is a good design.
David Hockney.

10.1 Introduction

For accurate amplification of the low-level signals of thermal sensors low noise is the main design parameter. This chapter discusses the design of low-noise CMOS amplifiers for high performance readout of thermopile arrays. The circuits have been used as the main amplifier in the readout system described in the previous chapter. Four amplifier configurations have been designed and fabricated:

1. A basic low-noise amplifier. The equivalent input thermal noise of the amplifier has been designed as 3.0 nV/√Hz and a gain of 40dB is required at a 20 kHz bandwidth.
2. The basic amplifier of 1.) has been extended with chopper switches to reduce flicker noise.
3. Auxiliary input stages have been added to 1.) for offset compensation.
4. Chopped auxiliary input stages have been added to 2.) for offset compensation.

The circuits will described in the next sections.
10.2 Folded Cascode Amplifier Analysis

The designs are based on the folded cascode amplifier topology which is a common implementation in high performance CMOS amplifiers [10.4]. Circuits using this topology have been described by [10.3] [10.4]. Figure 10-1 shows the basic circuit diagram.

![Basic circuit diagram of the amplifier.](image)

The main advantage of the folded cascode configuration, compared to e.g. current mirror loading, is that the drains of the input transistors are biased at the same voltage, resulting in better matching and an improved common mode rejection in the input stage. Another advantage of the folded cascode configuration compared with other configurations, such as the telescopic cascode, is that the input and output can be biased at independent voltage levels. Therefore the input and output voltage range of the folded cascode configuration is typically higher and includes the negative rail when using PMOS transistors as the input pair. Very large PMOS transistors M1 and M4 have been used for the input pair, since PMOS transistors exhibit less flicker noise than NMOS transistors [10.4]. The transistor parameters are given in Appendix D. For optimum noise performance the main input transistors are biased in weak inversion [10.4]. M1 and M4 have been designed as cross-coupled quads in the layout to improve
matching [10.9]. Current source loads using NMOS transistors M2 and M5 with a low \( W/L \) ratio and source degeneration resistors have been applied, since the use of resistors as loads reduces the gain of the input stage and also limits the input common mode voltage. Since M2 and M5 carry a large current, their contribution to the amplifier offset by mismatch cannot be neglected and these transistors and resistors have been implemented as cross-coupled quads also. The input stage is followed by the folded cascode current mirror of M7 and M9. A current mirror gives a single-ended output and a source follower output stage provides a low impedance output and isolates the gain stage from the output load. All bias currents in the circuit are coupled to a main bias current that can be externally set. This bias current can be changed a factor 3 to increase the noise performance and bandwidth at the expense of higher power dissipation.

### 10.2.1 DC Analysis

#### 10.2.1.1 Offset

The input stage consists of 4 groups of 4 large PMOS transistors in a common centroid layout for best matching and low initial offset. The current source transistors M2 and M5 carry relatively high currents, since they need to deliver the bias current for both the input pair and the folded current follower. A mismatch in both currents causes a large offset at the output, hence for good matching the load transistors and their source resistors \( R_o \) are large devices and have been designed in a common centroid layout. This results in an equivalent input offset of the amplifier of 0.5 mV typical.

#### 10.2.1.2 Gain

Figure 10-2 shows the basic amplifier. The gain of the input pair is high and the low source impedance of the cascode transistors ensures that all signal current from the input stage is transferred to the output. Assuming that the output buffer has a gain of unity the open loop DC gain of the main input of the amplifier equals:

\[
\frac{V_{out}}{V_i} = \frac{V_{in}}{V_i} = g_m R_{out} \tag{10-1}
\]
Low-Noise Amplifiers for Thermal Detectors

Where $R_{out}$ is the output resistance of the folded current follower. $R_{out}$ equals the output resistance of the current mirror in parallel with the output resistance of the cascode transistor M9:

$$R_{out} = g_{m10}R_{ds10}/g_{m9}R_{ds9}\{R_{ds1}/R_{ds2}\} \quad (10-2)$$

Since M2 has a much smaller $W/L$ ratio its output resistance is much smaller than the output resistance of M1, and we can write

$$R_{out} = g_{m10}R_{ds10}/g_{m9}R_{ds9}(R_{ds1}/R_{ds2} + g_{m9}R_{ds9}) \quad (10-3)$$

The output resistance of the folded current follower using the transistor data of Appendix D equals $R_{out} = 1.6 \times 10^6$ ohm and the gain of the amplifier is $A_m = 3280$.

10.2.2 AC Analysis

A detailed analysis of the AC small signal behaviour of the folded cascode amplifier has been described by [10.4] and [10.5]. From the extensive analysis presented in these papers it is shown that the small signal transfer function of a folded cascode amplifier can be simplified to the following two-pole equation

$$\frac{V_{sc}}{V_i} = \frac{A_{st}}{(1 + j \omega \tau_d)(1 + j \omega \tau_c)} \quad (10-4)$$

Here the dominant pole $\tau_d$ is always associated with the cascode output load $C_g$. The second pole $\tau_c$ is due to the cascode transistor where $C_g$ is the parasitic capacitance at the source of transistor M9

$$\tau_d = \frac{C_g}{g_{m1}}, \quad \text{and} \quad \tau_c = \frac{C_g}{g_{m9}} \quad (10-5)$$

For a gain setting $A_M = 100$ setting the compensation capacitor to $C_o = 32 \mu F$, results in an amplifier bandwidth of about $f_M = 100$ kHz, with a phase
margin of 60°. The transfer function of the output buffer equals:

\[
\frac{V_o}{V_{ic}} = \frac{Z_o}{(1 + g_{m1}Z_L)} \quad (10-6)
\]

If the load \(Z_L\) is capacitive this adds a pole: \(\omega_L = g_m1/C_L\). Generally this pole can be neglected.

### 10.2.3 Noise Analysis

The total output noise spectral density of a MOS transistor is given by the EKV model [10.4]:

\[
\frac{i_n^2}{\Delta f} = \frac{8kTg_m}{3} + \frac{K_i g_m^2}{W L C_m f} \quad (10-7)
\]

The first term is the white noise component, caused by the channel resistance of the device. The white noise output power is proportional to the MOS transistor transconductance. The second term is the flicker or \(1/f\) component. The flicker noise output power is proportional to the square of the transconductance and inversely proportional to the gate area. The origin of flicker noise in a MOS transistor is due to extra energy states existent at the boundary between the SiO\(_2\) and Si interface that can trap and release electrons in the channel. In CMOS amplifiers the noise is usually dominated by the \(1/f\) noise.

Sufficient gain is assumed in the input and intermediate stage of the amplifier in figure 10-1 so that the noise contributions of the output buffer stage can be neglected. The noise of transistors M11-M12 are common mode signals for the amplifier input stage and for reasons of symmetry they can be neglected.

The effective transconductance \(g_{mek}\) of the cascode transistor M7 equals

\[
g_{mek} = \frac{g_m}{1 + g_m (R_{ds1} + R_{ds2})} \quad (10-8)
\]
and transistor M8:

\[ g_{\text{ext}} = \frac{g_{m8}}{1 + g_{m8}r_{ds}} \]  \hspace{1cm} (10-9)

Since \( g_{m8}r_{ds} \ll 1 \) their effective transconductance is small and the contribution of the cascode transistors M7 and M8 to the output noise can be neglected also. Based on these assumptions the circuit of figure 10-1 can be transferred to the simplified equivalent circuit for noise analysis as shown in figure 10-2.

Figure 10-2 shows that the folded cascode amplifier topology forms a large current loop, in which the total equivalent output noise power can be easily found by adding the squared equivalent drain output current noise contributions of the individual transistors. Assuming all noise sources uncorrelated the total squared noise current at the output of the cascode in figure 10-2 can hence be written as

\[ \overline{\Delta I^2} = \overline{I_R^2} + \overline{I_L^2} + \overline{I_{M3}^2} \sum_{k,x} \overline{I_{W,x}^2} \]  \hspace{1cm} (10-10)
and since
\[ i_k^2 = i_{m_1}^2 + i_{m_2}^2 \quad \text{and} \quad i_k^2 = i_{m_4}^2 + i_{m_5}^2 \] (10-11)

it follows that
\[ \frac{i_k^2}{\Delta f} = \sum_{i=1}^{6} i_{m_i}^2 + i_{m_7}^2 + i_{m_9}^2 = \sum_{i=1}^{6} g_{m_i}^2 v_{m_i}^2 + g_{me_7}^2 v_{e_7}^2 + g_{me_9}^2 v_{e_9}^2 \] (10-12)

Where \( g_{mk} \) and \( v_{mk}^2 \) are the transconductance and input noise sources of the transistors and \( g_{me_7} \) and \( g_{me_9} \) are the equivalent transconductances of the cascode transistors. Assuming the transistors in both signal paths identical, i.e. \( M1-M3 \) equal \( M4-M6 \) and \( M7=M9 \) this yields
\[ \frac{i_k^2}{\Delta f} = 2 \left( \sum_{k=1}^{3} g_{mk}^2 v_{mk}^2 + g_{me_7}^2 v_{e_7}^2 \right) \] (10-13)

Hence the equivalent squared input noise voltage can be written as:
\[ \frac{v_{ni}^2}{g_{mni}} = 2 \left( \frac{v_{ni}^2}{g_{mni}} + \frac{g_{mc_2}^2 v_{c_2}^2}{g_{mc_2}} + \frac{g_{mc_3}^2 v_{c_3}^2}{g_{mc_3}} + \frac{g_{mc_9}^2 v_{c_9}^2}{g_{mc_9}} \right) \] (10-14)

The total squared equivalent input thermal noise voltage of the amplifier therefore equals:
\[ \frac{v_{ni}^2}{g_{mni}} = \frac{16kT}{3g_{mni}} \left( 1 + \frac{g_{mc_2}}{g_{mc_2}} + \frac{g_{mc_3}}{g_{mc_3}} + \frac{g_{mc_9}}{g_{mc_9}} \right) \] (10-15)

From this equation can be seen that the thermal noise contribution of the load transistors and the current mirror transistors can be small if their transconductance is lower than the transconductance of the input transistors. The transconductance of a MOS transistor in the saturation region is given by
\[ g_m^2 = 2 \mu C_{OX} \frac{W}{L} I_D \] (10-16)

The transconductance of the input stage increases with a larger \( W/L \) ratio of the input transistors. At small values of the current density in the channel (=large
Low-Noise Amplifiers for Thermal Detectors

W/L or low \( I_D \) the MOS transistor operates in weak inversion, resembling a voltage driven bipolar transistor. The transconductance of a MOS transistor in the weak inversion region is given by

\[
g_n = \frac{I_D}{2V_T} = \frac{I_D}{n kT} q, \quad \text{with } 1.5 < n < 2.5
\]

(10-17)

From equation (10-17) can be seen that the transconductance does not depend on the transistor parameters, as long as the device remains in weak inversion. Weak inversion operation gives the highest possible transconductance at a given value of \( I_D \). For optimal noise performance the input transistors should operate in weak inversion. The total equivalent input noise can be minimized therefore by increasing the drain current of the input pair, resulting in the typical trade-off between noise performance and power dissipation. Bandwidth and slew rate requirements cause the drain currents in all transistors to be in the same order of magnitude and therefore the transconductance of the load and current mirror stage should be decreased by choosing a low \( W/L \) ratio of the transistor M2-M3 (and M5-M6). The equivalent input flicker noise component is given by

\[
\sqrt{\frac{V_{n}}{2}} = \frac{2}{C_{OX} f} \left( \frac{K_{FP}}{W_1 L_1} + \frac{g_{m1}^2 K_{PN}}{g_{m1}^2 W_2 L_2} + \frac{g_{m3}^2 K_{PN}}{g_{m3}^2 W_3 L_3} \right)
\]

(10-18)

where \( K_{FP} \) is the flicker noise factor of the PMOS transistors and \( K_{FN} \) is the flicker noise factor of the NMOS transistors. Typically \( K_{FN} \) is a factor 10 higher that \( K_{FP} \) and the \( 1/f \) noise contribution of the N-type load transistors has to be taken into account. The thermal noise of the amplifier using the design parameters from Appendix D can be calculated from equation (10-18) as approximately 3.1 nV/√Hz.
10.2.3.1 Noise Reduction in Current Sources

In this paragraph it will be shown that adding degeneration resistors in MOS current sources, as shown in figure 10-4, can improve their noise behaviour.

The output noise current of the degenerated current source can be calculated by summing the individual noise contributions of the resistor \( V_{n2} \) and the MOS transistor \( V_{n1} \):

\[
\overline{i_n^2} = \overline{i_{n1}^2} + i_{nMOS}^2 = \frac{8 \cdot g_m^2}{1 + g_m^2 R_o} \left( 4 k T r + \frac{8 k T}{3 g_m} + \frac{K_f}{C_m W L f} \right)
\]  (10-19)

The equivalent input noise of the transistor-resistor combination can be rewritten as

\[
\overline{v_{mi}^2} = \frac{\overline{i_{n1}^2} + i_{nMOS}^2}{g_m^2} = \frac{8 k T}{3 g_m} + \xi_2 \left( \frac{K_f}{C_m W L f} \right)
\]  (10-20)

With: \( \xi_1 = \frac{3 g_m R_o + 2}{(1 + g_m R_o)^2} \) and: \( \xi_2 = \frac{1}{(1 + g_m R_o)^2} \)

If we compare this equation to the noise equation of the non-degenerated MOS (equation (10-7)) it can be seen that the thermal and 1/f noise components are reduced with a factor \( \xi_1 \) and \( \xi_2 \) respectively. Figure 10-4 shows the noise reduction as a function of the degeneration factor \( g_m R_o \). From both plots we can...
conclude that:

- Adding a source resistor $R_o$ decreases $\xi_1$ and decreases the $1/f$ noise of the circuit.
- Adding a source resistor $R_o$ decreases $\xi_2$ and decreases the thermal noise of the circuit.
- With increasing $R_o$, the $1/f$ noise decreases more than the thermal noise.

An important result from this is that both the thermal and the $1/f$ noise of current sources can be improved significantly by using large degeneration resistors. A large resistor value however limits the output voltage swing of the current source.

For thermal noise it can be seen from figure 10-4 that noise reduction is effective only if the source resistor $R$ is a factor 3, or more, higher compared to the transresistance of the MOS transistor. If the current source transistor is biased in weak inversion its transconductance follows from equation (10-21). Figure 10-4 shows that for a significant decrease (60%) in thermal noise of the current source the resistor $R_o$ should be larger than:

$$ R > \frac{3}{g_{m2}} > \frac{6V_T}{I_D} > \frac{160 \text{ mV}}{I_D} \quad (10-21) $$

In terms of biasing this means that the voltage drop across $R$ should be at least $6V_T = 160 \text{ mV}$. For a significant decrease (60%) in $1/f$ noise of the current source.
source the source resistor should be larger than:

$$R > \frac{0.8}{g_{m2}} \cdot \frac{1.6 \cdot V_T}{I_D} > \frac{42 \text{ mV}}{I_D}$$  \hspace{1cm} (10-22)$$

Again this means that the voltage drop across R should be at least $6 \cdot V_T = 42$ mV. Since the decrease in the $1/f$ component is higher than the thermal noise, the flicker or $1/f$ corner frequency will also change also by adding the resistor.

The $1/f$ corner frequency $f_c$ can be found by setting the flicker noise equal to the thermal noise component, i.e. combining equation (10-7) and equation (10-14). The $1/f$ corner frequency of the degenerated current source $f_{cR}$ hence follows from:

$$f_{cR} = \frac{\xi_1}{\xi_2} f_c = \frac{2}{3 \cdot g_{m2} R_o + 2} f_c$$  \hspace{1cm} (10-23)$$

The decrease of the $1/f$ corner frequency as a function of the product $g_{m2} R_o$ is plotted in figure 10-5.

$$\text{Fig. 10-5. Influence of the degeneration factor } g_{m2} R \text{ on the flicker noise corner frequency of a MOS current source.}$$

In current sources for biasing the reduction of the effective gain of the MOS transistor, by the use of degeneration resistors, is generally not an important factor. However significant improvements in the noise performance of these current sources can be obtained by adding degeneration resistors.
10.2.3.2 Flicker Noise

The flicker or 1/f corner frequency of the amplifier can be found by setting the 1/f noise equal to the thermal noise component. Taking only the noise contribution of the input transistors into account this results in a 1/f corner frequency of the amplifier:

\[ f_c = \frac{3qK_f}{16kT^2C_{ox}} \left( \frac{I_d}{W_1} \right) I_o \]  

(10-24)

Using the data of the AMIS-0.7μ CMOS process (Appendix D) on the input stage transistors biased in moderate inversion at \( I_d = 220 \mu A \) results a 1/f corner frequency of approximately \( f_c = 13 \text{ kHz} \). Figure 10-6 shows the simulation results of the noise spectrum of the amplifier.

![Simulated noise spectrum of the amplifier.](image)

At a chopper frequency of 44 Hz the equivalent input noise voltage of the amplifier is dominated by the 1/f noise component of 70 nV/Hz typically. The effect of the 1/f noise can be reduced by chopping, as will be discussed in the next
chapter. The simulation results of the circuit are summarized in Table 1.

**TABLE 1. Simulation results of the amplifier of figure 10-12**

<table>
<thead>
<tr>
<th>Condition</th>
<th>Condition</th>
</tr>
</thead>
<tbody>
<tr>
<td>Open loop Gain 3200</td>
<td>Bias =50uA</td>
</tr>
<tr>
<td>Open loop Gain 4100</td>
<td>Bias =100uA</td>
</tr>
<tr>
<td>Auxiliary input Gain 180</td>
<td>Bias =50uA</td>
</tr>
<tr>
<td>Dominant pole f = 1/(2πτD) 4 kHz</td>
<td>C₀ = 32 pF</td>
</tr>
<tr>
<td>Dominant pole f = 1/(2πτD) 44 kHz</td>
<td>C₀ =0 pF</td>
</tr>
<tr>
<td>Second pole f = 1/(2πτC) 11 MHz</td>
<td>Bias =50uA</td>
</tr>
<tr>
<td>Bias current range 10-120 μA</td>
<td>V cm =2 V</td>
</tr>
<tr>
<td>White noise 3.1 nV/√Hz 100 kHz</td>
<td></td>
</tr>
<tr>
<td>Flicker noise frequency 13 kHz</td>
<td>Gain 100</td>
</tr>
<tr>
<td>Bandwidth -3dB 110kHz</td>
<td>C₀ = 32 pF</td>
</tr>
</tbody>
</table>

The low frequency noise spectrum of the amplifier has been measured in an 40x gain configuration as shown in figure 10-7.

**Fig. 10-7. Test setup for measuring the low frequency noise spectrum of the amplifier.**

A Stanford Research SR560 low-noise amplifier set to a gain of 100 has been used to amplify the signal to typically 15 dB above the noise level of the data acquisition card. The data has been collected by the data acquisition card at a
10 μs sample interval. A dataset of about 1 million 16 bit samples has been processed by MATLAB to calculate the Fourier transform and power spectrum of the signal. In order to produce a more smooth noise power spectrum a moving average filter of 20 samples has been implemented before the data was plotted in figure 10-9.

The log scale shows the $1/f$ noise corner at around 10 kHz. A small 50 Hz mains component is visible. Also it can be seen that the noise in the 30Hz to 100Hz range of the spectrum is about 30dB higher than the thermal noise. For calibration a very small sinusoidal input signal (100 nV RMS) has been added to give a reference level at -80 dB. The spectral noise density in the 30Hz to 100Hz band has been measured using this reference level as $\sqrt{v_n^2} = 70 \text{nV}/\sqrt{\text{Hz}}$ typically.

### 10.3 Chopper Amplifier

The amplifier is chopped to reduce the effects of $1/f$ noise [10.12][10.13]. Figure 10-9 schematically shows how a chopper modulates the $1/f$ corner up and
the noise in the low frequency range is reduced.

![Fig. 10-9. Amplifier noise spectrum with and without chopping](image)

Low offset folded cascode chopper amplifier circuit implementations have been published by [10.4] [10.7] [10.8]. The amplifier chopper frequency is a compromise between the amplifier bandwidth and the flicker noise reduction. The amplifier with choppers is shown in figure 10-10.

![Fig. 10-10. Chopped folded cascode amplifier.](image)
Dynamic element matching of the current mirror transistors M3 and M6 is also implemented by an extra chopper. The gain of the amplifier has been chosen $A_M = 100$. For this gain setting and a phase margin of 60, by setting the compensation capacitor $C_o = 32 \, \mu F$, results in an amplifier bandwidth of about $f_M = 100 \, \text{kHz}$. The chopper frequency hence has been chosen at 25 kHz. At the amplifier main input NMOS transistors have been used as switches. The switches have been designed such that their “on” resistance does not contribute significantly to the noise of the amplifier.

![Spectral noise power (chopped at 25 kHz)](image)

*Fig. 10-11. Amplifier noise spectrum with chopping at 25 kHz.*

The noise spectrum has been measured using the setup of figure 10-7 and the resulting noise plot in figure 10-11 shows how the chopper modulates the $1/f$ peak to 25 kHz and the noise in the low frequency range is reduced. A comparison of the low frequency noise and offset of both amplifiers circuits is shown in Table 2. The main purpose of chopping the amplifier is low frequency noise reduction [10.12]. The amplifier is used as a low frequency AC amplifier in an optically chopped system, as described in chapter 9. In these systems the offset compensation is realized by an auxiliary input and this circuit implementation will be described in the next section.
The folded cascode amplifier as discussed in the previous sections has been extended with an auxiliary input stage. The auxiliary input stage is used for offset compensation as has been discussed in chapter 9. The amplifier circuit diagram is shown in figure 10-12.

The auxiliary input amplifier has already been discussed at the block diagram level in chapter 9. The signal gain of the auxiliary input can be derived from equation (9-9) in the previous chapter as

\[
\frac{V_A}{V_{AUX}} = -\left( \frac{g_{m1A}}{g_{m1}} \right) \left( \frac{1}{g_{m1}R_b} + \beta \right)
\]

(10-25)

since: \( \frac{1}{g_{m1}R_b} \ll \beta \) yields:

\[
\frac{V_A}{V_{AUX}} = -\frac{g_{m1A}}{\beta g_{m1}}
\]
Low-noise amplifiers, with- and without chopper switches, and or an auxiliary input stage have been fabricated in the AMIS 0.7μ CMOS process on a single die, as shown in figure 10-13.

10.5 Results and Conclusions

Low-noise amplifiers, with- and without chopper switches, and or an auxiliary input stage have been fabricated in the AMIS 0.7μ CMOS process on a single die, as shown in figure 10-13.
Results and Conclusions

The equivalent input thermal noise of the basic amplifier topology has been calculated as 3.0 nV/√Hz, from SPECTRE simulations a value of 3.1 nV/√Hz is found and a value of 4 nV/√Hz at 20 kHz has been measured. The calculated 1/f corner is 13 kHz and a value of 11 kHz results from SPECTRE simulations. The average input offset voltage is 0.51 mV measured over 25 samples.

The main target of chopping of the amplifier is noise reduction, since the amplifier is used as a low frequency AC amplifier in an optically chopped system, as described in chapter 9. Chopping at 4 kHz reduces the offset to 21 μV typical. The measured performance of the circuit is summarized in Table 3.

![Photograph of a fabricated CMOS die with the amplifiers.](image)

Fig. 10-13. Photograph of a fabricated CMOS die with the amplifiers.
Low-Noise Amplifiers for Thermal Detectors

TABLE 3. Measurement results of the amplifier of figure 10-1

<table>
<thead>
<tr>
<th>Condition</th>
<th>Open loop Gain</th>
<th>Auxiliary input Gain</th>
<th>Bandwidth -3dB</th>
<th>$C_o = 32 \text{ pF Gain } = 100$</th>
<th>Phase margin $C_o = 32 \text{ pF}$</th>
<th>Power consumption</th>
<th>Input voltage range</th>
<th>Output voltage range</th>
<th>White noise</th>
<th>Offset typical</th>
<th>Offset typical</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bias =50uA</td>
<td>3100</td>
<td>180</td>
<td>170 kHz</td>
<td>60°</td>
<td>2.5 mW</td>
<td>0 - 3.2V</td>
<td>Vdd = 5 V</td>
<td>Vdd = 5 V</td>
<td>3 nV/sqrt Hz</td>
<td>0.51 mV</td>
<td>21 μV</td>
</tr>
</tbody>
</table>

References


Results and Conclusions


Low-Noise Amplifiers for Thermal Detectors
11.1 Introduction

The design, fabrication and the results of a novel Mid-infrared microspectrometer based on Attenuated Total Reflection (ATR) will be discussed in this chapter. ATR infrared spectrometry is an important technique based on the interaction of a sample with the electric field of reflected light at the boundary of a medium. Unlike most other techniques, such as transmission and reflection spectroscopy, it is a surface measurement technique with the ability to measure highly absorbing samples. After a short introduction into the total reflection principle and Internal Reflection spectroscopy in general the MEMS fabricated silicon internal reflection element will be reported. Finally the MEMS complete spectrometer device will be presented, analysed and the measurement results of the fabricated device components will be given.

11.2 Total Reflection

Total reflection is an optical phenomenon where electromagnetic radiation (e.g. visible light) can, at certain angles, be totally reflected from an interface
Mid-infrared microspectrometer based on Attenuated Total Reflection

between two media [11.4]. Total internal reflection occurs when the light travels slower in the first medium (a larger refractive index) than the second medium and the incident angle is larger than the so-called critical angle. The critical angle for a silicon-air interface follows from Snell’s law:

\[ \sin \theta_c = \frac{n_2}{n_1} \quad \Rightarrow \quad \theta_c = \arcsin \left( \frac{n_{\text{air}}}{n_{\text{Si}}} \right) = \arcsin \left( \frac{1}{3.42} \right) = 17^\circ \quad (11-1) \]

An example of total reflection is shown in the photograph of figure 11-1. A laser beam is reflected twice at a 45 angle in a glass prism.

**Fig. 11-1.** Photograph showing Internal Reflection at a Glass-Air interface.

Reflection is 100% in case of no absorption in the other medium (Total Internal Reflection). Reflection of the beam at the boundary surface produces an electric field, called the evanescent wave near the boundary as shown in figure 11-2.

**Fig. 11-2.** Attenuated Total Reflection.
The presence of this electromagnetic field has been showed already in the
days of Newton by placing a second medium very close to the boundary surface
and then it can be observed that then the light can cross the gap, so then it is not
totally reflected. This effect is know as frustrated total reflection [11.4] and the
electromagnetic field at the boundary is called the evanescent field or wave,
presumably because it has been observed that the effect disappeared very rapidly
when moving away from the interface.

The evanescent field is well known in optics since it is essential in the
transmission of light in fibre optics. Fibre-optic sensors [11.1] and optical
chemical sensors based on waveguides [11.3][11.2] use total internal reflection
and the associated electric field for absorption measurements, but operate in the
visible or the near-infrared range. The reflected angle of a beam in fibres is
generally near the grazing angle, which causes a much smaller field and less
interaction with the sample than the device presented here, operating at a higher
internal reflection angle. Multiple reflections are used in waveguides or fibres to
increase the volume of the field interacting with the sample or curvatures are
fabricated to increase the internal reflection angle locally. The next paragraph
discusses the properties of the evanescent field more in detail.

11.2.1 The Evanescent Field

The evanescent field can be calculated by solving Maxwell’s equations at
the boundary interface between the two media. This analysis is beyond the scope
of this thesis and a full analysis for both parallel and perpendicular polarized light
covers several pages and has been given by several authors [11.10][11.16]. An
important result of this analysis is that the field intensity in the rarer medium is
non-zero and the field pattern is a non-transverse wave having components in all
directions. The decrease of the electric field in the rarer medium is exponentially
and is given by:

\[ E(y) = E_0 e^{-y/d_p} \quad \text{with:} \]

\[ d_p = \frac{\lambda}{2\pi \sqrt{n_1^2 \sin^2 \theta - n_2^2}} \]  \hspace{1cm} (11-2)

The factor \( d_p \) is called the penetration depth and depends on the wavelength,
the incident angle \( \theta \) and refractive indices of the high density medium \( n_1 \) and the
lower density medium $n_2$. Since the denominator of the equation for $d_p$ will be around unity, the penetration depth of the field will be always in the same order of magnitude as the wavelength of the radiation. Also there is a non-zero energy flow parallel to the surface. This field causes a lateral displacement between the incident and the reflected beam. This is known as the Goos-Hanchen shift [11.11] For parallel and perpendicular polarisation the lateral shift of the beam is given by [11.12]

$$x_{GH}^p = \frac{\lambda}{\pi n_1 \sqrt{\sin^2 \theta - \sin^2 \theta_c}} \tan \theta_c$$

and

$$x_{GH}^s = \frac{1}{\sin^2 \theta_c} x_{GH}^p,$$

(11-3)

where $\theta_c$ is the critical angle given by equation (11-1), $\theta$ the incident angle and $n_1$ and $n_2$ the refractive index of the high density and the lower density medium respectively. Both the penetration depth, from equation (11-2), and the lateral shift from equation (11-3) are in the same order of magnitude as the wavelength of the reflected light. The next section discusses the case that the rarer medium is not losless and Attenuated Total Reflection (ATR) occurs.

### 11.3 Internal Reflection Spectroscopy

Measuring absorption spectra, using the Attenuated Total Reflection (ATR) principle, has been independently published by both Fahrenfort [11.9] and Harrick [11.8] for material research in different disciplines and is a well-established technique nowadays in spectroscopy. Internal reflection spectrometry is based on the interaction of a sample with the electric field when total reflection of light occurs at the boundary of a medium, as shown in figure 11-3. A flow of energy only occurs in a direction parallel to the boundary and the amplitude of the standing wave pattern is damped in a direction normal to the boundary.
In losless media all incident rays above the so-called critical angle will be totally reflected back into the denser medium. However, in case of an absorbing sample, the intensity of the reflected light decreases due to an attenuation of the electric field that exists at the boundary surface (evanescent wave).

### 11.3.1 Attenuation of a single Internal Reflection

The relation between the incident and reflected intensity $R$ at a given wavelength depends on the refractive indices of both media, the incident angle of the light beam, and also on the polarization of the light, for parallel and perpendicular polarized light the attenuation can be calculated from the Fresnel equations [11.15]:

$$
T_{\|} = \frac{n_1 \cos \theta_1 - \sqrt{(n_2-i k_2)^2-n_1^2 \sin^2 \theta_1}}{n_1 \cos \theta_1 + \sqrt{(n_2-i k_2)^2-n_1^2 \sin^2 \theta_1}}^2
$$

$$
T_{\perp} = \frac{(n_2-i k_2)^2 \cos \theta_1 - n_1 \sqrt{(n_2-i k_2)^2-n_1^2 \sin^2 \theta_1}}{(n_2-i k_2)^2 \cos \theta_1 + n_1 \sqrt{(n_2-i k_2)^2-n_1^2 \sin^2 \theta_1}}^2
$$

(11-4)

In these equations $\theta$ is the incident angle, $n_1$ and $n_2$ the real parts of the refractive index of the high and the lower density medium ( =sample ) and $k_2$ the complex part of the refractive index of the lower density medium ( =sample ). Figure 11-4 shows, as an example, the calculated transmission as a function of the incident angle for both polarisation directions, using equation (11-4). for a silicon-air interface and a silicon-water interface assuming no loss in both
Mid-infrared microspectrometer based on Attenuated Total Reflection

materials.

Fig. 11-4. Transmission as a function of the incident angle.

\[ a) \] Silicon-to-Air interface (\(\parallel\) and \(\perp\) polarisation).

\[ b) \] Silicon-to-H\(_2\)O interface (\(\parallel\) and \(\perp\) polarisation).

From the plots in figure 11-4 a sharp rise in reflectance can be seen near the critical angle and total reflection occurs beyond the critical angle. A high sensitivity can be obtained therefore if the incident angle is near the critical angle. As an example figure 11-5 shows as the calculated transmission as a function of the incident angle if the sample medium (water) is absorbing \((k = 0.1)\).
The figure clearly shows a very large difference in transmission between an absorbing and a non-absorbing sample at angles just above the critical angle. The calculations given here are only valid for bulk materials. The analysis of very thin films is given by e.g. [11.13].

**11.4 Absorption Peak Detection**

As an example the absorption for a single reflection at the interface of silicon and water has been calculated. The real and imaginary part of the refractive index of water ($n_W$ and $k_W$) and silicon ($n_S$ and $k_S$) at different wavelengths in the Mid-infrared have been taken from [11.18], [11.19]and [11.20]. The values are plotted in figure 11-7.
Mid-infrared microspectrometer based on Attenuated Total Reflection

Fig. 11-6. Real and imaginary part of the refractive index of silicon ($n_{Si}$ and $k_{Si}$) and pure water ($n_{W}$ and $k_{W}$).

It can be seen that pure water has a high absorption or large $k_{W}$ at $\lambda = 3\mu m$.

Fig. 11-7. Mid-Infrared Internal Reflection of pure water using a silicon IRE.
The transmission spectrum in the Mid-infrared for a single reflection can be calculated using the data above and Fresnel equations for natural (unpolarized) light, and has been plotted in figure 11-7 at values $\phi_R = 19.5^\circ$ (critical angle), $45^\circ$ and $54.7^\circ$. This figure shows the highest transmission loss at the absorption peak ($\lambda = 3 \, \mu m$) at the critical angle. The absorption peak becomes less pronounced before and beyond the critical angle, however multiple internal reflections, as shown in figure 11-9, can increase the interaction with the sample and therefore the absorption.

The interaction of the evanescent field with the sample is highest near the critical angle, in practice this can not be easily achieved because of the dependency of the refractive index on the wavelength, alignment problems and collimation of the beam. In practice the reflected energy will be always smaller and less predictable. Other reasons for this are light scattering due to roughness at the interface and the non-ideal contact of the sample on the surface [11.15]. For optimum sensitivity the sample material must be placed in intimate contact with the surfaces where the total reflection takes place, i.e. on the top or on the reflecting V groove sidewalls. Preparation of samples is an important issue in spectroscopy in general [11.21] [11.6].

11.5 Properties of ATR Spectroscopy

This large difference in transmission and the unique properties of the evanescent field yield more pronounced peaks in the spectrum, as compared to measurements in transmission [11.15]. An important property of Attenuated Total Reflection Spectroscopy is, since it is a surface measurement technique, the ability to measure highly absorbing samples. Adding a thin surface layer on the IRE element, that selectively absorbs molecules of interest, enables the fabrication of many types of chemical or biological sensors. Internal reflection spectroscopy is a difficult technique for qualitative measurements in the visible and even more in the UV. In the infrared the demands on geometrical inaccuracies such as surface roughness are less critical, because at longer wavelengths surface scattering decreases, and the field depth and the interaction volume with the sample increases. Increasing the absorption by the sample can be easily realized by multiple internal reflections, as shown in the next paragraph.
11.6 Internal Reflection Elements.

Internal reflection spectrometry is implemented in conventional spectrometers by placing the sample on a so-called ATR or Internal Reflection Element (IRE) that is placed in the optical path of an existing spectrometer [11.14]. Figure 11-8 shows an ATR setup accessory of a commercial spectrometer system.

The Internal Reflection Elements used are macro-size, typically several centimetres in length and, depending on the application and spectral range, many different materials are used [11.7]. The dove-prism shape with multiple reflections as shown in figure 11-8b, is the most popular [11.8]. For mid- and far infrared spectrometers semiconductor materials like germanium and silicon are often used, because of their low absorption. Silicon also has a high hardness and low cost.

![Fig. 11-8. ATR option of a commercial spectrometer.](image)

_a). Optical path._

_b). Internal reflections in the IRE._

11.6.1 Micromachined Internal Reflection Elements.

In the spectrometer presented we have use anisotropic etching using KOH on double polished <100> wafers for fabrication of the internal reflection element.
Internal Reflection Elements.

As a consequence the etching angles are restricted for a <100> wafer to $\phi_e = 54.74^\circ$ and if the wafer is etched along the <110> plane, side wall angles of $\phi_e = 45^\circ$ can be fabricated also [11.17]. Including complementary angles typical reflection angles of $\phi_R = 19.5^\circ$, 35.2°, 45°, 54.7° and 70.5° can be used for internal reflection. Figure 11-9 shows some examples of how the infrared light can be guided in, internally reflected and coupled out in an etched silicon die.

Fig. 11-9. Light paths in the fabricated silicon IRE's.

a). Backside illumination.

b). Frontside illumination.

c). Frontside illumination with multiple reflections.

The high refractive index of silicon ensures that the infrared light can only exit at nearly the normal angle, as a consequence the light will be “trapped” inside the silicon and multiple reflections can be easily realized. The same KOH etching process as for the infrared source and detector, as has been discussed in chapter 5, has been used for the fabrication of various types of IRE elements.
11.7 Design of the ATR Microspectrometer

Figure 11-10 shows a cross-section of the fabricated spectrometer. The active elements in the spectrometer (the infrared source and detector array) are fabricated in the bottom die.

![Cross-section of the fabricated spectrometer](image)

The top die is basically a dove-prism ATR element defining the optical path by multiple reflections on the top, bottom and side wall surfaces. Figure 11-10 depicts the processing steps.

![Processing steps for the IRE element](image)
Design of the ATR Microspectrometer

In the double-polished top wafer, recesses of about $12 \mu m$ deep have been etched using a RIE etch step. Figure 11-8a shows a detail photo of a such a recess with a metal pattern grating.

![Fig. 11-12. Photos of the topwafer.](image)

*Fig. 11-12. Photos of the topwafer.*

- a). Recess with grating.
- b). Vgrooves by KOH etching.

Bulk micromachining has been used to etch the V-grooves and trapezium-shaped recesses on the other side of the top wafer. figure 11-13 shows a fabricated device on the bottom wafer.

![Fig. 11-13. Micro photograph of a spectrometer device on the bottom wafer.](image)

*Fig. 11-13. Micro photograph of a spectrometer device on the bottom wafer.*
Mid-infrared microspectrometer based on Attenuated Total Reflection

Wideband infrared radiation from the heater enters the top wafer through a slit. The infrared beam entering the die is multiple reflected at the bottom, top and etched V-groove surfaces and exits the silicon at the exit slit where the incident angle is less than the critical angle. The detector in this application consists of a single thermopile, fabricated in as described in chapter 5, thermally isolated on a cantilever beam. The thermopile detector is illuminated through the exit slit from the top wafer as shown in figure 11-14.

The maximum width size of the slit and the absorber area is the projection angle of IRE sidewall on the bottom surface of the top wafer. Using an anisotropic etched $<1,0,0>$ wafer with a thickness of $t_w = 525 \, \mu m$ the maximum exit slit width will be $W=381 \, \mu m$. Increasing the amount of optical power is therefore possible by increasing the length of the slit and the width of the absorber and thermopile accordingly. Scanning of the spectrum provides information about the type and amount of absorption in the sample as a function of the wavelength. For operation as a spectrometer optical interference filters [11.25][11.26] or spatial filters using transmission gratings [11.27] can be fabricated at the entrance or exit slit in the top wafer using standard IC technology processing steps.

The cavity between the two wafers must be evacuated to a pressure less than 1 mbar to minimize thermal leakage to enable the high temperature operation of the polysilicon heater wire. Aluminium has been used as the reflective material for the grating and the exit slit(s). Several different prototypes and individual parts of the spectrometer have been fabricated in a $15 \times 15 \, \text{mm}^2$ design area.

Fig. 11-14. Spectrometer optical path to the detector.
Polysilicon heaters on thin oxide-nitride bridges are used as infrared emitters and N+/P+ polysilicon thermocouples on thin membranes as infrared detectors [11.29].

### 11.8 Measurement Results

Assembly and testing of the complete system was not possible, due to a number of practical problems. The main problems encountered are the low yield of the large thermopile detectors (2x3mm²) that have been developed for this application and another major unsolved problem was the vacuum packaging of the heater element and the alignment.

Prototypes of infrared sources, infrared detector arrays based on thermopiles and many different IRE elements have been fabricated. Thermal infrared sources and thermopile infrared detectors arrays have been fabricated and tested for this application, these devices have been described more in detail in chapters 6 and 8.

The micromachined internal reflection elements, as described in section 11.6, have been tested using the setup shown in figure 11-7.

![Fig. 11-15. Photo of the test setup for measuring Internal Reflection.](image)

Infrared radiation from a large glowbar infrared source (Newport 6007) and a chopper was coupled into a fluoride-glass fiber using a set of two ZnSe lenses. The fiber illuminates the sidewall of the fabricated IRE from the bottom. Infrared
Mid-infrared microspectrometer based on Attenuated Total Reflection

radiation exited the silicon at the other sidewall and has been measured by placing another mid-infrared fiber coupled to a thermopile below the other sidewall. A large amount of the infrared light was lost by the inefficient optical coupling. A change of a few percent has been detected when a drop of liquid (water) was placed on the silicon IRE. A spectral scan was not possible due to alignment problems and light loss from the large beam of the available infrared monochromator (Jobin-Yvon Triax 180) resulting in very low optical power.

11.9 Conclusions

The concept and design of a novel integrated MEMS microspectrometer based on Internal Reflection in the bulk of a silicon die has been presented. In contrast to conventional spectrometers, the optical path is defined by wafer alignment, the accuracy of MEMS and integrated circuit lithography mainly. Chopping of the infrared source and detector signals is preferred the electrical domain, since then no moving parts are needed in the optical path.

Prototypes of infrared sources, infrared detector arrays based on thermopiles and different internal reflection elements (IRE’S) have been fabricated. Thermal infrared sources, based on thin polysilicon wires have been successfully fabricated and tested, the results have been presented in chapter 6. Thermopile infrared detectors arrays have been fabricated and tested [11.28]. The internal reflection effect in a silicon die has been demonstrated by applying water on the top die and measuring the change in transmission (a few %) using a wideband infrared source and infrared fibres for coupling to the IRE. A much higher change in transmission is expected if sharp mid-infrared filters, tuned to an absorption line of water, are included in the optical path. More qualitative research has to be done on the ATR effect in the top wafer.

For a complete microspectrometer system, many technological challenges such as the yield of the fabricated detectors, the integration of optical filters [11.24] and alignment, assembly (wafer bonding) and packaging problems still have to be solved. The presented Attenuated Total Reflection technique has, since it is a surface measurement technique, the ability to measure highly absorbing samples. By adding thin surface layers that selectively absorb molecules of interest many types of chemical or biological sensors can be fabricated. The unique properties of the evanescent field yield more pronounced peaks in the
spectrum, as compared to measurements in transmission or reflection [11.5].

11.10 Outlook

The presented concept allows integration of all critical components of the system: a thermal infrared source, optical filtering, the Internal Reflection Element, a detector array, and finally probably a signal processing chip. Also it will be possible to fabricate dual-beam systems where a sample can be compared with a reference. Dual beam systems are very popular in spectroscopy, since in these systems the spectral response of the source and the detector can be compensated.

A huge improvement in sensitivity can be obtained by the use of surface plasmons [11.2]. Surface plasmons are surface electromagnetic waves that propagate parallel along a metal/dielectric (or metal/air) interface [11.22][11.23]. They can be generated by infrared or visible light if the real dielectric constant of a metal is negative and larger than the positive dielectric constant of air or the dielectric. Typical metals that support surface plasmons are silver and gold, but metals such as copper, titanium, or chromium can also be used for surface plasmon generation. In a typical configuration, a metal film is evaporated onto the optical dense medium and light is totally reflected at the boundary and the evanescent wave penetrates through the metal film. The plasmons are excited at the outer side of the film. Since the wave is on the boundary of the metal and the external medium (air or water for example), these oscillations are very sensitive to any change of this boundary, such as the absorption of molecules to the metal surface.

11.11 References


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12 Microspectrometers based on Multilayer Interference Filters

12.1 Introduction

As has been discussed in chapter 4, microspectrometers based on mechanically tunable Fabry-Pérot filters or arrays of fixed Fabry-Pérot etalons are, due to their planar structure and their short optical pathlength, the most compatible optical filters in terms of MEMS and IC technology. Non-dispersive infrared (NDIR) spectrometers, as described in chapter 3, use Fabry-Pérot filters to filter the spectral components of interest.

The fabrication of Fabry-Pérot etalons involves the optical design and deposition of thin-film layers. Thin-film interference filters [12.4][12.6] are widely used in optics, since using many different materials [12.3], surfaces with virtually any desired transmittance or reflectance can be fabricated [12.1][12.2]. Practical applications of these filters are in anti-reflective coatings, spectral filtering and in heat reflecting or heat transmitting mirrors. IC-compatible fabrication of these filters, i.e. postprocessing of wafers after CMOS processing or micromachining, is also possible when suitable materials and deposition techniques are used.

The next section briefly discusses the basics of interference in thin-films.
and the design of highly reflective multilayer mirrors using this theory. Subsequently the design and fabrication of high-reflectance mid-infrared multilayers will be discussed. After this the design and fabrication of narrow band multilayer Fabry-Pérot filters using IC-compatible materials will be described. In the last sections of this chapter measurement results of the fabricated devices will be given and, based on these results, some applications and future work will be discussed.

12.2 Fabry-Pérot Filters

Fabry-Pérot interference filters, as described in chapter 2 and shown in figure 12-1, are the basic components for the realization of narrow band optical filters.

![Basic Fabry-Pérot structure](Fig. 12-1. Basic Fabry-Pérot structure.)

The first Fabry-Pérot filters were fabricated around the year 1900 using semi-transparent silver mirrors [12.7]. The progress that has been made in the last decades in the accurate computer controlled deposition of thin-film layers, new materials and new deposition techniques has resulted in a ongoing development of Fabry-Pérot types of filters. Most of the published recent research work on spectrometers use these devices for optical filtering. A complete 16 channel integrated microspectrometer system in the visible range has been fabricated by Correia et al. [12.8], by postprocessing of Fabry-Pérot filters on top of a CMOS wafer with an array of photodetectors and CMOS circuits for readout.

Mirrors fabricated by quarter-wavelength stacks of low loss materials with different refractive indices, as first described by Thelen [12.5], can produce a much higher reflectivity over a limited range of wavelengths. For this reason
these filters are commonly used in laser cavities [12.3], where high reflectance is required to obtain sufficient energy for stimulated emission.

The design of thin-film interference filters is a discipline in its own in the field of optics and several textbooks have been written on the design of many types of interference filters [12.1] [12.2] [12.11]. The next section briefly discusses the basics of thin-film interference and the design of highly reflective multilayer mirrors using this theory.

12.3 Stacked Bi-Layer Mirrors

The double bi-layer thin-film stack, as shown in figure 12-2, is a basic building block for many types of interference filters.

![Quarter wavelength structure](image)

*Fig. 12-2. Quarter wavelength structure.*

The thickness of these layers is in the same order of magnitude as the wavelength of the light. Layers with high (H) and low (L) refractive index are alternatingly deposited on a substrate. At each interface reflection occurs. No phase change occurs at the boundary of a low to a high refractive index material (LH interface) while a 180 phase shift occurs if light reflects from a HL interface [12.13].

If the optical thickness of the layers in the stack equals a quarter of the wavelength of the incident light (δ = λ/4) the various components of the incident light produced by the multiple reflections in the stratified medium will reappear at the front surface in phase. Constructive interference takes place...
and as a result the incident wavefront will be totally reflected [12.10]. For normally incident light the optical thickness $\delta$ of a layer is related to the physical thickness $d$ by:

$$\delta = n_s d,$$

(12-1)

with $n_s$ is the refractive index of the layer.

As an example figure 12-3 shows a simulation of the reflectance for normal incident light of different stacks of two different materials as a function of the wavelength. The reflectance of three different bi-layer stacks of polysilicon (H, $n = 3.51$) and silicon oxide (L, $n = 1.58$) are plotted in figure 12-3.

![Multilayer mirrors](image_url)

**Fig. 12-3.** mid-infrared reflectance of multiple quarter-wave layers.

a). PolySi - SiO (HL)$^4$.

b). PolySi - SiO (HL)$^8$.

c). PolySi - SiO (HL)$^{12}$.

Each layer pair is indicated by the value HL and a stack of $N$ layer pairs is indicated by $(HL)^N$. The quarter wave optical thickness $\delta$ of the H and L layer are tuned to $\lambda = 3\mu m$. From the figure it can be seen that the reflectance increases
with the number of layers in the stack and the reflectance remains high over a limited range of wavelengths, depending on the ratio of the low and the high refractive index [12.2].

Since these structures generally use non-conducting materials they are also named dielectric mirrors. Common low optical loss thin-film materials that can be accurately and uniformly deposited are fluorides (MgF$_2$, NaF, PbF$_2$), selenides (ZnSe), chlorides (PbCl$_2$), oxides (TO$_2$, Ta$_2$O$_5$, HfO$_2$, MgO, SiO$_2$) and many more [12.2]. Common low loss optical materials in the mid-infrared are (poly)Si, Ge and ZnSe.

12.4 Basic Thin-film Theory

A mathematical description of bilayer thin-film interference filters can be found in most textbooks on optics and thin-films [12.13][12.18][12.2]. The analysis involves the solution of Maxwell’s equations at the boundaries of the HL and LH interfaces and is due to its length beyond the scope of this thesis.

The final result however, shows that each layer can be elegantly represented by a 2x2 matrix [12.14][12.1]. This characteric matrix sets the relations between the input and output fields of that layer and involves only the material parameters and thickness of the layer:

$$
\begin{pmatrix}
E_1 \\
H_1
\end{pmatrix} =
\begin{pmatrix}
\cos \delta & i \sin \delta / n_s \\
i n_s \sin \delta & \cos \delta
\end{pmatrix}
\begin{pmatrix}
E_2 \\
H_2
\end{pmatrix},
$$

(12-2)

where $i$ is the complex operator, $n_s$ is the refractive index of the layer and $\delta$ the optical thickness of the layer. Matrix multiplication can be applied for the calculation of the transmission or reflection of multilayers and the characteristic matrix of a N layer pair, consisting of materials H and L can be generalized to:

$$M^N = (M_H M_L)^N$$

(12-3)

Once this matrix is calculated, it can be expressed as

$$M^N = \begin{bmatrix}
M_1 & M_2 \\
M_3 & M_4
\end{bmatrix},$$

(12-4)
from which the transmission and reflection coefficients of a multilayer stack can be calculated. The reflection coefficient equals [12.19]:

$$ r = \frac{(i M_i + n_l M_2) n_H - (i n_l M_4 - M_3)}{(i M_i + n_l M_2) n_H + (i n_l M_4 - M_3)} $$

(12-5)

Quarter wavelength layers, ie. $\delta = \lambda/4$, exhibit a high reflectance, as will be shown in the next section.

### 12.4.1 Quarter Wavelength Layers

For the special case of a quarter wave layer (QWOT layer = Quarter Wavelength Optical Thickness) $\cos(\delta) = 0$ and $\sin(\delta) = 1$ can be substituted in equation (12-2). The characteristic matrix of a quarter wavelength H layer can be simplified as:

$$ M_H = \begin{bmatrix} 0 & n_H / i \\ i n_H & 0 \end{bmatrix} $$

(12-6)

and the characteristic matrix of a quarter wavelength HL layer pair is found by matrix multiplication as in equation (12-3) for the H and L layer, yielding:

$$ M_{HL} = \begin{bmatrix} 0 & -n_H n_L \\ -n_H n_L & 0 \end{bmatrix} $$

(12-7)

The reflectance of a quarter wave HL layer pair can be derived from equation (12-5) as:

$$ R = |r|^2 = \frac{4 n_H^2 n_L^2}{(n_H n_L M_2 - M_3)^2 + (n_H M_4 + n_L M_1)^2} $$

(12-8)

Substitution of equation (12-7) in (12-8) yields for the reflection coefficient $R$ of a quarter wavelength HL layer:

$$ R = \frac{(n_H - n_L)^2}{(n_H + n_L)^2} $$

(12-9)
In case of large number $N$ of $HL$-layer pairs, the reflectance $R_N$ of the stack can be approximated by [12.1]

$$R_N = 1 - T_N = 1 - 4 \left( \frac{n_L}{n_H} \right)^N$$  \hspace{1cm} (12-10)

Using many layers, reflectances of more than 99.99 % can be obtained for quarter wavelength incident light.

12.4.2 Half Wavelength Layers

Another important building block is the half wavelength structure. If the optical thickness of a layer equals half the wavelength of the incident light a standing wave pattern occurs allowing the light to pass unattenuated. For a half wave layer we substitute $\cos(\delta) = 1$ and $\sin(\delta) = 0$ in equation (12-11). The characteristic matrix can now be written as:

$$M^N = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}^N$$  \hspace{1cm} (12-11)

The resulting matrix is the unity matrix, indicating no influence of this layer in a stack. A half-wave layer can be considered therefore as a neutral (or also called absentee) layer.

12.5 Compatible Thin-Film Materials

Some materials for thin-film deposition with low loss in the infrared are shown in Table 1. Thin-films of silicon oxide and silicon nitride have absorption peaks in the range $\lambda = 9-11 \, \mu m$ and low loss up to $\lambda = 8 \, \mu m$ [12.9]. Silicon oxide however, is better suited for the $L$ layer because of its lower refractive index ($n=1.5$) [12.9]. Silicon has low loss in the range $\lambda = 1.1-14 \, \mu m$ and is a good material for the $H$ layer.
12.6 Multilayer Filters

The analysis and design of practical multilayer filters is more complex and cannot be performed by hand. Dedicated software, such as TFCalc from Software Spectra [12.15], FilmStar from FTG software [12.16] and MacLeod from Thin Film Center [12.17] can be used for the design, calculation and optimization of thin-films.

12.7 Fabry-Pérot Etalons based on Multilayers

Fabry-Pérot etalons basically consist of two mirrors and a spacer layer forming a resonance cavity as shown in figure 12-1. In the previous section it has been shown that mid-infrared mirrors can be fabricated by stacking of quarter-wavelength layers of materials with different refractive indices, such as polysilicon and silicon dioxide, and spacer layers can be formed by half wave optical thickness (HWOT) layers. If many layers are used, the reflectivity can become very high [12.1]. For large numbers of layer pairs $N$ the reflectivity $R$ can be approximated by:

$$R = 1 - 4 \left( \frac{n_L}{n_H} \right)^N$$

(12-12)
The transmission of a Fabry-Pérot device is at a sharp maximum if constructive interference occurs in the spacer layer i.e. if the optical thickness of the spacer layer equals half the wavelength. A Fabry-Pérot etalon can be fabricated therefore in multilayer techniques by a half-wave (neutral) layer sandwiched between two quarter wave multi-layer mirrors as shown in figure 12-4.

![Fig. 12-4. Quarter wave multilayer Fabry-Pérot structure.](image)

As an example figure 12-5 shows a plot of the transmission of a thin-film Fabry-Pérot system, by the MacLeod thin-film software package.

![Fig. 12-5. Simulated transmission of a Multilayer Fabry-Pérot structure.](image)
The device is tuned for maximum transmission at 3 μm wavelength and thin-film layers of polysilicon (H, \(n=3.51\)) and silicon oxide (L, \(n=1.52\)) have been used. A detail of the plot of figure 12-5 is depicted in figure 12-6 and illustrates the resonance peak more clearly.

![Multilayer bandpass filter](image)

*Fig. 12-6. Detail of the resonance peak of figure 12-5.*

It must be noted that, if very sharp filters are used the transmitted mid-infrared signal energy to the detector will be very small. A better approach is therefore to use arrays of overlapping filters and find the spectrum by additional signal processing. From the simulations shown, it can be concluded that filters of 8 layers or more can provide adequate filtering for most mid-infrared applications. The next sections discuss the practical design issues and the fabrication process of the multilayer mid-infrared Fabry-Pérot filters.

### 12.8 Design of IC-compatible Multilayer Filters.

An array of Fabry-Pérot etalons has been designed [12.21] using TFCalc 3.3 [12.15]. The bottom and the top mirrors consist of 2 HL layer pairs. The spacer
layer is L material (SiO₂). The physical thickness of the layers is shown in Table 2.

### TABLE 2. Designed layer thicknesses for the filters of figure 12-7.

<table>
<thead>
<tr>
<th>Layer number</th>
<th>Material</th>
<th>Thickness [nm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Silicon (Substrate)</td>
<td>--</td>
</tr>
<tr>
<td>2</td>
<td>SiO₂</td>
<td>321</td>
</tr>
<tr>
<td>3</td>
<td>PolySi</td>
<td>144</td>
</tr>
<tr>
<td>4</td>
<td>SiO₂</td>
<td>321</td>
</tr>
<tr>
<td>5</td>
<td>PolySi</td>
<td>144</td>
</tr>
<tr>
<td>6</td>
<td>SiO₂</td>
<td>660-1020</td>
</tr>
<tr>
<td>7</td>
<td>PolySi</td>
<td>216</td>
</tr>
<tr>
<td>8</td>
<td>SiO₂</td>
<td>463</td>
</tr>
<tr>
<td>9</td>
<td>PolySi</td>
<td>144</td>
</tr>
<tr>
<td>10</td>
<td>SiO₂</td>
<td>321</td>
</tr>
<tr>
<td>11</td>
<td>PolySi</td>
<td>116</td>
</tr>
</tbody>
</table>

The thickness of the middle layer ranges from 660-1020 nm in 20 nm increments resulting in 20 resonance peaks from 2.2 μm to 2.86 μm with 30 nm wavelength increments..

*Fig. 12-7. Simulated transmission of the mid-infrared Fabry-Pérot filters.*
12.9 Fabrication of IC-compatible Multilayer Filters.

Common techniques for thin-film deposition are chemical vapor deposition, ion beam deposition, molecular beam epitaxy, and sputter deposition. The poly silicon and SiO$_2$ films for the designed Mid-infrared filters were deposited on silicon substrates by a computer controlled FHR MS150 sputter system at the LIMS Nanofabrication Laboratory MC2, of Chalmers University, Sweden. The system can handle substrates up to 150 mm and is equipped with four magnetron sputter cathodes and an optical spectrometer for on-line thickness measurement. The deposition of the two materials has been characterized and calibrated by measuring layer thickness and uniformity over a 6 inch wafer [12.21]. Samples of 10x10mm$^2$ of a diced silicon wafer have been prepared to deposit the layers for the filters.

12.10 Measurement Results

Optical measurements have been performed by an ellipsometer on the samples at three stages during deposition. Ellipsometric data analysis software was used to extract the thicknesses of the individual layers. The resulting measured physical layer thicknesses are shown in Table 3.

<table>
<thead>
<tr>
<th>Layer number</th>
<th>Material</th>
<th>Thickness [nm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Silicon (Substrate)</td>
<td>--</td>
</tr>
<tr>
<td>2</td>
<td>SiO$_2$</td>
<td>292</td>
</tr>
<tr>
<td>3</td>
<td>PolySi</td>
<td>145</td>
</tr>
<tr>
<td>4</td>
<td>SiO$_2$</td>
<td>472</td>
</tr>
<tr>
<td>5</td>
<td>PolySi</td>
<td>190</td>
</tr>
<tr>
<td>6</td>
<td>SiO$_2$</td>
<td>985</td>
</tr>
<tr>
<td>7</td>
<td>PolySi</td>
<td>197</td>
</tr>
<tr>
<td>8</td>
<td>SiO$_2$</td>
<td>491</td>
</tr>
<tr>
<td>9</td>
<td>PolySi</td>
<td>159</td>
</tr>
<tr>
<td>10</td>
<td>SiO$_2$</td>
<td>343</td>
</tr>
<tr>
<td>11</td>
<td>PolySi</td>
<td>130</td>
</tr>
<tr>
<td>12</td>
<td>SiO$_2$</td>
<td>7.1</td>
</tr>
</tbody>
</table>
The mid-infrared filter characteristics have been measured by a commercial Fourier Transform spectrometer at Chalmers University Sweden and are shown in figure 12-8.

The plot shows the transmission of the filters with a typical FWHM of 35 nm in the 2.4-3.2 μm infrared spectral range. The measured transmission peaks are shifted with respect to the designed wavelengths. Comparing the design data in Table 2 with the extracted data in Table 3 it can be found that maximum deviation occurs in the middle SiO₂ layer. The error is around 14 percent for this layer and this is the major reason for the wavelength shift of the filters. For polysilicon layers the thickness error is around 3 percent. It has been observed that the surface roughness of the layers increases with the thickness. Planarisation steps should be added to increase the quality of the filters if many layers are deposited.

12.11 Conclusions

This chapter has discussed the theory, the design and fabrication of thin-film
mirrors and Fabry-Pérot filters for use in Mid-Infrared microspectrometers. Sets of narrow band filters have been fabricated in the 2.4-3.2 μm infrared spectral range on 10x10 mm² silicon substrates. The FWHM of the filters is measured to be 35 nm. The most critical part in the fabrication of thin-film mirrors is precise thickness control. Theoretically it can be possible to control the thicknesses very accurately if each filter die is fabricated separately, however this is not practical or economical. More accurate calibration of the layer thicknesses, achieved by ellipsometric analysis, can help to much further improve precision in thickness control. Despite some shift in the response of the filter, narrowband silicon-compatible bandpass Fabry-Pérot filters have been fabricated, using multilayer thin-film mirrors.

12.12 Applications and Future Work

The filters are intended for use in Non-dispersive infrared (NDIR) mid-infrared spectrometers, as described in chapter 2. The dielectric thin-film filters discussed in this chapter, together with the integrated infrared thermal sources from chapter 6 and the integrated thermopile detector arrays described in chapter 8 form a relatively simple complete spectrometer system. Figure 12-8 shows a possible realisation of a fully integrated NDIR spectrometer for liquid samples.

![Fig. 12-9. Cross section of a fully integrated mid-infrared NDIR spectrometer.](image)

The transmission peaks of the optical filters should be designed to coincide with the absorption maxima and minima of the sample.
12.13 References

Microspectrometers based on Multilayer Interference Filters


Conclusions

"Writing a book is an adventure. To begin with, it is an amusement; then it becomes a mistress, next it becomes a master, and finally a tyrant.

Winston Churchill"

Although there is a growing trend towards miniaturisation, most spectrometers are still large and complex systems that would benefit greatly from miniaturisation. An overview of the state-of-art in infrared microspectrometers is given in chapter 4. A more widespread use of infrared spectrometers is mainly limited by the price and complexity of the available instruments. The competition of small portable microspectrometers, dedicated for specific applications, with the large bulky and expensive laboratory instruments is a great challenge.

The design and fabrication of a new mid-infrared microspectrometer based on Attenuated Total Reflection (ATR) was presented in chapter 11. The presented concept allows integration of all critical components of the spectrometer: a thermal infrared source, optical filtering, the Internal Reflection Element (IRE) and the detector array. The full potential of the complete ATR spectrometer has not yet been exploited, because the final processing steps have to be further developed. The fabrication and testing of the individual components however has shown promising results. For a complete microspectrometer system, many technological challenges such as the yield of the fabricated detectors, the integration of mid-infrared optical filters, alignment, assembly (wafer bonding) and vacuum packaging problems have to be further evaluated.
Conclusions

Narrow band Fabry-Pérot type of optical filters using multi-layer mirrors for the mid-infrared wavelength region were described in chapter 12. These filters are IC-compatible and have been successfully fabricated using thin poly-silicon and silicon-oxide layers. Sets of narrow band filters have been fabricated in the 2.4-3.2 \( \mu \)m infrared spectral range. The FWHM of the filters is measured to be 35 nm. Since the processing is IC-compatible these filters can be conveniently used in many types of mid-infrared spectrometers for narrow band optical filtering.

Since thermal isolation is a common issue for all thermal infrared detectors, improving thermal isolation can dramatically increase both the sensitivity and the detectivity of these devices. High thermal isolation can be realized in MEMS technology by placing the thermal sensor or actuator on thin micromachined membranes, as has been discussed in chapter 5.

An overview of uncooled infrared detectors and their properties was included in chapter 7. The design and optimization of the sensitivity and the signal-to-noise ratio of infrared detectors based on thermopiles was presented in chapter 8. Geometrical optimization rules for the design of infrared detectors based on thermopiles on cantilevers and bridges were also presented. The influence of: the size of the absorber area, the length and width of the thermopile legs, air conduction, planar and flat structures, etc. on the performance of the sensor was discussed also. The results of this thermal analysis are valid for cantilevers and bridges. Arrays of thermopile detectors were fabricated, but unfortunately most devices were broken after the final etch step. From these difficulties, that were encountered during several runs, we may conclude that the fabrication of cantilevers or bridges is much more critical than the fabrication of closed membranes. Large tensile stresses can easily cause failures in the fragile bridge and cantilever structures during the final etch step where the structures become freestanding. However the improvements in quality and reproducibility of the recent devices indicate good perspectives and a better yield of high sensitivity thermopile arrays using this technology in the future.

In most mid-infrared spectroscopy applications a well-defined infrared source is also needed. An infrared light source therefore should be considered an integral part of most spectrometer systems. An extensive analysis of the thermal properties of Mid-infrared emitters using polysilicon wires on thin membranes was presented in chapter 6. Their design and fabrication is also presented in chapter 6. IC-compatible thermal mid-infrared sources, operating as incandescent lamps, have been successfully fabricated using poly-silicon heater wires on thin
Mid-Infrared Microspectrometer Systems

silicon nitride bridges. Mid-infrared operation requires an operating temperature of around 700 °C. As in all types of thermal emitters the efficiency is only a few percent but nevertheless the total optical power can be higher than e.g. the output of mid-infrared LED’s fabricated by InAs heterostructures.

The integration of an electrically modulated infrared source to the basic microspectrometer, enables the integration of more signal processing in the design of microspectrometer systems. Signal processing, using chopping and coherent detection were discussed in chapter 9.

For accurate amplification of the low-level signals of thermal sensors low noise is the main design parameter. Low-noise amplifiers, both with- and without chopper switches, and an auxiliary input stage have been fabricated in the AMIS 0.7μ CMOS process and are discussed in chapter 10. The circuits will be applied for the readout of thermopile detectors in a miniaturized infrared spectrometer system.

A new method for the offset reduction of modulated signals will be introduced in this chapter. The method is based on a-priori information about the modulated signal, which is generally a square wave. A high-pass characteristic has been realized, without the use of large coupling capacitors, by offset-sampling and offset-subtraction within the optical chopper loop. Drift and other signals below the sampling frequency will be suppressed also. Fully differential circuit implementations of the presented circuits can result in a further reduction of the offset.

Two low-noise CMOS circuit implementations have been realized, an autozero circuit that removes DC signals at the input of CMOS amplifier and a circuit that uses DC feedback and an auxiliary input for offset compensation. A noise analysis of MOS transistor current sources and folded-cascode CMOS amplifiers was also presented in chapter 10.

Future work

Based on the research in this thesis Non-Dispersive InfraRed (NDIR) spectrometers, as described in chapter 3, are probably the most promising spectrometers for further research in the near future, because of their relatively straightforward construction. The integration of the infrared Fabry-Pérot filters, to filter the spectral components of interest, as discussed in chapter 12, together
Conclusions

with the integrated infrared thermal source of chapter 6 and the integrated thermopile detector arrays described in chapter 8 can result in NDIR spectrometer systems that are relatively simple, complete and low cost.

Research on spectrometers with electrically driven Micro Optical Micro Electro-Mechanical Systems ( MOMEMS ) such as micromirrors or comb drives with gratings is also promising, since these devices can be conveniently used for scanning or alignment in grating-based or Fourier transform based mid-infrared microspectrometers.

The combination of the shaping of silicon by microscale etch techniques (MEMS) and the light guiding properties of silicon in the mid-infrared has not been fully exploited and may result in much more interesting research on mid-infrared optical systems in the future.

Until now very few mid-infrared microspectrometers have been developed into viable commercial products. Therefore research and development on mid-infrared microspectrometers will remain an interesting field for many years in the future. The high number of applications with a large impact on society and the huge progress in micro mechanical systems ( MEMS ) will certainly drive further research on miniaturized spectrometers for many years to come.
Appendix A: Material properties

_Labor is work that leaves no trace behind when it is finished_
Mary McCarthy, writer.

The values in the following tables are indicative. References give more details, e.g. film thickness, temperature, deposition methods and/or doping levels.

**TABLE A1. Seebeck coefficients of some IC-compatible materials at 25 C.**

<table>
<thead>
<tr>
<th>IC-compatible Material</th>
<th>Seebeck coefficient $\alpha$ [ $\mu$V/K]</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>Silicon</td>
<td>440</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Poly N</td>
<td>-68</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>PolySi P</td>
<td>35 ... 22</td>
<td>[AA.9]</td>
</tr>
<tr>
<td>SiDiff P+</td>
<td>430</td>
<td>[AA.3]</td>
</tr>
<tr>
<td>SiDiff N+</td>
<td>-325</td>
<td>[AA.3]</td>
</tr>
<tr>
<td>PolySi N Mosis HP CMOS</td>
<td>-320</td>
<td>[AA.3]</td>
</tr>
<tr>
<td>PolySi N AMS 1.2u CMOS</td>
<td>-120</td>
<td>[AA.3]</td>
</tr>
<tr>
<td>PolySi N</td>
<td>-120</td>
<td>[AA.4]</td>
</tr>
<tr>
<td>PolySi P</td>
<td>170</td>
<td>[AA.4]</td>
</tr>
<tr>
<td>N-Poly SiGe</td>
<td>-120</td>
<td>[AA.37]</td>
</tr>
</tbody>
</table>
The thermal diffusivity $\alpha_c$ of a material is related to the specific heat $c_c$ by:

$$\alpha_c = \frac{\kappa}{\rho c_c} \text{[m}^2\text{s}^{-1}]$$

where:
- $\kappa$ is the thermal conductivity [Wm$^{-1}$K$^{-1}$].
- $c_c$ is the specific heat [J kg$^{-1}$ K$^{-1}$].
- $\rho$ is the density [kgm$^{-3}$].

For a homogeneous uniform beam the thermal conductance $G$ and the thermal capacitance $C$ equal:

$$G = \frac{\kappa}{\rho c_c} \frac{W d}{L} \text{[W/K]}$$

$$C = \rho c_c W L d \text{[K/W]}$$

**TABLE A1. Seebeck coefficients of some IC-compatible materials at 25 C.**

<table>
<thead>
<tr>
<th>IC-compatible Material</th>
<th>Seebeck coefficient $\alpha$ [ $\mu$V/K]</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>P-Poly SiGe</td>
<td>190</td>
<td>[AA.37]</td>
</tr>
<tr>
<td>Aluminium</td>
<td>-1.7</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Pt</td>
<td>-5.1</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Ni</td>
<td>-15</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Ge</td>
<td>300</td>
<td>[AA.1]</td>
</tr>
</tbody>
</table>

**TABLE A2. Seebeck coefficients of some other materials at 25 C.**

<table>
<thead>
<tr>
<th>IC-compatible Material</th>
<th>Seebeck coefficient $\alpha$ [ $\mu$V/K]</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bismuth Telluride (n)</td>
<td>240</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Bismuth Telluride (p)</td>
<td>162</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Au</td>
<td>6.5</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Ag</td>
<td>6.5</td>
<td>[AA.1]</td>
</tr>
<tr>
<td>Material</td>
<td>Density ρ [kg/m³]</td>
<td>Thermal conductivity κ [W/m K]</td>
</tr>
<tr>
<td>------------------</td>
<td>-------------------</td>
<td>--------------------------------</td>
</tr>
<tr>
<td>Bulk Crystal-line Si</td>
<td>2328</td>
<td>141</td>
</tr>
<tr>
<td>SiO</td>
<td></td>
<td>1.4</td>
</tr>
<tr>
<td>SiO₂ bulk</td>
<td>2220</td>
<td>1.38</td>
</tr>
<tr>
<td>PolySi</td>
<td></td>
<td>30-35</td>
</tr>
<tr>
<td>PolySi P</td>
<td></td>
<td>45</td>
</tr>
<tr>
<td>PolySi N</td>
<td></td>
<td>29</td>
</tr>
<tr>
<td>PolySi, P</td>
<td></td>
<td>34</td>
</tr>
<tr>
<td>SiN, 300nm</td>
<td>1.3 800nm</td>
<td>10,3</td>
</tr>
<tr>
<td>SiN, 500nm</td>
<td>1.3 800nm</td>
<td>9,6</td>
</tr>
<tr>
<td>SiN</td>
<td>3100 2400</td>
<td>3.2</td>
</tr>
<tr>
<td>SiO PSG</td>
<td></td>
<td>0.8</td>
</tr>
<tr>
<td>SiO BPSG</td>
<td></td>
<td>1.2</td>
</tr>
<tr>
<td>SiO LPCVD</td>
<td></td>
<td>1.3</td>
</tr>
<tr>
<td>Aluminium film</td>
<td></td>
<td>180</td>
</tr>
<tr>
<td>Aluminium bulk</td>
<td>2702</td>
<td>237</td>
</tr>
<tr>
<td>SiO 300nm</td>
<td></td>
<td>1.4</td>
</tr>
<tr>
<td>SiO 500nm</td>
<td></td>
<td>1.4</td>
</tr>
<tr>
<td>Gold</td>
<td>19300</td>
<td>296</td>
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TABLE A4. Mechanical properties of IC-compatible materials at 25 C.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
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<tr>
<td>Silicon crystalline</td>
<td>232</td>
<td></td>
<td></td>
<td>[AA.5]</td>
</tr>
<tr>
<td>Silicon bulk</td>
<td>130</td>
<td></td>
<td></td>
<td>[AA.5]</td>
</tr>
<tr>
<td>Si-film &lt;111&gt;</td>
<td>190</td>
<td>80</td>
<td></td>
<td>[AA.32]</td>
</tr>
<tr>
<td>Si-film &lt;110&gt;</td>
<td>161</td>
<td>50</td>
<td></td>
<td>[AA.34]</td>
</tr>
<tr>
<td>Si-film &lt;100&gt;</td>
<td>130</td>
<td></td>
<td></td>
<td>[AA.34]</td>
</tr>
<tr>
<td>Polysilicon film</td>
<td>151-171</td>
<td></td>
<td></td>
<td>[AA.36]</td>
</tr>
<tr>
<td>Polysilicon film</td>
<td>140</td>
<td></td>
<td></td>
<td>[AA.18]</td>
</tr>
<tr>
<td>SiO$_2$ Thermal</td>
<td>290</td>
<td></td>
<td></td>
<td>[AA.18]</td>
</tr>
<tr>
<td>SiO$_2$</td>
<td>74</td>
<td></td>
<td></td>
<td>[AA.29]</td>
</tr>
<tr>
<td>SiN$_x$</td>
<td>270</td>
<td></td>
<td></td>
<td>[AA.18]</td>
</tr>
<tr>
<td>SiN$_x$</td>
<td>230 +/- 16</td>
<td></td>
<td></td>
<td>[AA.26]</td>
</tr>
<tr>
<td>SiN$_x$ low-stress</td>
<td>230-265</td>
<td>2.2 -3.34</td>
<td></td>
<td>[AA.36]</td>
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<tr>
<td>SiN$_x$</td>
<td>280-320</td>
<td>11</td>
<td></td>
<td>[AA.30]</td>
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<tr>
<td>SiN$_x$ PECVD</td>
<td>134-142</td>
<td>--</td>
<td>1.80</td>
<td>[AA.28]</td>
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<tr>
<td>SiN$_x$</td>
<td>85-105</td>
<td></td>
<td></td>
<td>[AA.28]</td>
</tr>
<tr>
<td>Aluminium</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Nickel film</td>
<td>200</td>
<td>--</td>
<td></td>
<td>[AA.36]</td>
</tr>
<tr>
<td>TEOS</td>
<td></td>
<td></td>
<td>1.83</td>
<td>[AA.28]</td>
</tr>
<tr>
<td>(B)PSG</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
**TABLE A5. Infrared optical properties of some materials.**

<table>
<thead>
<tr>
<th>Material</th>
<th>Transmittance range [nm]</th>
<th>Refractive index n</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>IC-compatible Silicon</td>
<td>1100-15000</td>
<td>3.42</td>
<td>[AA.39]</td>
</tr>
<tr>
<td>Quartz</td>
<td>240-2500</td>
<td></td>
<td>[AA.46]</td>
</tr>
<tr>
<td>Polysilicon</td>
<td></td>
<td></td>
<td>[AA.44]</td>
</tr>
<tr>
<td>Thermal oxide (HIPOX)</td>
<td></td>
<td>1.44</td>
<td>[AA.41]</td>
</tr>
<tr>
<td>SiN (PECVD)</td>
<td>-4000</td>
<td>1.7-2.0</td>
<td>[AA.45]</td>
</tr>
<tr>
<td>SiO(PECVD)</td>
<td>-4000</td>
<td>1.47-1.68</td>
<td>[AA.45]</td>
</tr>
<tr>
<td>SiO (TEOS)</td>
<td></td>
<td>1.43</td>
<td>[AA.41]</td>
</tr>
<tr>
<td>SiO PSG</td>
<td></td>
<td>1.45</td>
<td>[AA.41]</td>
</tr>
<tr>
<td>SiC</td>
<td>220-4000</td>
<td>2.4</td>
<td>[AA.46]</td>
</tr>
<tr>
<td>Germanium</td>
<td>2-100u</td>
<td>4.0</td>
<td>[AA.46]</td>
</tr>
<tr>
<td>Other materials</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Fused silica</td>
<td>200-2500</td>
<td>1.46</td>
<td>[AA.46]</td>
</tr>
<tr>
<td>ZnSe</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>ZnS</td>
<td></td>
<td></td>
<td></td>
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<tr>
<td>ZrF</td>
<td>600-4760</td>
<td></td>
<td>[AA.46]</td>
</tr>
<tr>
<td>Chalcogenide</td>
<td>2200-11000</td>
<td></td>
<td>[AA.46]</td>
</tr>
<tr>
<td>Sapphire (Al2O3)</td>
<td>150-5000</td>
<td>1.76</td>
<td>[AA.46]</td>
</tr>
</tbody>
</table>

**References**


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Appendix B: Mathematica notebooks

There is a great satisfaction in building good tools for other people to use.
Freeman Dyson

Listings of some Mathematica notebooks have been added in this Appendix to show the reader how some of the equations in chapters 6, 8 and 11 have been evaluated and plotted. Furthermore they demonstrate the power of Mathematica in general and its ability to solve integrals and differential equations analytically in only a few lines. In my opinion these analytical solutions (if possible), illustrated by plots, provide much more insight than pure numerical solutions (e.g. by MATLAB or PSPICE) or finite element calculations.
IR source: Temperature distribution and Thermal time constant

This Mathematica 5.2 notebook solves the static and dynamic heat equation of a thin wire, whose ends are kept at ambient temperature. The static temperature profiles are calculated in case of conduction through the surrounding gas and through the wire only. Next the temperature decay when switched off is calculated and plotted from the dynamic heat equation. The dynamic IR radiation profile when switched off is plotted from the temperature decay.

Author: G. de Graaf, Delft University of Technology
Date: April 2008.

Some physical constants:

Plank’s constant: $h = 6.625 \times 10^{-34}$ [J s] Speed of light: $c = 2.998 \times 10^{14}$ [m/s]
Boltzmann constant: $k_B = 1.38 \times 10^{-23}$ [J/K] Stefan-Boltzmann constant: $\sigma = 5.75 \times 10^{-8}$ [W/m²K⁴]

In[17]:= $h = 6.625 \times 10^{-34}$; $c = 2.998 \times 10^{14}$;
$k_B = 1.38 \times 10^{-23}$; $\sigma = 5.75 \times 10^{-8}$;

Static with heat loss by conduction through surrounding air [*]:

Boundary conditions: the temperature at one end is zero $T_x[0] = 0$, constant heat flow by dissipation $q_e$. and the temperature flow at the other end ($x = L$) is zero because of symmetry. Heat flow through surrounding air by conduction term $\beta$ leads to a hyperbolic function:

In[19]:= $DSolve[$ \n$\{ \begin{array}{l}
T_x''[x] + \beta T_x'[x] = 0,
T_x'[0] = -q_e, T_x[0] = 0, \end{array} \}$, $T_x[x], \{x\}$]

Out[19]= $\{T_x[x] \rightarrow e^{-\frac{x}{\beta}(1. e^{0.014 (x - 0)} - 1. e^{0.014 (-x + 0)})}\}$

Static without heat loss through surrounding air [*]:

Boundary conditions: the temperature at one end is zero $T_x[0] = 0$, constant heat flow by dissipation $q_e$ and the temperature flow at the other end ($x = L$) is zero because of symmetry. leads to a parabolic function:
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\[ T_i(x, t) = \text{DSolve}[ \]
\[ \begin{align*}
    T_x''(x) &= -q_e, \\
    T_x'(L_b) &= 0, \\
    T_x(0) &= T_a
\end{align*} \]
\[ T(x) \]

\[ T_i(x) \text{ is the initial parabolic temperature distribution along the wire.} \]

Some data of the beam: \( L_b \text{ = length in meters, } T_m \text{ is the maximum temperature, } T_a \text{ is the ambient temperature} \)

\[ T_a = 400; q_e = 1000; L_b = 0.004; T_m = 700; \]
\[ \alpha = 0.001; t_e = 0.2 \times 10^{-2}; \]
\[ T_i(z) := \frac{T_m(4(L_b - z^2))}{L_b^2}; \]

\textbf{Transient \([\star]\):}

\( T_i(x) \) Initial distribution of temperature along the wire:

\[ \begin{align*}
    &u = T_i(x, t); \\
    &v6 = \text{NDSolve}[\{D_t u = \alpha D_x u, T[x, 0] := T_i(x), \\
                        T[0, t] := 0, T[L_b, t] := 0, T, \{x, 0, L_b\}, \{t, 0, t_e\}]\]
\]
In[35]:=
T3Dprofile = 
Plot3D[Evaluate[T[x, t] / First[v6]], {t, 0, te}, {x, 0, Lb}, 
PlotPoints → 40, Mesh → True, 
FaceGrids → {{-1, 0, 0}, {0, 1, 0}}, 
ViewPoint → {4.115, -3.347, 3.109}, ColorOutput → GrayLevel, 
TextStyle → {FontFamily → "Arial", FontSize → 10}, ImageSize → 
{300, 280}, PlotLabel → "Beam temperature T-f(x,t) ", 
AxesLabel → {"t\# s", "x", "T\# C"}, 
PlotRange → {{0, te}, {0, Lb}, {0, 700}}];

Beam temperature T-f(x,t)

W = 90 10^{-4}; e_a = 0.75; te = 0.1 10^{-3}; 
Evaluate[{2 e_a W L b o (T[x, t]^4 - T_a^4)} /. First[v6]];

{LightSources → {[[1.2, 1.], RGBColor[1, 0, 0]], 
[[1.0, 1.], RGBColor[0, 1, 0]], 
[[0.8, 1.], RGBColor[0, 0, 1]]}};

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In[32]:=
R3Dprofile = 
    Plot3D[Evaluate[(2 e_π Lb σ (T[x, t]^4 - Ta^4)) / . First[v6]], 
    {t, 0, te}, {x, 0, Lb}, PlotPoints -> 40, 
    Mesh -> True, FaceGrids -> {{-1, 0, 0}, {0, 1, 0}}, 
    ViewPoint -> {3.327, -3.541, 2.841}, ColorOutput -> GrayLevel, 
    TextStyle -> {FontFamily -> "Arial", FontSize -> 12}, ImageSize -> 
    {300, 240}, PlotLabel -> "Beam emitted Power P[t(x,t)"], 
    AxesLabel -> {"t(s)"", "x", "P"}, 
    PlotRange -> {{0, te}, {0, Lb}, {0, 0.008}}];

Beam emitted Power P[t(x,t)
Thin bridge and Cantilever stress

This Mathematica 5.2 notebook solves the differential equations for bending and stress of thin cantilever beams and bridges.

Author: G. de Graaf, Delft University of Technology
Date: april 2008. Only Mathematica 5.2 or higher can solve the set of equations with the given boundary conditions.

This is the general solution of the displacement function of a homogeneous uniform cantilever beam under a uniform load V:

\[
\text{Out}[1]= z = x^4 V L^2 24 M Y
\]

With Ym is Young's modulus and M is the area moment of inertia
Fill in the boundary conditions for the clamped beam at x = 0. z(0) = 0 finds C[1] to be zero and integration of z(x) and z'(0) = 0 finds C[2] = 0:

\[
\text{Out}[2]= z = x^4 V L^2 24 M Y
\]

This is the solution for z=0 clamped other end: no bending forces z''[L] = 0, z'''[L] = 0.

\[
\text{Out}[3]= \text{z[x]} := \frac{x^4 V L^2 24 M Y}{24 M Y}
\]

The maximum deflection at the tip for x=L.
In[5]:= \[Zm\] \[Zc\] \# x's.

L,M i \[tb3\] W ccccccccccccccc 12

Out[5]= \[Zc\] \# \[L]\'

This is the general solution of the displacement function of a homogeneous uniform bridge (dual clamped beam) with no load.

In[6]:= Assuming \[#x\]! 0 & \[L]\! 0 & \[Ym]\! 0 & \[Mi]\! 0, DSolve \[#z'''\]! \#/m 0, \[z], \[x]'';

DualClampedB = \[z] / . %[[1]]

Out[7]= Function\[#x\], C\#1\' C\#2\' x2 C\#3\' x3 C\#4\''

This is the solution of the displacement function of a homogeneous uniform bridge (dual clamped beam) under a point load P in the middle (x = \[L]/2):

In[8]:= Assuming \[x\] > 0 && \[L]\ > 0 && \[Ym]\ > 0 && \[Mi]\ > 0 && \[ol]\ > 0, DSolve\{\[Mi] \[Ym] (\[z'''\]! \[#x\]) = \[ol]\ = 0 , \[z]! 0 , \[z]'! 0 , \[z]'[L] = 0 , \[z]'\}[\[x\], \[z], \[x]\]'];

DualClampedB = \[z] / . %[[1]]

Out[9]= \{\[z] = Function\[#x\], \[L2\] x2 \[Vl\] 2\[L\] x3 \[Vl\] 3\[L\] 4 \[Vl\] 24 \[Mi] \[Ym]\}\}

The maximum deflection at the middle for x=L/2

In[10]:= DualClampedB \[L/2\]

Out[10]= \[L2\] \[ol\] / 384 \[Mi] \[Ym]\]

Give some data and plot the beam bending.
\( M_1 = \frac{\theta b^2 W}{2} \)

ATR calculation:

This Mathematica 5.2 notebook calculates and plots the transmission losses by absorption in water of a beam of infrared light when internally reflected at a Si-H20 interface, using the Fresnel equations.

Author: G. de Graaf, Delft University of Technology
Date: april 2008.

Some physical constants:

* Plank's constant: \( h = 6.625 \times 10^{-34} \text{ [J s]} \)
* Speed of light: \( c = 2.998 \times 10^{14} \text{ [\mu m / s]} \)
* Boltzmann constant: \( k_B = 1.38 \times 10^{-23} \text{ [J / K]} \)
* Stefan-Boltzmann constant: \( \sigma = 5.75 \times 10^{-8} \text{ [W / m}^2\text{K}^4] \)

Transmission loss at the silicon-air interface:

The reflection of light between two homogeneous media with refractive indices 1 and 2 follow from the Fresnel equations:

Reflected power:

\[
R_p(\theta) := \left\{ \frac{\cos(\theta) - \sqrt{(n_2 - 1) x_2}^2 - n_1^2 \cos(\theta)}{n_1 \cos(\theta) + \sqrt{(n_2 - 1) x_2}^2 - n_1^2 \cos(\theta)} \right\}^2;
\]

Absorbed power plot for single reflection as a function of the beam incident angle for unpolarized, p and s-polarization:

\( \theta_c \) is the critical angle. Data for water at 3\,um.
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\[
\begin{align*}
R_{\text{polar}}[\varphi] &= R_p[\varphi] / \left( n_2 = 1.35, n_3 = 0, n_1 = 3.42 \right); \\
\varphi &= \arcsin\left( \frac{n_3}{n_1} \right) / \left( n_1 = 3.436, n_2 = 1.35 \right); \\
R_{\text{loss}}[\varphi] &= R_L[\varphi] / \left( n_2 = 1.35, n_3 = 0.272, n_1 = 3.436 \right); \\
R_{\text{polar}}[\varphi]' &= R_p[\varphi]' / \left( n_2 = 1.35, n_3 = 0.272, n_1 = 3.436 \right); \\
R_{\text{loss}}[\varphi]' &= (R_{\text{polar}}[\varphi] + R_{\text{loss}}[\varphi]) / 2;
\end{align*}
\]

\textbf{Appendix 3.nb}

Alternated power plot for single reflection as a function of the beam incident angle for unpolarized, p and s-polarization. Data for water at 4.5 um.
\[ R_{\text{loss}} = \frac{R_{\text{loss}}}{M} \times \frac{R_{\text{loss}}}{M} \]

\[ M = \arcsin\left( \frac{n_2}{n_1} \right) \]

\[ R_{\text{loss}} = \frac{R_{\text{loss}}}{M} \times \frac{R_{\text{loss}}}{M} \]

\[ R_{\text{loss}} = \frac{R_{\text{loss}}}{M} \times \frac{R_{\text{loss}}}{M} \]

\[ \text{Out[54]} = 22.4116 \]

\[ \text{Out[55]} = \text{ReflectionPower} = \]

\[ \text{ParametricPlot:\{Evaluate[\phi / Degree], Evaluate[R_{\text{loss}}[\phi]], Evaluate[\phi / Degree], Evaluate[R_{\text{loss}}[\phi]], \{\phi, 0, \pi/2\}, PlotStyle\rightarrow\{\text{Thickness}[0.005], \text{GrayLevel}[0.2]\}, Background\rightarrow\text{GrayLevel}[0.98], DefaultFont\rightarrow\text{"Arial", 12}, Frame\rightarrow\text{True}, FrameLabel\rightarrow\text{"Incident angle \phi[Deg]", "Reflected power"}, FrameStyle\rightarrow\text{\{Thickness}[0.005], GridLines\rightarrow\text{Automatic}, PlotStyle\rightarrow\text{PointSize}[0.17], PlotLabel\rightarrow\text{"Total Reflection Silicon – Water\&4.5um n=1.31,k=0.0136"}, PlotLabel\rightarrow\text{FontForm["Title", \{"Font", size\}], PlotRange\rightarrow\text{\{0, 90\}, \{0, 1.0\}\}}] \]
Appendix C: Sample-and-Hold equations.

This thesis is rather long because I did not have the time to make it shorter.

Hold circuit

The Fourier transform of a hold circuit with a hold time $T_s$ equals:

$$ H(f) = \int_{-\infty}^{\infty} h(t) e^{-j2\pi ft} dt = -\frac{1}{j2\pi fT_s} \left[ e^{-j2\pi ft_s} - 1 \right] \left[ e^{-j2\pi ft_s} \right]_{t=0} $$

(16.1)

$$ H(f) = -\frac{1}{j2\pi fT_s} \left[ e^{-j2\pi ft_s} - 1 \right] = \frac{1 - e^{-j2\pi ft_s}}{j2\pi fT_s} $$

(16.2)

$$ H(f) = \frac{1 - e^{-j2\pi ft_s}}{2j\pi fT_s} = e^{-j2\pi ft_s} \left( \frac{e^{j\pi fT_s} - e^{-j\pi fT_s}}{2j} \right) \frac{1}{\pi fT_s} $$

(16.3)

$$ H(f) = e^{-j2\pi ft_s} \frac{\sin(\pi fT_s)}{\pi fT_s} = e^{-j2\pi ft_s} Sinc(\pi fT_s) $$
Sampling

The output signal of an ideal sampler can be written as the multiplication of the input signal with a series of Dirac impulses with a sampling period $T_s$:

$$v_s(t) = v_i(t) \sum_{m=-\infty}^{\infty} \delta(t - mT_s)$$  \hfill (16.4)

The Fourier series of the periodic pulses can be written as:

$$\sum_{m=-\infty}^{\infty} \delta(t - mT_s) = \sum_{k=-\infty}^{\infty} C_k e^{jk2\pi f_s}$$  \hfill (16.5)

The Fourier coefficients $C_k$ are given by:

$$C_k = \frac{1}{T_s} \int_{-\infty}^{\infty} \delta(t - mT_s)e^{-j2\pi f_s t} \, dt = \frac{1}{T_s}$$  \hfill (16.6)

Substitution of (16.6)) in (16.5)) gives:

$$\sum_{m=-\infty}^{\infty} \delta(t - mT_s) = \frac{1}{T_s} \sum_{k=-\infty}^{\infty} e^{jk2\pi f_s}$$  \hfill (16.7)

Using (16.7)) in (16.4)) yields the Fourier transform of $v_i(t)$:

$$v_s(f) = \frac{1}{T_s} \int_{-\infty}^{\infty} v_i(t) \sum_{k=-\infty}^{\infty} e^{-jkf_s} e^{-jft} \, dt$$  \hfill (16.8)

Or:

$$v_s(f) = \frac{1}{T_s} \sum_{k=-\infty}^{\infty} \int_{-\infty}^{\infty} v_i(t)e^{-(j2\pi f + j2\pi kf_s)} \, dt$$  \hfill (16.9)

$$v_s(f) = \frac{1}{T_s} \sum_{k=-\infty}^{\infty} v_i(j2\pi f - j2\pi kf_s) = \frac{1}{T_s} \sum_{k=-\infty}^{\infty} v_i(2\pi j(f - kf_s))$$  \hfill (16.10)
Equation (16.10)) represents a series of replicas of the input spectrum \( v_i(f) \) around multiples of the sampling frequency \( f_s \).

**References**


Appendix D: Low-Noise Amplifier Data.

The cure for boredom is curiosity. There is no cure for curiosity.
Dorothy Parker

Low noise CMOS amplifier

The transistor parameters are given in Table A1.

**TABLE A1. Parameters of the amplifier of figure 10.1**

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>M1-M4</td>
<td>2400</td>
<td>1</td>
<td>220</td>
<td>2000</td>
<td>2.4</td>
<td>417</td>
<td>10</td>
</tr>
<tr>
<td>M2-M5</td>
<td>500</td>
<td>28</td>
<td>530</td>
<td>540</td>
<td>1150</td>
<td>0.87</td>
<td>1</td>
</tr>
<tr>
<td>M3-M6</td>
<td>300</td>
<td>2</td>
<td>320</td>
<td>1600</td>
<td>17</td>
<td>130</td>
<td>5</td>
</tr>
<tr>
<td>M7-M9</td>
<td>300</td>
<td>4</td>
<td>320</td>
<td>3.2</td>
<td>7.5</td>
<td>590</td>
<td>10</td>
</tr>
<tr>
<td>M8-M10</td>
<td>300</td>
<td>2</td>
<td>320</td>
<td>1600</td>
<td>8.5</td>
<td>118</td>
<td>5</td>
</tr>
<tr>
<td>M11</td>
<td>360</td>
<td>2</td>
<td>440</td>
<td>96</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>M12</td>
<td>150</td>
<td>2</td>
<td>440</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Ma1-2</td>
<td>150</td>
<td>2</td>
<td>12</td>
<td>98</td>
<td>1.26</td>
<td>790</td>
<td>16</td>
</tr>
<tr>
<td>Ma3 (cas)</td>
<td>15</td>
<td>2</td>
<td>24</td>
<td>93</td>
<td>0.7</td>
<td>1400</td>
<td>4</td>
</tr>
</tbody>
</table>
Relevant process parameters of the CMOS process are given in Table A2.

<p>| Table A1. Parameters of the amplifier of figure 10.1 |
|----------------------------------|----------------|----------------|----------------|----------------|----------------|</p>
<table>
<thead>
<tr>
<th>Ibias = 50 µA</th>
<th>W [µm]</th>
<th>L [µm]</th>
<th>Id [µA]</th>
<th>gm [µA/V]</th>
<th>gds [µA/V]</th>
<th>rds [kΩ]</th>
<th>gm/Id</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ma4</td>
<td>15</td>
<td>2</td>
<td>24</td>
<td>100</td>
<td>4</td>
<td>250</td>
<td>3.8</td>
</tr>
<tr>
<td>M13</td>
<td>800</td>
<td>2</td>
<td>400</td>
<td>4700</td>
<td>10</td>
<td>100</td>
<td>12</td>
</tr>
</tbody>
</table>

| Table A2. CMOS Typical Design data |
|-----------------------------------|----------------|----------------|----------------|
| Mobility: m [cm²/Vs]              | NMOS | 440 | 150 |
| Threshold voltage Vₜ [V]          | PMOS | 0.77 | 1.01 |
| σ = Vₜ spread [mV]                |     | 3.5 | 1.1 |
| β [µA/V²]                        |     | 96 | 30 |
| Kₑ Flicker noise factor [V²/F]    |     | 3 x 10⁻²⁸ | 5 x 10⁻³⁰ |
| Cₒx [F/m²]                       |     | 20 x 10⁻¹⁶ | 20 x 10⁻¹⁶ |

Large devices
Summary

If it takes a lot of words to say what you have in mind, give it more thought.
Dennis Roth, writer.

This thesis discusses my research on mid-infrared spectrometers. Mid-infrared spectrometers are opto-electronic systems that measure the spectral energy distribution of light in the mid-infrared (1.4 μm to 10 μm) wavelength range. Completely integrated systems, fabricated in IC-compatible technology, are the ultimate goal. As a result the design of the first infrared microspectrometer based on Attenuated Total Reflection (ATR) in bulk silicon is presented in chapter 11.

Applications

The main applications of mid-infrared spectrometers are in emission and absorption spectroscopy. These techniques and their applications are described in chapter 3. Absorption or vibrational spectroscopy is an important technique that is used in many scientific, industrial and biomedical disciplines for measuring material compositions or for inspecting of physical processes. Nowadays, more widespread use of these instruments is mainly limited by the price and complexity of the available instruments. The competition of small portable microspectrometers, designed for specific applications is challenged by the large bulky size and the expense of the laboratory instruments.

Technology

Many new miniaturised optical and electro-optical components have been fabricated in the last decade as a result of the huge progress achieved in the development of technology for integrated-circuits (IC’s) and Micro Electro-
Mid-Infrared Microspectrometer Systems

Mechanical Systems (MEMS) technology. Micro Electro-Mechanical Systems (MEMS) are fabricated by etching and deposition techniques on a silicon substrate. Similar techniques are used for the fabrication of integrated circuits (IC’s). A short overview of the current research on spectrometers, especially infrared microspectrometers, based on IC and MEMS technology is given in chapter 4.

Spectrometer components

Spectrometers are comprised of several basic components. The main components are the optical filter and the infrared detector(s). Three basic principles for spectral filtering in the optical domain are discussed in chapter 2. An overview of infrared detectors and their properties is given in chapter 7. The low energy level of mid-infrared radiation results in low signal levels from the detectors. Some implementations of low-noise amplifier circuits in CMOS technology are given in chapter 10. Arrays of infrared detectors based on thermopiles are described in chapter 8 along with the geometrical optimization of the sensitivity and the signal-to-noise ratio of thermopiles, which is the main topic of chapter 8.

The spectrometer and its parts as described in this thesis, have been fabricated in an in-house technology developed by the DIMES institute of the Delft University of Technology. The fabrication is described in chapter 5 and the fabrication steps of a complete spectrometer system are discussed in more detail in chapter 11. New IC-compatible mid-infrared optical filters based on thin-film multi-layers have been developed and are discussed in chapter 12. In most mid-infrared spectroscopy applications a well-defined infrared source is also needed. An infrared light source should therefore be considered an integral part of most spectrometer systems. The design and fabrication of thermal infrared sources based on MEMS techniques is therefore described in chapter 6.

Generally, it can be stated that scaling down a spectrometer requires down scaling of the complete system: the light input coupling, the optical path and the detector and in many applications also the sample, the sample holder and the light source.

Microspectrometer systems

The integration of an electrically modulated infrared source to the basic
microspectrometer, enables the integration of more signal processing in the
design of microspectrometer systems. Signal processing, using chopping and
cohherent detection is discussed in chapter 9. A new method for offset reduction of
modulated signals is also introduced in this chapter.

**New concepts, original contributions and their results**

The main achievements reported in this thesis can be found in:

- The design and fabrication of a new mid-infrared microspectrometer based
  on Attenuated Total Reflection (ATR) presented in chapter 11:
  The presented concept allows the integration of all critical components of
  the spectrometer: a thermal infrared source, optical filtering, the Internal
  Reflection Element (IRE) and the detector array. The full potential of the
  complete ATR spectrometer has not yet been exploited, because the final
  processing steps have to be further developed. The fabrication and testing
  of the individual components however has shown promising results. For a
  complete microspectrometer system the technological problem of the large
  mechanical stress in detector arrays needs to be solved. After this, a
  number of extra steps, such as the integration of mid-infrared optical
  filters, alignment, assembly (wafer bonding) and the vacuum packaging,
  have to be further evaluated.

- The new offset and drift reduction technique for thermal sensor readout as
  presented in chapter 9.
  A sampling technique on modulated signals has been presented and
  implemented in CMOS circuits and has shown good results, which can be
  further adapted to the requirements of a specific application. Fully
  differential circuit implementations of the presented circuits can result in a
  further reduction of the offset.

- The design and optimization of the sensitivity and the signal-to-noise ratio
  of infrared detectors based on thermopiles, as presented in chapter 8.
  In chapter 8 guidelines are given for optimizing for the design of thermal
  infrared detectors based on thermopiles on thin film cantilevers and
  bridges. The influence of: the size of the absorber area, the length and
  width of the thermopile legs, air conduction, planar and flat structures etc.
  on the performance of the sensor is discussed here. Arrays of thermopile
  detectors have been fabricated. However the yield of several prototyping
  runs of the detector arrays was low due to excessive stress in the
cantilevers and bridges. The improvements in quality and reproducibility of the recent devices indicate better perspectives using this technology for high sensitivity thermopile arrays in the future.

- The design and fabrication of narrow band multilayer Fabry-Pérot filters using IC-compatible materials and technology as described in chapter 12. Multi-layer optical filters for the mid-infrared wavelength region have been successfully fabricated using thin polysilicon and siliconoxide layers. Since the processing is IC-compatible this technique can be used to produce optical filters for application in integrated spectrometers.
- An extensive analysis of thermal mid-infrared sources using polysilicon heater wires on thin membranes is presented in chapter 6.
- The design and fabrication of thermal infrared sources based on thin polysilicon wires, is also presented in chapter 6.
- IC-compatible thermal mid-infrared sources have been successfully fabricated using poly-silicon heater wires on thin silicon nitride bridges, operating as incandescent lamps.

Finally, some interesting circuit analyses and overviews that have resulted from the work in this thesis:
- The systematic noise analysis of folded-cascode CMOS amplifiers as described section 10.2.3 of chapter 10.
- The noise analysis of MOS transistor current sources as presented in section 10.2.3.1 of chapter 10.
- An overview of the state-of-art in infrared microspectrometers, which is given in chapter 4.
- An overview of uncooled infrared detectors and their properties, which is included in chapter 7.
- A comparison between the different principles of spectrometers and the effects of down scaling on the performance of spectrometers, which is included in chapter 2 of this thesis.

Outlook

Based on the research in this thesis Non-Dispersive InfraRed (NDIR) spectrometers, as described in chapter 3, are probably the most promising
Mid-Infrared Microspectrometer Systems

spectrometers for further research in the near future, because of their relatively straightforward construction. The integration of the infrared Fabry-Pérot filters, to filter the spectral components of interest, as discussed in chapter 12, together with the integrated infrared thermal source of chapter 6 and the integrated thermopile detector arrays described in chapter 8 can result in relatively simple, low cost, complete NDIR spectrometer systems.

Research on spectrometers with electrically driven Micro Optical Micro Electro-Mechanical Systems (MOMEMS) such as micromirrors or comb drives with gratings is also promising, since these devices can be conveniently used for either controlled scanning or alignment in the optical path of grating-based microspectrometers or for controlled displacement in interference based spectrometers, such as Fourier transform infrared microspectrometers.

Nowadays a great deal of effort is being devoted to the development and production of new microscale optical components using MEMS technology. Nevertheless, the combination the shaping of silicon by microscale etch techniques and the light guiding properties of silicon in the mid-infrared has not been fully exploited.

The extensive field of opto-electronics is also expanding to the mid-infrared. At present many new miniature components, such as new types of infrared detectors and powerful mid-infrared sources, as well as quantum cascade lasers, are being developed.

Very few mid-infrared microspectrometers have reached the stage of a viable commercial product until now. Therefore research and development on mid-infrared microspectrometers will remain a pre-requisite for many years in the future. The great progress and the large number of applications, with a significant impact on society, of the field of micro electro-mechanical systems (MEMS) will certainly drive further research on miniaturized spectrometers for many years to come.
Samenvatting

I try to leave out the parts that people skip.

Elmore Leonard, writer.

Dit proefschrift met de nederlandse titel "Mid-infrorood Microspectrometer Systemen" beschrijft mijn onderzoek aan infrarood spectrometers. Infrarood microspectrometers zijn zeer kleine opto-electronische systemen voor het meten van de spectrale energie verdeling van infrarood licht in het golflengtegebied van ongeveer 1.4μm tot 10μm. Geheel geïntegreerde, in IC technieken vervaardigde, systemen zijn het uiteindelijke doel van het onderzoek. Het belangrijkste resultaat van deze doelstelling is het ontwerp en de fabricage van de eerste miniatuur infrarood spectrometer op basis van Attenuated Total Reflection (ATR), zoals beschreven in hoofdstuk 11.

Toepassingen

De belangrijkste toepassingen van infrarood spectrometers zijn in de emissie- en absorptie-spectroscopie en deze technieken worden beschreven in hoofdstuk 3. Absorptie- of vibratie-spectroscopie is een belangrijke techniek voor wetenschappelijk onderzoek en in de industrie, toegepast voor materiaal onderzoek of voor het waarnemen van fysische processen. De prijs en de complexiteit van de huidige systemen voor spectrometrie beperken de mogelijkheden voor nieuw en meer toepassingen voor deze instrumenten. De concurrentie van deze kleine draagbare microspectrometers voor specifieke toepassingen met de universele, grote en dure laboratoriumapparatuur is een grote uitdaging.

Fabricage technieken
Als gevolg van de enorme ontwikkelingen op het gebied van de fabricage technieken voor geïntegreerde schakelingen (IC’s) en micromechanische fabricage technieken (MEMS) zijn in de laatste twee decennia zeer veel nieuwe optische en opto-electronische componenten op microschaal vervaardigd. Micro ElectroMechanische Systemen (MEMS) worden vervaardigd door depositie- en etstechnieken met silicium als dragermateriaal, technieken die ook gebruikt worden voor de vervaardiging van geïntegreerde circuits (IC’s). Een kort overzicht van de resultaten van het huidige onderzoek op het gebied van spectrometers, en in het bijzonder infrarood spectrometers, op basis van IC- en MEMS-technieken is gegeven in hoofdstuk 4.

**Basiscomponenten voor spectrometers**


**Microspectrometer systemen**

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De integratie van een, electrisch moduleerbare, infraroodbron in een spectrometer biedt nieuwe mogelijkheden voor electronische signaalverwerking in deze systemen. Signaalverwerking door middel van modulatie en coherente detectie wordt behandeld in hoofdstuk 9. Een nieuwe methode voor offset reductie van gemoduleerde signalen wordt eveneens geïntroduceerd in dit hoofdstuk.

**Nieuwe concepten, originele bijdragen en de resultaten daarvan**

De belangrijkste resultaten van het werk beschreven in dit proefschrift zijn:

- Het ontwerp en de fabricage van een nieuwe type infrarood spectrometer op basis van Attenuated Total Reflection (ATR), zoals beschreven in hoofdstuk 11.

Het gegeven ontwerp maakt integratie van alle belangrijke componenten van de spectrometer mogelijk: de infrarood bron, optische filtering, het Interne Reflectie Element (IRE) en de detectoren. Het samenvoegen tot een compleet spectrometer systeem is nog niet haalbaar gebleken. Goede resultaten zijn echter behaald bij de fabricage en tests van de individuele onderdelen. Voor de vervaardiging van een complete microspectrometer dient het technologische probleem van de hoge mechanische spanning in de detector arrays te worden opgelost. Een aantal extra stappen zoals de integratie van de optische filters, het samenvoegen van de beide silicium chips en de plaatsing in een vacuum behuizing dient dan ook nog te worden uitgevoerd.

- Een nieuwe techniek voor de reductie van drift en offset bij de uitlezing van thermische detectoren.

Synchrone bemonstering van gemoduleerde signalen, zoals geïntroduceerd in hoofdstuk 9 is met succes geïmplementeerd in CMOS circuits voor de reductie van drift en offset. Verbeterde prestaties zijn mogelijk indien de schakelingen worden ontworpen voor een specifieke sensor of een specifieke toepassing. Ook door de circuits geheel met verschillende versterkers en schakelaars uit te voeren kan de offset reductie verder verbeterd worden.

- Het ontwerp en de optimalisatie van de gevoeligheid en signaal-ruisverhouding van infrarood detectoren op basis van thermozuilen, gemaakt op dunne-film strips.
In hoofdstuk 8 worden richtlijnen voor het optimaliseren van de geometrie van thermische detectoren, gemaakt op dunne-film strips, gegeven. De invloed van de grootte van het absorberende oppervlak, de lengte en de breedte van de geleiders van de thermozuil, de thermisch geleiding van het omringende gas, vlakke of gestapelde thermozuilen enz. worden hier besproken. De opbrengst van de diverse series prototypen infrarood detectoren was zeer laag door zeer hoge mechanische spanningen in de membraanen. De verbetering van de kwaliteit de recent gefabriceerde series geven aan dat verdere verbetering in de toekomst mogelijk zal zijn.

- Het ontwerp en de fabricage van smalbandige Fabry-Pérot mid-infraroodfilters vervaardigd in IC compatibele technieken.
- Mid-infrarood optische filters, zoals beschreven in hoofdstuk 12, vervaardigd uit dunne lagen polysilicium en siliciumoxide op een silicium substraat, zijn met succes gerealiseerd. Daar deze filters met IC-compatible materialen in een IC compatibele techniek worden vervaardigd zijn zij zeer geschikt voor toepassing in microspectrometers.
- Een uitgebreide analyse en het ontwerp en de fabricage thermische infraroodbronnen op microschaal in IC compatibele techniek is gegeven in hoofdstuk 6. Mid-infrarood bronnen op basis van polysilicium weerstanden op dunne strips van siliciumnitride zijn met succes vervaardigd. Door hun kleine afmetingen en lage thermische massa zijn deze bronnen zeer geschikt voor geschakelde toepassingen.

Tenslotte heeft dit onderzoek geleid tot een aantal interessante analyses van electronische circuits en aantal overzichten op het gebied van spectrometers:

- Een systematische aanpak van de ruisanalyse van “folded-cascode” versterkers in paragraaf 10.2.3 van hoofdstuk 10.
- De ruisanalyse van CMOS stroombronnen in paragraaf 10.2.3.1 van hoofdstuk 10.
- Een overzicht van de huidige ontwikkelingen op het gebied van microspectrometers in hoofdstuk 4.
- Een overzicht van diverse soorten infrarooddetectoren en hun eigenschappen in hoofdstuk 7.
- Een vergelijking tussen de verschillende basisprincipes van spectrometers en hun prestaties, en de effecten daarvan bij het verkleinen van spectrometers naar microspectrometers.
Toekomstverwachtingen

Tot op heden zijn slechts enkele van de gepubliceerde microspectrometers ontwikkeld tot commerciële producten. Vele nieuwe gemeniaturiseerde componenten voor infrarood toepassingen, zoals nieuwe of goedkopere detectoren en krachtige mid-infrarood bronnen, zoals quantum-cascade lasers, zijn op dit moment in ontwikkeling. Mede dankzij de recente MEMS technologie komen vele nieuwe, zeer kleine, optische en opto-electrische componenten beschikbaar. Echter de combinatie van goede optische eigenschappen van silicium in het mid-infrarood en de vele mogelijkheden voor het maken van allerlei structuren met behulp van MEMS technieken zijn zeker nog niet ten volle benut.

Non-Dispersive Infrared (NDIR) spectrometers (hoofdstuk 3) zijn, gezien hun relatief eenvoudige opbouw en de resultaten van dit onderzoek, het meest geschikt voor verder onderzoek aan infrarood spectrometers in de nabije toekomst. Het samenvoegen van de Fabry-Pérot midinfrarood filters, zoals beschreven in hoofdstuk 12 met de thermozuil detectoren uit hoofdstuk 8 en de infraroodbron uit hoofdstuk 6 kan resulteren in een relatief eenvoudig, goedkoop en compleet mid-infrarood spectrometer systeem, zoals aangegeven in hoofdstuk 12 paragraaf 9.

De toepassing van Micro Optische MEchanische Micro Systemen (MOMEMS), zoals comb-drives en microspiegels bieden ook vele nieuwe mogelijkheden voor het maken van optische filters. Deze componenten kunnen met electrische besturing licht in het optisch pad gecontroleerd afbuigen, bijvoorbeeld door een tralie op deze component aan te brengen, of door middel door middel van zeer kleine verplaatsingen gecontroleerde interferentie van licht veroorzaken, zoals bij spectrometers op basis van Fourier Transformatie het geval is.

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Albert Schweitzer

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About the author

Happiness is a by-product, you cannot pursue it by itself.
Sam Levenson

Ger de Graaf was born in Delft, The Netherlands, on August 1, 1955. He has been a staff member at the Electronic Instrumentation Laboratory of the Faculty of Electrical Engineering at Delft University of Technology since 1976. He received his BSEE degree in electrical and control engineering from the Technische Hogeschool in Rotterdam in 1983. His graduation project was on the active feedback of piezo-electric accelerometers. From 1992 to 1996 he has had a part-time consultancy company specializing in computer controlled measurement systems, working on projects for the energy measurement of heating systems in apartment buildings. He is involved as a teacher in courses on instrumentation, measurement science and microsystems. He also has worked on measurement systems for remote experiments, via the Internet, supporting these courses. His main research interest are in analog and mixed-signal circuit design, opto-electronics, MEMS technology and sensors in general.
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There isn't any place to publish, in a dignified manner, what you actually did in order to get to do the work.
Richard Feynman

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