Towards a thermal noise limited μ -Coriolis mass flow sensor

Thomas Schut

**TOWARDS A THERMAL NOISE
LIMITED MICRO-CORIOLIS
MASS FLOW SENSOR**

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This work is part of the research program FLOW+ under project number 15019, which is co-funded by the Netherlands Organization for Scientific Research (NWO), Bronkhorst High-Tech and Krohne.

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TOWARDS A THERMAL NOISE LIMITED MICRO-CORIOLIS MASS FLOW SENSOR

Dissertation

to obtain
the degree of doctor at the University of Twente,
on the authority of the rector magnificus,
prof. dr. ir. A. Veldkamp,
on account of the decision of the Doctorate Board,
to be publicly defended
on Friday, 1 October 2021 at 12:45

by

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Abstract

Accurate measurement and control of fluid flow is critical in many applications, such as gas chromatography, wind sensing and biomedical applications. Therefore, flow sensors are often the crucial component in fluidic systems. Over the last decades, the research field of micro-machined flow sensors has been ever growing due to the advantages that come with miniaturization of such sensors. The main advantages of micro-machined flow sensors are their low power consumption and their ability to measure very low flow rates. Thermal flow sensors are widely used for their high resolution. However, thermal flow sensors have to be calibrated for a specific fluid, since each fluid has a different heat capacity. On the other hand, Coriolis mass flow sensors measure true mass flow, independent of fluid properties such as density and viscosity. This has made them of great importance in the field of mass flow measurement. The first micro-machined Coriolis mass flow sensors were presented almost two decades ago. Since then, much progress has been made to improve their performance. The aim of the research in this thesis is to push the resolution of the μ -Coriolis mass flow sensor (μ -CMFS) towards the limit associated with mechanical-thermal noise. The research in this thesis focuses on limitations to the performance of a μ -CMFS related to the mechanical design and how to address these. The research can be divided into two parts: 1) Definition of the limitations to the performance of a μ -CMFS and modelling of said sensor, 2) Presentation of methods and designs which address the aforementioned limitations. A μ -CMFS can in principle be seen as an inertial sensor, where a fluid flow through a micro-channel induces acceleration of the channel. Therefore performance limitations of micro-machined inertial sensors are reviewed, including (but not limited to) mechanical-thermal noise. For each limitation, the application to a μ -CMFS is described. To further the understanding of limitations to the performance of a μ -CMFS, the operating principle is discussed in detail. Furthermore, a model of the sensor was set up consisting of a finite element method (FEM) simulation in combination with an analytical model. The model allows for fast simulation of resonance frequencies, flow sensitivity and mechanical properties. The model is validated by designing several μ -CMFS devices with varying size and shape and comparing simulated and experimentally obtained parameters. After a model of the sensor is set up, limitations to actuation and readout methods are investigated.

Several transduction principles are discussed which can be exploited for actuation of the μ -CMFS. Improved magnet configurations for electromagnetic actuation of the μ -CMFS are presented. Their improved performance can enhance the signal-to-noise ratio of the overall sensor and reduce the amount of power required for actuation. Several readout schemes for a μ -CMFS are discussed to address the performance limitations presented throughout this thesis. Devices are realized based on two main readout methods: 1) A capacitive readout realized by comb-shaped electrodes attached to the micro-channel of the sensor, 2) A resistive readout based on strain gauges on top of or embedded in the micro-channel. The devices are fabricated using the so-called Surface Channel Technology fabrication process, which is described in detail in this thesis. This process allows for micro-channels to be created right underneath the surface of a silicon wafer. The fabricated devices have been characterized and several improvements have been achieved. μ -CMFS devices with a capacitive readout were realized which showed significant improved robustness against vibrations caused by environmental excitation or by water hammering (by which gas bubbles present in a liquid flow can cause in-plane forces on the suspended micro-channels). Furthermore, μ -CMFS devices with higher resonance frequencies and quality factors were realized. These devices should be more robust against external and in-plane vibrations while also increasing flow sensitivity. The latter was unfortunately not significantly increased because a secondary Coriolis vibration mode limited the displacement induced by fluid flow. In addition to devices with a capacitive readout, proof-of-principle μ -CMFS devices with a (piezo-)resistive readout were realized which can overcome limitations associated with a capacitive readout. First results are promising, although their resolution is still inferior to that of the capacitive readout. Finally, recommendations are given based on the characterized devices to push the resolution of the μ -Coriolis mass flow sensor further towards the thermal noise limit.

Contents

Abstract	I
Contents	III
1 Introduction	1
1.1 Background and motivation	2
1.2 Aim of the research	3
1.3 Dissertation outline	4
References	6
2 Performance limitations of Micro-machined Inertial Sensors	9
2.1 Introduction	10
2.2 Electrical noise	11
2.2.1 Electrical-Thermal noise	11
2.2.2 Flicker noise	12
2.3 Mechanical-Thermal Noise	14
2.4 Other limiting factors	18
2.4.1 Vibrations from external disturbances	18
2.4.2 Mode coupling	22
2.4.3 Transduction principle	23
2.5 Conclusion	24
References	25
3 Modelling of micro-Coriolis mass flow sensors	29
3.1 Introduction	30
3.2 Analytical model of a μ -Coriolis flow sensor	31
3.3 FEM model	35
3.3.1 Influence of channel loop size	36
3.3.2 Integration of spring structures	36
3.4 Mechanical-Thermal noise	38
3.5 Conclusion	39
References	40

4	Fabrication Methods	41
4.1	Introduction	42
4.2	Surface Channel Technology	43
4.2.1	Forming micro-channels	43
4.2.2	Fluidic inlets	43
4.2.3	Forming the micro-channel wall	44
4.2.4	Creating electrodes	44
4.2.5	Releasing the micro-channels from the Si bulk	47
4.3	Integration of Doped Silicon Structures	50
4.3.1	Implementing doped Si strips in conventional SCT	50
4.3.2	Adapted SCT: Doped poly-silicon structures on top of the micro-channel	52
4.3.3	Adapted SCT: Mono-crystalline silicon structures using silicon-on-insulator wafer with a doped device layer	54
4.4	Conclusion	57
	References	58
5	Actuation Methods	59
5.1	Introduction	60
5.2	Transduction principles	60
5.2.1	Electrostatic	60
5.2.2	Piezoelectric	62
5.2.3	Electrothermal	62
5.2.4	Electromagnetic	63
5.3	Lorentz actuation of a μ -Coriolis flow sensor	64
5.4	Improved magnet configurations	68
5.4.1	Configurations	68
5.4.2	Device and simulation	68
5.4.3	Results and discussion	70
5.5	Conclusion	74
	References	75
6	Readout Methods	77
6.1	Introduction	78
6.2	Transduction principles	79
6.2.1	Capacitive	79
6.2.2	(Piezo-)resistive	80
6.2.3	Piezoelectric	80
6.2.4	Optical	81
6.3	Capacitive readout of a μ -Coriolis flow sensor	82
6.3.1	Operating principle	82

6.3.2	FEM analysis	83
6.3.3	Modelling effects of non-linearity on the output signal	85
6.3.4	The effect of external disturbances	90
6.3.5	Differential Readout	90
6.3.6	Water hammer effect	94
6.4	Resistive readout	95
6.4.1	Motivation	95
6.4.2	Operating principle	96
6.5	Conclusion	98
	References	99
7	Design & Characterization	101
7.1	Introduction	102
7.2	Characterization methods	104
7.2.1	Resonance mode analysis by Laser Doppler Vibrometry	104
7.2.2	In-plane motion analysis by Stroboscopic Video Microscopy	106
7.2.3	Readout electronics	106
7.2.4	Flow measurement	108
7.3	Validation of the flow sensitivity model	110
7.3.1	Devices & Setup	110
7.3.2	Results & Discussion	112
7.3.3	Conclusion	114
7.4	μ -Coriolis flow sensor with differential readout	114
7.4.1	Design	115
7.4.2	Device & Setup	116
7.4.3	Results and discussion	118
7.4.4	Conclusion	122
7.5	Suspended capacitive readout	123
7.5.1	Design	123
7.5.2	Devices & Setup	126
7.5.3	Results & Discussion	128
7.5.4	Conclusion	133
7.6	μ -Coriolis flow sensors with increased stiffness	134
7.6.1	Motivation	134
7.6.2	Design	136
7.6.3	Devices & Setup	138
7.6.4	Results & Discussion	140
7.6.5	Conclusion	142
7.7	Resistive readout using metal strain gauges	144
7.7.1	Design	144

7.7.2	Devices & Setup	147
7.7.3	Results	149
7.7.4	Discussion & Conclusion	151
7.8	Piezo-resistive readout	152
7.8.1	Design	152
7.8.2	Device & Setup	153
7.8.3	Results	154
7.8.4	Discussion	158
7.8.5	Conclusion	158
7.9	Conclusion	159
	References	161
8	Conclusions and Recommendations	163
8.1	Conclusions	164
8.1.1	Definition of performance limitations and modelling of a μ -Coriolis mass flow sensor	164
8.1.2	Fabrication technology	164
8.1.3	Actuation and readout methods	164
8.1.4	Achieved improvements in performance of the μ -CMFS	165
8.2	Towards the thermal noise limit	166
8.3	Further recommendations	170
	References	171
A	Fabrication process flow	173
A.1	Conventional surface channel technology (SCT)	174
A.2	Adapted SCT: Embedded doped mono-crystalline silicon structures (based on silicon-on-insulator)	189
	Samenvatting	197
	Publications	199
	Acknowledgements	203
	About the author	205

1

Introduction

This chapter introduces the research of this thesis. The background and goals of the research are discussed after which an outline of the dissertation is given.

1

1.1 Background and motivation

Accurate measurement and control of fluid flow is critical in many applications, such as gas chromatography [1], wind sensing [2] and biomedical applications [3]. Therefore, flow sensors are often the crucial component in fluidic systems. Over the last decades, the research field of micro-machined flow sensors has been ever growing due to the advantages that come with miniaturization of such sensors. The main advantages of micro-machined flow sensors are their low power consumption and their ability to measure very low flow rates [4]. Thermal flow sensors are widely used for their high resolution [5, 6]. Their operation relies on the fact that when a fluid flows over a heater, a transfer of heat occurs, which can be translated to an electrical response. However, thermal flow sensors have to be calibrated for a specific fluid, since each fluid has a different heat capacity. On the other hand, Coriolis mass flow sensors measure true mass flow, independent of fluid properties such as density and viscosity. This has made them of great importance in the field of mass flow measurement [7, 8].

The operating principle of a Coriolis mass flow sensor is based on the generation of Coriolis forces on a flowing fluid. Figure 1.1 shows the basic structure of such a sensor and its operating principle. The structure consists of a suspended micro-channel following a rectangular loop. The channel is fixed at two points, such that it can vibrate in various modes. The channel of the sensor is brought into resonance in the actuation mode (Fig. 1.1a). When fluid flows through the channel with a mass flow rate ϕ_m , Coriolis forces are induced in the rotating channel section, see Fig. 1.1b. This causes additional vibration in the Coriolis/detection mode. The ratio between the detection and actuation vibration amplitudes is a measure for the mass flow rate.

Figure 1.1: Operation principle of a μ -Coriolis flow sensor. The ratio between the two modal vibration amplitudes is a measure for the flow rate.

Table 1.1: Specifications of the current state-of-the-art commercially available μ -CMFS (demonstrator, available upon request) [14] and target specifications.

Specification	Value	Target	Factor
Flow range	$10 \text{ mg h}^{-1} - 2 \text{ g h}^{-1}$	$10 \text{ } \mu\text{g h}^{-1} - 2 \text{ g h}^{-1}$	–
Zero-flow stability	$<\pm 2 \text{ mg h}^{-1}$	$<\pm 2 \text{ } \mu\text{g h}^{-1}$	1000

Haneveld et al. presented a highly sensitive μ -Coriolis mass flow sensor (μ -CMFS[†]) [9], making the first step towards reaching the same resolution as thermal flow sensors. Improvements have been made upon this sensor, which resulted in a more sensitive μ -CMFS presented by Groenesteijn et al. [10]. It had a zero-flow stability 40 times better than state of the art at that time. Alveringh et al. [11] showed that the resolution of the sensor in [10] can still be improved by a factor of 50 before reaching the fundamental limit associated with random thermal motion of the vibrating channel.

1

1.2 Aim of the research

At the Integrated Devices and Systems (IDS) group[‡] at the University of Twente, research has been done on μ -Coriolis mass flow sensors for over a decade. The research in this thesis builds upon that of Groenesteijn [12] and Alveringh [13]. Groenesteijn [12] (further) developed a microfluidic platform to realize micro-fluidic handling systems containing multiple sensors (μ -CMFS among others) and actuators. Furthermore, the μ -CMFS device presented by Haneveld et al. [9] was investigated and improved upon with respect to flow sensitivity and actuation/readout techniques. The work of Alveringh [13] included the derivation of the mechanical-thermal noise limit of the μ -CMFS presented by Groenesteijn et al. [10].

In this thesis, limitations to the performance of the μ -CMFS are analysed. The aim is to bring the resolution of the μ -CMFS towards the mechanical-thermal noise limit and address other limitations to its performance by investigating all these limits and designing novel devices to address these. Table 1.1 shows the specifications of the current[§] state-of-the-art commercially available μ -CMFS (demonstrator, available upon request) [14] and target specifications. The goal is to push the zero-flow stability of the μ -CMFS towards the thermal noise limit ($\sim 2 \text{ } \mu\text{g h}^{-1}$, see Tab. 1.1). Table 1.2 shows an overview of factors which limit the performance of the μ -CMFS. Each of the described limitations will be discussed and addressed in this thesis.

[†] **μ -Coriolis mass flow sensor** is abbreviated as **μ -CMFS** throughout this thesis.

[‡]The part of the group that is/was working on μ -Coriolis mass flow sensors was formerly called Micro Sensors and Systems (MSS) and before that Transducer Science and Technology (TST).

[§]Where *current* implies the state-of-the-art at the start of the research presented in this thesis.

Table 1.2: Overview of factors which limit the performance of the μ -CMFS.

Limiting factor	Effect
Flow sensitivity	Higher flow sensitivity can improve resolution
Channel geometry	Affects flow sensitivity and quality factor of the resonating channel
External vibrations	Can cause (vibration dependent) offset in the measured flow rate
Resonance mode	Mode shape and resonance frequency influence robustness against external vibrations and affect flow sensitivity
Noise in actuation/readout circuitry	Results in lower resolution through decreased signal-to-noise ratio
Actuation method	Improved actuation can lower power dissipation and improve signal-to-noise ratio in electronic circuitry for actuation
Non-linearity in readout	Causes/increases dependence on actuation amplitude and pressure

In addition to the work in this thesis, work is done at the Electronic Instrumentation Lab (EIL), at Delft University of Technology (TU Delft), by A.C. de Oliveira, to improve the actuation and readout electronic circuitry surrounding the μ -CMFS [15, 16]. The focus of the research presented in this thesis lies on tackling the performance limitations associated with the mechanical design of the μ -Coriolis mass flow sensor. The research outcome of this thesis and that of A.C. Oliveira will then be further built upon in another project to realize a demonstrator device with improved resolution through the combination of the μ -Coriolis mass flow sensor and a dedicated CMOS interface.

1.3 Dissertation outline

Chapter 2 gives an overview of the performance limitations associated with micro-machined inertial sensors in general. Sources of noise for these sensors are discussed as well as other factors which could limit their performance. For each limiting factor, the effect on (or application to) the μ -Coriolis mass flow sensor is briefly discussed.

A model of the μ -Coriolis mass flow sensor is set up and presented in Chapter 3, consisting of a combination of an analytical model and a finite-element-method (FEM) model. The model is able to swiftly determine resonance frequencies and flow sensitivity of μ -CMFS designs. Furthermore, the noise equivalent flow for a μ -CMFS related to mechanical-thermal noise is derived in this chapter to give an insight into the possible improvement.

Chapter 4 presents the fabrication technology of the devices presented in this thesis.

The technology builds further upon the microfluidic platform presented by Groenesteijn et al. [12]. Adapted processes are presented to allow for the integration of doped silicon structures in the micro-channels. This can be used for multiple applications, among which a piezo-resistive readout for a μ -CMFS.

Various transduction principles which can be utilized for actuation of a μ -CMFS are discussed in Chapter 5. Electromagnetic actuation by Lorentz forces is described in more detail and improved magnet configurations (with respect to those in [12]) are presented.

In Chapter 6, various transduction principles for the detection of the vibrations in a μ -CMFS are described. The limitations of a capacitive readout are investigated and a differential capacitive readout is proposed. Furthermore, a resistive readout is presented as a potential alternative to a capacitive readout.

Devices based on the model from Chapter 3 and the fabrication processes in Chapter 4 are presented in Chapter 7. These are designed to address the limitations discussed in Chapters 2 and 6. The most promising devices are highlighted and recommendations for moving further towards the thermal noise limit are given in Chapter 8.

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2

Performance limitations of Micro-machined Inertial Sensors

In this chapter, noise sources and other influences that may limit the performance of micro-machined inertial sensors are discussed. First, an overview is given of the main sources of noise in micro-machined inertial sensors. After which other limiting factors are discussed such as mode coupling, external disturbance vibrations and utilized transduction principle. The influence of each effect on the performance of a micro-Coriolis mass flow sensor (μ -CMFS) is briefly discussed.

2

2.1 Introduction

With the miniaturization of electronic devices and other equipment, micro-machined or MEMS (Micro Electro-Mechanical System) inertial sensors have become increasingly important [1]. The most well known of these device are accelerometers and gyroscopes. The advantage of MEMS devices is the fact that they can be integrated with CMOS technology. This can be done by creating the MEMS devices in the same fabrication process as the CMOS interface. When the fabrication processes of the MEMS device and CMOS interface are incompatible, a CMOS chip can be bonded to the MEMS chip or placed in the same package. Having the MEMS device and CMOS interface in close proximity to each other, or integrated on the same chip, can increase performance as well as reduce production costs.

A MEMS inertial sensor typically consists of one or more suspended masses and/or membranes. Miniaturization of these types of devices means the mass becomes increasingly smaller. The interaction of noise forces and environmental disturbances with such an extremely small mass can affect the performance of these inertial sensors [2]. Furthermore, MEMS inertial sensors are based on transduction between an electrical signal and mechanical motion. Thus, electrical noise is also of significant importance. The main noise sources in MEMS inertial sensors are [2]:

- Electrical-Thermal (Johnson-Nyquist) noise, caused by random thermal agitation of charge carriers in conductors
- Generation-Recombination noise, caused by random generation and recombination of electron-hole pairs in semi-conductors
- Flicker ($1/f$ noise), caused by random trapping and release of charge carriers in conductors
- Mechanical-Thermal noise, caused by temperature-induced excitation of atoms and molecules

Some of the noise sources listed above are mainly of significance in electronic readout circuit design and not so much in the design of the MEMS inertial sensor itself. The (intrinsic) limiting factor in the MEMS inertial sensor itself is mechanical-thermal noise. A summary of electrical noise sources closely related to the mechanical design of MEMS inertial sensors is presented in section 2.2 after which mechanical-thermal noise will be discussed in greater detail in section 2.3.

Apart from noise, there are also other effects to take into account. For example vibrations of the MEMS inertial sensor caused by external disturbances. Another well known issue in e.g. gyroscopes is mode coupling, where actuation of one mode can induce motion in another mode as well. Imperfect actuators or misalignment in a magnetic field can cause parasitic actuation of modes other than the mode that one desires to excite. This can cause motion, which may be detected as an acceleration or flow depending on the type of sensor. These effects are elaborated upon in section 2.4.

2.2 Electrical noise

2.2.1 Electrical-Thermal noise

Electrical-thermal noise is mostly of importance in electronic circuitry. However, since its origin and derivation is closely related to mechanical-thermal noise, it will be briefly discussed. In 1928 Johnson presented his discovery of an electromotive force in conductors related to the temperature of the conductor[3]. Johnson describes that this electromotive force is caused by thermal agitation of charge carriers in the conductor. Nyquist then theoretically calculated this electromotive force by using principles from thermodynamics and statistical mechanics [4]. It is often referred to as Johnson-Nyquist noise or thermal noise. Consider a resistor with resistance R at temperature T . If its two terminals are floating, there will be a noise voltage v_n across them. The power spectral density $\mathcal{V}_n$ of the noise voltage v_n can be expressed as [4]:

$$\mathcal{V}_n^2 = 4k_B T R \quad (2.1)$$

Where k_B is Boltzmann's constant. As can be seen from equation 2.1, the noise voltage power spectral density is constant over the frequency spectrum. For a certain measurement bandwidth Δf , the RMS noise voltage is:

$$v_n = \sqrt{4k_B T R \Delta f} \quad (2.2)$$

Taking into account this noise voltage, a resistor can be represented by an ideal resistor in series with a voltage source with voltage v_n , see Figure 2.1. To analyze the effect of thermal noise in an electric circuit, one simply has to replace all resistors in the circuit by the representation from Figure 2.1. No matter how complicated the circuit, the effect of thermal noise can always be analyzed in such a way. For a more detailed derivation, see [4].

Figure 2.1: Representation of electrical-thermal noise in a resistor. A resistor (a) can be represented by an ideal resistor in series with a noise voltage source v_n (b), see Equation 2.2.

When resistors are used in combination with a capacitor in a circuit, the bandwidth Δf will be defined by the effective noise bandwidth of the circuit [5]. For a first order RC circuit, this is:

$$\Delta f = \frac{1}{4RC} \quad (2.3)$$

The RMS noise output voltage of such a circuit can be found by substituting Eq. 2.3 into Eq. 2.2:

$$v_{n,RC} = \sqrt{\frac{4k_B T R}{4RC}} = \sqrt{\frac{k_B T}{C}} \quad (2.4)$$

The resistance R falls out of the equation, since higher R will reduce the bandwidth by the same amount as it increases the noise. The noise is then simply dependent on C . One can see that increasing C will reduce the amount of noise. The higher C , the lower the effective noise bandwidth and thus more noise is being filtered away. Since an ideal capacitor is inherently noiseless, only the bandwidth is lowered without adding more noise as would occur when increasing R .

MEMS inertial sensors are typically actuated and/or read out electrically. Examples of transduction principles are [6]:

- Thermal: Actuation or temperature sensing through an electrical conductor (heater/sensor)
- Piezoelectric: Inducing or sensing motion/deformation.
- (Piezo-)resistive: Sensing strain through a change in resistance.
- Electromagnetic: Actuation by Lorentz forces or sensing changes in magnetic field strength or inductance.

In all of these methods, electrical resistance plays a role and thus electrical-thermal noise is generated. Additionally, actuation and readout circuitry is necessary to make MEMS inertial sensors operable. This is another source of electrical-thermal noise.

2.2.2 Flicker noise

Flicker noise (or $1/f$ noise) is present in many physical systems such as signals in human cognition [7], DNA base sequences [8] and in MOSFETs [9]. The latter is most important for the subject of MEMS, since actuation and readout circuitry are mostly based on CMOS networks [1]. Flicker noise in MOSFETs originates from fluctuations in conductivity. Most theories state that it is caused by fluctuations in: mobility, the total number of carriers, or a combination of both [9]. The spectral density of the noise is approximately proportional to $1/f$, where f is the frequency. Since MEMS inertial sensors are typically read out electrically and their output signals amplified, $1/f$ noise can be of great importance for low frequency signals. For example, measuring low-frequency vibrations with an accelerometer or operating a MEMS

resonating sensor at a relatively low resonance frequency. Elaborating on the latter, designing a resonating MEMS inertial sensor with a higher resonance frequency can thus be beneficial. However, one must keep in mind the effect on other parameters. Taking a vibratory gyroscope as an example, the resonance frequencies of its modes can be increased by making the spring suspension more stiff or by decreasing its (movable) mass. These design changes of course affect its sensitivity and perhaps more importantly mechanical-thermal noise. One typical design choice to increase the SNR related to this noise is to choose a large proof mass. However, this decreases the resonance frequency, possibly increasing the influence of $1/f$ noise. Thus, there is always a trade-off. The same holds for a micro-Coriolis mass flow sensor where there is a trade-off between flow sensitivity, mechanical-thermal noise and resonance frequency. In Chapter 3, the various design parameters are discussed in detail, including the effect of resonance frequency.

2.3 Mechanical-Thermal Noise

In MEMS inertial sensors, there is one fundamental intrinsic noise mechanism, which is mechanical-thermal noise. The phenomenon is caused by random thermal agitation of atoms and molecules. Gabrielson gives a detailed description and derivation of mechanical-thermal noise in micro-machined acoustic and vibration sensors [10]. He states that a damped harmonic oscillator with mass m , damping coefficient r and spring constant k can be described by the following equation:

$$m \frac{d^2 x(t)}{dt^2} + r \frac{dx(t)}{dt} + kx(t) = F_n(t) \quad (2.5)$$

Where F_n is the fluctuation force, which represents the influence of random thermal agitation from the environment on the oscillator. Would this force not be present, the damper in the mass-spring-damper system would cause any oscillation to keep decreasing in amplitude, even the random vibration induced by molecular motion. According to the Fluctuation-Dissipation theorem[11], there needs to be a fluctuation force related to any dissipation/damping mechanism to guarantee that the temperature of a system does not drop below that of its surroundings. Similarly as Equation 2.5, a rotational mass-spring oscillator with mass moment of inertia J , damping coefficient R and rotational spring constant K can be described by:

$$J \frac{d^2 \theta(t)}{dt^2} + R \frac{d\theta(t)}{dt} + K\theta(t) = \tau_n(t) \quad (2.6)$$

Where θ is the angular displacement and τ_n is the fluctuation torque. The noise in a system at thermal equilibrium can be described by using the Equipartition theorem and the Nyquist relation [4]. The equipartition theorem states that in a system at thermal equilibrium, on average, the kinetic energy in the system is distributed evenly over the available degrees of freedom. Each degree of freedom can be seen as a storage of energy, with the amount of stored energy for each degree being equal to $\frac{1}{2}k_B T$, where k_B is Boltzmann's constant ($1.38 \cdot 10^{-23} \text{ J K}^{-1}$) and T is the absolute temperature. A stored energy is an energy form which depends on the square of a variable, e.g. kinetic ($\frac{1}{2}mv^2$), magnetic/inductive ($\frac{1}{2}Li^2$) or spring potential ($\frac{1}{2}kx^2$). For a rotational mass-spring oscillator, the kinetic energy caused by thermal agitation can thus be expressed as [10]:

$$E_n = \frac{1}{2} J \langle \mathcal{O}^2 \rangle = \frac{1}{2} k_B T \quad (2.7)$$

Where $\langle \mathcal{O}^2 \rangle$ is the average of the spectral density of the squared angular velocity over the full frequency spectrum.

To derive the noise torque power spectral density $\mathcal{T}_n$, Equation 2.6 is rewritten in terms of the complex amplitude of the angular velocity Ω , as a function of frequency:

$$j\omega J\Omega(\omega) + R\Omega(\omega) + \frac{1}{j\omega}K\Omega(\omega) = \tau_n(\omega) \quad (2.8)$$

Rewriting this in terms of the mean square angular velocity $\overline{\Omega^2}$:

$$\begin{aligned} \Omega(\omega) \left(j\omega J + R + \frac{1}{j\omega}K \right) &= \tau_n(\omega) \\ \Omega(\omega) &= \frac{\tau_n(\omega)}{j\omega J + R + \frac{1}{j\omega}K} \\ \overline{\Omega^2} &= \frac{\overline{\tau_n^2}}{R^2 + (\omega J - K/\omega)^2} \\ &= \frac{1}{R^2} \frac{\overline{\tau_n^2}}{1 + Q^2(\omega/\omega_0 - \omega_0/\omega)^2} \end{aligned} \quad (2.9)$$

E_n can be related to $\overline{\Omega^2}$ via Eq. 2.7:

$$\begin{aligned} E_n &= \frac{1}{2}J \langle \mathcal{O}^2 \rangle = \frac{1}{2}J \int_0^\infty \overline{\Omega^2} df \\ &= \frac{J}{2R^2} \int_0^\infty \frac{\overline{\tau_n^2}}{1 + Q^2(\omega/\omega_0 - \omega_0/\omega)^2} df \\ &= \frac{1}{4\pi R} \int_0^\infty \frac{\overline{\tau_n^2}Q}{1 + Q^2(f/f_0 - f_0/f)^2} d(f/f_0) \end{aligned} \quad (2.10)$$

As Nyquist described, the power transferred between dissipation mechanisms at thermal equilibrium is always of equal magnitude, no matter at what frequency the transfer occurs [4]. Similarly, $\overline{\tau_n^2}$ must be constant over the full frequency range. Thus E_n can be expressed as:

$$E_n = \frac{\overline{\tau_n^2}}{4\pi R} \int_0^\infty \frac{Q}{1 + Q^2(f/f_0 - f_0/f)^2} d(f/f_0) \quad (2.11)$$

Solving the integral in Equation 2.11 gives $\pi/2$ (the full derivation is a lengthy one and will not be presented here[†]). An expression for τ_n can then be derived:

$$E_n = \frac{\overline{\tau_n^2}}{4\pi R} \cdot \frac{\pi}{2} = \frac{\overline{\tau_n^2}}{8R} \quad (2.12)$$

$$\overline{\tau_n^2} = 8R \cdot E_n = 4k_B T R \quad (2.13)$$

[†]A possible approach is to start by factoring the denominator, then use partial fractions to split the integral into a sum of multiple integrals and solve these.

Figure 2.2: Representation of mechanical-thermal noise in a rotational mass-spring-damper oscillator. A damper (a) can be represented by an ideal damper with an additional applied noise torque τ_n (b), see Equation 2.15.

2

As stated earlier, $\overline{\tau_n^2}$ is constant over the full frequency range. Thus, Eq. 2.13 holds for all frequencies. The power spectral density of the noise torque is then:

$$\mathcal{T}_n^2 = 4k_B T R \quad (2.14)$$

The power spectral density is given with units $\text{N}^2 \text{m}^2 \text{Hz}^{-1}$. Thus, the RMS noise torque for a given bandwidth Δf is:

$$\tau_n = \sqrt{4k_B T R \Delta f} \quad (2.15)$$

This is essentially the same expression as for the RMS noise voltage associated with an electrical resistor (Equation 2.2). This is to be expected since the noise voltage and electrical resistor are electrical analogies of the noise torque and rotational damping coefficient. Just as for an electrical circuit, thermal noise in a rotational mechanical system can be analyzed by applying a noise torque to any damper in the system. Figure 2.2 shows how a mechanical rotational damped oscillator can be represented to take into account mechanical-thermal noise.

Figure 2.2b is simply a representation of Equation 2.6. Now that an expression for τ_n has been obtained, it can be used to derive e.g. the resulting noise angle θ_n . First, Equation 2.6 is converted to the frequency domain using Equation 2.15 for τ_n :

$$(j\omega)^2 J \theta_n(\omega) + j\omega R \theta_n(\omega) + K \theta_n(\omega) = \sqrt{4k_B T R \Delta f} \quad (2.16)$$

Now $\theta_n(\omega)$ can be expressed:

$$\begin{aligned}\theta_n(\omega) &= \frac{\sqrt{4k_B T R \Delta f}}{K - \omega^2 J + j\omega R} \\ |\theta_n(\omega)| &= \frac{\sqrt{4k_B T R \Delta f}}{K \sqrt{\left(1 - \left(\frac{\omega}{\omega_0}\right)^2\right)^2 + \left(\frac{\omega}{\omega_0 Q_c}\right)^2}}\end{aligned}\tag{2.17}$$

In a similar way one can derive the noise displacement in an accelerometer or pressure sensor. A derivation of the thermal noise equivalent flow for a micro-Coriolis mass flow sensor will be derived in Chapter 3 using the expression for the thermal noise torque from equation 2.15.

2.4 Other limiting factors

2.4.1 Vibrations from external disturbances

External disturbances or environmental excitation can cause vibrations of a MEMS inertial sensor. A distinction is made between accelerometers and gyroscopes, since an accelerometer is designed to measure vibrations whereas a gyroscope exploits vibrations to detect angular acceleration. In the former, external vibrations are thus simply the signal to be detected. In the case of gyroscopes however, these vibrations can potentially introduce errors, since their operation typically relies on detecting very small angular accelerations of micro-structures [1]. This can be of great importance when these devices are used in the automotive industry [12], in audio systems or in other situations where they might be in close proximity to vibrating parts. Let us take a simple vibratory gyroscope as an example, consisting of a proof mass attached to a rigid structure via springs, see Figure 2.3.

Figure 2.3: Schematic representation of a vibratory gyroscope where Ω represents the rotation that the sensor experiences.

A vibratory gyroscope as in Figure 2.3 operates by actuating the sensor at resonance using a driving force F_{drive} in the x -direction. When the sensor experiences a rotation Ω , Coriolis forces (F_{sense}) actuate the sensor along the y -axis. Since the sensor's operation relies on measuring small vibrations of the proof mass, environmental vibrations can cause errors in the measured angular rate. The sense mode vibrations of the proof mass can be described by a second order differential equation:

$$m \frac{d^2 y(t)}{dt^2} + r_y \frac{dy(t)}{dt} + k_y y(t) = F_{\text{sense}}(t) \quad (2.18)$$

Figure 2.4: The sense displacement $|Y|$ of a typical vibratory gyroscope with respect to the ratio of the sense force frequency ω and the sense mode resonance frequency ω_s (see Eq. 2.20).

2

The displacement y can then be derived in the same way as the noise angle in section 2.3 (Equation 2.17) by first converting Equation 2.18 to the frequency domain:

$$(j\omega)^2 mY(\omega) + j\omega rY(\omega) + kY(\omega) = F_{\text{sense}}(\omega) \quad (2.19)$$

Then expressing $|Y(\omega)|$:

$$|Y(\omega)| = \frac{|F_{\text{sense}}(\omega)|}{K \sqrt{\left(1 - \left(\frac{\omega}{\omega_s}\right)^2\right)^2 + \left(\frac{\omega}{\omega_s Q}\right)^2}} \quad (2.20)$$

Figure 2.4 shows the value of Equation 2.20 with respect to the frequency ratio ω/ω_s .

From this, it can be seen that forces close to the sense mode resonance frequency cause the largest displacement $|Y|$. This is why some gyroscopes are designed such that the sense and drive mode resonance frequencies are very close in value (mode-matched [1]). However, this does require strict control of device parameters (Q , mass, spring stiffness, etc.), where a small change in one of these parameters can cause a large change in sensitivity to rotation. Additionally it reduces the bandwidth of the device by a factor Q . Therefore mode-matching is commonly done when high sensitivity and low noise are crucial or high vibration frequencies are possible. An alternative operating method is mode-split operation, where the resonance frequencies of the device are purposefully mismatched so as to improve stability and increase bandwidth.

However, for both operation schemes (mode-matched or mode-split) environmental vibrations can cause issues. These vibrations induce an additional force F_{env} causing an additional displacement Y_{env} :

$$|Y_{\text{env}}(\omega)| = \frac{|F_{\text{env}}(\omega)|}{K \sqrt{\left(1 - \left(\frac{\omega}{\omega_s}\right)^2\right)^2 + \left(\frac{\omega}{\omega_s Q}\right)^2}} \quad (2.21)$$

$|Y_{\text{env}}|$ follows the same relation as displayed in Figure 2.4. This means that if environmental vibrations occur at frequencies close to ω_s they can cause significant displacements distorting the sense signal of the gyroscope. This can lead to an increased noise floor or disrupt operation of the sensor entirely. Methods to reduce the sensitivity to these vibrations include:

- Designing the sensor such that its resonance frequency is much higher than that of typical environmental vibrations.
- Operating the sensor at a higher order resonance mode which is not/less susceptible to the vibrations.
- Vibration isolation via additional spring suspension or active vibration control.
- Differential readout method where the vibrations are canceled out, e.g. via second mass vibrating with opposite phase.

Considering the first option, if the resonance frequency of the device is much higher than that of environmental vibrations, it would be less susceptible to these vibrations. This can be easily seen from Fig. 2.4, where low frequency forces cause relatively small displacement of the proof mass relative to forces at frequencies close to the resonance frequency. Instead of increasing the device's resonance frequency, an alternative would be to operate the device in a higher order resonance mode. This can be beneficial in two ways. Firstly, similar to increasing the resonance frequency of the device, using a higher order mode makes it less susceptible to low frequency environmental vibrations. Secondly, higher order mode shapes in combination with appropriate readout electrode positioning can allow for cancellation of motion caused by environmental vibrations. The latter is discussed in more detail further on in this section.

Another approach would be to isolate the MEMS device from these vibrations [13]. This type of method uses an isolating spring suspension [14–16] or active vibration control [17, 18] with e.g. electrostatic actuators. This type of vibration isolation has also been applied to Coriolis mass flow sensors on the macro scale [19]. A vibration isolator based on a (soft) spring suspension is essentially a mechanical low-pass filter. This filters out any environmental vibrations of frequencies higher than the device's resonance frequency. However, vibrations at frequencies around the resonance frequency are not filtered out using this method. Active vibration isolation can be a solution in this case. Figure 2.5 displays an example of an active vibration isolator for a MEMS device based on electrostatic force-feedback.

Figure 2.5: Schematic representation of an active vibration isolator for a MEMS device, consisting of an additional spring suspension and electrodes, which can be used for electrostatic force-feedback.

The MEMS device is suspended to an outer frame via springs. Electrodes at the bottom of the MEMS device and on the frame form a capacitor. When environmental vibrations excite the frame, the MEMS device will move, resulting in a change in capacitance. The capacitance can be measured and an appropriate feedback signal can be supplied to the electrodes to generate electrostatic forces, which can be used to tune the filter characteristics of the isolator [17]. This way, influence from environmental vibrations can be reduced more with respect to a passive isolator. The disadvantage of active vibration isolation is that it requires complex structures, making fabrication more cost/time intensive. It also increases power consumption and size of the device.

Finally, optimizing the readout method can allow for cancellation of motion caused by environmental vibrations. As aforementioned, this can be done by utilizing a higher order resonance mode or more specifically, a differential mode. A differential vibration mode is a mode where the vibrating masses/beams vibrate in opposite direction. By using a differential readout, common mode vibrations caused by environmental excitation can be canceled out. A well known example of an inertial sensor with a differential readout is a tuning fork gyroscope. Figure 2.6 shows a schematic representation of such a gyroscope. A drive force actuates the fingers of the tuning fork (Fig. 2.6a) in opposite directions. When the tuning fork experiences a rotation, Coriolis forces excite the fingers in a second vibration mode (again in opposite direction). Since environmental vibrations excite the entire device, vibration modes where both fingers vibrate in the same direction are much more affected. Figure 2.6b shows a MEMS implementation of a tuning fork gyroscope. It consists of two mechanically coupled proof masses. The masses are actuated in opposite direction just as for the classical tuning fork. When a rotation is applied, Coriolis forces cause a second vibration mode where the masses move perpendicular to the drive axis and once again in opposite direction. These movements are typically read out electrostatically. Using a differential actuation and readout such as depicted in Figure 2.6 reduces the effect of environmental vibrations. However, this requires good matching of the masses and springs. Moreover, for full cancellation the readout should be linear. A similar approach is applied to a μ -CMFS in Chapter 6 with measurement results in Chapter 7.

Figure 2.6: Schematic representation of a tuning fork gyroscope. Drive forces (F_{drive}) actuate the fingers/masses differentially. When the device experiences a rotation with angular velocity Ω , sense forces (F_{sense}) cause a differential vibration in a second vibration mode.

2

2.4.2 Mode coupling

MEMS resonators have numerous resonance modes. When these devices are based on a single mass and suspension, mechanical coupling between the modes can cause undesired vibration. This is a typical issue in MEMS vibratory gyroscopes. As briefly discussed in the previous section on external disturbances, MEMS gyroscopes can be operated in a mode-matched scheme or a mode-split scheme. The advantage of mode-matched operation is that the sensitivity to the angular rate is much larger when these frequencies are close together (Equation 2.20). However, when the same proof mass with spring suspension is used for the drive and sense modes, mechanical coupling can cause instability. A solution to this issue is to decouple the two modes from each other. This can be done by using separate spring suspensions for the two modes [20, 21]. Just as a vibratory gyroscope, the operation of a μ -CMFS is also based on drive and sense resonance modes. The main difference is that Coriolis forces are generated due to a fluid traveling through a vibrating channel instead of a rotation exerted on a vibrating proof mass. Decoupling the modes is more complicated however. As will be described in more detail in Chapter 3, a μ -CMFS consists of a suspended micro-channel. The actuation and detection vibrational modes are both out-of-plane modes. Where in the actuation mode, one side of the channel moves in opposite direction as the other and in the detection mode both sides move in the same direction. However, when viewing the respective sides of the micro-channel, both modes involve out-of-plane movement of these channel sections. Therefore, using separate suspensions to decouple the modes is not possible without changing the entire design and/or utilizing higher order vibrational modes. This can result in lower flow sensitivity and resolution of the sensor.

2.4.3 Transduction principle

MEMS inertial sensors are based on converting mechanical motion into an electrical signal, which is in turn translated to an output signal through electronic circuitry. The chosen transduction principle affects the performance of a device in terms of power consumption, stability, ease of implementation and conversion efficiency among others. Various transduction principles have been applied to MEMS inertial sensors such as capacitive, (piezo-)resistive, piezoelectric and optical [6, 22]. Capacitive transducers are most common in MEMS inertial sensors. This is thanks to the fact that they have high sensitivity, good noise performance, low drift, low temperature sensitivity and simple structure [1]. Since a capacitor itself is inherently noiseless, resolution limits associated with capacitive MEMS inertial sensors are caused by mechanical-thermal noise, noise in the electronic readout circuitry and mechanical sensitivity of the device. As discussed earlier, electrical-thermal noise in an RC circuit can be reduced by reducing the effective noise-bandwidth through increasing the capacitance C . Readout capacitors in a MEMS device form RC-circuits together with resistances of the connecting lines and the capacitor itself (since it is never ideal) and with the input impedance of the readout circuit. Thus, electrical-thermal noise is generated and consequently shaped by the readout capacitance. Increasing the readout capacitance can therefore reduce the electrical-thermal noise. Therefore in MEMS design, readout capacitors are generally designed with an as large capacitance as possible, e.g. by reducing the gap between capacitor electrodes. This is limited by what can be achieved in the fabrication process of the MEMS device. Limitations of a capacitive readout for a μ -CMFS are discussed in Chapter 6. Another transduction principle that is often used in MEMS inertial sensors is piezo-resistivity [23]. The first micro-machined accelerometer utilized piezo-resistivity for its operation [24]. Typically, these devices are based on doped silicon piezo-resistors, which are both used as suspension and readout structures. The main advantage of the use of piezo-resistors is their simple integration into silicon micro-machining. Another advantage is that readout circuitry for piezo-resistive sensing can be relatively simple due to the low output impedance voltage generated by a resistive bridge [1]. However, piezo-resistive inertial sensors suffer from larger temperature sensitivity as well as lower sensitivity to motion [22]. The latter depends on the type of inertial sensor and specific limitations to its motion, as will be discussed for the μ -CMFS in Chapter 6. Since a piezo-resistive readout is based on an electrical resistor, the main noise source in such devices (in addition to mechanical-thermal noise) is electrical-thermal (Johnson) noise, see section 2.2. The resolution limit of a piezo-resistive MEMS inertial sensor is defined by this noise as well as the relatively low sensitivity to motion (compared to a capacitive readout). Apart from capacitive and piezo-resistive devices, some other types of readout techniques are used for MEMS inertial sensors, such as piezoelectric [25] and optical [26]. The main disadvantage of these techniques is

complicated integration into silicon micro-machining. For certain applications, the use of these readout techniques can be beneficial. Piezoelectric materials can be used for both actuation and sensing (discussed in more detail in Chapters 5 and 6). The fundamental noise limit of piezoelectric MEMS inertial sensors is a combination of mechanical-thermal noise of the moving part of the device and electrical-thermal noise in the piezoelectric element [27]. Optical readout can be useful in environments where electromagnetic fields are present and other readouts such as capacitive/piezo-resistive cannot be applied due to the disturbance from electromagnetic interference [26]. Similarly to a capacitive readout, the fundamental noise limit is defined by mechanical-thermal noise, although fluctuations in wavelength and intensity of the used optical beam/laser can also be of influence [2, 28]. Chapter 6 discusses readout techniques for a μ -CMFS, focusing on capacitive and piezo-resistive detection. The fabrication processes to realize the readout structures are presented in Chapter 4. Piezoelectric and optical readout methods are out of the scope of this research. However, work is currently being done on piezoelectric and optical readout of a μ -CMFS at the Integrated Devices and Systems group at the University of Twente. [29, 30]

2

2.5 Conclusion

The influences of noise and other limitations on the performance of micro-machined inertial sensors have been discussed. Electrical noise sources such as electrical-thermal, generation-recombination and flicker ($1/f$) noise are mainly limiting the electronic readout circuitry surrounding these sensors. The focus of the research in this thesis lies on performance limitations related to mechanical design. Among these, mechanical-thermal noise limits the resolution of current state-of-the-art MEMS inertial sensors. The torque induced by mechanical-thermal noise in a rotational mass-spring-damper system has been derived. The effect of this noise torque on a μ -CMFS is discussed and the noise equivalent flow will be derived in Chapter 3. Limiting factors other than noise and their influence on a μ -CMFS have been discussed. Approaches on how to deal with these limits, resulting designs/devices and their experimental analyses will be presented in Chapters 6 and 7.

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3

Modelling of micro-Coriolis mass flow sensors

This chapter[†] starts with the derivation of an analytical model that describes the flow sensitivity of a μ -Coriolis mass flow sensor (μ -CMFS) and relates it to the dimensions and mechanical properties of the device structure. An FEM model is then set up in COMSOL Multiphysics[®], which is able to swiftly determine resonance frequencies and other properties for a range of design parameters. The resolution limit for flow sensing related to mechanical-thermal noise is briefly discussed to give an insight into the possible improvement. The combination of the analytical and FEM models is used to create new sensor designs, which are presented and characterized in chapter 7.

3

[†]This chapter is based on (part of) publication [1]:

- T. V. P. Schut, R. J. Wiegerink, and J. C. Lotters, “ μ -Coriolis mass flow sensor with improved flow sensitivity through modelling of the sensor,” *Microelectronic Engineering*, p. 111289, 2020

3.1 Introduction

Over the last decades, there has been much work on μ -Coriolis mass flow sensors (μ -CMFS), as discussed in chapter 1. However, only a number of them discuss modelling of such a sensor [2–5]. Of these, only Groenesteijn et al. [4] compare the simulation results with actual flow measurement results. The model by Groenesteijn et al. is based on the combination of SolidWorks and MATLAB®'s SPACAR package. Mechanical properties of the channel cross section are determined using SolidWorks, after which the channel structures are approximated as a finite number of line elements in SPACAR. Resonance frequencies and flow sensitivities are determined in SPACAR as well. This method allows for fast simulation of various sensor designs. With this model, resonance frequencies of various designs are predicted within 10% error. Errors of <26% are achieved for simulated flow sensitivities. As a part of the research in this thesis, a similar model has been set up based on finite element method simulation in COMSOL Multiphysics® in combination with an analytical model. The main advantage of this model over the one by Groenesteijn et al. is that it allows for fast simulation of a range of dimensions of the channel cross section in addition to the dimensions of the overall sensor structure (whereas in [4], channel cross section properties are determined separately in SolidWorks). Furthermore, elements with different cross sections than the main channel can be integrated as well without significant increase of simulation times. This allows for simulation of the effect of spring suspensions and supporting structures. First, an analytical model of the μ -CMFS is derived. Subsequently the FEM model is set up and simulation results are presented. μ -CMFS designs are made based on this model for validation of the simulations. These devices and experimental results are presented in Chapter 7 and a comparison between simulation and experiments is made. Furthermore, the flow resolution limit related to mechanical-thermal noise is derived and briefly discussed, after which this chapter is concluded.

3.2 Analytical model of a μ -Coriolis flow sensor

The operating principle of a Coriolis mass flow sensor is based on the generation of Coriolis forces on a flowing fluid. To generate these forces, angular motion is induced in a channel through which fluid flows. Figure 3.1 shows a schematic representation of how Coriolis forces are induced in such a channel. A fluid flows through a vibrating channel with length L (Fig. 3.1a). The path of a particle flowing through the channel is depicted. Viewed from the rotating reference frame, the path is straight and has length L . When viewing the path from the fixed reference frame (Fig. 3.1b), it can be observed that the particle not only travels along the x -axis, but also along the z -axis. This constantly changes the direction of velocity of the particle. It can be seen as a force applied by the channel wall to keep the particle in its path. As a result, the fluid particle exerts an opposing (Coriolis) force on the channel wall, perpendicular to the axis of rotation and the direction of flow.

$$\mathbf{F}_C = -2m\boldsymbol{\Omega} \times \mathbf{v} \quad (3.1)$$

Where $\mathbf{v}$ is the velocity of the particle. The Coriolis force (magnitude) on a channel with length L , vibrating with an angular velocity Ω , in terms of mass flow rate ϕ_m is:

$$F_C = -2L\phi_m\Omega \quad (3.2)$$

(a) A fluid flows through a vibrating channel, with angular velocity Ω . The path of a fluid particle is viewed from the rotating reference frame.

(b) The path of a fluid particle viewed from the fixed reference frame.

Figure 3.1: Schematic representation of how Coriolis forces are induced in a vibrating channel.

Figure 3.2: Operation principle of a μ -Coriolis flow sensor. The ratio between the two modal vibration amplitudes is a measure for the flow rate.

Thus the Coriolis force is linearly dependent on the mass flow rate through the channel. This effect is exploited in a Coriolis mass flow sensor. Figure 3.2 shows the basic structure of a μ -Coriolis mass flow sensor and its operating principle. The structure consists of a suspended micro-channel following a rectangular loop. The channel is fixed at two points, such that it can vibrate in various modes. The channel of the sensor is brought into resonance in the actuation mode at frequency ω_a (Fig. 3.2a). The channel can be actuated e.g. electrostatically, magnetically or piezoelectrically (discussed in more detail in Chapter 5). When fluid flows through the channel with a mass flow rate ϕ_m , Coriolis forces are induced in the rotating channel section, see Fig. 3.2b. This causes additional vibration in the Coriolis/detection mode with an angle θ_d and displacement Δz_d . The ratio between the detection and actuation vibration amplitudes is a measure for the mass flow rate. Sensitivity to flow is thus quantified as the ratio between the detection and actuation angles (θ_d/θ_a) per mass flow rate ϕ_m .

As aforementioned, Haneveld et al. [3] derived an analytical model describing the flow sensitivity of a μ -CMFS. They assumed that the actuation and detection resonance frequencies (ω_a and ω_d respectively) are far apart, such that the final equation can be simplified. However, this is not always true. Especially when simulating many different sets of dimensions, the resonance frequencies can be very close in value. Therefore, the flow sensitivity needs to be derived without making the assumption that $\omega_a \gg \omega_d$ or $\omega_a \ll \omega_d$.

Vibration of the sensor can be described by a second order differential equation:

$$J_{a,d} \frac{d^2 \theta_{a,d}(t)}{dt^2} + R_{a,d} \frac{d\theta_{a,d}(t)}{dt} + K_{a,d} \theta_{a,d}(t) = \tau_{a,d}(t) \quad (3.3)$$

Where the subscript a,d indicates either actuation or detection mode vibration. $J_{a,d}$ is the moment of inertia, $R_{a,d}$ is the damping coefficient, $K_{a,d}$ is the torsional

spring constant, $\theta_{a,d}$ is the angular displacement and finally $\tau_{a,d}$ is the actuation torque or detection torque induced by Coriolis forces.

The expression for the Coriolis force in a channel was given in Eq. 3.2. The μ -CMFS structure in Fig. 3.2 consists of multiple channel sections. When the device is actuated, all channel sections experience rotation around the axis perpendicular to the flow direction. Each channel section has its own length L_i and rotation angle θ_i (see Fig. 3.1), where the subscript i indicates the i^{th} channel section. The Coriolis force in each section can be defined as:

$$F_{C,i} = -2L_i\phi_m \frac{d\theta_i}{dt} \quad (3.4)$$

The angle θ_i for each section is dependent on the overall actuation angle θ_a (see Fig. 3.2a). Thus, the total Coriolis force on the channel induces a detection torque $\tau_d(t)$, proportional to the actuation angular velocity $d\theta_a/dt$:

$$\tau_d \propto \frac{d\theta_a}{dt} \quad (3.5)$$

The actuation angle and angular velocity will be of the form:

$$\theta_a(t) = \alpha \sin(\omega_a t) \quad (3.6)$$

$$\frac{d\theta_a(t)}{dt} = \omega_a \cdot \alpha \cos(\omega_a t) \quad (3.7)$$

Where α is the amplitude of the overall actuation angle (see Fig. 3.2a). From Equation 3.5 it then follows that $\tau_d(t)$ is of the same form as $d\theta_a/dt$:

$$\tau_d(t) = T_d \cos(\omega_a t) \quad (3.8)$$

Where T_d is the amplitude of $\tau_d(t)$, being linearly dependent on ω_a , α and ϕ_m . Applying $\tau_d(t)$ to Equation 3.3 gives a detection angle of the following form:

$$\theta_d(t) = \beta \sin(\omega_a t) + \gamma \cos(\omega_a t) \quad (3.9)$$

Inserting this in equation 3.3 and solving for β and γ gives:

$$\beta = \gamma \frac{\omega_a \omega_d}{(\omega_d^2 - \omega_a^2) Q_d} \quad (3.10)$$

$$\gamma = \frac{T_d \omega_d^2 Q_d^2 (\omega_d^2 - \omega_a^2)}{K_d (Q_d^2 (\omega_d^2 - \omega_a^2)^2 + \omega_a^2 \omega_d^2)} \quad (3.11)$$

Where Q_d is the detection mode quality factor. Now that we have obtained expressions for γ and β , we can derive the ratio between the actuation and detection angle amplitudes:

$$\begin{aligned}\theta_d(t) &= \beta \sin(\omega_a t) + \gamma \cos(\omega_a t) \\ &= \sqrt{\beta^2 + \gamma^2} \cdot \sin\left(\omega_a t + \arctan\left(\frac{\gamma}{\beta}\right)\right)\end{aligned}\quad (3.12)$$

$$\frac{\hat{\theta}_d}{\hat{\theta}_a} = \frac{\sqrt{\beta^2 + \gamma^2}}{\alpha} \quad (3.13)$$

With equations 3.10 and 3.11 this gives:

$$\frac{\hat{\theta}_d}{\hat{\theta}_a} = \frac{T_d \omega_d^2 Q_d^2 (\omega_d^2 - \omega_a^2)}{\alpha K_d (Q_d^2 (\omega_d^2 - \omega_a^2)^2 + \omega_a^2 \omega_d^2)} \sqrt{1 + \frac{\omega_a \omega_d}{(\omega_d^2 - \omega_a^2) Q_d}} \quad (3.14)$$

Since what is finally measured is the displacement of the channel in the z direction Δz_d (see Figure 3.2), the flow sensitivity is finally quantified as:

$$\begin{aligned}\frac{\Delta \hat{z}_d}{\hat{\theta}_a} &= L_y \cdot \frac{\hat{\theta}_d}{\hat{\theta}_a} \\ \frac{\Delta \hat{z}_d}{\hat{\theta}_a} &= \frac{L_y T_d \omega_d^2 Q_d^2 (\omega_d^2 - \omega_a^2)}{\alpha K_d (Q_d^2 (\omega_d^2 - \omega_a^2)^2 + \omega_a^2 \omega_d^2)} \sqrt{1 + \frac{\omega_a \omega_d}{(\omega_d^2 - \omega_a^2) Q_d}}\end{aligned}\quad (3.15)$$

Where L_y is the distance between the detection mode rotational axis and the channel section where $\Delta \hat{z}_d$ is measured (Indicated in Fig. 3.2a). Equation 3.15 holds for small angles θ_d . From equation 3.15, it seems as if the length of the channel section where Coriolis forces are induced (L_x , see Fig. 3.2a) has no influence on the sensitivity. However, the detection torque T_d is directly dependent on L_x and other parameters (ω_a , ω_d , K_d and Q_d) are affected by L_x as well. Since many of the variables involved are related to each other, T_d is not expressed and is determined in the FEM simulation.

3.3 FEM model

The FEM model consists of two simulation steps. Firstly, a 2D simulation step of the channel cross-section (Figure 3.3a) calculates mechanical properties such as second moment of area and torsion constant. These are then used in a following 3D eigen frequency simulation step (Figure 3.3b). The 3D structure in Figure 3.3b consists of a collection of 1D line elements in 3D space. Different cross-sectional geometries can be coupled to the various line elements to simulate the influence of e.g. spring suspensions or channel sections with different diameters. Figure 3.4 shows the simulated actuation and detection modes of the geometry in Fig. 3.3b.

Figure 3.3: Geometries in COMSOL[®] used for 2-step FEM simulation of the μ -CMFS, to determine resonance frequencies, Coriolis forces and torsional springs constants.

Figure 3.4: Actuation and detection mode shapes of a typical μ -CMFS geometry (Fig. 3.3b) simulated in COMSOL[®], with N_2 as flow medium. The color scheme indicates mechanical stress from low (blue) to high (red).

The 3D eigen frequency simulation step determines the resonance frequencies of the device. Coriolis forces are calculated by evaluating Equation 3.2 for each channel section. The total Coriolis/detection mode torque amplitude T_d is then derived by evaluating Eq. 3.4 for each channel section and multiplying with the distance to the actuation mode rotational axis. Additionally, ω_a , ω_d , α and K_d are calculated in the simulation. Combining these parameters with Equation 3.15 and the quality factor Q_d gives the simulated flow sensitivity. Q_d is estimated based on previously characterized devices. Simulating the aforementioned parameters for one set of dimensions can be done in around 30 seconds. Thus, many different dimensions and designs can be compared relatively fast by sweeping variables.

3.3.1 Influence of channel loop size

The influence of the channel loop size is investigated by varying L_x and L_y . Figure 3.5 shows the simulated flow sensitivity for a range of values of L_x and L_y . As can be seen from the figure, the highest sensitivity is found when L_x is large and L_y is small. However, this is also the range where the actuation and detection mode resonance frequencies are close in value. This is an unstable region, since a small change in Q_d or ω_a/ω_d can cause a significant change in flow sensitivity. For larger L_y , the resonance frequencies are further apart in value and the sensitivity is still relatively high (indicated in Figure 3.5). A disadvantage of such a large channel loop, is that the pressure drop over the device becomes larger, which can be unfavourable for certain applications. However, since there are large differences between sensitivities of smaller and larger channel loop devices, 3 designs are made based on these measurements to validate the model. These designs and a comparison of experimental and modelled parameters is presented in chapter 7.

3.3.2 Integration of spring structures

As aforementioned, the model allows for different cross section properties to be coupled to the 1D line elements. This means that spring suspensions and other structures attached to the micro-channel can be easily incorporated in the simulation. Such spring suspensions can be useful for suppressing in-plane motion of the sensor structure. Figure 3.6 shows an example of how such a spring suspension can be implemented and how it affects the actuation and detection vibrational modes. When comparing the resonance mode shapes and frequencies to those in Figure 3.4, there is only a small change due to the spring suspension. However, in-plane motion will be heavily suppressed with such a suspension. More detailed discussion and characterization on in-plane motion suppression will be presented in chapter 7.

Figure 3.5: Simulated flow sensitivity in relation to L_x and L_y (see Figure 3.3) with N_2 as the flow medium. Large L_x and small L_y give a high sensitivity but low stability since ω_{a} and ω_{d} are close in value in this range. Small changes in $\omega_{\text{a}}/\omega_{\text{d}}$ or Q_{d} can cause significant changes in flow sensitivity. Large L_x and L_y gives relatively high flow sensitivity and better stability.

3

(a) Actuation/*Twist* vibrational mode,
 $\omega_{\text{a}} = 3291.1 \text{ Hz}$.

(b) Detection/*Swing* vibrational mode,
 $\omega_{\text{d}} = 1863.2 \text{ Hz}$.

Figure 3.6: Actuation and detection mode shapes of a $\mu\text{-CMFS}$ geometry with spring suspension, consisting of SiRN membranes with a cross section of $82 \times 2.5 \mu\text{m}$ to suppress in-plane motion. The flow medium is N_2 .

3.4 Mechanical-Thermal noise

In Chapter 2, mechanical-thermal noise in micro-machined inertial sensors was discussed. The noise torque induced by mechanical-thermal noise in a rotational mass-spring-damper system was derived (Equation 2.15). A μ -CMFS can also be viewed as a rotational mass-spring-damper system (Equation 3.3). The detection torque amplitude T_d induced by Coriolis forces is linearly proportional to the mass flow rate ϕ_m . The detection torque per mass flow rate (T_d/ϕ_m) is calculated in the FEM simulation by evaluating Equation 3.2 for each 1D line element. To view the effect of a noise torque induced by mechanical-thermal noise, T_d is equated to Equation 2.15 to find the noise equivalent flow ϕ_n :

$$\begin{aligned} \frac{T_d}{\phi_m} = \frac{\tau_n}{\phi_n} &= \frac{\sqrt{4k_B T R_d \Delta f}}{\phi_n} \\ \phi_n &= \frac{\sqrt{4k_B T R_d \Delta f}}{T_d/\phi_m} = \frac{\sqrt{4k_B T (K_d/\omega_d Q_d) \Delta f}}{T_d/\phi_m} \end{aligned} \quad (3.16)$$

Alveringh et al. presented a similar derivation of the noise torque in a μ -CMFS [6]. The model was confirmed experimentally by measuring the displacement of a μ -CFMS at different temperatures using Laser Doppler Vibrometry (this technique will be explained in more detail in Chapter 7). An expression for the noise equivalent flow is also derived in [6]. However, only the Coriolis forces generated in a single channel section (section with length L_x in Fig. 3.3b) are taken into account. Whereas in Eq. 3.16, T_d is derived in the FEM model by calculating the contribution of each channel section to the total Coriolis torque. Therefore the expression for the noise equivalent flow in Eq. 3.16 is more complete with respect to the one presented in [6].

Oliveira et al. [7] presented the current state of the art combination of μ -CMFS and CMOS based actuation/readout electronics, achieving a zero-flow stability of $\pm 0.6 \text{ mg h}^{-1}$ (For N_2 as flow medium) with a measurement bandwidth of 3 Hz. Inserting the dimensions of the device presented in [7] into the FEM model and evaluating Equation 3.16, gives a noise equivalent flow of 0.45 ng s^{-1} . This means that the mechanical-thermal noise floor lies a factor of ~ 480 below the current state of the art zero-flow stability (for a MEMS/CMOS combination).

Groenesteijn et al. previously achieved a zero-flow stability of 14 ng s^{-1} [8], which is a factor 12 better than the device in [7]. The high resolution was attained by improving the flow sensitivity through decreasing the micro-channel diameter and using overlapping readout electrodes to decrease sensitivity to actuation motion. However, this gives some disadvantages. A smaller channel diameter increases the pressure drop over the micro-channel significantly. In addition, overlapping readout electrodes cause increased dependence of the flow sensitivity on the exact position of

the actuation mode rotational axis. The latter translates to larger deviation between devices as well as causing effects from pressure and flow medium, which can cause a shift of the rotational axis. For these reasons, the device presented by Oliveira et al. is taken as a reference throughout this thesis.

3.5 Conclusion

A model has been set up to predict resonance behaviour and flow sensitivity values for a μ -Coriolis mass flow sensor. Mechanical properties and resonance frequencies are determined using FEM in COMSOL[®] after which flow sensitivity is calculated using an analytical solution. The analytical model was derived without assuming that the actuation and detection mode resonance frequencies are far apart (as opposed to [3, 9]), giving a more thorough representation of the μ -CMFS. The FEM model allows for simple integration of spring structures and fast simulation (30 seconds for 1 set of design parameters). The model can be used as a platform for designing new μ -CMFS devices. In Chapter 7, devices with different channel loop dimensions are presented and characterized to validate the model. Furthermore, other devices have been designed and are presented in Chapter 7 as well. As much improvement can still be made before reaching the resolution limit related to mechanical-thermal noise, designs in this thesis are based on improving flow sensitivity and actuation/readout methods of the μ -CMFS. The latter is thoroughly discussed in Chapters 5 and 6. Furthermore, all devices in this thesis are designed with a channel diameter in the range of 48–55 μm (depending on the specific fabrication run), to limit the pressure drop needed to operate the sensor. Additionally, the focus lies on improving performance of the sensor without increasing dependence on other parameters, like the position of the rotational axis [8] or the pressure inside the micro-channel.

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4

Fabrication Methods

This chapter[†] introduces the fabrication process (Surface Channel Technology, SCT) which is used to realize the μ -Coriolis mass flow sensor devices presented in this thesis. First, the basic SCT is described in detail. Subsequently, several methods for integration of doped silicon structures in SCT are discussed.

[†]This chapter is partially based on publication [1]:

- Y. Zhao, H.-W. Veltkamp, T. V. P. Schut, R. G. P. Sanders, B. Breazu, J. Groenesteijn, M. J. de Boer, R. J. Wiegink, and J. C. Lötters, “Heavily-Doped Bulk Silicon Sidewall Electrodes Embedded between Free-Hanging Microfluidic Channels by Modified Surface Channel Technology,” *Micromachines*, vol. 11, no. 6, p. 561, 2020

4.1 Introduction

There is a great variety of MEMS devices and many of them have their own dedicated fabrication process [2]. The reason for this is that these devices typically have very specific requirements depending on their application. The same holds for the μ -Coriolis flow sensor. The features that are required for this type of device are:

- Suspended micro-channels
- Accessibility to the channels (in-/outlets)
- Electrodes on top of channels
- Readout structures (capacitive/resistive)

A fabrication process that satisfies these requirements was developed by Dijkstra et al. [3]. The process was based on a preceding technology to create buried micro-channels by Tjerkstra et al. [4]. The technology was then further developed by Groenesteijn et al. [5, 6] and the latest version is presented in this thesis. All devices shown in this thesis are fabricated using this technology, as described in this chapter. An elaborate process flow with all fabrication steps and details can be found in Appendix A.

4.2 Surface Channel Technology

The microfluidic devices presented in this dissertation are all fabricated based on the same technology, namely Surface Channel Technology (SCT). The name derives from the fact that the channels are created at the surface of a silicon wafer. Table 4.1 lists the used materials and their color as depicted in the figures in this chapter.

Table 4.1: Used materials

Material/Abbreviation	Description
Si	Silicon
SiRN	Silicon-rich silicon nitride
SiO ₂	Silicon dioxide
Photoresist	Positive UV Photoresist
Metal	Metal stack of Ta, Pt and Au
Al ₂ O ₃	Aluminium oxide

4.2.1 Forming micro-channels

Figure 4.1 shows the fabrication outline for forming the micro-channels. First, 500 nm of SiRN is deposited on a silicon wafer by means of Low Pressure Chemical Vapour Deposition (LPCVD). 500 nm of SiO₂ is deposited subsequently, once again by LPCVD (Figure 4.1a). This layer serves as a hard mask while etching the micro-channel to protect the SiRN layer. Photoresist is deposited and developed with a pattern of 5 by 2 μm slits. The pattern is then transferred into the SiRN and SiO₂ layers by directional plasma etching (Fig. 4.1b). The silicon underneath is etched semi-isotropically by a SF₆ plasma. During this etch, a (semi-)spherical shape is formed in the silicon under each slit. After a certain duration of etching, these (semi-)spheres connect together to form a channel (Fig. 4.1c). The shape and dimensions of the channel can be adjusted by using multiple rows of slits and/or changing the slit size and spacing [6].

4.2.2 Fluidic inlets

Once the micro-channels are formed, the fluidic inlets are created. This process starts with the deposition of another layer of SiO₂ (1.2 μm , LPCVD), see Fig. 4.2a. This layer serves as an etch stop during the inlet etch as well as to protect the micro-channels on the topside of the wafer while the backside is being processed. Photoresist is deposited on the backside of the wafer and developed to create a pattern of fluidic inlets. A directional plasma etch then transfers this pattern into the SiO₂ and SiRN layers (Fig. 4.2b). A deep reactive ion etch (DRIE, Bosch process) is then used to etch

Figure 4.1: Fabrication steps to form the surface micro-channels.

the fluidic inlet up to the SiO₂ channel wall (Fig. 4.2c). Subsequently, the fluorocarbons deposited during the inlet etch are removed by an O₂ plasma followed by stripping of any remaining photoresist (O₂/Ar plasma). Finally the SiO₂ layer is removed by means of wet etching in a 50% HF solution (Fig. 4.2d).

4.2.3 Forming the micro-channel wall

Now that the fluidic inlets have been created, the channel wall can be formed. this is done by depositing a thick layer of SiRN by means of LPCVD. Figure 4.3 shows the topside view (a) and backside view (b) after deposition of SiRN.

The layer should be thick enough to close the 2 μm wide slits and provide a structurally rigid channel roof. Thus, typically the layer thickness is chosen to be in the range of 1.7–2 μm .

4.2.4 Creating electrodes

After the SiRN micro-channel has been formed, the following step is to create electrodes for actuation and readout of the sensor. To reduce cross-talk in the actuation and readout of the sensor, it is important to ground the silicon substrate. Figure 4.4 shows the process of creating electrical contact between the silicon bulk and the surface of the wafer.

The contact windows are first defined in photoresist (Fig. 4.4a) after which the windows are etched into the SiRN layer below (Fig. 4.4b). The last step is to deposit a metal layer by sputtering (Fig. 4.4c). The metals in this layer can be chosen depending on the application. For example, for high temperature applications: platinum with a tantalum adhesion layer, for higher conductivity: gold with a chromium adhesion layer. In the case of a μ -Coriolis mass flow sensor, durability as well as high conductivity are important. Therefore, a combination of platinum and gold with a tantalum adhesion layer is used. Preceding the deposition of the Ta/Pt/Au stack is sputtering of a thin

Figure 4.2: Fabrication steps to create fluidic inlets to the micro-channels.

4

Figure 4.3: Forming of the micro-channel walls by LPCVD of SiRN.

Figure 4.4: Fabrication steps to create electrical contact between the silicon bulk and surface of the wafer.

Figure 4.5: Fabrication steps to pattern and cap electrodes.

layer (5 nm) of platinum. This is then annealed at 400 °C for 30 minutes in a N_2 environment to form a thin conductive layer of PtSi for good Ohmic contact to the silicon bulk. Tantalum, platinum and gold (10 nm Ta, 20 nm Pt, 200 nm Au) are then deposited consecutively by sputtering, without breaking vacuum.

After electrical contact to the bulk has been realized, the electrodes can be patterned. A photoresist layer is deposited and developed to pattern electrodes (Fig. 4.5a). The electrodes are then etched by means of ion beam etching (Fig. 4.5b). The metal layer is then covered by a layer of aluminium oxide (Al_2O_3 , 400 nm) by electron beam evaporation. This layer serves as capping for the metals such that they are more robust to high temperatures and as a hard mask during the release etch, where the channels are etched free from the silicon bulk. After the metals have been capped, they are annealed at 400 °C for 4 h in a N_2 environment to increase durability and resilience to high temperatures.

Figure 4.6: Partially etching an opening to the bond pads.

4.2.5 Releasing the micro-channels from the Si bulk

Now that the electrodes have been defined, the next and final step is to release the micro-channels by etching them free from the silicon bulk. This step consists of multiple sub-steps. Before the etch windows for the release etch are created, the bond pads which are used for electrical connection between the micro-chip and external electronics have to be made accessible. This is done by etching an opening into the Al_2O_3 capping. Figure 4.6 shows this process. Photoresist is first deposited and developed (Fig. 4.6). To make the bond pads accessible for external contact, the Al_2O_3 is then etched partially (Fig. 4.6). The reason for this is that the bond pads have to be protected during the following etching steps. However, after the micro-channels are released, photo lithography using photoresist is no longer possible. There is no way to define an opening to access the bond pads in that stage. Thus, a thin layer of 50-80 nm of Al_2O_3 is left, so that the bond pads can be opened up at the end of the entire process.

Before the channel can be released from the silicon bulk, etch windows need to be defined. Additionally, comb readout electrodes are shaped in the same step as well. Photoresist is deposited and developed, defining the etch windows as well as comb fingers for the readout electrodes (Fig. 4.7a). The Al_2O_3 layer is etched with a directional plasma etch (Fig. 4.7b) after which the fingers of the comb readout electrodes are formed in the metal layer by IBE (Fig. 4.7c). The reason why the comb electrodes are defined in this step and not in the previous IBE step (4.5b) is the fact that they are now perfectly aligned to the release etch windows. If done in the previous IBE step, there could be misalignment between the comb shape in the metal layer and the SiRN layer. A small misalignment can have a significant effect, since the comb fingers and gap between them are only a few micron wide.

Now that the comb readout electrodes have been shaped, the micro-channels can be released. Figure 4.8 shows the final steps in the SCT fabrication process. First, the SiRN layer is etched with Al_2O_3 as hard mask (Fig. 4.8a). The photoresist is etched in this process as well. Then the silicon bulk is semi-isotropically etched using an SF_6

(a) Patterning release etch windows and readout electrodes.

(b) Directional plasma etching of Al_2O_3 .

(c) IBE of comb-shaped readout electrodes.

Figure 4.7: Fabrication steps to open etch windows for the release etch and defining comb electrodes.

(a) Directional plasma etching of SiRN layer

(b) Releasing the micro-channels by plasma etching of Si.

(c) Etching Al_2O_3 layer to reveal bond pads.

Figure 4.8: Releasing of the micro-channels from the silicon bulk.

plasma (Fig. 4.8b). This is done in cycles of 5 minutes to prevent overheating of the micro-channels. The silicon is etched to a depth of 250 μm , approximately halfway through the wafer. This allows the wafer to be broken along grooves etched in between devices to separate it into chips. Before doing this, the entire surface of the Al_2O_3 layer is etched by the thickness of the thin layer left on the bond pads. (50-80 nm left in Fig. 4.6b subtracted by the amount etched in step (a) and (b) from Fig. 4.8). This will reveal the bond pads while leaving the capping intact on the rest of the surface. Figure 4.9 shows some SEM images of structures made by SCT.

(a) Micro-channel formed in Si wafer by semi-isotropic SF_6 plasma etching.

(b) Released SiRN micro-channel with metal tracks on top.

(c) Patterned metal structures in proximity to a released micro-channel.

(d) Comb-shaped electrodes for capacitive readout of moving structures.

Figure 4.9: SEM images of structures formed by SCT.

4.3 Integration of Doped Silicon Structures

The conventional SCT process allows for the integration of metal electrodes on and surrounding the micro-channels. This can be used to realize heaters and/or temperature sensors[7], readout capacitors [8] and strain gauges [9, 10] among others. Since thin film metal heaters can become unstable at high temperatures [1], it can be beneficial to implement doped silicon heaters. In the case of the μ -Coriolis mass flow sensor, a capacitive readout has certain limitations (discussed in detail in Chapter 6). Therefore a resistive readout based on strain gauges is a good alternative [10]. Doped silicon strain gauges implemented on top of the micro-channel could improve the performance of such a readout greatly. Implementing doped silicon into the SCT process would give great benefits. Three approaches to achieving this are presented in the following subsections:

- Implementing doped Si strips in the conventional SCT
- Depositing doped poly-silicon onto the SiRN micro-channel
- Using a SOI wafer with a mono-crystalline doped Si device layer

4.3.1 Implementing doped Si strips in conventional SCT

In [1], a method is presented to implement doped mono-crystalline silicon strips in the conventional SCT process without the need to adapt any of the fabrication steps. This is achieved by clever adaptation of the mask design. Typically, the micro-channel slits and release openings are designed such that where the micro-channel should be released, all silicon around and underneath the channel is etched away. However, when the release opening is too far from the micro-channel or when multiple micro-channels are directly next to each other, some silicon can remain attached to the micro-channel. One can control this by adapting the duration of the release etch or by optimizing the dimensions and distance of the release windows. Another effect is that when two rows of slits that define a micro-channel are in close proximity to each other, the channels start to overlap and merge into one. By controlling the duration of the channel etch and the distance between the rows of slits precisely, some silicon can be trapped. Figure 4.10 shows how these two methods can be used to embed a doped silicon strip in a (set of) micro-channel(s).

The first method starts with defining two rows of slits at a distance slightly larger than the desired channel radius. Then, two parallel channels are etched such that a small amount of silicon (a few μm in width) is left in between them (4.10a.1). In the following step, the SiRN channel wall is deposited (4.10b.1). Finally the channel is released, leaving a silicon strip in between the two micro-channels (4.10c.1). Figure 4.11a shows an SEM image of the resulting structure. The dimensions of the strip can be determined by precisely controlling the distance between the rows of slits, the

(a.1) Etching two closely adjacent channels.

(a.2) Etching two micro-channels until they overlap and merge.

(b.1) Deposition of SiRN channel wall.

(b.2) Deposition of SiRN channel wall.

(c.1) Release etch with optimized duration.

(c.2) Doped Si strip is protected from the release etch by the SiRN layer.

Figure 4.10: Two methods to embed a doped silicon strip in a (set of) micro-channel(s) using conventional SCT.

(a) Doped Si strip in between two non-overlapping micro-channels. (b) Doped Si strip in micro-channel formed by two merged channels.

Figure 4.11: SEM images of a doped Si strip embedded in a micro-channel using conventional SCT (resulting from the method in Fig. 4.10).

distance and dimensions of the release etch windows and the channel and release etch duration. The second method starts with defining two rows of slits at a distance slightly smaller than the desired radius of a single channel. Two parallel micro-channels are then etched until they start to overlap and merge into one (4.10a.2). This leaves a small amount of silicon at the top of the channel. SiRN is deposited to form the channel wall, trapping the silicon strip (4.10b.2). Finally, the channel is released by semi-isotropically etching the surrounding silicon (4.10c.2). The trapped silicon strip is protected during the release etch by the SiRN channel wall and thus remains. figure 4.11 shows an SEM image of the resulting structure. This method does not rely on precise control of the release etch and can thus be easier to implement. However, the distance between the rows of slits and the channel etch duration are critical. Typically, the channel radius is chosen to be in the range of 40-100 μm . A 5% deviation in the channel radius can already make the difference between a silicon strip being trapped or not, or even between it being present or etched away completely. Advantages of both variants of this method are the fact that it can be implemented in conventional SCT and that the silicon strips can be embedded on the channel sidewall (beneficial for heating and temperature sensing). However, the main disadvantage is that it requires very precise control and relies heavily on consistency and reliability of multiple fabrication steps and uniformity over the wafer. Another disadvantage is that freedom of design in the sense of different structures and dimensions is very limited.

4.3.2 Adapted SCT: Doped poly-silicon structures on top of the micro-channel

Implementation of doped silicon strips by adapting the SCT process can lead to much improved control of dimensions and freedom of design. This can be achieved by including poly-silicon deposition, doping and patterning steps. Figure 4.12 shows the process outline which can be used to realize doped poly-silicon strips on top of surface

Figure 4.12: Embedding doped poly-silicon structures on top of surface micro-channels by adaptation of the SCT process.

channels.

The process up to and including deposition of the channel wall (Fig. 4.3) remains the same as for the conventional SCT process. A poly-silicon layer is then deposited and doped by deposition of a diborate (B_2O_5) layer followed by a drive-in step (Fig. 4.12a). The chosen doping level depends on the application of the poly-Si structures. For temperature sensors, the TCR (temperature coefficient of resistance) should be optimized. For strain gauges, there is a trade-off between gauge factor, resistivity and TCR. After doping the poly-silicon layer, it is patterned by directional plasma etching (Fig. 4.12b). Next, the structures are capped with a SiO_2 layer by means of LPCVD (Fig. 4.12c). This layer serves as protection for the poly-silicon structures during IBE of the metal electrodes in a later step as well as during the release etch. Openings are then etched through the SiO_2 layer to the poly-silicon structures (Fig. 4.12d). Subsequently, a metal layer (e.g. Al) is deposited by sputtering and patterned by means of IBE (Fig. 4.12e). This step serves to form electrical contact to the poly-silicon structures. After this step, there is an option to cap the metal structures with an Al_2O_3 layer as done in Fig. 4.5c. However, step coverage might become an issue, depending on the chosen thickness of the poly-silicon layer. It may be required to cap the metals, since exposed

Figure 4.13: Patterning device and BOX layers of a SOI wafer to form highly doped mono-crystalline structures. SiRN and SiO₂ are then consecutively deposited to serve as the channel roof material and hard mask during the channel etch respectively.

metals are not always allowed in certain plasma etching machines. The final step is to etch windows through the SiO₂ and SiRN layers and releasing the micro-channels (Fig. 4.12f). During this etch, the SiO₂ will serve as a hard mask. When capping the metals with Al₂O₃, this layer can serve as hard mask just as in Fig. 4.8.

There are some advantages and disadvantages of this method. First of all, after deposition of the channel wall, any wet processing is preferably avoided since liquids can enter the channel inlets on the backside of the wafer through capillary action. This can be difficult to evaporate once again, since the topside of the channel is closed. The deposition and doping of the poly-silicon layer requires multiple preceding/succeeding wet processing steps. Another disadvantage is that poly-silicon has a considerably lower gauge factor than mono-crystalline silicon and it depends strongly upon the structure of the material and thus on processing [11]. A significant advantage of this method is that the thickness, dimensions and location of the poly-silicon structures can be defined entirely to suit the application. This is a huge increase in freedom of design compared to the method described in section 4.3.1.

4.3.3 Adapted SCT: Mono-crystalline silicon structures using silicon-on-insulator wafer with a doped device layer

Ideally, one would like to utilize the relatively higher gauge factor of mono-crystalline silicon while keeping the design freedom which comes with using LPCVD poly-silicon. This can be achieved by using a silicon-on-insulator wafer with a highly doped device layer. Structures can be patterned in the device layer and incorporated in the micro-channels. These can then serve as strain gauges, heaters and temperature sensors, among others.

Figure 4.13 shows the first steps in the process, to pattern the highly doped mono-

Figure 4.14: Formation of the micro-channel. During the channel etch, mono-Si structures are protected by SiRN and SiO₂ (BOX) layers. Mono-Si structures are encapsulated by another layer of SiRN, embedding them in the roof of the micro-channel.

crystalline silicon structures and cap them consecutively with SiRN and SiO₂. The process starts with a SOI wafer with a thin highly doped device layer (Fig. 4.13a). Structures are patterned in the device and BOX layers using a directional plasma etch (Fig. 4.13b). SiRN and SiO₂ layers are then conformally deposited by means of LPCVD. This is the same step as in Figure 4.1a. The SiRN layer will form the roof of the micro-channel whereas the SiO₂ layer is used as a hard mask during the channel etch. From here on out, the process is in principal identical as the standard SCT described in Section 4.2.

The next step is to form the micro-channel, depicted in Figure 4.14. Just as for the standard SCT, slits are etched into the SiO₂ and SiRN layers by a directional plasma etch. Then the channel is etched semi-isotropically with an SF₆ plasma (Fig. 4.14a). During this step, the mono-crystalline silicon structures are protected from the top and sides by SiRN and from the bottom by SiO₂ (BOX layer). After the channel has been formed, fluidic inlets are created following the same steps as in Figure 4.2. Once the inlets have been formed, SiO₂ is stripped (Fig. 4.14b) and the SiRN channel wall is deposited by LPCVD (Fig. 4.14). In this step, the mono-Si structures are encapsulated and embedded in the micro-channel roof.

The final steps in the process are to create electrodes for actuation, readout, etc. and release the micro-channel. To make electrical contact with the mono-Si structures, openings are first etched in the SiRN layer (Fig. 4.15a). Metal is deposited by means of sputtering and subsequently patterned to form electrodes (Fig. 4.15b). The electrodes are then capped with an Al₂O₃ layer (Fig. 4.15c), which serves as a hard mask for the channel release etch. Finally, openings are etched in the Al₂O₃ and SiRN layers through which the bulk silicon is etched to release the micro-channel. The steps shown in Figure 4.15 are identical to those displayed in Figures 4.4 to 4.8. Some sub-steps

Figure 4.15: Forming electrical contact to the mono-Si structures followed by capping of the metals and release of the micro-channel.

4

are not shown, since they are the same as for the standard SCT and are not affected by the addition of the mono-Si structures. As shown, this fabrication method allows for the embedding of highly doped mono-crystalline silicon structures in the roof of the micro-channel with minimum adjustment to the standard SCT process.

4.4 Conclusion

Multiple variants of the SCT process have been discussed in this chapter. The conventional SCT can be used to fabricate suspended SiRN micro-channels with electrodes on the channel roof. This allows for the realization of various types of micro-fluidic devices such as μ -Coriolis mass flow sensors, thermal flow sensors, pressure sensors, etc. An adapted SCT process in which doped Si structures are embedded in the micro-channel roof is described as well. This adaptation extends the design possibilities since it can be used to create heaters, temperature sensors and piezo-resistive strain gauges all in close proximity to the fluid traveling through the micro-channels. All of the devices presented in this thesis are fabricated using the technology described in either Section 4.2 or 4.3.3. As aforementioned, an elaborate process flow with all fabrication steps and details can be found in Appendix A.

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5

Actuation Methods

In this chapter[†], various possible transduction principles for actuation of MEMS resonators are introduced and compared qualitatively. Subsequently, the Lorentz actuation method for actuation of a μ -Coriolis mass flow sensor is described in more detail. Then, improved magnet configurations are presented to achieve stronger magnetic fields for more efficient actuation.

[†]This chapter is based on publication [1]:

- T. V. P. Schut, Y. P. Klein, R. J. Wiegerink, and J. C. Lotters, “Magnetic field strength improvement for Lorentz actuation of a μ -Coriolis mass flow sensor,” *Microelectronic Engineering*, vol. 226, p. 111236, 2020

5.1 Introduction

The operating principle of a μ -Coriolis mass flow sensor (μ -CMFS) has been discussed in detail in Chapter 3. The operation of the sensor is entirely dependent on the fact that the micro-channel structure is consistently vibrating at resonance. To achieve this, a certain actuation force has to be exerted on the structure. Additionally, vibration is only desired in one particular vibrational mode and thus the actuation force should be exerted in a specific position and direction. There are various methods to achieve this, which are used in numerous types of MEMS resonators [2]. Various transduction principles and how they can be utilized for actuation of a μ -CMFS will be discussed in the following section. Afterwards the preferred method will be described in more detail and improvements are presented.

5.2 Transduction principles

A μ -CMFS can be simply viewed as a MEMS resonator. The basic principle of a MEMS resonator is that a mechanical vibration is induced by means of an electrical signal. There are multiple transduction principles that can be utilized to achieve this. The most common methods of transduction for actuation are:

- Electrostatic
- Piezoelectric
- Electrothermal
- Electromagnetic

Each method will be briefly discussed and their advantages and disadvantages will be highlighted.

5.2.1 Electrostatic

Electrostatic actuation relies on the fact that a voltage applied between two conducting structures induces a force pulling them together, assuming they are free to move. Figure 5.1 shows a schematic representation of this. This principle is broadly applied in MEMS resonators due to its simplicity of integration in micro-machining. Doped silicon is often used as electrode material, since typically the structure that is brought into resonance is made of silicon. Another material that can be used is metal, that can be deposited and patterned as desired. Electrostatic actuation can be used to induce both in-plane and out-of-plane motion. For in-plane motion, an electrode structure that is commonly used is a comb drive actuator [3]. These actuators are versatile and can be used to induce controlled motion in a specific direction. Motion is typically induced in a direction parallel to the interdigitated plates and thus the gap between the plates remains constant. This means that relatively large displacements

Figure 5.1: Schematic representation of electrostatic actuation. A voltage v_{act} applied across two conductive structures generates charges, which pull on each other, inducing a force F .

can be achieved compared to parallel plate capacitors, where it is limited by the pull-in effect [4]. Out-of-plane MEMS resonators are less common. This is due to the nature of silicon micro-machining, where structures can be easily patterned and defined in the horizontal plane, but not in the vertical plane. For out-of-plane actuation, a capacitor structure with a vertical gap needs to be realized. However, this gap then directly defines the maximum displacement. Alternatively, a regular comb drive structure can be used if there is an initial vertical gap between the two comb electrodes (e.g. due to residual stress in spring suspension).

To summarize, the main advantages of electrostatic actuation are versatility and ease of implementation. A disadvantage is the relatively small capacitance associated with micro-structures. The capacitance can be increased by narrowing the gap between electrodes or increasing electrode area. However, this may not always be straightforward. Another way of increasing the induced displacement is by increasing the actuation voltage. However, this means that vacuum packaging may be necessary to prevent electrical breakdown. This complicates the fabrication process.

Electrostatic actuation for a μ -CMFS is complicated due to the disadvantages aforementioned. The fabrication process necessary for creating the micro-channels (SCT) limits the freedom of design for actuation electrodes. As described in Chapter 4, the silicon around the micro-channel is entirely etched away to release it. This means that doped silicon electrodes attached to the micro-channel cannot be realized without significantly complicating the fabrication process. Metal electrodes can be defined on top of the micro-channel or attached membranes. However, the small thickness of this layer significantly limits the (vertical) electrode area resulting in a very small capacitance. This means that very large voltages would be required to achieve reasonable displacement of the micro-channel. Since vacuum packaging of the device would be costly and complex, electrostatic actuation is not considered as a viable method for the μ -CMFS devices presented in this thesis.

(a) A structure consisting of a stack of piezoelectric and conductive layers. (b) When a voltage is applied over the piezoelectric material, it contracts.

Figure 5.2: Schematic representation of how the piezoelectric effect can be used to induce bending.

5.2.2 Piezoelectric

As aforementioned, piezoelectric materials can be used for actuation as well as sensing. Figure 5.2 shows a schematic representation how a piezoelectric material can be used to induce motion. When applying a voltage over a piezoelectric layer, it expands or contracts, depending on the polarity. This can be used to induce bending. The main issue with integrating piezoelectric materials into MEMS resonators is the technical difficulty to incorporate these materials into conventional silicon micro-machining processes. However, first steps have been made to apply these materials for the actuation of a μ -CMFS by Zeng et al. [5, 6] (IDS group, University of Twente). The use of an external piezo ceramic results in high noise and possible difficulties with packaging [5]. After this proof of principle, more work was done to integrate a thin film piezoelectric material [6]. Relatively high displacements were achieved with actuation voltages in the range of 0.1–2V. However, some non-linearity was observed in the relation between the displacement and actuation voltage. Additionally there are issues with degradation of the piezoelectric thin films over time when being actuated. The quality of these thin films need to be improved before they can be used as a viable method of actuation for a μ -CMFS.

5.2.3 Electrothermal

Most micro-machining processes allow for patterning of conductors (e.g. doped silicon or metal) on/in the micro-structures. Since a conductive material is all that is required for thermal actuation, it is an attractive and straightforward method. Figure 5.3 shows how electrothermal actuation can be applied to a suspended beam to induce bending. By passing an alternating current through a conductive material, a dynamic heating power is generated through resistive heating. This power results

(a) A structure consisting of a conductive track (e.g. metal) patterned on top of a beam (e.g. silicon). (b) When a voltage is applied over the track, it expands at a different rate than the underlying beam material, inducing bending.

Figure 5.3: Schematic representation of how electrothermal actuation can be used to induce bending in a suspended beam.

in local dynamic expansion of the conductive material and the structure on which the conductor resides. A mismatch between the thermal expansions coefficient of the conductor and the structure on which the conductor resides induces mechanical stress resulting in bending (see Fig. 5.3b). By matching the frequency of the alternating driving current to the resonance frequency of one of the mechanical vibrational modes, the micro-structure can be actuated and brought into resonance. The main advantages of thermal actuation are relatively high excitation forces at low voltages and simplicity of integration in micro-machining [7]. However, the high power consumption make it an undesirable method in many applications. In addition, large temperature differences are induced in the heater. For MEMS resonators this can affect suspension stiffness. This influences the resonance frequency and in case of inertial sensors, the sensitivity to the measured parameter. Specifically for a μ -CMFS, flow through the micro-channel would change the temperature distribution affecting the actuation and creating additional dependence on flow rate. Therefore this method of actuation is not practically considered in the research presented in this thesis.

5.2.4 Electromagnetic

Electromagnetic actuation exploits the (Lorentz) force exerted on charges when they flow through a conductor in the presence of a magnetic field. The concept will be explained in more detail in the next section. Electromagnetic actuation and sensing is widely applied on the macro scale, e.g. motors or generators. On the micro-scale, electromagnetic transduction is mainly used for actuation, since induction voltages induced by motion of micro-structures are typically very small. Electromagnetic actuation of MEMS requires external magnets, which are often large in comparison to the micro-chip containing the MEMS device being actuated. Thus for applications where miniaturization is crucial, this technique is not suitable. However, there are

advantages of using electromagnetic actuation, such as: ease of integration in micro-machining, relatively large displacements and low actuation voltages (compared to electrostatic actuation). For these reasons and the disadvantages of the previously discussed techniques, Haneveld et al. used electromagnetic actuation for their highly sensitive μ -CMFS [8]. All devices presented in this thesis stem from the device presented by Haneveld et al. and use electromagnetic/Lorentz actuation. The next section describes Lorentz actuation of a μ -CMFS in detail.

5.3 Lorentz actuation of a μ -Coriolis flow sensor

For the advantages discussed in the previous section, Lorentz actuation is the preferred method for actuation of the μ -Coriolis mass flow sensors presented in this thesis. Figure 5.4 shows a μ -CMFS chip mounted on a PCB with external permanent magnets. This type of setup has been used for these devices ever since Haneveld presented the first electromagnetically actuated μ -CMFS [8]. Alveringh et al. presented an improved setup where external magnets were no longer glued to the chip PCB but to a larger PCB to which chips can be connected in a plug and play fashion [9].

To explain how the μ -CMFS is actuated using Lorentz forces, this force first needs to be described. The Lorentz force exerted on flowing charges is defined as:

$$\mathbf{F}_L = q\mathbf{v} \times \mathbf{B} \quad (5.1)$$

Where q is the charge, $\mathbf{v}$ the velocity vector of the flowing charges and $\mathbf{B}$ the magnetic field vector. For a current $\mathbf{i}$ flowing through a conductor with length L , residing in a magnetic field, $\mathbf{F}_L$ can be written as:

$$\mathbf{F}_L = L\mathbf{i} \times \mathbf{B} \quad (5.2)$$

Figure 5.5 shows a schematic representation of the generation of Lorentz force in

Figure 5.4: μ -Coriolis mass flow sensor chip mounted on PCB with external permanent magnets for generating a magnetic field.

Figure 5.5: Schematic representation of the generation of Lorentz force on a conductor. Charges flowing with velocity v or a current i through the conductor, generates a Lorentz force perpendicular to the flow direction and orientation of the magnetic field B .

Figure 5.6: An alternating current i (indicated by yellow arrows) is fed through a metal track on top of the micro-channel. In combination with a magnetic field B , this generates Lorentz forces in the channel sections perpendicular to B . These forces actuate the sensor in the desired mode, depending on the particular design and orientation of the magnetic field.

such a conductor. In the case of the μ -CMFS, a current i flows through a metal track on top of the micro-channel. Since the magnitude of F_L depends on the orientation of the magnetic field relative to the micro-channel, force can be exerted onto specific sections. Figure 5.6 shows how this can be used for actuation of a μ -CMFS. An alternating current i flows through a metal track on top of the channel. Lorentz forces (F_L) are generated in channel sections where the current flows perpendicular to the magnetic field B . By matching the frequency of i to the mechanical resonance frequency of a vibrational mode and choosing the appropriate orientation of B , one can actuate the channel loop in either the *Twist* mode (Fig. 5.6a) or *Swing* mode (Fig. 5.6b).

During operation, the μ -CMFS is actuated at one of the resonance modes. The resonance frequency is dependent on the flow medium, pressure, temperature etc. Therefore, to keep the sensor always at resonance, an actuation loop can be used. Figure 5.7 shows a schematic representation of the sensor and closed amplification loop which is used to bring it into resonance (in the *Twist* mode). There are two metal tracks on top of the microfluidic channel of the sensor. An actuation voltage v_{act} is supplied to one track, generating a current which induces Lorentz forces as

Figure 5.7: Schematic representation of the actuation loop which is used to bring the μ -Coriolis mass flow sensor into resonance. Where v_{EMF} is the generated electromotive force and v_{act} is the actuation voltage used to generate Lorentz forces in combination with a magnetic field $\mathbf{B}$.

depicted in Fig. 5.6. Due to vibration of the channel, an electromotive force (EMF) is generated on the other track. This EMF (v_{EMF}) is then fed into an amplifier circuit. The amplified signal is fed back to the actuation track of the sensor. This closed loop circuit drives the sensor into resonance.

The vibrations of the sensor during actuation (in the *Twist* mode) can be described by a second order differential equation:

$$J\theta''(t) + R\theta'(t) + K\theta(t) = \tau_{\text{ext}}(t) \quad (5.3)$$

Where J is the moment of inertia, R is the damping coefficient, K is the torsional spring constant, $\tau_{\text{ext}}(t)$ is the external torque induced by Lorentz forces and $\theta(t)$ is the actuation mode angle (θ_t in Fig. 5.6a). By rewriting Equation 5.2 and substituting $i = v_{\text{act}}/R_{\text{track}}$, $\tau_{\text{ext}}(t)$ can be approximated as:

$$\tau_{\text{ext}}(t) = \frac{BLd}{R_{\text{track}}} \cdot v_{\text{act}}(t) \quad (5.4)$$

Where B is the magnetic field strength, L is the length of metal track where Lorentz forces are generated, d is the distance between these track sections and R_{track} is the resistance of the metal track (see Fig. 5.7). The induction voltage generated on the second metal track is dependent on the velocity of the channel in the z direction:

$$v_{\text{EMF}} = 2BL \cdot z'(t) \quad (5.5)$$

This can then be related to the actuation angle θ (for small value of θ):

$$z(t) = \frac{d}{2} \cdot \sin(\theta(t)) \approx \frac{d}{2} \theta(t) \quad (5.6)$$

$$\begin{aligned} v_{\text{EMF}}(\omega) &= 2BL \cdot j\omega Z(\omega) \\ &= BLd \cdot j\omega \theta(\omega) \end{aligned} \quad (5.7)$$

$$\theta(\omega) = \frac{v_{\text{EMF}}(\omega)}{j\omega BLd} \quad (5.8)$$

Now converting Equation 5.3 to the frequency domain and substituting Equations 5.4 and 5.8:

$$\begin{aligned} J(j\omega)^2 \theta(\omega) + Rj\omega \theta(\omega) + K\theta(\omega) &= \tau_{\text{ext}}(\omega) \\ \theta(\omega) \cdot \left((j\omega)^2 J + j\omega R + K \right) &= \frac{BLd}{R_{\text{track}}} \cdot v_{\text{act}}(\omega) \\ \frac{v_{\text{EMF}}(\omega)}{j\omega BLd} \cdot \left((j\omega)^2 J + j\omega R + K \right) &= \frac{BLd}{R_{\text{track}}} \cdot v_{\text{act}}(\omega) \end{aligned} \quad (5.9)$$

Finally, the transfer function from v_{act} to v_{EMF} becomes:

$$\begin{aligned} \frac{v_{\text{EMF}}}{v_{\text{act}}} &= \frac{j\omega B^2 L^2 d^2}{R_{\text{track}} \cdot \left((j\omega)^2 J + j\omega R + K \right)} \\ &= \frac{j\omega B^2 L^2 d^2 / K}{R_{\text{track}} \cdot \left(1 + j \frac{\omega}{\omega_0} \frac{1}{Q} + \left(j \frac{\omega}{\omega_0} \right)^2 \right)} \end{aligned} \quad (5.10)$$

Where ω_0 is the resonance frequency and Q is the quality factor:

$$Q = \frac{\sqrt{KJ}}{R} \quad (5.11)$$

$$\omega_0 = \sqrt{\frac{K}{J}} \quad (5.12)$$

Equation 5.10 shows that the transfer can be optimized by decreasing R_{track} or increasing B . The other parameters are correlated and typically optimized for flow sensitivity. Furthermore, R_{track} is dependent on the thickness, width and material of the conductive track on the micro-channel, which are defined by the fabrication process. The most efficient way to increase the transfer is thus to increase B . The next section discusses improved magnet configurations to realize this.

5.4 Improved magnet configurations

Groenesteijn et al. [10] have presented improved magnet configurations for actuation of a μ -Coriolis mass flow sensor. Miniature magnets are placed on-chip and with their magnetic poles all in the same direction. This showed improved field strengths but has large disadvantages, since magnets have to be somehow mounted on-chip without damaging the sensor device itself and/or without detaching over time. To further improve upon the research presented by Groenesteijn et al., new magnet configurations have been realized which give a large increase in the generated magnetic field strength.

5.4.1 Configurations

Figure 5.8 shows three configurations that are chosen and compared. The first uses the field between two parallel cylindrical magnets (Fig. 5.8a), the second uses the field within a Halbach ring based configuration (Fig. 5.8b) and the third uses the field in close proximity to a cubic magnet (Fig. 5.8c). The Halbach configuration is a special magnet arrangement which reduces the stray field while enhancing the field at the ring-center. The Halbach magnet configuration has been invented by K. Halbach [11, 12] and is still used today, e.g. for several machines [13] or mobile NMR applications [14].

The cylindrical configuration consists of two cylindrical, axially magnetized magnets (radius: 3mm, length: 13mm, material: NdFeB N42, obtained from www.supermagnete.nl) arranged parallel with a distance of 19mm between them. The chip has been placed in the middle with an offset of 3mm in the z -direction, as shown in Fig. 5.8a. Our Halbach-ring (Fig. 5.8b) consists of 8 magnets (material NdFeB N45, obtained from www.supermagnete.nl). 4 of them, placed in the corners, cubic with an edge length of 7 mm, the other ones are 7x7x14 mm. The cubic configuration (Fig. 5.8c) consists of a cubic magnet with an edge length of 12 mm (material NdFeB N45, obtained from www.supermagnete.nl) placed parallel, directly above the chip.

5.4.2 Device and simulation

Figure 5.9 shows an SEM image of the sensor used to characterize the magnet configurations. Multiple metal tracks can be seen on the channel, for capacitive readout and temperature measurement among others. In this paper only two metal tracks are actually used: one to provide the current for actuation of the sensor and one for measurement of the induction voltage generated by vibration of the channel. In all magnet configurations in Figure 5.8, the channels of interest have a length of 2.25 mm and are located at $x = -2$ mm and 2 mm (see Fig. 5.9).

The magnetic field strengths for the different magnet configurations are simulated. The simulations have been done using CADES simulation software, entirely described

Figure 5.8: μ -Coriolis flow sensor chip on PCB with different magnet configurations. The poles of the magnets are defined as: $\begin{bmatrix} \text{S} & \text{N} \end{bmatrix}$.

Figure 5.9: SEM image of the sensor type used for all simulations and measurements in this paper.

by Delinchant et al. [15]. Magnetic interactions are modelled with the MacMMems tool, which uses the Coulombian equivalent charge method to generate a semi-analytic model. This model is used by the CADES framework (component calculator, component optimizer) to calculate the designs. Equation 5.10 is used in combination with the simulated magnetic field strengths to generate transfer curves for the various magnet configurations. These are compared with transfer curves measured using a gain-phase analyser. The measurement setup is effectively as shown in Figure 5.7 where the amplifier circuit is then substituted by an HP 4194A gain-phase analyser.

5

5.4.3 Results and discussion

The magnetic field strength of three different magnet configurations have been simulated. For all configurations the y -field shows a uniform profile along the channel, see Figure 5.10. The Halbach configuration reaches a strength 0.3T, the single cuboid configuration 0.25T. The cylindrical configuration has the lowest field with 0.05T.

The x -field of the configuration along the y -axis is displayed in Figure 5.11. The field is anti-symmetric around $x = 0$ due to symmetry of the magnet configurations. This means additional Lorentz forces are generated in the channel section perpendicular to the y -axis. Due to the anti-symmetry of the x -field this will cause additional vibration in the actuation mode. The Coriolis mode is unaffected by this field.

The stray magnetic field strengths of the three configurations were simulated at a distance of 10mm from the outer magnet surface, see Table 5.1. The Halbach configuration shows the lowest stray field strength of 0.006T, compared to 0.025T for the cylindrical configuration and 0.081T for the cubic configuration.

The magnitude and phase transfer between the actuation voltage v_{act} and induction voltage v_{EMF} was measured for the three magnet configurations in Figure 5.8. The

Figure 5.10: Simulated magnetic field strength in the y -direction along the 2.25 mm side channel section (figure 5.9).

Table 5.1: Stray magnetic field strength along y -axis of the three configurations at a distance of 10mm from the outer magnet surface.

Configuration	Stray field [T]
Cylindrical	0.025
Halbach	0.006
Cube	0.081

magnitude and phase transfer are displayed in Figures 5.12a and 5.12b respectively. The analytical curves are generated using equation 5.10 in combination with the simulated magnetic field strengths for the three configurations (see Figure 5.10). For both figures the x -axis is centered around the resonance frequency, which lies around 3.8kHz varying slightly with temperature, pressure and from chip to chip.

The measured magnitude transfers for the 3 configurations are; Halbach: -59.1dB, Cubic: -58.0dB, Cylindrical: -84.6dB. As expected, both the Halbach and single cube configuration show a higher magnitude transfer than the cylindrical configuration, around 26.5dB. For the cubic configuration, this corresponds well with the analytical curve. The measured value is slightly higher than simulated. This could be due to deviation of the position of the magnet along the z -axis where it is actually closer to the sensor chip than simulated. Other causes could be a deviation in the quality factor or resonance frequency of the μ -Coriolis mass flow sensor. For the Halbach

Figure 5.11: Simulated magnetic field strength in the x -direction along the top 4 mm channel section (figure 5.9).

configuration, the simulated value of the magnitude transfer lies 4.4 dB higher than the measured value. This is most likely due to misalignment of the magnets in the Halbach configuration. Since the fields of the magnets in the configuration oppose each other, it is difficult to accurately place and align them. Misalignment of the sensor chip to the centre of the magnets along the z -axis (where the magnetic field is the strongest) could be another cause of the lower measured transfer.

The phase transfer of all the configurations should in principle be a curve going from $+90^\circ$ to -90° , eventually going back to $+90^\circ$ due to cross-talk via parasitic capacitance between the actuation and induction tracks. At resonance, the phase shift should be 0° . However, since the signal in case of the cylindrical configuration is so small, the phase shift is heavily affected by cross-talk (see Figure 5.12b). The phase shift does not even reach 0° . The EMF signal is too small to overcome the cross-talk via parasitic capacitance between the two metal tracks. We can see a similar effect in the magnitude transfer of the cylindrical configuration where the resonance peak barely reaches out above the noise floor. To stably bring the sensor into resonance, it is critical that this 0° point can be detected. For the other two configurations, the phase transfer is as expected. For these, the sensor should be able to be brought stably into resonance with an amplification loop with relatively low gain. This can enhance the signal-to-noise ratio of the overall sensor. Especially for fluid density measurements this can improve the accuracy significantly, since the resonance frequency is directly related to the

(a) The measured magnitude transfers for the 3 magnet configurations.

(b) The measured phase transfers for the 3 magnet configurations

Figure 5.12: Magnitude and phase transfers between the actuation voltage v_{act} and induction voltage v_{EMF} .

density of the fluid flowing through the channel. Furthermore, the actuation current can be lower while still achieving a higher displacement amplitude. This can be beneficial for reducing resistive heating of the micro-channel. Heating up of the micro-channel induces stress, resulting in small changes in the mechanical behaviour of the μ -CMFS. By reducing resistive heating, temperature effects can be limited. Groenesteijn et al. [16] showed the relation between the dissipated actuation power and the resulting increase in temperature of the micro-channel of the μ -CMFS for a flow range of 0–1 g h⁻¹. In this range, the temperature increase is 0.05–0.11 °C/mW of dissipated power. For the device in Fig. 5.9, around 4 mW of power is dissipated to reach a typical actuation angle of $\theta_a = 5$ mrad (θ_t in Fig. 5.6). This would mean a temperature increase of 0.2–0.44 °C. For the presented Halbach and cube configurations, the increase in transfer of 26.5 dB means that the actuation current can be lowered by a factor of 4.6 to achieve the same actuation angle with respect to the cylindrical configuration. This means that the dissipated power can be reduced by a factor of 21 (0.2 mW). The temperature increase would then be reduced to 0.01–0.02 °C.

5.5 Conclusion

Various transduction principles for actuation of MEMS resonators and in particular the μ -CMFS have been discussed. Advantages and disadvantages of each method were highlighted. The preferred method of actuation was chosen to be Lorentz actuation for its simplicity of integration into micro-machining, low excitation voltages and large displacements. A comparison was made between three magnet configurations to increase magnetic field strength and optimize actuation of the μ -CMFS. Magnitude transfer measurements between the actuation and induction voltages for the cubic and Halbach configurations show a significantly higher transfer (26.5 dB higher) than for the cylindrical configuration. Phase transfer measurements show that the Halbach and cubic configurations generate transfer plots according to simulation. The low EMF signal amplitude for the cylindrical configuration causes the phase to only reach down to 45° before going back up to 90°. This means the Halbach and cubic configurations are more suitable for use with an amplification loop to bring it into stable resonance, which is important for overall performance of the μ -CMFS. Furthermore, with these configurations, the temperature increase of the micro-channel due to resistive heating [16] can be reduced by a factor 21 (from 0.2–0.44 °C to 0.01–0.02 °C, in a flow range of 0–1 g h⁻¹, with a typical actuation angle of 5 mrad (θ_t in Fig. 5.6)). To conclude, the Halbach configuration has the most promising results regarding its field strength and stray-field but is hard to assemble due to the repelling forces of the magnets. Therefore, the best option is to use the configuration utilizing a single large cubic magnet. This combines a high magnetic field strength and a practical way to assemble the final sensor.

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6

Readout Methods

This chapter[†] introduces various transduction principles that are used in MEMS resonators. Advantages and disadvantages of each method are highlighted and their potential application to a μ -Coriolis mass flow sensor is discussed. A capacitive readout method is described in more detail in the section that follows. Non-linearity of the readout capacitance is modelled and investigated. Other limitations are discussed after which differential and resistive readout methods are presented to overcome these.

[†]This chapter is based on parts of publications [1, 2]:

- T. V. P. Schut, R. J. Wiegerink, and J. C. Lötters, “ μ -Coriolis Mass Flow Sensor With Differential Capacitive Readout,” *IEEE Sensors Journal*, vol. 21, no. 5, pp. 5886–5894, 2020
- T. V. P. Schut, R. J. Wiegerink, and J. C. Lötters, “ μ -Coriolis Mass Flow Sensor with Resistive Readout,” *Micromachines*, vol. 11, no. 2, p. 184, 2020

6.1 Introduction

Most MEMS sensors convert a physical phenomenon such as pressure, acceleration or flow to a mechanical movement or strain, which in turn is converted into an electrical signal. This also holds for resonating MEMS devices/sensors, where resonant motion needs to be monitored/detected. A certain detection scheme is necessary to convert the physical quantity that one wants to measure to an electrical output signal. This commonly consists of two steps. First, a transduction takes place to convert a physical quantity to a mechanical effect. For example, acceleration of a suspended proof mass causes displacement of that mass (basic principle of an accelerometer). The mechanical displacement or strain then needs to be converted to an electrical output signal. This is achieved via a certain readout method, e.g. electrostatic or (piezo-)resistive. The first μ -Coriolis mass flow sensor (μ -CMFS) used external optical detection to convert vibrations due to Coriolis forces to an electrical signal [3]. The first integrated readout for such a device was based on capacitive detection [4]. The devices presented in this thesis stem from the sensor presented by Haneveld et al. [5, 6], which started off with an external optical readout [5] after which work was done to integrate comb electrodes for capacitive detection [6]. In the next section, the potential transduction principles that can be utilized for readout of MEMS resonators and in particular μ -CMFS are discussed. The sections that follow provide more detail on capacitive and (piezo-)resistive readout methods and their (dis)advantages.

6.2 Transduction principles

Just as for actuation of MEMS resonators (see Chapter 5), various transduction principles can be utilized for detection of the vibrations involved in their operation. Most common transduction principles for readout are:

- Capacitive
- (Piezo-)resistive
- Piezoelectric
- Optical

Each method will be briefly discussed and their advantages and disadvantages will be highlighted.

6.2.1 Capacitive

Capacitive sensing is based on the measurement of changes in capacitance between a fixed (stator) electrode and moving (rotor) electrode. Figure 6.1 shows an example structure based on two parallel plate electrodes. When the rotor plate moves, the distance between the plates d changes. This in turn causes the capacitance C to change, which can be then be detected. An example of how this principle can be used, is to sense the displacement of a proof mass in accelerometers and gyroscopes or to detect the deflection of a membrane in pressure sensors. It is the most common readout method in MEMS [7], mainly due to its simple integration in micro-machining. Further advantages of a capacitive readout are its high sensitivity, low power consumption, design simplicity, low drift, low temperature dependency and good noise performance. These are the reasons why the μ -CMFS device presented by Haneveld et al. [6] used such a readout. A capacitive readout method for a μ -CMFS will be described later in this chapter. Devices utilizing this type of readout are presented in Chapter 7.

Figure 6.1: Schematic representation of a capacitive readout based on two parallel plate electrodes. A displacement Δz changes the distance between the plates d . This change in d affects the capacitance C between the plates, which can be measured.

Figure 6.2: Schematic representation of a basic (piezo-)resistive readout realized by a strain gauge on top of a suspended beam.

6.2.2 (Piezo-)resistive

Resistive sensing relies on a change in electrical resistance of a material when subjected to strain. This can be due to a change in dimensions of a conductor (e.g. metal strain gauge) or a change in the resistivity of a material (piezo-resistivity). Figure 6.2 shows an example of how a strain gauge can be used to detect bending of a suspended beam. Piezo-resistive sensing is common in MEMS, once again due to the relative ease of integration into silicon micro-machining. Implementation is simple, since doped silicon can be used as a piezo-resistor. The main disadvantage of a (piezo-)resistive readout is temperature dependence of the resistivity of a material. However, this can be reduced by utilizing a Wheatstone bridge configuration with four identical resistors [8]. This type of readout can be used for detection of vibrations in a μ -CMFS. Different versions of the SCT fabrication process were presented in Chapter 4. Metal strain gauges can be realized in conventional SCT. Piezo-resistive strain gauges can be realized by the adapted SCT process, integrating doped silicon strain gauges into the micro-channel roof. Application of such a readout to a μ -CMFS is described in more detail further on in this chapter, devices and measurements are presented in Chapter 7.

6.2.3 Piezoelectric

Piezoelectric actuation was discussed in Chapter 5. A voltage applied across the piezoelectric material results in contraction/expansions of the layer (for positive/negative applied voltage). Similarly, when a piezoelectric material experiences compression or elongations, charges are generated, which can be translated to an output voltage. Figure 6.3 shows an example of how this effect can be utilized to detect deflection of a suspended beam. Charges are generated dynamically, meaning that a

(a) When no force is applied, $v_{out} = 0$. (b) When a force induces compression of the piezoelectric layer, a potential is generated. (c) When a force induces elongation of the piezoelectric layer, v_{out} is of opposite sign.

Figure 6.3: Schematic representation of piezoelectric sensing of deflection of a suspended beam. A dynamic potential v_{out} is generated when the piezoelectric material is compressed or elongated.

static compression or elongation will not result in an output voltage. Only dynamic motion can thus be detected. Integration of piezoelectric material in a μ -CMFS was discussed in Chapter 5. As aforementioned, the main bottleneck is currently the quality of the piezoelectric materials and the difficult integration into the SCT process. Research on integration of piezoelectric material into μ -CMFS devices fabricated by SCT is currently ongoing and is not in the scope of the research presented in this thesis. Therefore, a piezoelectric readout method for sensing vibrations in a μ -CMFS is not practically considered.

6.2.4 Optical

Optical detection relies on the difference between transmitted and received light. Optical transduction in MEMS is mainly used for communication [9]. An external optical readout based on reflection of light on a moving surface can be used for μ -CMFS devices [3, 5, 10]. However, these optical systems are often large and difficult to align to the MEMS device. Integration of waveguides into μ -CMFS devices could overcome these disadvantages. Research on integration of waveguides into the SCT process is currently being done at the IDS group at the University of Twente. This research is in its early stages and thus an optical readout will not be in the scope of this thesis.

6.3 Capacitive readout of a μ -Coriolis flow sensor

6.3.1 Operating principle

Vibrations of the channel of the μ -Coriolis mass flow sensor are typically detected by a set of capacitive readout structures. Figure 6.4 shows an SEM image of a μ -CMFS with capacitive readout, indicating the readout structures. One comb is attached to the moving channel while the other is fixed to the substrate. Stresses developed in the layer stack during the fabrication process cause the suspended channel to bend slightly upward when released. This translates to an initial distance between the two comb electrodes in the order of 1 to 2 μm (along the z -axis). When the channel is vibrating, the comb fingers move towards and away from each other, generating a change in capacitance. By using two capacitors, actuation and detection mode vibrations can be distinguished. Figure 6.5 shows a schematic representation describing the operating principle of the capacitive readout.

The readout electrodes are centered around the *Twist* mode rotational axis, optimized for actuation in that mode. When actuated, the capacitance signals over time are 180° out of phase. When a fluid flows through the channel, *Swing* mode vibrations are induced through Coriolis forces, causing common mode capacitance signals. Adding the capacitance changes due to actuation and flow together gives two signals that have a phase difference φ dependent on the mass flow rate ϕ_m .

Figure 6.4: SEM image of a μ -Coriolis mass flow sensor with capacitive readout. The capacitive readout consists of two comb capacitor structures used for measuring displacement of the micro-channel.

Figure 6.5: Operating principle of the capacitive readout. When actuated in *Twist* mode, the capacitance signals over time are 180° out of phase. *Swing* mode vibration induced by a mass flow through the channel causes common mode signals. When actuating with flow, the capacitance signals will have a phase difference φ dependent on the mass flow rate ϕ_m .

Figure 6.6: SEM image of the readout structures which are often used with a μ -CMFS. The structures consist of two comb-shaped metal electrodes on top of a SiRN membrane connected to the micro-channel.

6.3.2 FEM analysis

As stated before, the readout capacitors consist of two comb shaped electrodes. The electrodes consist of $2.5\ \mu\text{m}$ SiRN with a $200\ \text{nm}$ metal layer (typically gold and/or platinum) deposited on top, see Fig. 6.6. There is an initial distance of $1\text{--}2\ \mu\text{m}$ between the fixed and moving combs. The change in capacitance resulting from the combs moving toward each other is non-linear with the deflection of the micro-channel. This means that the sensitivity of the capacitive readout is not constant, which can generate errors in reading the mass flow rate. A FEM model is set up in COMSOL Multiphysics® to investigate the non-linearity of the readout capacitors.

Figure 6.7: Geometry of one capacitive finger pair, as realised in COMSOL[®]. Parameters are indicated in the figure. An initial offset between the two combs is defined as depicted in (c).

Table 6.1: Dimensions of the readout capacitor geometry in COMSOL[®], see Figure 6.7.

Parameter	Symbol	Magnitude (μm)
Thickness of metal layer	t_{metal}	0.2
Thickness of SiRN layer	t_{SiRN}	2.5
Comb finger length	L_f	50
Comb finger width	w_f	7
Gap between comb fingers	g_x	5
Gap between comb finger and comb base	g_y	6
Displacement between combs along z -axis	d	0–10

6

The comb fixed to the moving micro-channel of the μ -CMFS experiences rotational vibrations around different axes depending on the vibrational mode. Simulating the movements of the electrodes for different combinations of vibrational modes is complex and very time intensive. To reduce simulation times and increase the flexibility of the model, a single finger pair is simulated. Figure 6.7 displays the geometry of the comb finger pair as realized in COMSOL[®], indicated dimensions are listed in Table 6.1. By using continuous boundary conditions, the simulated capacitance of the finger pair accurately represents that of a pair within the comb-shaped readout capacitor.

In the simulation, d is varied from 0 to 10 μm in steps of 0.1 μm . For each value of d , the capacitance is calculated. The resulting relation between d and the readout capacitance is displayed in Figure 6.8. As can be seen, the relation is highly non-linear. Typically, the aforementioned initial offset (along the z -axis) between the combs is

Figure 6.8: Capacitance of one comb finger pair with respect to the displacement d , simulated in COMSOL[®]. The relation can be approximated by a 4th order polynomial function.

in the order of 1–2 μm . The displacement due to actuation is in the order of 1–3 μm , depending on the actuation amplitude and the position of the comb finger pair. Viewing Figure 6.8, this would clearly cause d to vary within a range that cannot be considered linear. As explained earlier, the phase difference between the two readout capacitance signals is a measure for the mass flow rate. By using the 4th order polynomial fit from Figure 6.8, the readout capacitors of the μ -CMFS can be modelled.

6.3.3 Modelling effects of non-linearity on the output signal

Now that the relation between capacitance and displacement of one comb finger pair has been found, it can be used to describe the entire capacitive readout. This can be done by defining the positions of all the finger pairs in the readout capacitors over time. Then by calculating the capacitance of each comb finger pair at their various positions and displacement d , the total readout capacitance over time can be derived. The capacitance of one comb finger pair can be expressed as a 4th order polynomial function of the displacement d :

$$C_i(t) = \alpha \cdot d(t)^4 + \beta \cdot d(t)^3 + \gamma \cdot d(t)^2 + \delta \cdot d(t) + \epsilon \quad (6.1)$$

Where i indicates the i^{th} finger pair of the capacitor. The total capacitance is then:

$$C(t) = \sum_{i=1}^{N_f} C_i(t) \quad (6.2)$$

Where N_f is the number of finger pairs in one readout capacitor.

Figure 6.9: Illustration of the effect of the two main vibrational modes on distance d between the readout electrodes. Top: *Twist* mode induces rotation of the channel, Bottom: *Swing* mode induces up-/downwards motion. d is a function of position x along the channel and time t .

To calculate the total capacitance, d must be defined. To do this, the two main vibrational modes are viewed. Figure 6.9 shows these modes and how they affect d . As a result of the vibrating channel, d is a function of position x along the channel and time t . To express $d(x, t)$, the separate contributions of the two modes need to be derived. When actuating the sensor at the *Twist* mode resonance frequency with an actuation angle θ_t , the contribution of the *Twist* mode can be expressed as:

$$\begin{aligned}
 \theta_t(t) &= \hat{\theta}_t \cdot \cos(\omega_t t) \\
 d_t(x, t) &= x \cdot \sin(\theta_t(t)) \\
 &= x \cdot \sin(\hat{\theta}_t \cdot \cos(\omega_t t)) \\
 &\approx x \cdot \hat{\theta}_t \cdot \cos(\omega_t t)
 \end{aligned} \tag{6.3}$$

6

Where $\hat{\theta}_t$ is the *Twist* mode actuation angle amplitude, which is typically in the order of a few mrad. When there is a flow through the channel, Coriolis forces are generated and cause a vibration in the *Swing* mode (at the actuation/*Twist* mode resonance frequency). The *Swing* mode displacement is constant along the channel sections where the readout electrodes reside:

$$d_s(t) = \hat{d}_s \cdot \sin(\omega_t t) \tag{6.4}$$

Where $\hat{d}_s$ is the *Swing* mode displacement amplitude. The total displacement $d(x, t)$ then becomes:

$$d(x, t) = d_t(x, t) + d_s(t) + d_0 = x \cdot \hat{\theta}_t \cdot \cos(\omega_t t) + \hat{d}_s \cdot \sin(\omega_t t) + d_0 \tag{6.5}$$

Where d_0 is the distance between the readout electrodes at rest.

Figure 6.10: Modelled readout capacitance over time in relation to the initial distance d_0 between the readout electrodes. $d_s = 0$ and $\hat{\theta}_t = 5$ mrad, corresponding to 10 μm displacement in the corners of the channel loop. Non-linearity can be clearly observed at lower values of d_0 .

As aforementioned, this initial distance is a result of residual stress developed in the materials of the micro-channel during fabrication. By substituting Eq. 6.5 into Eq. 6.1, the capacitance over time of one finger pair can be calculated and then summed using Eq. 6.2. These calculations are done in MATLAB for various ranges and combinations of values for $\hat{\theta}_t$, $\hat{d}_s$ and d_0 to be able to analyze their effect in detail.

Firstly, the effect of the initial distance d_0 between the readout electrodes is studied. For this analysis, the capacitance over time is calculated for $d_0 = 1, 3, 5$ μm . θ_t is set to be 5 mrad (typical actuation amplitude) and $d_s = 0$ (only actuation). Figure 6.10 shows the modelled capacitance over time for the two readout capacitors. When comparing the left and right readout capacitances, they are 180° out of phase. This is to be expected since there is no *Swing* mode vibration due to flow, which would induce an additional phase shift. For $d_0 = 5$ μm the capacitance is a clear sinusoidal signal. When d_0 is lower however, the signal flattens at the peaks or even shows a dip. This effect can be attributed to the fact that the readout electrodes are coming close to each other or passing through each other. This corresponds well with Figure 6.8, where it shows that for lower values of d , the capacitance becomes more non-linear and less sensitive to changes in d . The same effect can be seen when keeping d_0 constant while increasing the actuation angle θ_t . Figure 6.11 shows the readout capacitance in relation to the actuation angle. The signal amplitude increases with θ_t until it reaches a maximum and eventually forms a dip as in Figure 6.10. Since it involves a rotation instead of an up-/downwards movement of the entire electrode, finger pairs closer to the rotational axis will be moving in a different range of the curve in Figure 6.8 than finger pairs further from the rotational axis. This results in a slightly different non-linearity effect than that of varying d_0 .

Figure 6.11: Modelled readout capacitance over time in relation to the actuation angle θ_t . $d_s = 0$ and $d_0 = 3 \mu\text{m}$.

Next, the effect of this phenomenon on the flow sensitivity will be viewed. As described earlier, the phase difference φ_{out} between the two capacitance signals is a measure for the flow rate. More precisely, the ratio between the detection mode displacement and the actuation mode angle is linearly dependent on the flow rate (see Chapter 3, Eq. 3.15). φ_{out} represents this ratio, as depicted in 6.5. However, the non-linearity effect displayed in Figs. 6.10 and 6.11 affects the measured phase difference φ_{out} . φ_{out} is calculated from the in-phase and quadrature components of the capacitance signals, just as is done in the readout electronics (which will be presented in Chapter 7). Figure 6.12 shows the relation between φ_{out} and the ratio d_s/θ_t for different values of d_0 (corresponding to Fig. 6.10). A linear relation can be seen for each value of d_0 (Fig. 6.12a). However, the slope of the lines is different for each value of d_0 resulting from the non-linearity of the readout capacitance. This translates to different values of the sensitivity $\varphi_{\text{out}}/(d_s/\theta_t)$ with respect to d_0 . A difference of $1 \mu\text{m}$ in d_0 results in approximately a 1% change in sensitivity, which would mean a potential error of 1% in the measured flow rate. The curve seems to flatten when d_0 becomes larger, corresponding with the curve in Fig. 6.8.

Again, a similar effect can be observed when varying θ_t . Figure 6.13 shows the relation between φ_{out} and the ratio d_s/θ_t for different values of θ_t (corresponding to Fig. 6.11). Just as in Fig. 6.12, the relation, although linear, has a slightly different slope for each value of θ_t (Fig. 6.13a). The sensitivity $\varphi_{\text{out}}/(d_s/\theta_t)$ varies with θ_t , see Fig. 6.13b. A difference of 1 mrad in θ_t results in approximately a 1% change in sensitivity as well. The slope of the curve becomes higher for larger actuation amplitudes. This corresponds with Fig. 6.8, since a larger actuation amplitude would mean that d would get closer to 0 where the curve is more non-linear.

(a) Output phase difference φ_{out} between capacitance signals in relation to the detection mode amplitude d_s . φ_{out} is simulated for different values of d_0 , $\theta_t = 5 \text{ mrad}$. (b) Sensitivity of the output phase difference φ_{out} in relation to d_0 .

Figure 6.12: Dependence of the output phase difference φ_{out} on the initial displacement d_0 between the readout electrodes.

(a) Output phase difference φ_{out} between capacitance signals in relation to the detection mode amplitude d_s . φ_{out} is simulated for different values of θ_t , $d_0 = 3 \mu\text{m}$. (b) Sensitivity of the output phase difference φ_{out} in relation to θ_t .

Figure 6.13: Dependence of the output phase difference φ_{out} and its sensitivity on the actuation angle θ_t .

6.3.4 The effect of external disturbances

External disturbances, such as sound waves generated by machinery in close proximity to the sensor, can cause vibrations of the micro-channel interfering with its operation. In principle, signals at frequencies other than the actuation frequency can be distinguished from the output signal due to flow. As will be discussed in Chapter 7, a lock-in amplifier is used to filter out any signals at other frequencies. However, due to the non-linear readout capacitors, signals at other frequencies can actually cause disturbance at the actuation frequency. This can be derived from Equation 6.1. In the case that there is a disturbance at an arbitrary frequency ω_{dist} , $d(t)$ will be a sum of signals at the two frequencies. Let us assume a common mode disturbance, where the channel section on which the readout electrodes reside experiences up-/downward vibration without rotation. In this case $d(t)$ is similar to Equation 6.5:

$$d(x, t) = x \cdot \hat{\theta}_t \cdot \cos(\omega_t t) + \hat{d}_{\text{dist}} \cdot \sin(\omega_{\text{dist}} t) + d_0 \quad (6.6)$$

Where the only difference is that there is now a common disturbance at frequency ω_{dist} instead of *Swing* mode motion at ω_t due to Coriolis forces. When squaring such an expression as Eq. 6.6, the outcome contains terms at the sum and difference frequencies $(\omega_t + \omega_{\text{dist}})$ and $(\omega_t - \omega_{\text{dist}})$ as well as terms at ω_t and ω_{dist} . Since Eq. 6.1 is a 4th order polynomial function, substituting Eq. 6.6 into 6.1 will give an expression with countless terms of sinusoidal functions. Among these, there are terms with frequency ω_t of which the amplitude depends on $\hat{d}_{\text{dist}}$. This is due to the fact that when terms with frequency $(\omega_t + \omega_{\text{dist}})$ are multiplied with terms of frequency ω_{dist} , one of the outcomes is at frequency ω_t . However, the amplitude of the term depends on $\hat{d}_{\text{dist}}$. Similarly to this example, there are more terms at ω_t depending on $\hat{d}_{\text{dist}}$. Thus, a common mode vibration at an arbitrary frequency can cause a disturbance of the readout capacitance at the actuation mode resonance frequency. The effect from these disturbances can be reduced by using a differential readout method.

6

6.3.5 Differential Readout

Differential capacitive sensing is widely applied in the field of micro-machined sensors [11–14]. One of the main advantages of such a readout method is that it can greatly reduce sensitivity to other vibration modes/directions in inertial sensors [15]. This advantage can be exploited and applied to a μ -Coriolis flow sensor to reduce the effect of common mode vibrational disturbances.

External disturbances can cause vibration of the sensor chip resulting in upwards and downwards movement of the suspended channel. These vibrations will be most present at the detection mode resonance frequency. Coriolis forces are induced at the actuation mode resonance frequency. This means that in principle, disturbance

Figure 6.14: Operating principle of a μ -CMFS with differential readout. The effect of common mode vibrations due to external disturbances can be reduced by using 2 readout capacitors on each side of the channel loop.

vibrations and Coriolis vibrations can be distinguished from each other. However, as previously described, the capacitive readout is non-linear for large deflections. This non-linearity is typically not an issue, since Coriolis deflections are in the order of 10–100 nm. When external disturbances cause deflections in the order of microns, non-linearity will start to have its effect. As mentioned in the previous section, a disturbance at an arbitrary frequency (in this case the detection mode resonance frequency) can cause a disturbance at the actuation mode resonance frequency. As a result, there will be an unpredictable offset in the measured flow caused by these disturbance vibrations. This can be avoided by using a differential readout method, based on a double channel loop device with additional capacitive readout structures.

Fig. 6.14 shows the operating principle of a μ -Coriolis mass flow sensor with differential capacitive detection. Similarly as for a single loop device (see Fig. 6.5), the sensor is brought into resonance in the actuation mode. A mass flow now follows a flow path through two mechanically connected channel loops. Coriolis forces are generated

in each of the loops in opposite direction, causing a differential Coriolis/detection vibration mode. By using a set of comb capacitors on each side of the device, common mode terms in the capacitance signals can now be (partially) cancelled out. Due to non-linearity of the readout capacitances, the influence of common mode disturbance signals is not the same on each side of the sensor. This means that the effect of disturbance signals will be reduced but not eliminated.

Fig. 6.15 shows how the differential capacitive readout operates and can reduce the effect of common mode disturbances. Actuation vibration detected by 4 capacitors C_1 , C_2 , C_3 and C_4 gives two signals of opposite phase for each capacitor pair ($C_1(t)$ & $C_2(t)$ and $C_3(t)$ & $C_4(t)$). Coriolis vibration due to a mass flow through the channel causes in-phase signals for each set of capacitors. However, $C_1(t)$ & $C_2(t)$ are 180° out of phase with $C_3(t)$ & $C_4(t)$. Viewing operation with flow, the total capacitance change due to actuation and Coriolis vibration gives two phase differences φ_a and φ_b dependent on the mass flow through the channel of the sensor. As aforementioned, common mode vibrations due to external disturbances cause additional capacitance variations at the actuation resonance frequency. This causes a shift in the phase shifts φ_a and φ_b in opposite direction. For a single channel loop device with just two readout comb capacitors, this change in phase shift would be impossible to distinguish from a change in flow. This is the case for previously presented devices [4, 6, 16, 17]. However, by using 4 readout capacitors, one can obtain a phase difference φ_{ab} (see Fig. 6.15) dependent on the mass flow rate and less dependent on external disturbances. A device with such a differential capacitive readout is presented and characterized in Chapter 7.

Figure 6.15: Schematic representation of the differential capacitive detection scheme. For clarity, only resulting signals at the actuation mode resonance frequency are viewed. Signals at other frequencies can be filtered out. φ_{ab} has a reduced dependence on external disturbances compared to φ_a and φ_b (or φ in Fig. 6.5). The effect of external disturbances cannot be completely eliminated due to non-linearity of the readout capacitors.

6.3.6 Water hammer effect

To optimize the sensitivity and noise performance of the capacitive readout, the gap between the fingers of the comb structures is chosen as small as possible, see Fig. 6.7 and Tab. 6.1. Typically, this gap is chosen to be around 6 μm . Since the sensor is actuated at the *Twist* mode resonance frequency, there is little to no in-plane motion of the channel loop. However, when gas bubbles are present in the fluid, they can cause a sudden local peak in flow velocity. This can be explained by viewing the equation for mass flow rate:

$$\phi_m = \rho v A \quad (6.7)$$

Where ρ is the density of the fluid, v is the flow velocity and A is the cross sectional area of the channel. When a gas bubble enters a steady flow of a liquid, locally ρ will decrease significantly. Since A and ϕ_m are constant throughout the channel, this will lead to a local increase of flow velocity v . This is then followed by a sudden drop in v once the gas bubble passes. Similarly, the pressure locally drops and surges, causing the liquid following the gas bubble to 'hammer' on the channel. This can cause a large in-plane force on the micro-channel. Since the gap between the fingers of the comb electrodes is only a few μm , this can result in collision of the electrodes, leading to damage to the readout structures or them getting stuck to each other. Figure 6.16 shows a microscope photo of a $\mu\text{-CMFS}$ device after this phenomenon has occurred.

There are various ways to approach this problem:

- Suppress in-plane motion (e.g. by spring suspension or changed channel design).
- Suspend the capacitive readout structures to allow them to move in-plane.
- Use a different type of readout that allows in-plane motion, e.g. (piezo-)resistive.

In Chapter 7, devices are presented based on these approaches. Furthermore, application of a resistive readout to a $\mu\text{-CMFS}$ is described in the following section.

6

Figure 6.16: Microscope photo of a $\mu\text{-CMFS}$ where the readout electrodes are stuck due to in-plane motion from water hammer effect.

6.4 Resistive readout

6.4.1 Motivation

As aforementioned, a capacitive readout has certain limitations. For one, inherent non-linearity causes a dependence of the sensitivity (to Coriolis vibrations) on the actuation amplitude and initial distance between the fixed and suspended combs. This also allows for external vibrations to cause an offset in the measured flow rate.

An additional effect of non-linearity of the capacitive readout is that *Swing* mode actuation is not feasible. For *Twist* mode actuation, the capacitive readout structures are put as close to the *Twist* mode rotational axis as possible to maximize sensitivity to *Swing* mode Coriolis vibrations. This also limits the displacement of the moving comb electrodes, since displacements due to actuation are relatively small in this position along the channel. Thus, a relatively large actuation amplitude can be achieved without the comb electrodes surpassing each other. This would cause changes in sensitivity such as depicted in Fig. 6.13 or larger. For *Swing* mode actuation, one would like to maximize sensitivity for *Twist* mode Coriolis vibrations. However, the largest displacements in both modes occur in the top corners of the channel loop (see Figure 6.5). Putting the readout structures here would limit the actuation amplitude significantly, since relatively large displacements due to actuation occur here. Therefore, *Twist* mode actuation is typically used for these devices. However, external disturbances causing up-/downwards motion of the whole sensor chip, can more easily induce vibrations in the *Swing* mode. When the *Swing* mode is used as Coriolis mode, small disturbances can have a significant effect on the output signal, since Coriolis displacements are relatively small (in the order of 10–100 nm). Because of this fact, it would be preferable to actuate the sensor in the *Swing* mode and detect *Twist* mode Coriolis forces. Then disturbances causing *Swing* mode vibrations will simply in-/decrease the actuation amplitude or shift its phase slightly. Since Coriolis forces are a result of actuation vibrations (in combination with a flow), this change in amplitude or phase will affect both the actuation and detection modes. The ratio between the amplitudes of the two modes is not affected by this.

Another effect is that pressure inside the channel causes the channel to bend upwards, due to its non-circular cross-section. This results in a change in distance between the capacitor combs, which shifts the point of non-linearity (see Figure 6.12). Furthermore, relative permittivity of the fluid flowing through the channel influences the measured capacitance. Finally, the water hammer effect described earlier can damage the readout structures. A (piezo-)resistive readout would not suffer from all these limitations and could be operated using either *Twist* or *Swing* mode actuation.

6.4.2 Operating principle

The resistance of a rectangular conductor is defined as:

$$R = \rho \frac{l}{A}, \quad (6.8)$$

Where ρ is the resistivity of the conductor material, l is the length of the conductor, and A is its cross-sectional area. When the conductor is deformed due to an applied strain, its resistance changes. For a metal conductor, ρ remains constant with deformation. The change in the ratio l/A solely determines the change in resistance. In the case of piezo-resistive materials, ρ changes with deformation of the material as well. The relative change in resistance due to an applied strain can be expressed as [18]:

$$\frac{\Delta R}{R_0} = \frac{\Delta l}{l_0} - \frac{\Delta A}{A_0} + \frac{\Delta \rho}{\rho_0} \quad (6.9)$$

The relative change in cross-sectional area of the conductor is related to the applied strain ($\Delta l/l_0$) via Poisson's ratio λ :

$$\frac{\Delta A}{A_0} = -2\lambda \frac{\Delta l}{l_0} \quad (6.10)$$

The sensitivity to strain is commonly called gauge factor. It can be expressed as:

$$G = \frac{\Delta R/R_0}{\Delta l/l_0} = 1 + 2\lambda + \frac{\Delta \rho/\rho_0}{\Delta l/l_0} \quad (6.11)$$

Poisson's ratio for most solid materials such as metals, polymers and ceramics is in the range of $0.25 < \lambda < 0.35$ [19]. In the case of metal, the third term in Equation 6.11 goes to zero and G is determined by λ . For piezo-resistive materials, the third term in Equation 6.11 is dominant. In the case of p -doped mono-crystalline silicon, G ranges from -75 to +150, whereas for p -doped poly-silicon, the gauge factor ranges from -10 to +30 [20, 21]. The exact value of G depends on the amount of dopant. Many MEMS devices are fabricated using a combination of silicon, metals and other materials. This allows for relatively simple implementation of strain gauges. Metal strain gauges have been integrated with a μ -Coriolis mass flow sensor in the past for pressure sensing [22, 23]. Static changes in the resistance of the strain gauge were related to deformation of the micro-channel roof induced by an applied pressure inside the channel. This showed that possibly the same principle could be used to detect vibration of the resonating micro-channel of the μ -CMFS itself. Figure 6.17 shows the implementation of such a readout to detect mass flow. The strain gauges in this case, consist of metal tracks patterned on top of the micro-channel. However, they can also be realized by using p -doped silicon following the fabrication process presented in 4.3.3. In the example in Figure 6.17, the sensor is actuated in the *Swing* mode.

Figure 6.17: Operating principle of a resistive readout implemented in a μ -Coriolis mass flow sensor. When actuated in the *Swing* mode, both strain gauges will be compressed or elongated at the same time. This results in equal resistance change. In the *Twist* mode, one strain gauge is elongated while the other is compressed. This results in opposite change in resistance. Combination of the two modes gives two resistance signals with a phase shift φ dependent on the mass flow rate ϕ_m .

Left in Fig. 6.17, the configuration of the strain gauges when the sensor is not being actuated is depicted. When the sensor is actuated, both strain gauges are elongated at the same time and compressed at the same time. The resistances over time of the two strain gauges are thus in phase for the two strain gauges. When there is a flow through the channel, Coriolis forces induce vibration in the *Twist* mode. This vibration is 90 degrees out of phase relative to the actuation (see Chapter 3). One strain gauge is elongated while the other is compressed. The resistances over time of the strain gauges are thus in opposite phase and ± 90 degrees out of phase with the resistance change due to actuation. When combining actuation and Coriolis vibrations, the resistances of the strain gauges will have a phase difference which depends on the mass flow rate through the channel. This method of readout is advantageous, since it does not depend on the amplitudes of the resistive signals but only on the phase. Temperature dependence in the amplitudes of the signal could be more significant than for a capacitive readout. However, since the phase of the signals is measured, temperature dependence of the resistive readout should be comparable to that of a capacitive readout. The same readout method can also be used while operating the sensor in the *Twist* mode. In this case, the *Swing* and *Twist* modes in Fig. 6.17 would simply be interchanged.

6.5 Conclusion

In this chapter, various potential transduction principles for the detection of the vibrations in a μ -Coriolis mass flow sensor were introduced. Capacitive and (piezo-)resistive readout methods have been described in more detail. A model was set up to investigate the limitations of a capacitive readout. The inherent non-linearity causes dependence of the flow sensitivity on the actuation amplitude, pressure, initial distance (along the z axis) between the readout electrodes and external disturbances. Additionally, operation of the μ -CMFS using *Swing* mode actuation is not feasible. A differential capacitive readout was proposed to limit the effect from external disturbances. Furthermore, a resistive readout was presented which could be used to overcome the limitations of a capacitive readout as well as the so-called water hammer effect. In Chapter 7, various devices are presented based on:

- Differential capacitive readout
- Suspension to overcome water hammer effect
- Devices with larger stiffness
- Resistive readout utilizing metal strain gauges
- Piezo-resistive readout using doped mono-crystalline silicon

The devices are characterized and a more quantitative comparison of the readout methods will be given, as opposed to the qualitative analysis in this chapter.

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Design & Characterization

This chapter[†] discusses μ -Coriolis mass flow sensor designs that have been realized based on the analyses presented in this thesis. Characterization methods are first discussed, after which the model presented in Chapter 3 is validated experimentally. Subsequently, several improved μ -CMFS designs are presented and characterization results are discussed.

[†]This chapter is based on (parts of) publications [1–3]:

- T. V. P. Schut, R. J. Wiegerink, and J. C. Lötters, “ μ -Coriolis mass flow sensor with improved flow sensitivity through modelling of the sensor,” *Microelectronic Engineering*, p. 111289, 2020
- T. V. P. Schut, R. J. Wiegerink, and J. C. Lötters, “ μ -Coriolis Mass Flow Sensor With Differential Capacitive Readout,” *IEEE Sensors Journal*, vol. 21, no. 5, pp. 5886–5894, 2020
- T. V. P. Schut, R. J. Wiegerink, and J. C. Lötters, “ μ -Coriolis Mass Flow Sensor with Resistive Readout,” *Micromachines*, vol. 11, no. 2, p. 184, 2020

7.1 Introduction

In the preceding chapters, various aspects and limitations of the μ -Coriolis mass flow sensor (μ -CMFS) have been discussed. This chapter presents novel designs based on the analyses in these chapters. Each device was designed to address one or more of the limitations that were discussed. Table 7.1 gives an overview of the devices to be presented, divided by method of readout, with their novelty and main goal indicated. Below, an outline of the devices and corresponding sections is given, briefly describing each type of device and its motivation:

- In Fig. 3.5 (Chapter 3), it was shown that devices with a larger channel loop size should show higher flow sensitivity. Therefore, μ -CMFS devices with varying channel loop size were designed to 1) validate the model and 2) increase flow sensitivity of the μ -CMFS. The devices are presented and characterized in **section 7.3**.
- In Chapter 2, the concept of a differential readout for MEMS inertial sensors was introduced. The idea behind such a readout is that the effect of vibrations induced by external disturbances can be (partially) cancelled out. A differential capacitive readout for a μ -CMFS was proposed in Chapter 6 and the operating principle (Fig. 6.14) and readout scheme (Fig. 6.15) were discussed. Based on this proposed readout, devices were designed and fabricated. The devices have been characterized by generating artificial external vibrations and investigating the ability of the device to reduce disturbances in the output signal due to these vibrations. The results of this characterization are presented and discussed in **section 7.4**.
- The water hammer effect was introduced in Chapter 6, where gas bubbles in a liquid flow can cause sudden in-plane motion of the micro-channel of the μ -CMFS. This can cause breakage of the capacitive readout structures, since the gap between the comb electrodes is only a few micron. To address this problem, devices with a suspended capacitive readout are presented in **section 7.5** that should allow for larger in-plane motion. As such, in-plane vibration is not suppressed but the readout is made flexible in the in-plane direction to prevent breakage of the readout electrodes.
- In **section 7.6**, μ -CMFS devices with a more stiff suspension are presented. Primarily, this increases the resonance frequencies and Q factors of the devices. This should lead to a higher flow sensitivity, as will be explained. Additionally, $1/f$ noise in the readout electronics (see Chapter 2) could be reduced, since the signals involved in operating the μ -CMFS are at a higher frequency. Finally, the suspension is also more stiff in the in-plane direction, reducing the magnitude of vibrations induced by water hammering.

Table 7.1: Overview of the novelties and goals of the devices presented in this chapter.

Readout	Section	Novelty	Main goal
Capacitive	7.3	Varying channel loop size	Validation of the model from Chapter 3
	7.4	Differential readout	Reduce influence of external vibrations
	7.5	Suspended readout electrodes	Increase robustness against water hammer effect
	7.6	Increased resonance frequency and Q factor	Increase flow sensitivity and resolution
Resistive	7.7	Metal strain gauges	Provide alternative to capacitive readout that does not suffer from water hammer effect
	7.8	Piezo-resistive strain gauges	Improve resolution with respect to metal strain gauge readout

- Disadvantages of a capacitive readout for a μ -CMFS were discussed in Chapter 6. The largest disadvantage is the aforementioned water hammer effect. An alternative readout method that does not risk breakage due this effect is a resistive readout. Such a readout based on strain gauges was described in Chapter 6. In **section 7.7**, novel devices are presented that integrate metal strain gauges on top of the micro-channel of a μ -CMFS. After a proof-of-concept, the flow sensitivity and resolution of the devices are investigated.
- As a more sensitive alternative to the μ -CMFS devices with metal strain gauges in 7.7, the integration of piezo-resistive strain gauges in a μ -CMFS is researched. Proof-of-principle devices are presented in **section 7.8** and their performance is compared to that of the devices in 7.7.

To characterize the devices listed above, a certain set of experiments have to be executed. Characterization methods for these experiments are described in the following section.

7.2 Characterization methods

To obtain the characterization results that are presented in this chapter, a certain set of experiments is executed. In this section, the following methods/aspects will be discussed:

- *Analysis of out-of-plane resonant behaviour by Laser Doppler Vibrometry:*
To confirm whether the μ -CMFS operates mechanically as expected. In the case of validation of the flow sensitivity model, this method is used in combination with a flow measurement.
- *Analysis of in-plane vibration by Stroboscopic Video Microscopy:*
To analyse robustness against the water hammer effect.
- *(Zero-)Flow measurement:*
To determine the flow sensitivity and resolution of the μ -CMFS.
- *Electronic circuitry used for capacitive and resistive readout methods.*

7.2.1 Resonance mode analysis by Laser Doppler Vibrometry

Laser Doppler Vibrometry (LDV) has been used for optical characterization of resonant MEMS devices since the 1990's [4]. Advantages of LDV are that a device can be characterized without necessity of physical contact to the micro-structure. Furthermore, it is a very accurate method and can be used to detect vibrations down to the picometer range. Figure 7.1 shows the principle of operation of a Laser Doppler Vibrometer. A laser beam is split into two paths. One travels via a mirror directly to a photo-detector. The second beam is focused on a vibrating sample through a lens. The laser beam is then reflected. The frequency and phase of the reflected light is affected by the vibrating sample, as a result of the Doppler effect. The reflected light then travels to the photo-detector as well, where the two beams interfere and create a modulated signal dependent on the (out-of-plane) velocity of the vibrating sample [5].

For characterization of devices in this thesis, a Polytec MSA-400 system with an integrated LDV is used. Figure 7.2 shows a schematic representation of a measurement setup to characterize resonant behaviour of a μ -CMFS. The sensor is actuated by means of Lorentz actuation (see Chapter 5) using the function generator of the MSA-400 in combination with magnets residing on the chip PCB. The laser spot is swept through multiple positions on the channel as to characterize the different modes of the device. First, a periodic chirp[†] is used as actuation signal to get the frequency response of the μ -CMFS over a specified range. From the resulting response, the resonance frequencies and the Q factors of the different vibrational modes can be determined. Subsequently, the sensor can be driven at a specific resonance frequency using a sinusoidal signal to determine the relation between actuation voltage and displacement amplitude.

[†]A sinusoidal signal of which the frequency is swept through a specified range.

Figure 7.1: Principle of operation of a Laser Doppler Vibrometer. A laser beam is split into two paths. One beam travels directly to a photo-detector. The other is reflected on a vibrating sample, affecting the phase and frequency of the light through the Doppler effect. It then travels to the detector, where the two beams interfere and create a modulated signal dependent on the vibrating sample's velocity.

Figure 7.2: Schematic representation of a setup to characterize resonant behaviour of a μ -CMFS using LDV on a Polytec MSA-400.

Figure 7.3: Schematic representation of a setup to characterize in-plane vibrations of a μ -CMFS using SVM on a Polytec MSA-400.

7.2.2 In-plane motion analysis by Stroboscopic Video Microscopy

In addition to LDV, the Polytec MSA-400 also integrates Stroboscopic Video Microscopy (SVM) to analyze in-plane motion of micro-structures. SVM works by flashing light pulses on a vibrating sample and capturing the reflected light with a camera. Vibrations at frequencies higher than the frequency of flashing can be analysed by collating images captured over multiple periods of vibration [6]. This technique can be used for in-plane motion measurements down to picometer resolution [7]. This is of particular interest for the μ -CMFS to investigate the effect of in-plane motion of the device and to evaluate robustness against the water hammer effect. The measurement setup for these types of measurements is displayed in Fig. 7.3. The setup is very similar to that for LDV measurements (Fig. 7.2). The main difference is that a magnet now resides underneath the chipboard to generate a magnetic field perpendicular to the substrate of the μ -CMFS chip. Furthermore, the SVM interface of the MSA-400 is used as opposed to the LDV interface.

7

7.2.3 Readout electronics

Capacitive and resistive readout methods were presented in Chapter 6. In both cases the phase difference between the readout capacitances/resistances over time is a measure for the mass flow rate. To determine this phase difference, dedicated electronic readout circuitry is used.

Capacitive

Two topologies are commonly used to measure capacitance changes: an impedance bridge, or a capacitance controlled oscillator [8]. The impedance bridge provides the advantage that the parasitic capacitance to ground can be eliminated while keeping

Figure 7.4: Schematic representation of the electronic circuit for a capacitive readout. Single ended readout: Readout capacitance C_1 (see Fig. 6.15) is converted to a voltage using a charge amplifier via a 1 MHz carrier signal v_s and a reference capacitor C_{ref} . v_s is mixed out after which an SR860 lock-in amplifier determines the magnitude and phase of the signal (with actuation voltage $v_{\text{act}}(t)$ as a reference). **Differential readout:** Additional readout capacitance C_4 is connected such that $C_1 - C_4$ is converted to a voltage. A duplicate of this circuit is used for readout capacitance C_2 (and C_3 in case of differential readout).

the circuitry relatively simple and compact. This can be beneficial for micro-machined sensors where the readout electronics may be integrated in the sensor die and the parasitic capacitance is often larger than the sensing capacitance. This method is therefore used to read out the μ -CMFS. Fig. 7.4 shows a simplified schematic of the electronic readout circuit (based on the readout presented by Haneveld et al. [9]). Both single ended and differential readout topologies are shown. For a single ended readout: The readout electrode residing on the resonating micro-channel is supplied with a high frequency ($\sim 1\text{MHz}$) sinusoidal carrier signal v_s . C_1 is taken as example (see Fig. 6.5, Chapter 6). A duplicate of the circuit is used for reading out C_2 . The readout capacitance is converted to a voltage via a charge amplifier and reference capacitor C_{ref} . Subsequently, the signal is multiplied by the carrier signal v_s such that the resulting signal contains components at the actuation mode resonance frequency ω_a and at the sum/difference of the double carrier frequency and the actuation frequency ($2\omega_s \pm \omega_a$). The higher frequency components are then filtered out by a low-pass filter. Finally, a lock-in amplifier is used to obtain the amplitude and phase of the signal with respect to the actuation signal. The difference between the output phases of the two circuits (one for C_1 , one for C_2) is a measure for the mass flow rate.

For a differential readout: An additional readout capacitor (C_4 is taken as an example, see Fig. 6.15) is supplied with the inverse of the carrier signal ($-v_s$), indicated in blue in Fig. 7.4. As a result, the difference between the two readout capacitances $C_1 - C_4$ is converted to a voltage. Once again, the carrier signal is mixed out and a lock-in amplifier determines the magnitude and phase of the signal. In this case a

Figure 7.5: Schematic representation of the electronic circuit for a resistive readout. The resistance of one strain gauge $R_{1/2}$ is converted to a voltage via a 1 MHz carrier signal v_s and a reference resistor R_{Ref} . v_s is mixed out after which an SR860 lock-in amplifier determines the magnitude and phase of the signal (with actuation voltage v_{act} as a frequency reference).

duplicate of the circuit is used to determine $C_2 - C_3$. The phase difference between the output phases of the two circuits is a measure for the mass flow rate and less influenced by external disturbances (see Chapter 6).

Resistive

As presented in Chapter 6, the resistive readout method for a μ -Coriolis mass flow sensor has a very similar operating principle to that of a capacitive readout. Therefore the readout circuit that is used is very similar as well. Figure 7.5 shows a schematic representation of the readout topology used to determine the magnitude and phase of the readout resistances. $R_{1/2}$ represents either the left or right strain gauge of the μ -CMFS (see Fig. 6.17). The resistance of one strain gauge $R_{1/2}$ is converted to a voltage via a high frequency (~ 1 MHz) carrier signal and an on-chip reference resistor R_{Ref} . The reason why a high frequency carrier signal is used, is because the actuation signal induces voltages at the actuation mode resonance frequency over the strain gauges through cross-talk. By using a carrier signal, only variations in the resistance of the strain gauges (which modulate the carrier signal) due to motion of the channel will be viewed. Once again, the signal is demodulated by multiplying with the carrier signal and filtering out high frequency components. Finally, a lock-in amplifier determines the magnitude and phase of the signal. Each strain gauge is read out with a circuit such as depicted in Figure 7.5. The phase difference between the two resulting signals is a measure for the mass flow rate.

7.2.4 Flow measurement

The primary criteria in the performance of a micro-CMFS mass flow sensor are its sensitivity to flow and the lowest measurable flow rate or zero-flow stability.

Figure 7.6: Schematic representation of a flow measurement setup to characterize performance of a μ -Coriolis mass flow sensor. Nitrogen gas is fed through the micro-channel of the device via an input pressure P_{in} . A Bronkhorst® EL-FLOW® flow controller is used to set the flow rate.

Therefore, the performance of the μ -CMFS devices in this chapter is characterized through a flow measurement and a zero-flow measurement. In the first, the output signal is measured at different flow rates to determine flow sensitivity of the device. A schematic representation of a flow measurement setup is displayed in Figure 7.6.

Nitrogen gas is fed through the micro-channel of the device via an input pressure P_{in} of 5 bar (gauge pressure). Blue arrows indicate the flow path through the setup. The mass flow rate is controlled at the outlet of the device by a Bronkhorst® EL-FLOW® flow controller. The microfluidic setup is based on that presented by Alveringh et al. [10]. The μ -CMFS chip is mounted on a PCB where a cubical magnet resides in a 3D printed dedicated holder above the chip. The cubical magnet configuration (presented in Chapter 5) is used for all measurements presented in this thesis, unless stated otherwise. At the bottom of the chip PCB, the in-/outlets of the μ -CMFS chip can be accessed by 3D printed fluid connectors. The chip can then be read out either by LDV or by a capacitive/resistive readout in combination with the electronic circuitry presented in Figs. 7.4 and 7.5. The electronic circuits are realized on so-called modules which can be connected to the μ -CMFS chip [10].

During a flow measurement, the flow rate is controlled in steps through a certain flow range. The step size and flow range are given for each device in the following sections. The output phase difference of the electronic readout is then related to the applied flow to calculate flow sensitivity. For a zero flow measurement, the micro-channel of the μ -CMFS is kept at a constant pressure applied to its inlet while the outlet is closed off. The standard deviation of the output signal is then determined over a certain time period. By dividing this standard deviation by the flow sensitivity, a zero flow stability in g/h is obtained. This value defines the resolution of the device.

During all measurements with capacitive/resistive readout, the integration constant of the low-pass filter on the SR860 lock-in amplifier is set to 100 ms and its slope is set to 24 dB/decade. After sampling the data, it is averaged with a moving average over a time interval of 1s.

7.3 Validation of the flow sensitivity model

In Chapter 3, a model was presented that predicts flow sensitivity and resonant behaviour of μ -CMFS designs. Fig. 3.5 showed that devices with a larger channel loop should have higher flow sensitivity. In addition, changing the channel loop size, significantly changes the resonance frequencies of the device. Therefore, devices with varying channel loop size are very suitable to verify the accuracy of the model derived in Chapter 3 with respect to the prediction of resonance frequencies and flow sensitivity. Thus, such devices were designed and fabricated to 1) validate the model and 2) realize higher flow sensitivity. As was explained in Chapter 3, the ratio between the detection/Coriolis mode displacement amplitude $\Delta \hat{z}_d$ and the actuation angle amplitude $\hat{\theta}_a$ is a measure for the mass flow rate ϕ_m through the micro-channel of the μ -CMFS. Therefore, flow sensitivity is defined as this ratio divided by the mass flow rate ($(\Delta \hat{z}_d / \hat{\theta}_a) / \phi_m$). Fig. 3.2 is repeated as Fig. 7.7 to indicate $\hat{\theta}_a$ and $\Delta \hat{z}_d$. The devices will be compared by determining the flow sensitivity of each device. A comparison will also be made to modelled parameters.

7.3.1 Devices & Setup

Three types of devices were designed based on the simulation results presented in Fig. 3.5 of Chapter 3. Figure 7.8 shows microscopic photographs of the fabricated devices. Fig. 7.8a shows a device with typical dimensions based on the sensor in [11] (Type A). This device type serves as a reference to which the other two are compared. Fig. 7.8b shows a device with a significantly increased channel loop size (Type B). A variation of this device with an increased in-/outlet channel length is shown in Fig. 7.8c (Type C).

Figure 7.7: Schematic representation of the actuation/detection modes of the μ -CMFS. Reproduced from Fig. 3.2 for clarity.

(a) Device type A with typical dimensions.

(b) Device type B with large channel loop size.

(c) Device type C with large loop size and longer in-/outlet.

Figure 7.8: Microscope images of the fabricated devices. Important dimensions are indicated in the figure. Laser Doppler Vibrometer measurement points are located on the channel at the rotational axis of the actuation mode and 700 μ m to the left and right of this axis.

Figure 7.9: Schematic representation of the measurement setup. N_2 is fed through the sensor chip with an input pressure of 5 bar (gauge pressure). A Polytec MSA-400 Laser Doppler Vibrometer is used to measure the vibrations of the μ -Coriolis mass flow sensor.

Figure 7.9 shows the measurement setup. Nitrogen gas is fed through the sensor at an input pressure of $P_{in} = 5$ bar while the sensor is being actuated. The mass flow is incremented in steps of 0.15 g/h (for type A) or 0.075 g/h (for type B and C) from 0 to 1.2 g/h and decremented back to 0 g/h. A Polytec MSA-400 is used to measure the displacement of the channel at the measurement points indicated in Figure 7.8. From these measurements, the flow sensitivity $(\Delta \hat{z}_d / \hat{\theta}_a) / \phi_m$ is determined for each device.

7.3.2 Results & Discussion

The ratio $\Delta \hat{z}_d / \hat{\theta}_a$ was measured over a range of flows for the various devices (to later determine the flow sensitivity $(\Delta \hat{z}_d / \hat{\theta}_a) / \phi_m$). Figure 7.10 shows the measurement results as well as the response predicted by Eq. 3.15. As can be seen, the devices with a larger channel loop area have a higher flow sensitivity. A longer in-/outlet length also increases sensitivity, but there is more spread in the measurement points. This is most likely due to the fact that the actuation and detection frequencies are much closer to each other in value, which makes the sensitivity more dependent on the detection mode quality factor Q_d (see Equation 3.15). Q_d changes with temperature, air pressure, humidity etc., causing the output signal to drift.

The modelled and measured parameters of the sensors are compared in table 7.2. The model predicts the resonance frequencies within 10% error for all except Type A, Chip 2. There is a spread in the resonance frequencies of the fabricated devices and this device in particular is an outlier as can be seen from the table. The modelled

Figure 7.10: Measured and modelled flow sensitivities for the devices in Figure 7.8. Modelled flow sensitivities are estimated by combining output parameters of the FEM model with Equation 3.15.

		Type A			Type B	Type C	
		Chip 1	Chip 2	Chip 3	Chip 1	Chip 1	Chip 2
Actuation mode frequency f_a [Hz]	FEM Model	2602	2602	2602	576	394	394
	Measured	2382	2131	2486	544	386	398
	Error	+9.2%	+22%	+4.7%	+5.9%	+2.1%	-1.0%
Detection mode frequency f_d [Hz]	FEM Model	1593	1593	1593	359	309	309
	Measured	1491	1249	1547	350	301	306
	Error	+6.8%	+28%	+3.0%	+2.6%	+2.6%	+1.0%
Sensitivity $(\Delta z_d / \theta_a) / \phi_m$ [($\mu\text{m/rad}$) / (g/h)]	Model*	17.9	17.9	17.9	96.1	227	227
	Measured	17.0	18.6	-	67.2	188	-
	Error	+5.3%	-3.8%	-	+43%	+21%	-

Table 7.2: Comparison of the modelled and measured parameters of the sensor devices with N_2 as flow medium. Resonance frequencies f_a and f_d are simulated in the FEM model. '-' indicates that this parameter could not be measured. *The flow sensitivity is calculated by combining the output parameters of the FEM model and Equation 3.15.

sensitivities for type A are within 6% accurate. For Type B and C, the modelled sensitivity is predicted with approximately 20-40% error. This large error may be caused by the fact that there is a much larger pressure difference between the different sections of these large loop channel devices compared to a smaller loop device. A difference in pressure can cause a difference in stiffness of the channel sections [12]. Due to this and the fact that the density of the gas inside is also dependent on pressure, the shape of the resonance mode can also be influenced. Device B is 3.8 times more sensitive than the reference device A. Device C is 10.6 times more sensitive than Device A, but its output signal shows more spread.

7.3.3 Conclusion

Three types of μ -CMFS devices were designed and fabricated to validate the model presented in Chapter 3 and increase flow sensitivity. Flow sensitivity of the devices has been improved approximately 4-11 times with respect to a reference device. The FEM model predicts the resonance frequencies of the devices within 10% error for all except 1 (out of 6) device(s). The predicted sensitivities are within 6-40% accurate depending on the device type. Although flow sensitivity of the devices was improved, the most sensitive device (C) shows more drift in the output signal. This is most likely from dependence on the detection mode quality factor Q_d , since the actuation/detection mode resonance frequencies are closer together. Furthermore, device B showed higher flow sensitivity, but not as high as was expected. As briefly discussed in Chapter 3, a larger channel loop increases the required pressure drop over the device. Due to this fact and the results presented in this chapter, it can be concluded that the benefits of the larger channel loop devices do not outweigh the disadvantages. Most importantly however, the model has been proven to be able to predict resonance frequencies and flow sensitivity of μ -Coriolis mass flow sensors within a desired accuracy range. After validation through these experiments, it has been used to design novel μ -CMFS devices, which are presented in the following sections.

7

7.4 μ -Coriolis flow sensor with differential readout

As introduced in Chapter 2, external disturbances can induce vibrations in MEMS inertial sensors. This can cause errors in the detection of angular acceleration in a gyroscope or in the measured flow rate in the case of a μ -CMFS. As described in detail in Chapter 6, vibrations at an arbitrary frequency can cause a disturbance at the actuation mode resonance frequency due to non-linearity of the capacitive readout. This can cause a phase shift in the output signal translating to an unpredictable offset in the measured flow rate (see Fig. 6.15). A differential capacitive readout for a μ -CMFS was proposed in Chapter 6 to reduce this effect. By utilizing a double channel

loop design, Coriolis forces will be generated differentially. This way, common mode disturbances can be (partially) cancelled. A device was realized to investigate (the reduction of) the effect of external disturbances on the performance of the μ -CMFS.

7.4.1 Design

Figure 7.11 shows a schematic representation of the μ -CMFS with differential readout. Two identical suspended rectangular channel loops are mechanically connected through four membranes. This ensures that they act together as a single mechanical resonator. Metal tracks on top of the channel are used for actuation and readout of the sensor. Comb-shaped capacitive readout structures reside on two sides of the sensor for detecting actuation and Coriolis vibrations. The flow path is indicated in the figure by black arrows. This flow path ensures that Coriolis forces will be opposite in each channel loop. Thus, common mode vibrations can be partially cancelled out as described in Chapter 6 (Fig. 6.15).

The sensor design has been implemented in COMSOL Multiphysics[®] following the method presented in Chapter 3, to view the different modes of resonance. The fluid inside the channel is nitrogen gas. Fig. 7.12 shows the simulated resonance shapes of the modes of interest.

Figure 7.11: Schematic representation of a μ -CMFS with differential readout. Two mechanically connected rectangular channel loops act as a single mechanical resonator. Comb structures for readout reside on two sides of the sensor to detect and distinguish actuation/Coriolis vibrations and external disturbances. The flow path is indicated by black arrows.

Figure 7.12: Resonance modes of the devices simulated in COMSOL®, with N_2 as flow medium. The two mechanically connected channel loops move together as one mechanical resonator. The common and detection mode frequencies are close together due to the symmetrical design.

The actuation (Fig. 7.12a) and detection modes (Fig. 7.12b) are both a type of *Twist* mode, where each side of the channel loop moves in opposite direction. These mode shapes allow for the cancellation of common mode vibrations (Fig. 7.12c) using the detection scheme presented in Fig. 6.15 of Chapter 6.

7.4.2 Device & Setup

A microscope image of the fabricated device is presented in Fig. 7.13. The device is fabricated according to the process described in 4.2.

Figure 7.14 shows the measurement setup. It is based on the setup discussed in section 7.2 (Fig. 7.6). The typical flow range of a μ -CMFS is $0\text{--}1\text{ g h}^{-1}$. Accurate measurement of flow rate is most important in the lower flow range. The mass flow rate is controlled from $0\text{--}0.6\text{ g h}^{-1}$ in steps of 60 mg h^{-1} . A piezo-electric resonator is mounted on top of the printed circuit board (PCB) on which the sensor chip resides. This resonator is used to generate artificial external disturbances that cause vibration of the sensor chip. The resulting deflection amplitude of the channel is in the order of $1\text{ }\mu\text{m}$ (at the position of the readout capacitors), which is in the same order of magnitude as the actuation vibration and 1 to 2 orders of magnitude higher than the Coriolis vibration due to flow. As explained in Chapter 6, common mode vibrations of the sensor will cause an offset in the measured mass flow rate. To be able to distinguish the aforementioned offset from changes in flow rate or other possible causes of an offset, the piezo-electric resonator is driven with an amplitude modulated signal. The signal has a base frequency of around $1.6\text{--}1.7\text{ kHz}$, which is the resonance frequency of the common vibrational mode (Fig. 7.12c). This signal is modulated with a 0.1 Hz sine wave, such that the piezo-electric resonator goes from its off-state to fully exciting and back to off in a period of 10 s . Thus, the effect of the external disturbance will be clearly visible in the output phase signal.

Figure 7.13: Microscope photograph of the double channel loop μ -Coriolis mass flow sensor with differential readout. The readout capacitors are indicated, their numbers correspond to those in Fig. 6.15.

Figure 7.14: Schematic representation of the measurement setup. The setup is similar to that in Fig. 7.6. A piezo-electric resonator is mounted on top of the printed circuit board (PCB) with the sensor chip.

Figure 7.15: Deflection of the μ -Coriolis mass flow sensor measured using a Polytec MSA-400 LDV. The sensor is actuated in actuation and detection & common mode by rotating the magnetic field 90° in between measurements. The sensor is also actuated by driving the piezo-resonator to view in which frequency range disturbances would mainly occur.

7.4.3 Results and discussion

The resonance frequencies of the device are determined using a Polytec MSA-400 LDV. By doing this, it may be determined if the two mechanically coupled channel loops act as a single resonator. Additionally, the influence of artificial disturbance can be observed. The sensor is actuated in the actuation mode (Fig. 7.12a) and by rotating the magnetic field 90 degrees it is also actuated in the detection mode (Fig. 7.12b) and common/disturbance mode (Fig. 7.12c). To view the influence of external disturbances, the sensor is also actuated by solely driving the piezo-electric resonator attached to the chip board. The sensor is actuated by a periodic chirp signal, sweeping through a frequency range of 1–3.5 kHz. The Polytec MSA-400 is used to measure the deflection at the corner of the channel loop. Fig. 7.15 shows the obtained deflection spectra. To be able to properly compare the various spectra, the normalized deflection is plotted.

For each resonance mode, one clear peak can be observed. This indicates that there is in fact a single resonance mode for the two mechanically connected channel loops (for each mode respectively). The resonance frequencies of the three modes are $f_a = 2598$ Hz, $f_d = 1675$ Hz and $f_c = 1667$ Hz for the actuation, detection and common modes respectively. This corresponds to the simulated resonance frequencies (see Fig. 7.12). The deflection peak for when the sensor is driven by the piezo-resonator is almost fully overlapping with the common mode peak, indicating that most disturbance is generated in the common vibrational mode.

Figure 7.16: Phase difference output signal in relation to the applied flow rate for single ended and differential readout with and without artificially generated disturbance. Each data point represents the average phase shift over 200s at the indicated mass flow rate.

After confirming that the sensor acts as a single resonator, flow measurements are executed. A total of four measurements are done. Two measurements where only one set of capacitors is read out (top and bottom of Fig. 6.15, equivalent to the readout method in [9, 11, 13]) and two measurements where the sensor is read out differentially. For each set of measurements, one is executed with a piezo-electric resonator generating artificial external disturbances and one without, see Fig. 7.14. The resulting relation between the output phase shift and mass flow rate is displayed in Figure 7.16. The four measurements show equal flow sensitivity. This is to be expected, since the relation between the output phase shift and the detection mode vibration amplitude (dependent on the mass flow rate) does not change when switching from single ended readout to differential readout (see Fig. 6.15). Furthermore, the external disturbance generated by the piezo-resonator does not affect the flow sensitivity.

However, it does introduce an offset. As explained earlier, the offset introduced by external disturbances is visualized by using an amplitude modulated signal to drive the piezo-resonator. This causes an increase in mean absolute deviation (MAD) instead of an offset, see Fig. 7.17. When no external disturbance is induced, the differential and single ended methods show a nearly identical MAD ($0.6\text{--}1\text{ mg h}^{-1}$). However, when inducing external disturbances, the MAD is 60-70% lower when using a differential readout ($1.5\text{--}2.2\text{ mg h}^{-1}$) as opposed to single ended ($5.3\text{--}6\text{ mg h}^{-1}$).

Figure 7.17: Mean absolute deviation (MAD) of the phase difference output signal at each measured flow rate. Each data point represents the MAD over 200s at the indicated mass flow rate.

To more clearly view the effect of external disturbance on the phase shift output signal, a zero flow measurement is executed. During this measurement, the channel of the sensor is kept at a constant gauge pressure of 5 bar. The outlet of the channel is closed off, such that there is no flow. The output phase shift is measured for a duration of 1800s. Figure 7.18 shows the measurement results.

First, the output signal is viewed when no external disturbances are induced. The noise level of the output signal is similar for both readout methods. When viewing the output signal while external disturbances are induced, their effect can be clearly observed. The 0.1Hz modulated signal is clearly recognisable from the output phase. Thus external disturbances most definitely have a large influence on the performance of the μ -CMFS. A Fast Fourier Transform (FFT) of the output phase signal shows the magnitude of the 0.1 Hz modulated disturbance more clearly.

Figure 7.19 shows the power spectral density of the output phase signal. Here it can be observed that the differential readout reduces the effect of the artificial external disturbance by 64%. There are several effects which may explain why the disturbance is not fully cancelled out. As described in Chapter 6, the readout capacitance is non-linear with respect to deflection of the channel. There is always a combination of the actuation, detection and common mode vibrations. Thus the moving readout

Figure 7.18: Measured phase shift in zero-flow conditions for 1800s. Influence of external disturbance is clearly recognisable from the 0.1Hz modulation.

Figure 7.19: Normalized power spectral density of the zero-flow phase shift signals (Fig. 7.18). 64% reduction of the 0.1 Hz modulated disturbance can be observed.

electrodes on one side of the sensor do not experience the same movement with respect to their fixed counterparts as on the opposite side. Since the capacitance change with deflection is non-linear, this means that full cancellation is not possible. Another cause could be that since the common and detection modes are very close in frequency, the generated artificial disturbance may also partially excite the detection mode. Since in this mode the channel is vibrating in opposite directions on both sides of the sensor, this movement is not cancelled. Further improvements could increase the amount of reduction. Non-linearity could be reduced by increasing the gap (along z -axis) between the fixed and moving combs of the readout capacitors. If the combs are further away from each other, the relation between capacitance and deflection is linear for a larger range of deflection. Another option is to develop a differential readout based on strain gauges on top of or embedded in the micro-channel [3, 14]. These would suffer less from non-linearity issues than the current capacitive readout structures. Thus, disturbance signals at other frequencies than the actuation mode resonance frequency should no longer induce an offset. Furthermore, since the readout is linear in this case, full cancellation would be possible theoretically.

7.4.4 Conclusion

A μ -Coriolis mass flow sensor with differential readout was designed, fabricated and characterised. A comparison was made between the differential readout and a single ended readout (equivalent to the readout method in [9, 11, 15]). The differential readout reduces the effect of external disturbances (causing common mode vibrations of the micro-channel) by 60–70% in a flow range of 0–0.6 g h⁻¹. The effect of the external disturbances on the zero-flow stability is reduced by approximately 64%. Full elimination of external disturbances is not possible due to non-linearity of the readout capacitors. Reducing non-linearity of the existing capacitive readout or implementing (piezo-resistive) strain gauges (which suffer much less from non-linearity issues) would reduce/cancel the offset induced by external disturbances and improve cancellation of any disturbance signals close to or at the actuation frequency.

7.5 Suspended capacitive readout

In Chapter 6, the water hammer effect was introduced, by which gas bubbles present in a liquid flow can cause sudden in-plane motion of the μ -CMFS. The gap between the fingers of the readout electrodes is typically designed to be $5\text{ }\mu\text{m}$ (see Fig. 6.7 & Tab. 6.1 in Chapter 6). Thus, any vibration that results in the channel moving more than this distance (in-plane) at the position of the readout electrodes, can cause damage or breakage of the readout structures. Since this is a significant limitation to the performance of the μ -CMFS, this issue needs to be addressed. In this section, a solution to this problem is proposed. By suspending the capacitive readout structures via springs, the electrodes can move with the micro-channel in-plane. The suspension needs to be designed in such a way, that only the electrode attached to the main channel (through which fluid flows) moves during out-of-plane vibrations. The μ -CMFS can then be operated as usual, but would allow for the channel to move in-plane without damaging the capacitive readout structures. Devices with such a suspended readout were designed and characterization results will be presented.

7.5.1 Design

In the SCT fabrication process, there are two main ways to realize spring structures. One can use the micro-channel itself or use thin SiRN membranes. The micro-channel diameter can be chosen by using one or more rows of slits (see Chapter 4). This way the channel can be made more or less stiff. A combination of micro-channels and membranes as suspension have been used to design two devices. In both devices, the micro-channel and spring structures are suspended from the same position on the silicon bulk. This is to avoid stress being developed due to the difference between thermal expansion coefficients of Si and SiRN. Figure 7.20 shows design A. A second channel loop is positioned inside the main channel loop through which fluid flows. $50\text{ }\mu\text{m}$ wide SiRN membranes connect the two channel loops such that they move together when in-plane forces are exerted. The diameter of the inner channel is designed to be $100\text{ }\mu\text{m}$ by using three rows of slits for the channel etch as opposed to a single row of slits for the outer channel (resulting in a diameter of $55\text{ }\mu\text{m}$). This should make the inner channel stiffer such that it will not move with the outer loop in the *Swing* and *Twist* modes. Readout electrodes reside on the outer channel loop (outer electrodes) and on the inner channel loop (inner electrodes). Figure 7.21 shows design B. A similar approach has been taken as for design A. A structure is realized within the main channel loop, once again connected by $50\text{ }\mu\text{m}$ wide membranes. It consists of a $82\text{ }\mu\text{m}$ channel, splitting into two $55\text{ }\mu\text{m}$ channels to which the inner readout electrodes are attached. The different shape and high stiffness of the structure ensures that it does not vibrate with the main channel loop in the *Swing* and *Twist* modes.

Figure 7.20: Schematic representation of design A. An inner channel loop consisting of a 100 μm wide channel serves as suspension for the inner readout electrodes. Thin SiRN membranes connect the two channel loops. For clarity, wiring needed for actuation is not shown.

Figure 7.21: Schematic representation of design B. The channel loop is connected to an inner suspension. The suspension consists of an 82 μm wide channel, which splits into two 55 μm channels. For clarity, wiring needed for actuation is not shown.

Figure 7.22: Vibrational modes of device type A simulated in COMSOL®, with air as flow medium. The actuated channel loop resonates separately in the *Twist* and *Swing* modes. For the in-plane mode, both channel loops move together such that there is (virtually) no relative displacement between the readout electrodes.

Figure 7.23: Vibrational modes of device type B simulated in COMSOL®, with air as flow medium. The actuated channel loop resonates separately in the *Twist* and *Swing* modes. For the in-plane mode, the channel loop and suspension move together such that there is (virtually) no relative displacement between the readout electrodes.

The designs are analyzed in COMSOL Multiphysics® using the model presented in Chapter 3. Figure 7.22 shows the simulated vibrational modes of interest for design A. For the *Twist* mode (Fig. 7.22a), the outer channel loop clearly moves separately from the inner loop. The connecting SiRN membranes act as flexible springs, allowing for the outer loop to move freely. To simulate the effect of Coriolis forces, a force is exerted on the channel section to which the outer electrodes are attached, at the *Twist* mode resonance frequency (Fig. 7.22b). The inner channel loop moves only slightly while the outer loop is actuated. However, when viewing the in-plane mode (Fig. 7.22c), the connecting springs force the two channel loops to move together. Thus, the readout electrodes will move together as well, avoiding collision. Figure 7.23 shows the simulated vibrational modes of design B. Once again, the outer channel loop can vibrate separately from the inner suspension in the *Twist* (Fig. 7.23a) and *Swing* (Fig. 7.23b) modes. For the in-plane vibrational mode, the inner suspension moves with the outer channel loop due to the connecting spring structures. Just as for design A, this should avoid relative displacement between the readout electrodes when a sudden in-plane force is exerted on the channel.

Figure 7.24: Microscope photos of the fabricated device type A. Important features are highlighted and enlarged for clarity.

When simulating the resonance modes with water as flow medium, all resonance frequencies shift downwards by approximately 25–30%. Furthermore, the mode shapes remain the same as shown in Figs. 7.22 and 7.23. Therefore, it can be assumed that the suspension operates similarly while using a liquid or gas as flow medium.

7

7.5.2 Devices & Setup

Figure 7.24 shows the fabricated device of type A (corresponding to design A, Fig. 7.20). The membrane springs and readout electrodes are highlighted in Fig. 7.24a and enlarged in Fig. 7.24b and 7.24c respectively. The readout electrodes are connected to the channels by thin membranes to allow for sufficiently large windows through which the Si bulk can be etched away during release of the device (Appendix A.1, step 40).

Figure 7.25 shows the fabricated device of type B (corresponding to design B, Fig. 7.21). The point where the inner suspension channel splits into two separate channels

(a) Overview of the device.

(b) Inner suspension channel splitting into two.

(c) Spring connecting the channel loop and inner suspension.

Figure 7.25: Microscope photos of the fabricated device type B. Important features are highlighted and enlarged for clarity.

is highlighted in Fig. 7.25a and enlarged in Fig. 7.25b. One can see that the channel starts out with two rows of slits that split into two separate rows, forming smaller channels. The membrane springs are highlighted and enlarged as well (Fig. 7.25c). They connect the outer channel loop to the inner suspension at the point where the inner readout electrodes reside.

The devices are first characterized by means of out-of-plane resonant analysis using LDV to characterize the *Twist* and *Swing* vibrational modes. Subsequently, in-plane motion analysis is done using SVM. Both out-of-plane and in-plane analyses are done with a Polytec MSA-400. Finally, flow measurements are executed to determine performance of the sensors. See 7.2 for more details on the characterization methods.

Figure 7.26: Out-of-plane deflection measured by LDV. The devices are actuated in the actuation and detection modes by rotating the magnetic field 90° in between measurements. The channels are filled with air.

7.5.3 Results & Discussion

The actuation/*Twist* and detection/*Swing* modes of the two devices are characterized by applying a periodic chirp actuation signal sweeping through a frequency range of 1–3 kHz and measuring the deflection using LDV. The magnetic field is rotated 90° between measurements to be able to actuate both modes. During the measurements, the channels are filled with air. The measured responses are shown in Figure 7.26.

The main *Swing* mode resonance frequency is 1.52 kHz. This differs around 20% from the modelled resonance frequency (1.87 kHz). As discussed in 7.3, there is always a spread in parameters between devices, wafers and fabrication runs. If the channel dimensions are slightly different than the ones used for simulation, this can already change the resonance frequencies by 10–20% (see Tab. 7.2). A second peak can be observed at a frequency of 2.82 kHz, which is where the two channel loops vibrate in opposite direction. The actuation mode resonance frequency is considerably lower than the simulated value (Fig. 7.22a). When viewing the measured response of device B (Fig.

(a) *Twist*/Actuation mode, $f_0 = 2.43$ kHz.

(b) *Swing*/Coriolis mode, actuated at $f_0 = 2.43$ kHz.

Figure 7.27: Vibrational mode shapes of device type A obtained by LDV. The micro-channels are filled with air.

7.26a), the same can be observed. The actuation mode resonance frequency is much lower than simulated (Fig. 7.23a). As aforementioned, a 10–20% error in resonance frequency with respect to the model is not out of the ordinary. Furthermore, since the FEM model is made up of 1D line elements (Fig. 3.3), the connections between the spring structures and channels might not be accurately represented. This could further increase the error. The measured *Swing* mode resonance frequency for device B is 1.91 kHz, which corresponds well to the simulated value of 1.87 kHz.

In addition to the frequency response, the mode shapes of the devices are also viewed. First the devices are actuated in the *Twist* mode. To see the effect of Coriolis forces, the magnetic field is then rotated 90° and the devices are once again actuated at the *Twist* mode resonance frequency (since Coriolis forces are also induced at this frequency). The observed mode shapes for device A are displayed in Figure 7.27. For the actuation/*Twist* mode, the deflection of the outer channel loop is much larger than that of the inner channel loop. However, for the *Swing* mode (at the *Twist* mode resonance frequency), the inner channel loop vibrates with the outer loop. This is not as it shows in simulation, see Fig. 7.22b. This could be due to the same effect that has

Figure 7.28: Vibrational mode shapes of device type B obtained by LDV. The micro-channels are filled with air.

caused a large difference between the simulated and measured *Twist* mode resonance frequencies. The fact that both the inner and outer channel loops move, will result in a different change in readout capacitance due to Coriolis forces than for a device with a fixed capacitive readout. The observed mode shapes for device B are displayed in Figure 7.28. For both modes, the channel loop moves almost entirely separate from the inner suspension. Thus, vibrations in both modes should have a similar effect on the readout capacitance as for a typical μ -CMFS device with fixed readout electrodes.

Now that the *Swing* and *Twist* modes have been characterized, the in-plane mode will be viewed. The devices are actuated by inducing in-plane Lorentz forces through an out-of-plane magnetic field. The resulting vibrations are viewed by SVM using a Polytec MSA-400. See Fig. 7.3 for the measurement setup. The resonance frequencies of the two in-plane modes are found to be 4.36 kHz for device A and 4.55 kHz for device B. This is again significantly lower than the simulated frequencies (Fig. 7.22c and Fig. 7.23c), most likely due to the same effect that caused the mismatch in simulated and measured resonance frequencies for the other modes. The positions of the readout electrodes are monitored while actuating the devices in the in-plane mode.

(a) In-plane displacement of the two sets of readout electrodes. (b) Relative displacement between the inner and outer readout electrodes.

Figure 7.29: In-plane displacement of the inner and outer readout electrodes for device A. A reduction of 96% and 40%, along x and y axis respectively, can be observed when comparing the relative displacement between electrodes to absolute displacement of the channel loop.

Since the actuation mode resonance frequencies of the devices are much lower than the in-plane mode resonance frequencies, in reality, potential in-plane forces will be induced at frequencies much lower than the in-plane mode resonance frequency. In this frequency range, the in-plane mode will be excited quasi-statically. However, during the measurements, the devices are actuated at resonance to induce a large enough vibrational amplitude to be able to observe the motion of the channels and suspension.

Figure 7.29 shows the in-plane displacement of the readout electrodes for device A over one period T . A maximum displacement along the x -axis (Δx) of just over 5 μm is reached for the channel loop where the outer electrodes reside, see Fig. 7.29a. The inner electrodes show an almost identical magnitude. Displacement along the y -axis can also be observed, most likely due to another in-plane vibrational mode relatively close in frequency to the actuated mode. This is confirmed by the fact that the two vibrations are 90° out of phase. For displacement in the y -direction (Δy), the difference between the electrodes is larger than for displacement along the x -axis. However, the magnitude of Δy is relatively small compared to Δx . The relative displacement between the readout electrodes is plotted in Fig. 7.29b. Along the x -axis, the maximum displacement is only 0.2 μm , for the y -axis it is 0.8 μm . This is a 96% and 40% reduction, along x and y axis respectively, compared to the absolute in-plane displacement of the channel loop. As mentioned in Chapter 6, the typical gap between the comb fingers of the readout electrodes is around 5 μm and 6 μm between the comb fingers and the opposite comb base (see Fig. 6.7). This would mean that the channel loop can withstand in-plane deflections of up to 135 μm (along the x -axis) before the

(a) In-plane displacement of the two sets of readout electrodes. (b) Relative displacement between the inner and outer readout electrodes.

Figure 7.30: In-plane displacement of the inner and outer readout electrodes for device B. A reduction of 99% and 63%, along x and y axis respectively, can be observed when comparing relative displacement between the electrodes to absolute displacement of the channel loop.

electrode fingers collide. However, the comb finger and opposite comb base would collide at an in-plane displacement of the channel loop of 40 μm along the x -axis. Since the readout capacitance is mainly defined by the gap between the fingers, the gap in the y direction could be made larger by design. This would allow for more in-plane deflection of the channel loop without drastically reducing the readout capacitance.

The measured in-plane displacement for device B is shown in Figure 7.30. For Δx , its magnitude is nearly identical for both electrodes (see Fig. 7.30a). The difference for Δy is larger, but still relatively small. The vibrations along the two axes are 180° out of phase, indicating that they are caused by one and the same resonance mode. The relative displacement between the readout electrodes is depicted in Fig. 7.30b. A reduction of 99% and 63% is achieved, along x and y axis respectively, compared to the absolute in-plane displacement of the channel loop. This translates to a maximum in-plane displacement for the channel loop along the x -axis of around 340 μm, when viewing only collision between the electrode fingers. Once again, the deflection in the y -direction is the bottleneck with a maximum in-plane displacement of the channel loop along the x -axis of 68 μm. As aforementioned, this could be solved by designing the gap between the electrodes along the y -axis larger. However, a potentially better solution would be to apply similar spring structures (as have been used to reduce the relative displacement in the x -direction) to significantly reduce relative y displacement. This could be a relatively simple way to achieve similar reductions of up to 99%.

Figure 7.31: Phase difference output signal in relation to the applied flow rate for device A. Each data point represents the average phase shift over 200s at the indicated mass flow rate.

Now that out-of-plane and in-plane resonance modes have been characterized, flow measurements are executed. The mass flow rate is controlled from $0\text{--}0.6\text{ g h}^{-1}$ in steps of 60 mg h^{-1} . Figure 7.31 shows the resulting output phase difference in relation to the applied flow rate for both device types. Device A shows a flow sensitivity of roughly a factor 4 lower than that of device B. This can be attributed to the fact that *Swing* mode vibrations at the actuation frequency cause both the inner and outer channel loop to move (see Fig. 7.27a). This causes a smaller change in readout capacitance due to Coriolis forces. For device B, the flow sensitivity is comparable to other μ -CMFS devices, e.g. the differential device discussed in the previous section (Fig. 7.16) or the device presented in [15].

7.5.4 Conclusion

As discussed in the beginning of this section, the water hammer effect is one of the most significant limitations for the performance of the μ -Coriolis mass flow sensor. To address this issue, the allowed magnitude of in-plane vibration of the micro-channel ($5\text{ }\mu\text{m}$ for μ -CMFS device with fixed capacitive readout [11, 15]) must be increased significantly. Two μ -CMFS devices have been realized that can withstand in-plane vibrations of up to 40 and $68\text{ }\mu\text{m}$ (along the x -axis) for the respective device types. This is approximately a factor 8 and 14 improvement respectively. The maximum allowed

in-plane displacement of the channel can potentially be increased to 140–340 μm by adapting or integrating additional spring structures to suppress relative displacement of the readout electrodes along the y -axis or increasing the gap between readout electrodes in the y -direction. This would be a potential improvement of 30–60 with respect to a device with a fixed capacitive readout (e.g. the devices in [11, 15]). Flow measurements have shown that these devices with suspended capacitive readout can operate as usual, although device A has shown some decreased performance. Device B in particular displays promising performance. With small alterations to the design, it would potentially be able to withstand in-plane deflections of the channel loop up to 340 μm , without compromising its performance as a flow sensor.

7.6 μ -Coriolis flow sensors with increased stiffness

In the previous section, μ -CMFS devices with suspended capacitive readout were presented to overcome breakage by sudden in-plane motion of the micro-channel. Another way to adapt to this effect, is to increase the stiffness of the channel loop. Other advantages of larger stiffness are that it increases the resonance frequency and Q factor of the device. The first can be beneficial for robustness against external vibrations and noise performance of the device (related to $1/f$ noise, see Chapter 2). Increasing Q can lead to higher flow sensitivity and stability. Devices were designed and fabricated with larger stiffness, ω_0 and Q . In this section the design process and characterization results are presented.

7.6.1 Motivation

In Chapter 3, it was shown that the highest flow sensitivity is achieved when the actuation and detection mode resonance frequencies of the μ -CMFS are close together in value (see Eq. 3.15 and Fig. 3.5). However, as was stated, in this region a small change in Q_d or ω_d/ω_a can cause a significant change in the flow sensitivity. This effect was discussed by Haneveld et al. [9]. They chose the resonance frequencies such that the ratio between the two (ω_a/ω_d) is larger than 1.4. In this way a change in the Q factor hardly affects the flow sensitivity as long as it is higher than 30. Instead of designing the device such that $\omega_a/\omega_d > 1.4$, one can also try to increase Q such that the resonance peaks can be closer together. The Q factor is defined as:

$$Q = \frac{f_0}{\Delta f} \quad (7.1)$$

Where f_0 is the resonance frequency and Δf is the width of the resonance peak at half of its maximum value (Full Width at Half Maximum, FWHM). Figure 7.32 shows how Q and f_0 influence the shape of the resonance peaks of a μ -Coriolis mass flow

(a) Typical resonance frequencies and Q for a μ -Coriolis mass flow sensor device. (b) Increasing the Q factor makes the resonance peaks sharper.

(c) When increasing Q while increasing f_0 by same factor, resonance peaks retain shape.

Figure 7.32: Representation of the effect of the Q factor and resonance frequency on the deflection spectrum of a μ -Coriolis mass flow sensor.

sensor. The normalized deflection is calculated for different values of Q and f_0 using a basic 2nd order transfer:

$$|\theta_{a,d}(\omega)| = \frac{|\tau_{a,d}(\omega)|}{K_{a,d} \sqrt{\left(1 - \left(\frac{\omega}{\omega_{a,d}}\right)^2\right)^2 + \left(\frac{\omega}{\omega_{a,d} Q_{a,d}}\right)^2}} \quad (7.2)$$

Where the subscript a,d indicates the actuation or detection mode respectively. Fig. 7.32a shows the deflection spectrum for typical values of f_d , f_a , Q_d and Q_a . It can be observed that the position on the detection mode curve where Coriolis forces are induced (indicated by dashed line) is relatively far away from the peak. Where a change in Q_d will affect the deflection closer to the peak significantly, at this position, the

magnitude will barely change. When increasing the Q factor (Fig. 7.32b), the resonance peaks become sharper. This allows to design them closer together to increase the flow sensitivity while achieving similar dependency on Q_d as for Fig. 7.32a. Alternatively, one could increase the resonance frequencies of the device (see Fig. 7.32c). If the resonance frequency is increased by the same factor as Q , Δf remains the same. Thus, the absolute difference between the resonance frequencies can be kept the same while also retaining the same dependency on Q_d . Although the absolute difference between ω_d and ω_a is the same, the ratio ω_d/ω_a becomes smaller. This theoretically should increase the flow sensitivity (see Chapter 3, Eq. 3.15).

As briefly discussed in Chapter 2, an increased resonance frequency can also reduce the effect of vibrations induced by environmental excitation. Since external vibrations involved with most applications (automotive and aerospace among others) are in a frequency range <2 kHz [16], increasing the resonance frequency of the μ -CMFS well above this range should lead to improved robustness against these vibrations.

Finally, as aforementioned, increasing stiffness of the μ -CMFS device structure can improve suppression of in-plane motion. This can prevent breakage of the capacitive readout structures. Increasing stiffness goes together with higher resonance frequency and Q factor. Thus, a design can be made with higher stiffness, leading to improved suppression of in-plane vibration and the previously discussed additional benefits.

7.6.2 Design

As could be seen in the μ -CMFS designs presented so far, the channels of the device serve as its suspension. If focusing only on suppression of in-plane motion of a suspended channel, the ideal design would be a channel fixed at both ends. However, this of course does not allow for rotation of the channel around the axis perpendicular to the direction of flow. To allow for this rotation, the suspension would ideally be infinitely stiff in the in-plane direction while having finite stiffness in the out-of-plane direction. A micro-channel following a circular path can be used to achieve a relatively stiff in-plane suspension, while being more flexible out-of-plane. On this basis, a design of a μ -CMFS was made in COMSOL Multiphysics®. Figure 7.33 shows the basic design of the large stiffness μ -CMFS realized in COMSOL®. The design consists of a straight channel with length L , which splits into two channels on both sides. These channels follow a circular path (with radius r) and are fixed at their ends such that they serve as springs. This double circular suspension is very stiff in the in-plane direction, while allowing for the center channel to move up/down and rotate around the axis perpendicular to the direction of flow (indicated by blue arrows in Fig. 7.33).

The modes of interest are simulated in COMSOL® and are displayed in Figure 7.34. For this simulation, a channel radius of $55\text{ }\mu\text{m}$ is chosen. Furthermore, $L = 4000\text{ }\mu\text{m}$ and $r = 800\text{ }\mu\text{m}$. This set of dimensions is an example to show the resonant behaviour of

Figure 7.33: Basic design of a large stiffness μ -CMFS made in COMSOL[®]. The double circular suspension suppresses in-plane motion while being relatively more flexible out-of-plane. r and L can be varied to control the resonance frequencies of the device. Blue arrows indicate the flow path.

(a) Detection/*Swing* mode, $f_0 = 8.7$ kHz. (b) Actuation/*Twist* mode, $f_0 = 9.6$ kHz. (c) In-plane mode, $f_0 = 22.2$ kHz.

Figure 7.34: Simulated vibrational modes of the μ -CMFS design in Fig. 7.33 with $L = 4000$ μm , $r = 800$ μm and a channel radius of 55 μm . N_2 is used as flow medium.

such a geometry. In the *Swing* mode, the entire center channel vibrates up-/downwards (Fig. 7.34a). This shows that the circular suspension is working as desired. In the *Twist* mode (Fig. 7.34b), the center channel rotates around the axis perpendicular to the flow direction. Thus when actuating this mode and a fluid flows through the channel, Coriolis forces will be generated in the center channel, exciting the *Swing* mode (Fig. 7.34a). Fig. 7.34c shows that the in-plane mode is at a much higher frequency and thus the suspension is significantly more stiff in this mode.

As aforementioned, the expansion coefficients of SiRN and Si do not have the same value. Thus, when the temperature of the structure increases, the SiRN channel will expand at a different rate than the Si substrate it is attached to. To reduce this effect, the points where the structure is fixed (suspension points) to the substrate are typically designed in close proximity to each other. For the design in Fig. 7.33, this is not the case. However, it can be achieved by using larger micro-channels that act as a rigid connector. Figure 7.35 shows an adapted design where this technique is applied. Three parallel channels with a radius of 100 μm (as opposed to 55 μm for the rest of the structure) connect the circular suspension to a central suspension

Figure 7.35: Adapted design of the one depicted in Fig. 7.33 with reduced distance between suspension points. A rigid suspension is formed by three parallel channels with a radius of 100 μm .

Figure 7.36: Simulated vibrational modes of the μ -CMFS design in Fig. 7.35 with $L = 4000$ μm and $r = 800$ μm . The resonance frequencies differ only slightly from those in Fig. 7.34. N_2 is used as flow medium.

point. To show that these structures indeed act as a rigid connector, the adapted design is simulated in COMSOL[®]. Figure 7.36 shows the simulated modes of interest. The vibrational mode shape of each mode is nearly identical to those in Fig. 7.34. The parallel channel suspension does (virtually) not move with the rest of the channel structure. The resonance frequencies vary slightly when compared to those in Fig. 7.34, showing that the parallel channels do not act as an ideal rigid connector. However, they do show to be significantly more stiff than the circular suspension.

7

7.6.3 Devices & Setup

Based on the basic design presented in Fig. 7.35, four devices are designed and fabricated. Microscope photos of the four fabricated devices are displayed in Figure 7.37. The design parameters are given in Table 7.3. Various combinations of L and r are chosen such that the devices show a spread in resonance frequency as well as in the relative difference between the *Twist* and *Swing* mode resonance frequencies. The concept is that devices A and B would be in a slightly higher frequency range than devices C and D to see the effect of frequency on Q . Originally, devices A and C were designed to have a relatively larger difference between f_t and f_s than devices B

Table 7.3: Design parameters of the devices.

Device	L (μm)	r (μm)	f_s (kHz)	f_t (kHz)
A	4800	600	7.7	9.2
B	3050	600	9.8	10.1
C	4400	800	5.4	5.6
D	2650	800	5.9	5.9

and D. However, there are slight differences in layer thicknesses and channel radii of the designed and fabricated devices. Therefore, the simulated frequencies are closer together than initially designed, in particular for devices B, C and D. This means that these devices will most likely have overlapping resonance peaks. As discussed before, this results in a high dependency of the flow sensitivity on Q . However, the simulated resonance frequencies of device A show a similar relative difference as depicted in Fig. 7.32. Thus, this device seems a promising candidate for testing if increased stiffness does indeed lead to increased Q and flow sensitivity (as discussed in 7.6.2).

(a) Device A

(b) Device B

(c) Device C

(d) Device D

Figure 7.37: Microscope photos of the fabricated devices. Four devices are realized with the design parameters in Tab. 7.3.

To characterize the devices, resonant analysis is first executed by Laser Doppler Vibrometry (LDV). The LDV measurement setup is depicted in Fig. 7.2. After the resonant analysis, flow measurements are performed. The flow measurement setup is shown in Fig. 7.6. See 7.2 for more details on the characterization methods.

(a) Device A, $f_d = 6.93$ kHz, $f_a = 8.04$ kHz. (b) Device B, $f_d = 8.71$ kHz, $f_a = 8.90$ kHz.

(c) Device C, $f_d = 4.64$ kHz, $f_a = 4.71$ kHz. (d) Device D, $f_d = 5.07$ kHz, $f_a = 4.89$ kHz.

Figure 7.38: Deflection spectra of the devices, measured by LDV. During the measurements, the micro-channels are filled with air.

7.6.4 Results & Discussion

The actuation/*Twist* and detection/*Swing* modes of the devices are characterized by applying a periodic chirp actuation signal sweeping through a frequency range of 1–12 kHz and measuring the deflection by LDV. The magnetic field is rotated 90° between measurements to be able to actuate both modes. The obtained deflection spectra are displayed in Figure 7.38. As predicted, all devices except A have overlapping resonance peaks. As aforementioned, this is due to slight differences between dimensions of the designs and fabricated devices. Focusing on device A, the Q factors of the two modes are higher than typical values (Fig. 7.32a). Therefore, the peaks are

Figure 7.39: Output of the capacitive readout in relation to the applied flow for device A. Each data point is an average of the output signal over a period of 200 seconds at a stable flow. The flow sensitivity is $3.77\text{ }^{\circ}/(\text{g h}^{-1})$.

relatively sharper and do not overlap. The relative factor between the resonance frequencies is similar to that in Fig. 7.32. Another important observation is that there is a secondary *Swing* mode close to the actuation mode resonance frequency. This is similar as for the device discussed in 7.5 (Fig. 7.26a). In this mode, the center channel moves in opposite direction as the circular suspension. Coriolis forces may also induce (counteracting) vibrations in this mode. This can cause a reduced net vibration amplitude due to Coriolis forces (from flow) and therefore result in a lower flow sensitivity. As aforementioned, environmental vibrations in most applications typically occur in a frequency range $<2\text{ kHz}$. In 7.4 (Fig. 7.17), it was shown that external vibrations at the frequency of one of the resonance modes can cause errors in the measured flow rate. Since the resonance modes of the devices presented in this section are well above 2 kHz , environmental disturbances in this frequency range would cause vibration amplitudes (and errors in flow rate) of 1–2 orders of magnitude lower than for the device in 7.4 (when using single ended readout).

To investigate the sensor's performance, flow measurements are executed. As aforementioned, all devices except device A have overlapping resonance peaks. This causes both significant *Swing* and *Twist* mode vibrations when the devices are actuated. The capacitive readout is designed for *Twist* mode actuation with very small *Swing* mode vibrations due to Coriolis forces. With the current design of the capacitive readout, the devices with overlapping resonance frequencies cannot be characterized. Additionally (as discussed earlier), these devices will have a high dependency on Q . For these reasons, only device A is characterized by execution of flow measurements.

Figure 7.40: Phase difference output of the capacitive readout over a period of 1800 seconds for device A. The sensor’s channel is kept at a constant pressure of 5 bar (gauge pressure) while its outlet is closed off.

The mass flow rate is controlled from $0\text{--}0.6\text{ g h}^{-1}$ in steps of 60 mg h^{-1} . Figure 7.39 shows the resulting output phase difference in relation to the applied flow rate for device A. The flow sensitivity of the device is $3.77\text{ }^{\circ}/(\text{g h}^{-1})$. This is lower than for a μ -Coriolis mass flow sensor of typical dimensions, e.g. device B in Fig. 7.31. This is most likely due to counteracting Coriolis vibrations in the secondary *Swing* mode. To view the zero stability of the device, a zero flow measurement is executed. During this measurement, the channel of the sensor is kept at a constant gauge pressure of 5 bar. The outlet of the channel is closed off, such that there is no flow. The output phase shift is measured for a duration of 1800s. Figure 7.40 shows the measurement results.

The standard deviation of the output phase difference over this period is 0.002° . This translates to a zero-flow stability (using the flow sensitivity of $3.77\text{ }^{\circ}/(\text{g h}^{-1})$) of 0.64 mg h^{-1} . This is very similar to that presented in [15] for nitrogen. Combining device A with the electronic readout presented in [15] could lead to an improved zero-flow stability due to lower $1/f$ noise and lower drift of the phase difference output signal. Another option would be to redesign the device such that the secondary *Swing* mode is at a much higher frequency. This way, a higher flow sensitivity could be achieved. If the same zero stability in terms of phase shift can be kept, this would improve the zero-flow stability in terms of flow rate.

7

7.6.5 Conclusion

Large stiffness μ -CMFS devices have been designed, fabricated and characterized with higher resonance frequencies and Q factors. Due to differences between designed and fabricated dimensions, only one device could be fully characterized. A relatively low flow sensitivity ($3.77\text{ }^{\circ}/(\text{g h}^{-1})$) was observed, most likely due to the presence of a secondary *Swing* mode in which Coriolis forces induce counteracting vibrations. The obtained zero-flow stability of 0.64 mg h^{-1} is similar to that achieved before by

Oliveira et al. [15] (0.6 mg h^{-1} for N_2). Combining the device with the dedicated CMOS interface from [15] could lead to better zero-flow stability. Furthermore, redesigning the device such that the secondary *Swing* mode does not interfere with its operation, could lead to a much higher flow sensitivity and zero-flow stability. Additionally, due to the larger stiffness and higher resonance frequencies, the device should have improved robustness against in-plane vibrations due to the water hammer effect as well as against environmental vibrations (which occur in a frequency range of $<2 \text{ kHz}$ for most applications [16]). This could be confirmed in the future through similar methods as presented in 7.5 and 7.4.

7.7 Resistive readout using metal strain gauges

A capacitive readout for a μ -Coriolis mass flow sensor has certain limitations. As was shown in 7.4, external disturbances can cause an unpredictable offset in the measured flow rate. Furthermore, non-linearity of the readout causes a dependence of the sensitivity (to displacement of the micro-channel) on the actuation amplitude, initial distance between the electrodes and pressure (see Chapter 6). In addition, one of the most significant limitations is the aforementioned water hammer effect, which can cause breakage of the readout structures. In Chapter 6, an alternative readout was proposed based on strain gauges on top of or embedded in the micro-channel. This type of readout would be more linear and does not risk breakage as a result of in-plane vibration of the channel (due to the water hammer effect). There are two ways of realizing such a readout. One is to use the conventional SCT process (see 4.2) and pattern metal strain gauges on top of the micro-channel. Another way is to embed doped silicon strain gauges in the micro-channel roof by the process described in 4.3.1 or 4.3.3. Incorporation of metal strain gauges does not require adaptation of the conventional SCT fabrication process. Therefore this option was explored first. μ -CMFS devices with a resistive readout based on metal strain gauges are presented in this section. Devices with silicon strain gauges are presented in Section 7.8.

7.7.1 Design

Three μ -CMFS designs with a resistive readout (based on metal strain gauges) were realized. Two of the designs are based on the same channel loop structure, with different strain gauge designs. The third design exploits a different channel loop geometry. Figures 7.41 and 7.42 show the two channel structures.

Both designs are based on a channel loop of $L_x \times L_y = 5 \times 3$ mm, which are typical dimensions of a μ -CMFS (e.g. the devices in 7.5). Designs A and B share the same channel loop structure (Figure 7.41). The strain gauges in these designs reside on the channel sections with length l_1 . Additional membranes with length l_2 connect the channel to the substrate to allow for the actuation track. For design A, the strain gauges are patterned at a 45° angle to measure torsion of the channel section with length l_1 . This makes the design relatively more sensitive to *Swing* mode vibrations. Thus, this design is designed for *Twist* mode actuation (*Swing* mode Coriolis forces). For Design B, the strain gauges are patterned parallel to the channel section with length l_1 to measure bending along its axis. This makes the device relatively more sensitive to *Twist* mode vibration. Thus, it is designed for *Swing* mode actuation (*Twist* mode Coriolis forces). Microscopic photographs of the fabricated devices presented later on in Fig. 7.44 will show the difference between the two designs more clearly. Design C is based on a T-junction channel structure connected through two membranes to

Figure 7.41: Design A/B: Two strain gauges are positioned on the in-/outlet channel sections. For design A, the strain gauge is designed to measure torsion of the channel section with length l_1 around its axis. For design B, the strain gauge is designed to measure bending of the same channel section along its axis. Two membranes with length l_2 connect the channel loop to the substrate to allow for the metal actuation track.

Figure 7.42: Design C: A T-junction channel structure is connected through two membranes to the main channel loop. Four strain gauges reside on the membranes with length l_3 . Two experience elongate, while the other two experience compression.

Figure 7.43: Vibrational modes for the different device types, simulated in COMSOL®. The normalized stress is displayed to show where it is highest. Based on this, the locations of the strain gauges were chosen. In the simulation, the micro-channels are filled with N₂.

the main channel (Fig. 7.42). This way, stress is introduced in a specific location by design. Four strain gauges reside on the two membranes with length l_3 . Two will elongate, while the other two are compressed. This design can be used for both modes of actuation. See 6.4 for more details on the operating principle of the resistive readout.

The vibrational modes for the devices were simulated in COMSOL Multiphysics® and are depicted in Figure 7.43. The resonance frequencies for device A&B are $f_t = 2426$ Hz and $f_s = 1312$ Hz for the *Twist* and *Swing* mode respectively. The stress is highest in the channel sections with length l_1 . Therefore, the two designs were made as aforementioned, with A optimized for *Twist* mode actuation and B optimized for *Swing* mode actuation. For design C, the resonance frequencies are $f_t = 2519$ Hz and $f_s = 1699$ Hz for the *Twist* and *Swing* mode respectively (Fig. 7.43c&d). The *Swing* mode resonance frequency is most affected by the additional T-shaped channel structure and suspension. This indicates that the membranes with length l_3 (see Fig. 7.42) increased the overall stiffness of the device in this mode. Additionally, in this mode, stress is highest where the two membranes are attached to the main channel loop and the T-junction channel structure. In the *Twist* mode, the stress is relatively

lower and the resonance frequency is less affected. Therefore, one would expect for these dimensions that the device is more sensitive to *Swing* mode vibration than *Twist* mode vibration. The sensitivity to either mode of actuation could be tuned by increasing/decreasing the distance between the two membranes, increasing/decreasing the sensitivity to *Twist* mode vibration.

7.7.2 Devices & Setup

Figure 7.44 shows microscope images of the three sensor types that were designed and fabricated. The position of the strain gauges and reference resistors are indicated in the figure. The resistance value ratios between the devices are 1:0.73:0.37 (A:B:C) based on designed geometry. For all three devices, the sensor can in principle be actuated in either the *Twist* or *Swing* mode. However, sensitivity to flow differs throughout the devices and between actuation modes. As previously mentioned, device A (Figure 7.44a&b) is designed for *Twist* mode actuation. Coriolis vibrations in the *Swing* mode are expected to induce torsion of the channel sections with length l_1 , see Fig. 7.43a&b. The strain gauges are patterned at a 45° angle with respect to the channel, in order to measure this torsion. Device B (Fig. 7.44c&d) is designed for *Swing* mode actuation. Coriolis vibrations in the *Twist* mode will cause bending of the channel sections as depicted in Figures 6.17 and 7.43a&b. Finally, device C (Fig. 7.44e&f) is designed to be used in either mode of actuation. Both *Twist* and *Swing* mode vibrations will cause elongation/compression of the strain gauges in a similar way, see Fig. 7.43c&d. Therefore, the flow sensitivity of the device is expected to be similar for both modes of actuation. The sensitivity can be tuned by changing the distance between the two membranes on which the strain gauges reside. The larger this distance, the larger the sensitivity to *Twist* mode vibrations and thus relatively lower sensitivity to *Swing* mode vibrations. The devices are characterized by executing (zero-)flow measurements as described in 7.2 (see Fig. 7.6 for measurement setup). The strain gauges are read out using the circuit in Fig. 7.5 where the reference resistors are the ones indicated in Fig. 7.44.

(a) Sensor type A with 2 strain gauges close to the fixed in-/outlet channels.

(b) Strain gauge of sensor type A.

(c) Sensor type B with 2 strain gauges close to the fixed in-/outlet channels.

(d) Strain gauge of sensor type B.

(e) Sensor type C. A T-shaped channel structure is connected to the main channel by 2 membranes, on which the strain gauges reside.

(f) Strain gauges of sensor type C. Each membrane has two strain gauges.

Figure 7.44: Microscope images of the three sensor types. Strain gauges and reference resistors are indicated in the subfigures. The metal tracks of the gauges are 4 μm wide. The mask design of the strain gauges and reference resistors are identical.

Figure 7.45: Deflection of channel section of device A where strain gauge resides see Fig. (7.44a&b), measured by LDV. Each circle represents one measurement point of the LDV. It can be seen that mostly bending along the channel's axis occurs, and only limited torsion around its axis.

7.7.3 Results

The resistances of the strain gauges were measured and are $\sim 1100/800/460 \, \Omega$ for design A/B/C. With ratios between the resistances of 1:0.73:0.42 (A:B:C), this corresponds quite well to the theoretical resistance ratios presented earlier (1:0.73:0.37). As aforementioned, device A was designed for *Twist* mode actuation. However, initial measurements showed that the device was not sensitive to *Swing* mode Coriolis vibrations. The expectation was that torsion of the channel sections on which the strain gauges reside would induce enough strain to be able to measure this with the strain gauges at a 45° angle (see Fig. 7.44b). To investigate why the strain gauges are not sensitive to *Swing* mode vibrations, an analysis is done by LDV (see Fig. 7.2 for measurement setup). The device is actuated in the *Swing* mode and the deflection of the channel section on which the strain gauge resides (see Fig. 7.44a) is viewed.

Figure 7.45 shows a map of the measured deflection for a range of measurement positions (of the laser of the LDV) on the channel section. It can be observed that mostly bending along the channel's axis occurs, as opposed to torsion around its axis. There is only limited torsion of the channel, which can be seen from the fact that the deflection at position $x = 2250 \, \mu\text{m}$, only differs slightly between the two sides of the channel roof. This small amount of torsion is evidently not enough to produce a measurable strain (using this particular strain gauge design).

The strain gauges of design A show a higher sensitivity to *Twist* mode vibrations, even though they were not designed for this purpose. This causes device A to be sensitive to flow only when actuated in the *Swing* mode. For device B, the sensor was only sensitive to flow when actuated in the *Swing* mode, as expected. Device C was actuated in both modes and showed sensitivity to flow for both the *Twist* and *Swing* mode. Flow measurements are executed with all three device types. The mass flow rate is controlled from $0\text{--}0.6 \, \text{g h}^{-1}$ in steps of $60 \, \text{mg h}^{-1}$. Figure 7.46 shows the resulting output phase difference in relation to the applied flow rate.

Figure 7.46: Output of the resistive readout in relation to the applied flow. Each shown measurement point is an average of the output signal over a period of 200 seconds at a stable flow. The error bars represent the standard deviation, which is determined over the same time period.

There is a linear relation between the output and mass flow rate for each device. As can be seen from Figure 7.46, device type B shows the lowest standard deviation and also shows the highest sensitivity to flow. For each device, a zero-flow measurement has been executed to determine the devices' zero stability over time. The outlet of the flow sensor was closed off and the channel of the sensor was kept at a constant pressure of 5 bar (gauge pressure). The results are displayed in Figure 7.47. Device B shows the highest stability whereas the stability of device A is an order of magnitude lower. The latter was to be expected since device A was in principle designed to be actuated in the *Twist* mode and is thus not optimized for *Swing* mode actuation.

All experimental parameters of the devices are displayed in Table 7.4. When comparing all devices, it is immediately clear that device B shows the best overall performance. Its zero-flow stability is a factor 5 better than for device C, indicating strain during operation is much higher in the channel sections with length l_1 than the membranes with l_3 (see Figs. 7.41 and 7.42). For device C, flow sensitivity for *Twist* mode actuation is approximately a factor 2 higher than for *Swing* mode actuation. However, the standard deviation and zero-flow stability of the phase output signal is around two times higher as well, compensating for the larger flow sensitivity (see Figure 7.47). This results in nearly equal standard deviation and zero-flow stability of the flow signal (in g h^{-1}). Since the zero-flow stability and standard deviation of the

Figure 7.47: Phase difference output of the resistive readout with no flow over a period of 1800 seconds. The sensor’s channel is kept at a constant pressure of 5 bar (gauge pressure) while its outlet is closed off.

Table 7.4: Experimentally obtained values for the devices, determined from the (zero-)flow measurements. Zero-flow stability is defined as the standard deviation of the flow rate over a period of 1800 s (see Fig. 7.47).

Device	Actuation	Resonance Frequency (<i>model</i> /Measured) [Hz]	Flow Sensitivity [°/(g h ⁻¹)]	Standard Deviation [g h ⁻¹]	Zero-Flow Stability [g h ⁻¹]
A	<i>Twist</i>	2426 / 2113	-	-	-
	<i>Swing</i>	1312 / 1184	1.08	0.12	0.19
B	<i>Twist</i>	2426 / 2156	-	-	-
	<i>Swing</i>	1312 / 1203	2.22	0.02	0.02
C	<i>Twist</i>	2519 / 2254	1.23	0.10	0.11
	<i>Swing</i>	1699 / 1624	0.63	0.10	0.09

phase output signal are superior for *Swing* mode actuation, one could try to improve the sensitivity to *Twist* mode Coriolis forces to improve the performance of the device. This can be done by increasing the distance between the membranes on which the strain gauges reside, as previously mentioned. However, the flow sensitivity would need to be increased by a factor 4.5 to match the zero-flow stability of device B. Since the distance between the membranes on which the strain gauges reside can only be increased by approximately a factor 2, device B would still outperform device C.

7.7.4 Discussion & Conclusion

The resistive readout shows great promise for improving the resolution and accuracy of the μ -Coriolis mass flow sensor. As aforementioned, Oliveira et al. presented a zero-flow stability of 0.6 mg h⁻¹ (for N₂) for a μ -CMFS with a capacitive readout [15]. The zero-flow stability of the resistive readout (20 mg h⁻¹) is thus still a factor 33 worse. Optimization of the device could lead it to match the performance of (or outperform) the capacitive readout in the future. This could be achieved by optimizing the design of the sensor and the strain gauges. Another option is to use a piezo-resistive material

to increase the sensitivity to strain significantly, e.g., doped mono-crystalline Silicon as proposed in [14]. Doped poly-silicon strain gauges on top of the channel of the sensor are also an option. The gauge factor of p-doped mono crystalline silicon ranges from -75 to +150 [17], whereas, for doped poly-silicon, the gauge factor ranges from -10 to +30 [17, 18] depending on the doping concentrations. Therefore a resistive readout based on doped mono-crystalline strain gauges would be relatively more sensitive. μ -CMFS devices with such a readout are presented in the following section.

7.8 Piezo-resistive readout

In the previous section, μ -Coriolis mass flow sensor devices with a resistive readout based on metal strain gauges were presented. The best achieved zero-flow stability was 20 mg h^{-1} as opposed to 0.6 mg h^{-1} for a capacitive readout [15]. As mentioned, the readout can be improved by using doped mono-crystalline silicon as a strain gauge material. This would increase sensitivity to strain and could bring the zero-flow stability closer to that of a capacitive readout. A method of embedding such strain gauges in the roof of a surface channel has been described in 4.3.3. Using this process, a μ -CMFS device with integrated *p*-doped mono-crystalline strain gauges was successfully fabricated. Characterization results are presented in this section.

7.8.1 Design

To be able to compare the piezo-resistive readout as closely as possible to the readout based on metal strain gauges, a nearly identical[†] channel loop design is chosen (as in Fig. 7.41). Figure 7.48 shows a schematic representation of the device structure.

The dimensions of the channel loop are equal to those in Fig. 7.41 ($L_x \times L_y = 5 \times 3 \text{ mm}$). The strain gauges consist of two strips of silicon ($l \times w \times h = 200\mu\text{m} \times 20\mu\text{m} \times 340\text{nm}$)[‡] embedded in the channel roof. These strips are connected by a metal track to form a single strain gauge. The reason for this is that the silicon cannot be embedded too close to (or at the position of) the slits through which the micro-channel is etched (see 4.3.3, Figs. 4.13 and 4.14). This could cause issues with accurately defining the slits by photo-lithography. Thus, contact openings are etched to the doped silicon strips and metal is deposited on top to later form the connecting track (just as in Fig. 4.15a&b). Furthermore, the strain gauges are shorter than the metal strain gauges in 7.41, to limit their resistance. Additionally, the channel is fixed right next to the position of the strain gauges to maximize the strain in this area.

[†]For optimal comparison, an additional design was made, where the channel loop geometry was completely identical to the one in Fig. 7.41. However, due to fabrication complications, none of the devices based on this design could be characterized.

[‡] $l \times w \times h = \text{length} \times \text{width} \times \text{height}$

Figure 7.48: μ -CMFS design with piezo-resistive strain gauges. Two strain gauges are positioned on the in-/outlet channel sections. For clarity, the silicon strain gauges are shown on top of the channel, although they are in fact covered by SiRN. The design is based on that in Fig. 7.41.

7.8.2 Device & Setup

Figure 7.49 shows microscope images of the fabricated device. The position of the strain gauges and reference resistors are indicated (Fig. 7.49a). A close-up of the strain gauge (Fig. 7.49b) shows the two doped silicon strips embedded in the channel roof. Metal tracks are connected to the strips via contact openings. The resistance of each strain gauge is approximately $800\ \Omega$. This is the same as for the device with metal strain gauges in Fig. 7.44b and should thus give a similar electrical-thermal noise floor (see Chapter 2). Therefore, the performance of the two devices (device B from Fig. 7.44b and the device presented in this section) can be compared well.

The device is characterized by executing (zero-)flow measurements as described in 7.2 (see Fig. 7.6 for measurement setup). The strain gauges are read out using the circuit in Fig. 7.5, where the reference resistors are the ones indicated in Fig. 7.49a. Furthermore, the device is compared to device B in Fig. 7.44b, from now on simply referred to as the device with metal strain gauges. Both devices are actuated in the *Swing* mode, inducing *Twist* mode Coriolis forces due to flow. First, a comparison is done by varying the actuation displacement amplitude Δz_{act} and bridge supply voltage v_s (see Fig. 7.5) and viewing their effect on the output signal. Afterwards, flow sensitivity and zero-stability of the devices is determined.

(a) Sensor type B with 2 strain gauges close to the fixed in-/outlet channels. (b) Close-up of the *p*-doped mono-crystalline silicon strain gauge.

Figure 7.49: Microscope images of the device. The locations of the strain gauges and reference resistors are indicated. The mask design of the strain gauges and reference resistors are identical.

7.8.3 Results

To compare the performance of the piezo-resistive and metal strain gauges, the output signal for both devices is viewed. Figures 7.50 and 7.51 show how the actuation displacement amplitude Δz_{act} (displacement of the channel section with length L_x , Figs. 7.41 & 7.48) and supply voltage v_s of the resistive bridge affect the output signal (see Fig. 7.5 for the readout circuit). First, the output voltage amplitude v_{out} is viewed (Fig. 7.50). For both devices, the output voltage amplitude changes linearly with v_s and Δz_{act} . For the piezo-resistive device, the two strain gauges (left and right) show a slightly different slope. This can be attributed to the fact that their resistance and that of the reference resistors are not identical (up to 10% difference). This affects the sensitivity of the readout circuit to a resistance change due to strain. Furthermore, the derivative of v_{out} with respect to both v_s and Δz_{act} is approximately a factor 7.5 higher for the piezo-resistive readout than for the resistive readout based on metal strain gauges. Since the actuation displacement amplitude is the same, this shows that the piezo-resistive strain gauges are more sensitive to strain.

Flow rate is determined by measuring the phase difference φ_{diff} between the two output signals (left and right strain gauge), see 7.2. Thus, to compare resolution of the two devices, the standard deviation of the output phase difference is viewed, see Fig. 7.51. It can be clearly observed that for both devices the standard deviation of φ_{diff} goes down when increasing v_s as well as when increasing Δz_{act} . This is to be expected, since in both cases the signal amplitude (due to strain) increases while mechanical-thermal and electrical-thermal noise remain the same. However, the curves

(a) v_{out} in relation to v_s , $\Delta z_{\text{act}} = 100 \mu\text{m}$. (b) v_{out} in relation to Δz_{act} , $v_s = 1 \text{ V}$.

Figure 7.50: Influence of the bridge supply voltage v_s and actuation amplitude Δz_{act} on the output voltage amplitude v_{out} . Each data point represents the mean over a period of 60 seconds. During the measurements, the micro-channels are filled with N_2 .

(a) SD of φ_{diff} in relation to v_s , $\Delta z_{\text{act}} = 100 \mu\text{m}$. (b) SD of φ_{diff} in relation to Δz_{act} , $v_s = 1 \text{ V}$.

Figure 7.51: Influence of the bridge supply voltage v_s and actuation amplitude Δz_{act} on the standard deviation (SD) of the output phase difference φ_{diff} . Each data point represents the SD over a period of 60 s. During the measurements, the micro-channels are filled with N_2 .

Table 7.5: Experimental parameters of the piezo-resistive and metal strain gauge readout, extracted from the (zero-)flow measurements.

Device	Flow sensitivity [$^{\circ}/(\text{g h}^{-1})$]	Zero-flow SD [$^{\circ}$]	Zero-flow stability [mg h^{-1}]
Piezo	2.51	0.025	9.75
Metal	2.23	0.045	20.3

plateau once v_s and Δz_{act} reach a certain value. This seems to indicate that another noise limit is reached. The difference between the value of the standard deviation where the plateaus occur is approximately 2. Thus, the increase in sensitivity of a factor 7.5 for the piezo-resistive readout with respect to the metal strain gauge readout does not lead to a direct increase in SNR by the same factor.

Next, performance of the two devices is compared by executing (zero-)flow measurements. The mass flow rate is controlled from 0 – 0.6 g h^{-1} in steps of 60 mg h^{-1} . During the measurement, $v_s = 1 \text{ V}$ and $\Delta z_{\text{act}} = 100 \text{ }\mu\text{m}$. Figure 7.52 shows the resulting output phase difference in relation to the applied flow rate for both devices. Both devices show a linear relation between the phase shift and the applied flow. However, the output signal of the piezo-resistive readout shows an offset at zero flow. This is most likely due to a very small leak in the micro-channel of the device or where the sensor chip is bonded to its PCB (see 7.2). During flow measurements, the flow is measured before and after the device. These measured flow rates showed a constant offset between each other for the piezo-resistive device and not for the metal strain gauge device. This confirms that a leak is present in the piezo-resistive device. Unfortunately, due to fabrication complications, the device presented here is the only functional one of its type. Since the offset in flow rate is constant, the applied flow rate is still valid.

Going back to Fig. 7.52, the flow sensitivity of the piezo-resistive readout is only slightly higher than that of the metal strain gauge readout. A similar sensitivity was expected, since the channel loop designs are nearly identical (Figs. 7.41 and 7.48). Just as was shown in Figs. 7.50 and 7.51, the standard deviation of the phase shift is lower for the piezo-resistive readout. To more clearly see the stability of the output phase shift of the two devices, a zero-flow measurement was executed. Figure 7.53 shows the resulting phase difference output over a time period of 1800 seconds. Once again, it can be observed that the piezo-resistive readout is more stable than the metal strain gauge readout. Table 7.5 displays the experimental parameters of the two devices. The standard deviation (SD) over the duration of the zero-flow measurement is 1.8 times higher for the metal strain gauge readout than the piezo-resistive readout. However, due to the difference in flow sensitivity, the zero-flow stability differs by a factor 2.1 between the devices.

Figure 7.52: Phase difference output of the piezo-resistive and metal strain gauge devices in relation to the applied flow. Each data point is an average of the output signal over a period of 200 seconds at a stable flow. The error bars represent the standard deviation, which is determined over the same time period. During the measurement, $v_s = 1$ V and $\Delta z_{\text{act}} = 100$ μm .

Figure 7.53: Phase difference output of the piezo-resistive readout with no flow over a period of 1800 seconds. The sensor's channel is kept at a constant pressure of 5 bar (gauge pressure) while its outlet is closed off.

7.8.4 Discussion

As could be seen from the experimental results, the piezo-resistive readout shows an improved performance with respect to the metal strain gauge readout. However, the increase in SNR is not as high as the increase in sensitivity to strain. From this fact, it seems that another noise limit is reached that is related to the actual strain that is measured. Another option would be that the readout electronics are limiting the resolution. The bridge formed by the strain gauge and on-chip reference resistor is fed by a differential supply ($\pm v_s$, see Fig. 7.5). This is realized by two Agilent 33220A waveform generators that are synchronized by a 10 MHz reference signal. The standard deviation of the phase difference output signal is dependent on the stability of the supply signals. When increasing the output voltage amplitude of the Agilent 33220A, it automatically optimizes the settings for the output amplifier and attenuators [19]. In Fig. 7.51a, this occurs at $v_s = 0.7$ V and $v_s = 1.5$ V, where a jump in the standard deviation can be observed. Therefore it is clear that the stability of the supply voltage has a large influence on the overall performance. Generating a fully differential dedicated supply voltage could possibly lead to better performance. Further improvements to the readout electronics could be realized by creating a dedicated CMOS interface similarly as was done by Oliveira et al. [15] for a μ -CMFS with a capacitive readout. Furthermore, the doping of the silicon device layer from which the strain gauges are formed could be adjusted (see 4.3.3). This way, the gauge factor of the strain gauges could be increased. The effect of this on the limit that could be observed in Fig. 7.51 can then be investigated. As of yet, the resolution of the piezo-resistive readout is still inferior to that of a capacitive readout (zero-flow stability of 9.75 mg h^{-1} as opposed to 0.6 mg h^{-1} [15], a factor of 16). The previously mentioned improvements could push the performance to that of a capacitive readout.

7.8.5 Conclusion

A μ -Coriolis mass flow sensor with a piezo-resistive readout was realized. The p -doped mono-crystalline silicon strain gauges have been proven to increase the sensitivity to strain with respect to a metal strain gauge. The zero-flow stability of the μ -CMFS with piezo-resistive readout (9.75 mg h^{-1}) is approximately a factor of 2 better than the metal strain gauge readout (20.3 mg h^{-1}). However, this is still a factor of 16 worse than that of a capacitive readout (0.6 mg h^{-1} for N_2 [15]). Limits to the resolution of the device were shown and discussed. Lowering these limits could be achieved by improvements to the readout electronics and by increasing the gauge factor of the strain gauges. The latter can be done by adjustment of the doping concentration of the silicon device layer (see 4.3.3). These improvements could lead the piezo-resistive readout to match the performance of the capacitive readout without the disadvantages discussed in Chapter 6.

7.9 Conclusion

Several devices were presented, each addressing an issue or effect that limits the performance of the μ -Coriolis mass flow sensor. An overview of the devices, their main goal and achievement/conclusion is given in Table 7.6 divided by their method of readout. A summary of the results for each device is outlined in more detail below:

- The model in Chapter 3 was validated through μ -CMFS devices with varying channel loop size. Resonance frequencies were predicted within 10% error for all except 1 (out of 6) device(s). The predicted flow sensitivities were within 6–40% accurate. The model was used to design further μ -CMFS devices.
- A differential capacitive readout was shown to be able to reduce the effect of vibrations from environmental disturbances by 60-70% compared to a single ended readout.
- Devices with suspended capacitive readout structures were presented, which can withstand in-plane vibrations of up to 68 μm (at the position of the readout structures). This can be potentially increased to 340 μm by making minor design adjustments. A μ -CMFS with a fixed capacitive readout such as in [15] would only allow in-plane vibrations up to 5 μm .
- A novel device with a more stiff suspension was shown to increase the resonance frequency of the undesired in-plane mode significantly. Increased Q factors allowed the resonance peaks to be designed closer together, theoretically increasing flow sensitivity. The zero-flow stability was 0.64 mg h^{-1} , similar to that presented in [15] (0.6 mg h^{-1} for N_2). This is most likely limited by an undesired secondary *Swing* mode. The design could be adapted taking into account this secondary mode, so that the flow sensitivity and resolution can be increased. Furthermore, the higher resonance frequency should make the device more robust against external vibrations (typically in a frequency range $<2 \text{ kHz}$ [16]). This could be confirmed by applying a similar method as in 7.4.
- μ -CMFS devices with a resistive readout based on metal strain gauges were presented. The resolution of the device with the best performance was 20 mg h^{-1} , still a factor 33 worse than that of the device presented in [15]. However, the device does not suffer from the disadvantages of capacitive sensing discussed in Chapter 6, most importantly the water hammer effect, that can lead to breakage of the capacitive readout.
- To improve upon the device with a metal strain gauge readout, piezo-resistive strain gauges were embedded in the micro-channel of the μ -CMFS. The sensitivity to strain was increased significantly while the zero-flow stability was improved by a factor 2 (9.8 mg h^{-1}). Further improvements were discussed that could lead to the device matching the resolution of a μ -CMFS with a capacitive readout (0.6 mg h^{-1} for N_2 [15]).

Table 7.6: Overview of the main goals and achievements/conclusions of the devices presented in this chapter.

Readout	Section	Main goal	Achievement/Conclusion
Capacitive	[15]	<i>Reference device</i>	<i>Zero-flow stability of 0.6 mg h^{-1} for N_2, max. in-plane vibration amplitude of $5 \text{ }\mu\text{m}$.</i>
	7.3	Validation of the model in Chapter 3 through μ -CMFS devices with varying channel loop size.	Predicted resonance frequencies within 10% accurate except for 1 (out of 6) device(s), flow sensitivity within 6–40% accurate.
	7.4	Reduce influence of external vibrations with a differential readout.	60–70% reduction of offset in measured flow rate due to external vibrations.
	7.5	Increase robustness against water hammer effect by a suspended capacitive readout.	Allowed in-plane vibration increased to $68 \text{ }\mu\text{m}$, can be potentially further increased to $340 \text{ }\mu\text{m}$.
	7.6	Increase flow sensitivity and resolution through larger stiffness suspension.	zero-flow stability of 0.64 mg h^{-1} , limited by secondary <i>Swing</i> mode. Design changes and dedicated CMOS interface such as in [15] could increase resolution.
Resistive	7.7	Integrate metal strain gauges with a μ -CMFS to provide an alternative to a capacitive readout that does not suffer from the water hammer effect.	Proof-of-principle, zero-flow stability of 20 mg h^{-1} .
	7.8	Improve resolution with respect to metal strain gauge readout by realizing embedded doped-silicon piezo-resistive strain gauges in the micro-channel of the μ -CMFS.	Proof-of-principle, zero-flow stability of 9.8 mg h^{-1} . Could be increased by improved readout electronics and increasing gauge factor.

Various performance limitations were addressed in this chapter. However, the resolution in terms of flow rate was unfortunately not significantly improved. This could be done by combining the devices with a dedicated CMOS interface such as presented in [15]. Furthermore, design changes to push the resolution of the μ -Coriolis mass flow sensor towards the thermal noise limit will be proposed in the next and final chapter.

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8

Conclusions and Recommendations

This chapter gives a summary of all the conclusions of the preceding chapters. The most promising results from Chapter 7 are highlighted and design recommendations are given for a future device to push the performance of the μ -Coriolis mass flow sensor further towards the mechanical-thermal noise limit. Finally, some additional recommendations are given.

8.1 Conclusions

The aim of this research was to push the resolution of the μ -Coriolis mass flow sensor towards the mechanical-thermal noise limit as well as address several other performance limitations. The work is divided between optimization of the mechanical design (this thesis) and realization of a dedicated CMOS interface for the μ -CMFS. On the latter, research is still ongoing (at the time of finalizing this dissertation) at the Delft University of Technology by A.C. de Oliveira. The approach for optimizing the mechanical design of the μ -CMFS has been to start by defining the performance limitations. These limitations were then addressed one by one throughout the chapters of this thesis.

8.1.1 Definition of performance limitations and modelling of a μ -Coriolis mass flow sensor

The influences of noise and other limitations on the performance of micro-machined inertial sensors in general have been discussed in Chapter 2. Additionally, for each limit, the effect on the performance of a μ -CMFS was briefly described. In Chapter 3, a model was set up that predicts flow sensitivity and other parameters of the μ -CMFS. Furthermore, the noise equivalent flow related to mechanical-thermal noise was calculated and compared to the current state-of-the-art resolution for a μ -CMFS with dedicated CMOS interface [1]. It was concluded that the resolution could still be improved by a factor 480 before reaching the mechanical-thermal noise limit (0.45 ng s^{-1} for the device in [1]).

8.1.2 Fabrication technology

The Surface Channel Technology (SCT) fabrication process was presented in Chapter 4. The technology is an updated version of that presented in [2]. Adapted processes have been developed to realize doped silicon structures embedded in the micro-channels. This can be used for local heating, temperature sensing and piezo-resistive strain gauges among others. All devices in this thesis have been fabricated using one of the technologies described in Chapter 4.

8.1.3 Actuation and readout methods

Transduction principles for actuation and readout of a μ -Coriolis mass flow sensor have been discussed in Chapters 5 and 6. Lorentz actuation remains the preferred method of actuation for its simple integration into silicon micro-machining, low actuation voltages and relative large displacements. Improved magnet configurations have been presented that realize a higher transfer (26.5 dB) between the inductive

tracks of the μ -CMFS. This can be utilized to reduce the needed actuation power and to reduce heating up of the micro-channel and flow medium through resistive heating. With the presented improved magnet configurations, the dissipated power required to actuate the sensor can be reduced by a factor 21 (4 mW to 0.2 mW) to achieve the same actuation angle (5 mrad, θ_t in Fig. 5.6) with respect to the conventional cylindrical configuration. This translates to a reduction of the induced increase in temperature of the channel from 0.2–0.44 °C to 0.01–0.02 °C in a flow range of 0–1 g h⁻¹. Furthermore, it can lead to a reduction of noise generated in the actuation circuitry, since the required amplification to keep the sensor at resonance is lower. Two main readout methods were chosen to realize novel μ -CMFS devices: *capacitive* and *resistive*. Limitations and possible improvements to the readout methods are discussed in Chapter 6 after which devices based on these methods are presented in Chapter 7.

8.1.4 Achieved improvements in performance of the μ -CMFS

The achieved improvements to the performance of the μ -Coriolis mass flow sensor can be divided based on the utilized readout method:

Capacitive

- A differential readout was shown to be able to reduce the effect of vibrations from environmental disturbances by 60-70% with respect to a single ended readout.
- A suspended capacitive readout was realized, that can withstand in-plane vibrations of the micro-channel (which can be caused by the water hammer effect, see Chapter 6) up to 68 μ m. This can potentially be increased to 340 μ m by minor design changes. The devices presented by Groenesteijn et al. [2] would only be able to withstand in-plane vibrations up to a few micron, depending on the design of the capacitive readout.
- A device with a more stiff suspension was shown to increase the resonance frequency and Q factor. A zero flow stability of 0.64 mg h⁻¹ was achieved, similar to that presented in [1]. Combining the device with the CMOS interface from [1] could lead to a better resolution due to reduced $1/f$ noise. Furthermore, the higher resonance frequency should make the device more robust against external vibrations (typically in a frequency range <2 kHz [3]). This could be confirmed by applying a similar method as in 7.4.

Resistive

- Multiple μ -CMFS devices with a resistive readout based on metal strain gauges were developed. The device with the best performance had a zero-flow stability of 20 mg h^{-1} , a factor 25 worse than that of the device presented in [1]. However, the readout does not risk breakage due to the water hammer effect, as opposed to the capacitive readout.
- The first μ -Coriolis mass flow sensor with a piezo-resistive readout was fabricated using the fabrication process presented in 4.3.3. The device shows a zero-flow stability of 9.8 mg h^{-1} , approximately 2 times better than the readout based on metal strain gauges. Electronics optimized for resistive readout and increasing of the gauge factor of the strain gauges could lead to the performance matching (or even surpassing) that of the capacitive readout.

8.2 Towards the thermal noise limit

The most important limitations of the state-of-the-art μ -Coriolis mass flow sensor (concerning the mechanical design) are the flow sensitivity and the fragility of the capacitive readout with respect to in-plane motion of the micro-channel caused by the water hammer effect (see Chapter 6). Furthermore, the gap between the readout electrodes of the device could be reduced. This would increase the capacitance and reduce electrical-thermal noise (see Chapter 2). However, this would make it even more susceptible to breakage from the water hammer effect. Therefore, the most important challenges are to create a device which is robust to in-plane motion and to achieve an increased flow sensitivity and decreased noise floor at the same time. For this, there are basically two pathways: 1) To create a μ -CMFS with a capacitive readout which can withstand large in-plane vibrations and 2) To use a piezo-resistive readout, which does not risk breakage due to in-plane motion.

Adapted suspended capacitive readout

As was shown in 7.5, a suspended capacitive readout can increase the allowed in-plane motion from $5 \text{ }\mu\text{m}$ to $68 \text{ }\mu\text{m}$. Fig. 8.1a shows a schematic representation of the readout electrodes used for the devices in 7.5. The bottleneck for this device (Fig. 7.25) was the gap between the readout electrodes along the y -axis (g_y). The allowed in-plane motion can be easily increased by increasing g_y . When viewing only the effect of the gap between the electrodes along the x -axis (g_x), the allowed in-plane motion in the x -direction is around $340 \text{ }\mu\text{m}$ (see 7.5). This means that there is room for decreasing g_x . This would increase the readout capacitance, potentially decreasing the electrical-thermal noise floor (see Chapter 2). By increasing g_y by 25%, the allowed in-plane vibration amplitude would become $1.25 \cdot 68 \text{ }\mu\text{m} \approx 85 \text{ }\mu\text{m}$. g_x can then be

Figure 8.1: Schematic representation of adapted capacitive readout electrodes. a) Dimensions used for the device in 7.5, b) Adapted dimensions to increase capacitance and allowed in-plane motion.

Table 8.1: Overview of the simulated capacitances of the readout electrode designs in Fig. 8.1 for an equal total electrode width of $700 \mu\text{m}$. The potential decrease of the Electrical-Thermal noise floor is given as well as the increase in allowed in-plane motion of the micro-channel at the position of the readout electrodes (along the x -axis).

Design	Simulated capacitance [fF]	Electrical-Thermal noise floor [factor]	Allowed in-plane motion [μm]
Fig. 8.1a (from 7.5)	49.0	1	68
Fig. 8.1b (adapted)	111	0.66	85

decreased by a factor 4, still allowing for $85 \mu\text{m}$ in-plane motion along the x -axis ($340 \mu\text{m}/4 = 85 \mu\text{m}$). Fig. 8.1b shows a schematic representation of a capacitive readout with these adapted dimensions. Not only does g_x decrease, but the total amount of finger pairs will increase (when using the same total electrode width). The capacitance of both readout electrode designs from Fig. 8.1 are simulated in COMSOL Multiphysics® using the model from Chapter 6 (Fig. 6.7). The total readout electrode width is chosen the same for both simulations ($700 \mu\text{m}$). The resulting capacitances and the consequential change in electrical-thermal noise and allowed in-plane motion is given in Table 8.1.

The capacitance would more than double for such an adapted electrode design. The electrical-thermal noise floor would theoretically be a factor 0.66 of the noise floor for the original design (see Chapter 2, Eq. 2.4). Additionally, the allowed in-plane vibration amplitude (along the x axis) of the channel at the position of the readout electrodes

Figure 8.2: Proposed design realized in COMSOL[®], with $L = 3000 \mu\text{m}$ and $r = 600 \mu\text{m}$. The rigid suspension has been simplified with respect to the design in 7.35.

potentially increases by a factor 2, as aforementioned. This readout would improve the resolution of the μ -Coriolis mass flow sensor as well as increase its robustness to the water hammer effect (even further).

Large stiffness capacitive μ -CMFS devices

Another way to suppress in-plane vibrations is to increase the stiffness of the suspension of the μ -CMFS. In 7.6, devices were presented which realized a larger stiffness. Apart from increased robustness against in-plane vibrations, the devices should also be less susceptible to external vibrations, which occur in a lower frequency range ($< 2 \text{ kHz}$) for most applications [3]. Furthermore, $1/f$ noise generated in the actuation/readout circuitry could be reduced through the higher resonance frequency. Finally, the higher Q factor would allow for the resonance peaks of the operating modes to be designed closer together (since they are sharper) without increasing dependence on the Q factor. The latter theoretically increases the flow sensitivity of the device (see Chapter 3, Eq. 3.15). The resulting flow sensitivity of the characterized μ -CMFS device in 7.6 was lower than expected, since a secondary Coriolis mode limited the displacement induced by Coriolis forces. Redesigning the device such that this secondary mode is at a much higher frequency should lead to a higher flow sensitivity and through that a higher zero-flow stability. This can be done by optimizing the radius of the circular suspension and the length of the center channel. Figure 8.2 shows a proposed novel design. L and r are chosen similar to that of device B in 7.6. With these values of L and r , the rigid suspension can be simplified by using a single channel with a diameter of $100 \mu\text{m}$ for each circular spring. The vibrational modes are simulated in COMSOL Multiphysics[®]. The mode shapes and their resonance frequencies are shown in Figure 8.3.

Figure 8.3: Simulated vibrational modes of the proposed μ -CMFS design in Fig. 8.2, with N_2 as flow medium. The secondary Coriolis/*Swing* and in-plane modes are at much higher frequencies than the operating modes.

As can be seen, the *Twist* mode and primary *Swing* mode are at frequencies of 12.4 kHz and 13.3 kHz, with an absolute difference in frequency of just below 1 kHz. As discussed in 7.6, keeping the absolute frequency difference the same while increasing the respective resonance frequencies can lead to higher flow sensitivity (Fig. 7.32). The resonance frequencies of the secondary *Swing* mode and the first in-plane mode are much higher than the operating modes. This should 1) lead to the device being more robust to in-plane vibrations induced by the water hammer effect, 2) decrease the excitation of the secondary *Swing* mode by Coriolis forces at the *Twist* mode resonance frequency. The latter should result in a higher flow sensitivity if the hypothesis proposed in 7.6 (that the secondary *Swing* mode was limiting the displacement induced by Coriolis forces) is correct. Furthermore, if the effect of in-plane mode vibrations is sufficiently reduced, one could combine the design with the proposed adapted readout presented in Fig. 8.1b. The recommendation would be to realize this design with several values of L (see Fig. 8.2) as to prevent overlapping of the resonance peaks. This is because there can always be differences between designed channel dimensions (roof thickness, channel radius, etc.) and dimensions of the fabricated device. These errors affect the resonance frequencies and can cause the resonance peaks to overlap.

Improved piezo-resistive readout

Another way to address the limitation introduced by the water hammer effect is to use a resistive readout. A proof-of-principle for such a readout based on metal strain gauges was presented in 7.7 and for a piezo-resistive readout in 7.8. The piezo-resistive readout showed a better resolution (9.8 mg h^{-1}) than the metal resistive readout (20.3 mg h^{-1}) in terms of zero-flow stability. However, this is still inferior to the capacitive readout (0.6 mg h^{-1} [1]). Significant improvement is necessary for the piezo-resistive readout to be a viable alternative to the capacitive readout. The resolution of the piezo-resistive readout could be improved in several ways:

- Using a dedicated CMOS interface as was done for a capacitive readout by Oliveira et al. [1].
- Increase the sensitivity to strain of the piezo-resistor by increasing the gauge factor through the use of a lower dopant concentration.
- Decreasing the resistance of the piezo-resistor to limit electrical-thermal noise, for example by changing the mask design or using a larger layer thickness.

Lowering the dopant concentration would increase the gauge factor [4] of the piezo-resistor as well as its electrical resistance [5]. The first is favourable, since it will improve the sensitivity to strain. A higher resistance could potentially increase electrical-thermal noise however (see Chapter 2). By using a larger layer thickness and/or adjusting the mask design of the piezo-resistor, the resistance can be decreased. Therefore, a recommendation for a potential next generation of μ -CMFS with piezo-resistive readout, would be to apply all three of the above mentioned potential improvements. Using a dedicated CMOS interface together with a higher gauge factor and limited/decreased resistance could lead to an improved resolution of the μ -CMFS.

8.3 Further recommendations

The model from Chapter 3 does not predict the Q factor of the simulated designs. Experimentally obtained Q factors were used to calculate the flow sensitivity of the devices presented in this thesis. Future research into the various sources of damping could give more insight. If the Q factor of the designs could be predicted through simulation, this would give a powerful tool to optimize the flow sensitivity through the method discussed in 7.6.

In applications where external vibrations are present, such as in vehicles or audio systems, the output signal of the μ -CMFS can be affected. It was shown in 7.4 that external vibrations cause an offset in the measured flow rate, which can be (partially) cancelled by a differential readout. An improved differential readout could be realized through the use of (piezo-resistive) strain gauges. This would give a more linear response to displacement of the channel and therefore improved cancellation.

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A

Fabrication process flow

This appendix contains a summary of all the fabrication steps for the two types surface channel technology presented in this thesis.

A

A.1 Conventional surface channel technology (SCT)

Table A.1: Used materials and abbreviations

Material/Abbreviation	Description
Si	Silicon
SiRN	Silicon-rich silicon nitride
SiO ₂	Silicon dioxide
Photoresist	Positive UV Photoresist
Metal	Metal stack of Ta, Pt and Au
Al ₂ O ₃	Aluminium oxide
LPCVD	Low Pressure Chemical Vapour Deposition
TEOS	Tetraethyl orthosilicate
HMDS	Hexamethyl disilazane
ICP	Inductively Coupled Power (plasma)
CCP	Capacitively Coupled Power (plasma)
RF	Radio Frequency
DI water	De-Ionized water
SIMS	Secondary Ion Mass Spectrometry

A.1.1 Substrate

Step 1 - Substrate selection

- Wafer type: Monocrystalline Silicon
- Thickness: 525±25 µm
- Orientation: <100>
- Dopant: P-type, Boron
- Resistivity: <0.05 Ω cm
- Double side polished

A.1.2 Deposition of SiRN and SiO₂

Step 2 - Pre-furnace cleaning

- Clean organic traces: 10 min in 99% HNO₃
- Quick dump rinse
- Clean metallic traces: 10 min in 69% HNO₃ at 95 °C
- Quick dump rinse
- Etch native oxide: 1 min in 1% HF
- Quick dump rinse
- Spin dry

Step 3 - LPCVD of SiRN

- Equipment: Tempress Systems Furnace
- Aimed Thickness: 500 nm
- Deposition rate: 4 nm min⁻¹
- SiH₂Cl₂ flow: 77.5 sccm
- NH₃ flow: 20 sccm
- N₂ flow: 250 sccm
- Temperature: 820/850/870 °C (Zone 1/Zone 2/Zone 3)
- Pressure: 150 mTorr
- Check layer thickness using ellipsometry
- Check for particles, scratches or haze in the layer using a cold light source

Step 4 - LPCVD of SiO₂

- Equipment: Tempress Systems Furnace
- Aimed Thickness: 500 nm
- Deposition rate: 10 nm min⁻¹
- TEOS flow: 40 sccm
- N₂ flow: 30 sccm
- Temperature: 710/725/740 °C (Zone 1/Zone 2/Zone 3)
- Pressure: 400 mTorr
- Check layer thickness using ellipsometry
- Check for particles, scratches or haze in the layer using a cold light source

A.1.3 Forming micro-channels

Step 5 - Lithography of channels

- Dehydration bake: 5 min at 120 °C
- Spin coat HMDS, 30 s at 4000 rpm
- Spin coat Olin OIR 908-17, 30 s at 4000 rpm, for 1.7 μm thickness
- Pre-exposure bake: 90 s at 95 °C
- Expose: 3.5 s at 12 mW cm^{-2} , vacuum contact, pre-exposure delay of 120s
- Post-exposure bake: 1 min at 120 °C
- Develop resist: 60 s in OPD 4262
- Quick dump rinse
- Spin dry
- Check result under microscope

Step 6 - Reactive ion etching of SiRN/SiO₂

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- CHF₃ flow: 100 sccm
- Ar flow: 100 sccm
- Pressure: Valve 100%
- ICP: 1500 W
- CCP-RF: 150 W
- Table temperature: 20 °C
- He backside pressure: 10 Torr

Step 7 - Reactive ion etching of Si

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- SF₆ flow: 800 sccm
- Pressure: 90 mTorr
- ICP: 4000 W
- CCP-RF: 0 W
- Table temperature: 20 °C
- He backside pressure: 10 Torr

Step 8 - Stripping of any remaining resist

- Equipment: PVA TePla GIGAbatch 360
- O₂ flow: 360 sccm
- Ar flow: 160 sccm
- Pressure: 0.6 mbar
- Power: 800 W
- Time: 40-60 min, depends on amount of wafers

Step 9 - Removal of metal traces

- Solution: HCl:H₂O₂:H₂O (1:1:5 vol%)
- Temperature: 70 °C
- Time: 15 min

A.1.4 Inlets on backside

Step 10 - Pre-furnace cleaning

- Clean organic traces: 10 min in 99% HNO₃
- Quick dump rinse
- Clean metallic traces: 10 min in 69% HNO₃ at 95 °C
- Quick dump rinse
- Submerge wafers in DI water for 2 h to diffuse any remaining HNO₃
- Quick dump rinse
- Spin dry
- Dry 12 h in N₂ purged environment at 40 °C

Step 11 - LPCVD of SiO₂

- Equipment: Tempres Systems Furnace
- Aimed Thickness: 1200 nm
- Deposition rate: 10 nm min⁻¹
- TEOS flow: 40 sccm
- N₂ flow: 30 sccm
- Temperature: 710/725/740 °C (Zone 1/Zone 2/Zone 3)
- Pressure: 400 mTorr
- Check layer thickness using ellipsometry
- Check for particles, scratches or haze in the layer using a cold light source

Step 12 - Lithography of inlets (backside)

- Dehydration bake: 5 min at 120 °C
- Spin coat HMDS, 30 s at 2500 rpm
- Spin coat Olin OIR 908-35, 30 s at 2500 rpm, for 4.5 μm thickness
- Leave 10 h in low-UV area to outgas resist
- Expose: 12 s at 12 mW cm^{-2} , soft contact
- Develop resist: 60 s in OPD 4262
- Quick dump rinse
- Spin dry
- Leave 10 h in low-UV area to outgas resist
- Hard bake: 2 h at 120 °C
- Check result under microscope

Step 13 - Reactive ion etching of SiRN/SiO₂

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- CHF₃ flow: 100 sccm
- Ar flow: 100 sccm
- Pressure: Valve 100%
- ICP: 1500 W
- CCP-RF: 150 W
- Table temperature: 20 °C
- He backside pressure: 10 Torr

Step 14 - Deep reactive ion etching of Si

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- Process: High rate Bosch (3 step)
- Cycles: Depends on loading, aspect ratio, etc. Determine nr. of cycles with dummy wafer
- Settings:

	Dep.	Break	Etch
Time (s)	0.6	0.6	1.8
ICP (W)	2500	1750	3500
CCP-RF (W)	0	95	0
Pressure (mTorr)	90	25	120
C ₄ F ₈ flow (sccm)	200	10	10
SF ₆ flow (sccm)	10	200	800

Step 15 - Stripping of fluorocarbon

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- O₂ flow: 100sccm
- Pressure 10mTorr
- ICP: 2000W
- Table temperature: 10 °C
- He backside pressure: 10 Torr
- Time: 5 min

Step 16 - Stripping of resist

- Equipment: PVA TePla GIGAbatch 360
- O₂ flow: 360 sccm
- Ar flow: 160 sccm
- Pressure: 0.6 mbar
- Power: 800 W
- Time: 40-60 min, depends on amount of wafers

Step 17 - Wet etching of SiO_2

- Submerge wafers in 50% HF (etch rate($2.5\mu\text{m min}^{-1}$))
- Quick dump rinse
- Submerge wafers in DI water for 2 h to diffuse any remaining HF
- Quick dump rinse
- Spin dry (or directly move to next step)

Step 18 - Removal of metal traces

- Solution: $\text{HCl}:\text{H}_2\text{O}_2:\text{H}_2\text{O}$ (1:1:5 vol%)
- Temperature: 70°C
- Time: 15 min

A.1.5 Forming micro-channel wall

Step 19 - Pre-furnace cleaning

- Clean organic traces: 10 min in 99% HNO_3
- Quick dump rinse
- Clean metallic traces: 10 min in 69% HNO_3 at 95 °C
- Quick dump rinse
- Submerge wafers in DI water for 2 h to diffuse any remaining HNO_3
- Quick dump rinse
- Spin dry
- Dry 12 h in N_2 purged environment at 40 °C

Step 20 - LPCVD of SiRN

- Equipment: Tempres Systems Furnace
- Aimed Thickness: 2000 nm
- Deposition rate: 4 nm min⁻¹
- SiH_2Cl_2 flow: 77.5 sccm
- NH_3 flow: 20 sccm
- N_2 flow: 250 sccm
- Temperature: 820/850/870 °C (Zone 1/Zone 2/Zone 3)
- Pressure: 150 mTorr
- Check layer thickness using ellipsometry
- Check for particles, scratches or haze in the layer using a cold light source

A.1.6 Forming electrodes

Step 21 - Lithography of contact windows

- Dehydration bake: 5 min at 120 °C
- Spin coat HMDS, 30 s at 4000 rpm
- Spin coat Olin OIR 908-35, 30 s at 4000 rpm, for 3.5 μm thickness
- Pre-exposure bake: 90 s at 95 °C
- Expose: 9 s at 12 mW cm^{-2} , soft contact
- Post-exposure bake: 1 min at 120 °C
- Apply dicing foil on backside to prevent liquids from entering channels
- Develop resist: 60 s in OPD 4262
- Quick dump rinse
- Spin dry
- Remove dicing foil
- Hard bake: 10 min at 120 °C
- Check result under microscope

Step 22 - Reactive ion etching of SiRN/SiO₂

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- CHF₃ flow: 100 sccm
- Ar flow: 100 sccm
- Pressure: Valve 100%
- ICP: 1500 W
- CCP-RF: 150 W
- Table temperature: 20 °C
- He backside pressure: 10 Torr

Step 23 - Stripping of resist

- Equipment: PVA TePla GIGAbatch 360
- O₂ flow: 360 sccm
- Ar flow: 160 sccm
- Pressure: 0.6 mbar
- Power: 800 W
- Time: 40-60 min, depends on amount of wafers

Step 24 - Gas phase etching of native SiO₂

- Equipment: Idonus HF VPE
- Pre-bake: 2 min at 120 °C
- Electrostatic clamping to chuck
- Chuck temperature: 35 °C
- Time: 1-2 min
- Proceed to metal deposition immediately to avoid forming of new native oxide layer

Step 25 - Sputtering of Pt

- Equipment: TCOathy
- Aimed thickness: 5 nm
- Base pressure: $< 10^{-6}$ mbar
- Sputter pressure: $6.6 \cdot 10^{-3}$ mbar
- Power: 200 W
- Deposition rate: 24 nm min^{-1}

Step 26 - Annealing to form PtSi for good ohmic contact to Si

- Equipment: Tempress Systems Furnace
- N₂ flow: $4 \cdot 10^3$ sccm
- Temperature: 400 °C
- Time: 30 min

Step 27 - Sputtering of Ta/Pt/Au stack

- Equipment: TCOathy
- Aimed thickness: 10/20/200 nm
- Base pressure: $< 10^{-6}$ mbar
- Sputter pressure: $6.6 \cdot 10^{-3}$ mbar
- Power: 200 W
- Deposition rate: $9.4/24/41.5 \text{ nm min}^{-1}$

Step 28 - Lithography of electrodes

- Dehydration bake: 5 min at 120 °C
- Spin coat HMDS, 30 s at 4000 rpm
- Spin coat Olin OIR 908-17, 30 s at 4000 rpm, for 1.7 μm thickness
- Pre-exposure bake: 90 s at 95 °C
- Expose: 4 s at 12 mW cm^{-2} , hard contact
- Post-exposure bake: 1 min at 120 °C
- Apply dicing foil on backside to prevent liquids from entering channels
- Develop resist: 60 s in OPD 4262
- Quick dump rinse
- Spin dry
- Remove dicing foil
- Hard bake: 10 min at 120 °C
- Check result under microscope

Step 29 - Ion beam etching of metals

- Equipment: Oxford Ionfab 300 Plus
- Neutralizer current: 100 mA
- RF Generator Power: 300 W
- Beam Current: 50 mA
- Beam Voltage: 300 V
- Beam Accelerator voltage: 300 V
- Cool gas pressure: 5 Torr
- Platen temperature: 15 °C, active cooling)
- Platen drive: 5 rpm
- Platen position: -20 °
- Ar flow (neutralizer): 5 sccm
- Ar flow (beam): 5 sccm
- Time: Determine with built-in Hiden SIMS Probe End Point Detector

Step 30 - Stripping of resist

- Equipment: PVA TePla GIGAbatch 360
- O₂ flow: 360 sccm
- Ar flow: 160 sccm
- Pressure: 0.6 mbar
- Power: 800 W
- Time: 40-60 min, depends on amount of wafers

A.1.7 Capping & Annealing of metals

Step 31 - Evaporation of Al_2O_3

- Equipment: Balzers BAK 600
- Aimed thickness: 400 nm
- Base pressure: $< 10^{-6}$ mbar
- Emission current: check log book
- Deposition rate: check log book

Step 32 - Annealing of capped metals

- Equipment: Tempress Systems Furnace
- N_2 flow: $4 \cdot 10^3$ sccm
- Temperature: 400 °C
- Time: 4 h

A.1.8 Creating window openings to bond pads

Step 33 - Lithography of bond pad windows

- Dehydration bake: 5 min at 120 °C
- Spin coat HMDS, 30 s at 4000 rpm
- Spin coat Olin OIR 908-17, 30 s at 4000 rpm, for 1.7 μm thickness
- Pre-exposure bake: 90 s at 95 °C
- Expose: 4 s at 12 mW cm^{-2} , soft contact
- Post-exposure bake: 1 min at 120 °C
- Apply dicing foil on backside to prevent liquids from entering channels
- Develop resist: 60 s in OPD 4262
- Quick dump rinse
- Spin dry
- Remove dicing foil
- Hard bake: 10 min at 120 °C
- Check result under microscope

Step 34 - Reactive ion etching of Al_2O_3

- Equipment: Oxford Instruments PlasmaPro 100 Cobra
- BCl_3 flow: 25 sccm
- HBr flow: 10 sccm
- Pressure: 3 mTorr
- ICP: 1750 W
- CCP-RF: 20 W
- Table temperature: 2 °C
- He backside pressure: 10 Torr
- Time: Etch until 50-100 nm Al_2O_3 is left (as protection during release etch (step 40))

Step 35 - Stripping of resist

- Equipment: PVA TePla GIGAbatch 360
- O_2 flow: 360 sccm
- Ar flow: 160 sccm
- Pressure: 0.6 mbar
- Power: 800 W
- Time: 40-60 min, depends on amount of wafers

A.1.9 Releasing channels and forming comb readout electrodes

Step 36 - Lithography of release windows

- Dehydration bake: 5 min at 120 °C
- Spin coat HMDS, 30 s at 4000 rpm
- Spin coat Olin OIR 908-17, 30 s at 4000 rpm, for 1.7 μm thickness
- Pre-exposure bake: 90 s at 95 °C
- Expose: 4 s at 12 mW cm^{-2} , hard contact
- Post-exposure bake: 1 min at 120 °C
- Apply dicing foil on backside to prevent liquids from entering channels
- Develop resist: 60 s in OPD 4262
- Quick dump rinse
- Spin dry
- Remove dicing foil
- Check result under microscope

Step 37 - Reactive ion etching of Al_2O_3

- Equipment: Oxford Instruments PlasmaPro 100 Cobra
- BCl_3 flow: 25 sccm
- HBr flow: 10 sccm
- Pressure: 3 mTorr
- ICP: 1750 W
- CCP-RF: 20 W
- Table temperature: 2 °C
- He backside pressure: 10 Torr
- Time: Etch until 50-100 nm Al_2O_3 is left (as protection during release etch)

Step 38 - Ion beam etching of readout combs

- Equipment: Oxford Ionfab 300 Plus
- Neutralizer current: 100 mA
- RF Generator Power: 300 W
- Beam Current: 50 mA
- Beam Voltage: 300 V
- Beam Accelerator voltage: 300 V
- Cool gas pressure: 5 Torr
- Platen temperature: 15 °C (active cooling)
- Platen drive: 5 rpm
- Platen position: -20 °
- Ar flow (neutralizer): 5 sccm
- Ar flow (beam): 5 sccm
- Time: Use time determined in step 29

Step 39 - Reactive ion etching of SiRN

- Equipment: Oxford Instruments PlasmaPro 100 Cobra
- SF_6 flow: 50 sccm
- O_2 flow: 20 sccm
- Pressure: 20 mTorr
- ICP: 1750 W
- CCP-RF: 50 W
- Table temperature: 10 °C
- He backside pressure: 10 Torr

Step 40 - Deep reactive ion etching of Si

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- Cycles: 7
- Settings:

	Strike	Etch	Wait
Time (s)	1	300	120
ICP (W)	1000	1000	0
CCP-RF (W)	50	0	0
Pressure (%/mTorr)	15%	90	100%
SF ₆ flow (sccm)	200	200	0

A.1.10 Opening bond pads

Step 41 - Reactive ion etching of Al₂O₃

- Equipment: Oxford Instruments PlasmaPro 100 Cobra
- BCl₃ flow: 25 sccm
- HBr flow: 10 sccm
- Pressure: 3 mTorr
- ICP: 1750 W
- CCP-RF: 20 W
- Table temperature: 2 °C
- He backside pressure: 10 Torr
- Important note: Etches all Al₂O₃, etch just long enough to open bond pads

Step 42 - Stripping of any remaining resist

- Equipment: PVA TePla GIGAbatch 360
- O₂ flow: 360 sccm
- Ar flow: 160 sccm
- Pressure: 0.6 mbar
- Power: 800 W
- Time: 40-60 min, depends on amount of wafers

Step 43 - Dice chips manually

- Equipment: Diamond pen
- Note: Use grooves etched during release etch to break wafer into separate chips

A.2 Adapted SCT: Embedded doped mono-crystalline silicon structures (based on silicon-on-insulator)

Table A.2: Used materials and abbreviations, in addition to Table A.1.

Material/Abbreviation	Description
■ Si (doped)	Highly doped (P-type) silicon
TCR	Temperature coefficient resistance

A.2.1 Substrate

Step 1 - Substrate selection

- Wafer type: Silicon-on-insulator
- Device layer: 300–500 nm, depending on desired thickness of doped silicon structures.
- Buried oxide layer: 400 nm
- Handle layer: 525 ± 25 μm
- Device and handle layer orientation: $\langle 100 \rangle$
- Device layer dopant: P-type, Boron
- Device layer resistivity: 0.001–0.01 Ωcm , depending on desired TCR, gauge factor, etc.
- Double side polished

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A.2.2 Patterning device layer

Step 2 - Lithography of doped silicon structures

- Dehydration bake: 5 min at 120 °C
- Spin coat HMDS, 30 s at 4000 rpm
- Spin coat Olin OIR 908-17, 30 s at 4000 rpm, for 1.7 μm thickness
- Pre-exposure bake: 90 s at 95 °C
- Expose: 3.5 s at 12 mW cm^{-2} , vacuum contact, pre-exposure delay of 120s
- Post-exposure bake: 1 min at 120 °C
- Develop resist: 60 s in OPD 4262
- Quick dump rinse
- Spin dry
- Check result under microscope

Step 3 - Reactive ion etching of Si and SiO₂

- Equipment: Oxford Instruments PlasmaPro 100 Estrelas
- CHF₃ flow: 100 sccm
- Ar flow: 100 sccm
- Pressure: Valve 100%
- ICP: 1500 W
- CCP-RF: 150 W
- Table temperature: 20 °C
- He backside pressure: 10 Torr

Step 4 - Stripping of resist

- Equipment: PVA TePla GIGAbatch 360
- O₂ flow: 360 sccm
- Ar flow: 160 sccm
- Pressure: 0.6 mbar
- Power: 800 W
- Time: 40-60 min, depends on amount of wafers

A.2.3 Deposition of SiRN and SiO₂

Step 5 - Pre-furnace cleaning

- See section A.1.2, step 2.

Step 6 - LPCVD of SiRN

- See section A.1.2, step 3.

Step 7 - LPCVD of SiO₂

- See section A.1.2, step 4.

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A.2.4 Forming micro-channels

Step 8 - Lithography of channels

- See section A.1.3, step 5.

Step 9 - Reactive ion etching of SiRN/SiO₂

- See section A.1.3, step 6.

Step 10 - Reactive ion etching of Si

- See section A.1.3, step 7.

Step 11 - Stripping of any remaining resist

- See section A.1.3, step 8.

Step 12 - Removal of metal traces

- See section A.1.3, step 9.

A.2.5 Inlets on backside

Step 13 - Pre-furnace cleaning

- See section A.1.4, step 10.

Step 14 - LPCVD of SiO_2

- See section A.1.4, step 11.

Step 15 - Lithography of inlets (backside)

- See section A.1.4, step 12.

Step 16 - Reactive ion etching of SiRN/SiO_2

- See section A.1.4, step 13.

Step 17 - Deep reactive ion etching of Si

- See section A.1.4, step 14.

Step 18 - Stripping of fluorocarbon

- See section A.1.4, step 15.

Step 19 - Stripping of resist

- See section A.1.4, step 16.

Step 20 - Wet etching of SiO_2

- See section A.1.4, step 17.

Step 21 - Removal of metal traces

- See section A.1.4, step 18.

A.2.6 Forming micro-channel wall

Step 22 - Pre-furnace cleaning

- See section A.1.5, step 19.

Step 23 - LPCVD of SiRN

- See section A.1.5, step 20.

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A.2.7 Forming electrodes

Step 24 - Lithography of contact windows

- See section A.1.6, step 21.

Step 25 - Reactive ion etching of SiRN/SiO₂

- See section A.1.6, step 22.

Step 26 - Stripping of resist

- See section A.1.6, step 23.

Step 27 - Gas phase etching of native SiO_2

- See section A.1.6, step 24.

Step 28 - Sputtering of Pt

- See section A.1.6, step 25.

Step 29 - Annealing to form PtSi for good ohmic contact to Si

- See section A.1.6, step 26.

Step 30 - Sputtering of Ta/Pt/Au stack

- See section A.1.6, step 27.

Step 31 - Lithography of electrodes

- See section A.1.6, step 28.

Step 32 - Ion beam etching of metals

- See section A.1.6, step 29.

Step 33 - Stripping of resist

- See section A.1.6, step 30.

A.2.8 Capping & Annealing of metals

Step 34 - Evaporation of Al_2O_3

- See section A.1.7, step 31.

Step 35 - Annealing of capped metals

- See section A.1.7, step 32.

A.2.9 Creating window openings to bond pads

See section A.1.8 for graphical representation.

Step 36 - Lithography of bond pad windows

- See section A.1.8, step 33.

Step 37 - Reactive ion etching of Al_2O_3

- See section A.1.8, step 34.

Step 38 - Stripping of resist

- See section A.1.8, step 35.

A.2.10 Releasing channels (and forming comb readout electrodes)

Step 39 - Lithography of release windows

- See section A.1.9, step 36.

Step 40 - Reactive ion etching of Al_2O_3

- See section A.1.9, step 37.

Step 41 - Ion beam etching of readout combs

- See section A.1.9, step 38. Skip step if no comb electrodes or other metal structures need to be formed using the release mask.

Step 42 - Reactive ion etching of SiRN

- See section A.1.9, step 39.

Step 43 - Deep reactive ion etching of Si

- See section A.1.9, step 40.

A.2.11 Opening bond pads

See section A.1.8 for graphical representation.

Step 44 - Reactive ion etching of Al_2O_3

- See section A.1.10, step 41.

Step 45 - Stripping of any remaining resist

- See section A.1.10, step 42.

Step 46 - Dice chips manually

- See section A.1.10, step 43.

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Samenvatting

Het nauwkeurig meten en regelen van debiet in een vloeistof- of gasstroom is in vele toepassingen van groot belang. Voorbeelden van toepassingen zijn: gaschromatografie, meting van windsnelheid en biomedische toepassingen. Zodoende zijn debietmeters vaak een cruciaal onderdeel in een fluïdisch systeem. Gedurende de afgelopen decennia is het onderzoeksveld van micro-debietmeters altijd groeiende geweest, vanwege de voordelen die miniaturisatie van dit soort sensoren met zich mee brengt. De belangrijkste voordelen van micro-debietmeters zijn laag vermogensverbruik en de mogelijkheid om zeer lage debieten te meten. Thermische debietmeters worden veel toegepast vanwege hun hoge resolutie. Echter moeten thermische debietmeters voor een specifiek(e) vloeistof of gas worden gecalibreerd. Dit is omdat ieder fluïdum een andere warmtecapaciteit heeft. Daarentegen meten Coriolis debietmeters direct massa-debiet, onafhankelijk van eigenschappen van het fluïdum zoals dichtheid en viscositeit. Om deze reden zijn ze uitgegroeid tot een belangrijk onderdeel van het onderzoeksveld dat zich richt op massa-debietmeting. De eerste Coriolis massa-debietmeters die waren gefabriceerd door middel van micro-technologie, werden bijna twee decennia geleden gepresenteerd. Sindsdien is er veel vooruitgang geboekt om hun prestatie te verbeteren. Het doel van het onderzoek in dit proefschrift is om de resolutie van de μ -Coriolis massa-debietmeter (μ -CMDM) naar het limiet gerelateerd aan mechanisch-thermische ruis te drijven. Het onderzoek in dit proefschrift richt zich op beperkingen van de prestatie van een μ -CMDM gerelateerd aan het mechanische ontwerp en hoe men deze kan aanpakken. Het onderzoek kan onderverdeeld worden in twee delen: 1) Definitie van de beperkingen van de prestatie van een μ -CMDM en het modelleren van de sensor, 2) Presentatie van methodes en ontwerpen om de voorgenoemde beperkingen aan te pakken. Een μ -CMDM kan in principe worden gezien als een traagheidsmeter, waar een debiet door een micro-kanaal versnelling van het kanaal veroorzaakt. Om deze reden worden beperkingen van de prestatie van micro-traagheidsmeters besproken, inclusief (maar niet beperkt tot) mechanisch-thermische ruis. Voor elke beperking wordt de toepassing op een μ -CMDM beschreven. Om het begrip van de beperkingen van de prestatie van een μ -CMDM te bevorderen, wordt het werkingsprincipe in detail besproken. Verder is er een model van de sensor opgezet, bestaande uit een Eindige Elementen Methode simulatie in combinatie met een analytisch model. Het model

maakt snelle simulatie van resonantiefrequenties, debietsgevoeligheid en mechanische eigenschappen mogelijk. Het model wordt gevalideerd door verscheidende μ -CMDM ontwerpen te realiseren met variërende vorm en grootte en vervolgens gesimuleerde en experimenteel verkregen parameters te vergelijken. Nadat een model van de sensor is opgezet, worden beperkingen van de aandrijvings- en uitlezingsmethodes onderzocht. Verscheidene transductieprincipes worden besproken die kunnen worden gebruikt voor aandrijving van de μ -CMDM. Verbeterde magneetconfiguraties voor elektromagnetische aandrijving van de μ -CMDM worden gepresenteerd. De hierdoor verbeterde prestatie kan de signaal-ruisverhouding van de gehele sensor verbeteren en de hoeveelheid vermogen die nodig is voor aandrijving verlagen. Verscheidene uitlezingsmethodes voor een μ -CMDM worden besproken om de prestatiebeperkingen die in dit proefschrift worden gepresenteerd aan te pakken. Ontwerpen zijn gerealiseerd op basis van twee algemene uitlezingsmethodes: 1) Een capacitieve uitlezing gerealiseerd door kamvormige elektrodes die zijn bevestigd aan het micro-kanaal van de sensor, 2) Een resistieve uitlezing gebaseerd op rekstrookjes bovenop of geïntegreerd in het micro-kanaal. De sensoren zijn gefabriceerd door middel van het zogenoemde *Surface Channel Technology* (oppervlak-kanaal-technologie) fabricage proces, dat in detail wordt beschreven in dit proefschrift. Dit proces maakt het mogelijk om micro-kanalen direct onder het oppervlak van een silicium *wafer* (schijf) te creëren. De gefabriceerde sensoren zijn gekarakteriseerd en verscheidene verbeteringen zijn behaald. μ -CMDM ontwerpen met een capacitieve uitlezing zijn gerealiseerd, die significant verbeterde bestendigheid lieten zien tegen trillingen die veroorzaakt worden door externe bronnen in de nabije omgeving van de sensor of door waterslag (waar bubbels in een vloeistofstroom in-het-vlak trillingen van de micro-kanalen kan veroorzaken). Verder zijn er μ -CMDM ontwerpen gerealiseerd met hogere resonantiefrequenties en kwaliteitsfactoren. Deze sensoren zouden beter bestendig moeten zijn tegen externe en in-het-vlak trillingen, terwijl de debietsgevoeligheid ook hoger wordt. Het laatste is helaas niet significant verhoogd, doordat een secundaire Coriolis trillingsmodus de uitwijking van het kanaal (veroorzaakt door debiet) heeft beperkt. Naast ontwerpen met een capacitieve uitlezing, werden er *proof-of-principle*[†] ontwerpen met een (piezo-)resistieve uitlezing gerealiseerd, die de beperkingen gerelateerd aan een capacitieve uitlezing kunnen overkomen. Eerste resultaten zijn veelbelovend, hoewel de behaalde resolutie nog steeds inferieur is aan die van de capacitieve uitlezing. Tenslotte worden er aanbevelingen gegeven, gebaseerd op de gekarakteriseerde sensoren, om de resolutie van de μ -Coriolis massa-debietmeter verder naar het thermische ruislimiet te drijven.

[†]het aantonen dat een nieuwe technologie of methode daadwerkelijk functioneert zoals verwacht.

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Acknowledgements

This dissertation is a result of four years of research. Although my name is the only one on the cover, of course I could not have done this by myself. I would like to thank everyone that has contributed to this work and/or supported me over the last few years.

First of all I would like to thank my supervisors Remco Wiegerink and Joost Lötters. Remco, you have supervised me through my bachelor assignment, master assignment and PhD. Over all these years, your office door was always open so that I could walk in any moment to ask a question or to discuss something (or to complain about bugs in CleWin). Your advice was always useful and nearly always helped me solve the problem that I came to you for. Joost, your suggestions and ideas always drove me to not only think "How can I improve my devices?", but also to ask myself "What else can we do with this?". Your connection with industry led me to always think about application and additional functionality.

I would also like to thank the other members of the IDS group. The fun group outings and activities were a great way to get to know fellow group members and talk to each other about matters other than work. Special thanks to my (ex-)office mates, Anneirudh, Dennis, Haye, Henk-Willem, Mahdieh, Qihui, Sabrine, Yaxiang and Yiyuan. Our office was always a pleasant place to work and also to have a drink together every now and then on a Friday afternoon. Of course I also want to thank Pino for his help with technical issues and the occasional courgette/zucchini cake.

During my PhD I have fabricated many devices. Without the help of others, this would never have been possible. My thanks to Jarno, Jack and Kechun for all the help and support. Furthermore, I would like to thank the MESA+ nanolab staff for their help, and in particular Meint for the interesting and often long discussions about new fabrication ideas.

This project would have never happened without the support of the members of my user committee and the Dutch research council (NWO). Many thanks to the people from Bronkhorst High-Tech and Krohne for their interest and support in my work. Many thanks as well to my collaborators at the Electronic Instrumentation Lab (EIL) of Delft University of Technology (TU Delft). Special thanks to Yannick and Luca for the collaborative projects we did/started together.

I would like to thank my family and friends for their support during and before my PhD so that I could even make it this far. Finally, there is one last person that I would like to thank. Maria, thank you for your love and support over the last few years, and of course for introducing Spanish cuisine into my life (just kidding). On to our next adventure!

Thomas Schut

Enschede, September 2021

About the author

Thomas Schut was born on the 7th of June, 1994 in Goes, the Netherlands. During his primary and secondary education, he showed an interest in the more technological aspects of life and also developed a passion for music. After carefully considering a career path in technology or becoming the new Angus Young, he started his BSc. in Electrical Engineering at the University of Twente, Enschede, The Netherlands (2012–2015). He finalized his bachelor with an assignment on a micro angular position sensor at the Transducer Science and Technology (TST) group. He then followed an MSc. in Electrical Engineering, moving towards the field of Micro Electro-Mechanical Systems (MEMS) (2015–2017). He finished his studies with an assignment on a micro-Coriolis mass flow sensor with integrated pressure sensors at the Micro Sensors and Systems group (MSS, formerly TST). Subsequently, he started as a PhD candidate at the MSS group (now IDS, Integrated Devices and Systems) working on the subject presented in this dissertation (2017–2021).

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