High Precision MEMS Displacement Sensors

Device Techniques and Readout Circuits

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To my parents
Declaration

The thesis contains no material which has been accepted for the award of any other degree or diploma in any university or other tertiary institution and, to the best of my knowledge and belief, contains no material previously published or written by another person, except where due reference has been made in the text. I give consent to the final version of my thesis being made available worldwide when deposited in the University’s Digital Repository, subject to the provisions of the Copyright Act 1968.

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March, 2014
Statement of Collaboration

I hereby certify that some of the work embodied in this thesis has been done in collaboration with other researchers. I have included as part of the thesis a statement clearly outlining the extent of collaboration, with whom and under what auspices.

- In Chapter 5, the design and characterisation of the 2-DoF nanopositioner, the design and implementation of its feedback control system and the AFM imaging were performed in collaboration with Mr. Anthony Fowler and Dr. Yuen Yong (of the University of Newcastle).

- In Chapter 6, the design and characterisation of the piezoresistive sensors was performed in collaboration with Dr. Ali Bazaei and Mr. Mohammad Marofi (of the University of Newcastle).

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Abstract

Displacement sensors are key components in the closed loop control of nanopositioning systems, producing repeatable motions with nanometer resolution. Microelectromechanical systems (MEMS) offer a miniaturised solution for implementing low cost and high speed nanopositioners with integrated sensors. Among the MEMS displacement sensing techniques, only the electrothermal and piezoresistive sensors can be implemented for this purpose within a reasonable footprint. The performance of the aforementioned sensors is severely influenced by their inherent high noise levels, which limits the displacement resolution of the MEMS nanopositioner. As such, this thesis presents several displacement sensing techniques to improve the performance of the MEMS nanopositioners.

In the first approach, a readout circuit is presented to increase the sensitivity of the electrothermal sensors. The sensor is coupled to a ring oscillator power supply by a ratiometric interface in order to convert the small resistive changes (390–400Ω) to the wide frequency variations (350–550kHz). Subsequently, the frequency demodulator circuits are designed to produce a voltage output. The experiments are carried out on a 1-Degree of Freedom (1-DoF) nanopositioner with a thermal actuator. In addition to the high sensitivity achievement, the ring oscillator nonlinearity is designed to cancel the actuator nonlinearity to produce a linear input-output transfer function, which otherwise needs a lookup table for calibration and closed loop control.
The MEMS sensors are fabricated in doped silicon that inherently generates the flicker and thermal noise. The second approach in this thesis consists of two distinct solutions to mitigate the noise contribution. The flicker noise is inversely proportional to the applied heating signal frequency. Hence, a new excitation and readout technique is presented to drive the sensor with a high frequency voltage. The experiments are successfully conducted on a 1-DoF MEMS nanopositioner, which demonstrate an 8dB flicker noise reduction, when compared to the conventional dc (direct current) excitation. In order to alleviate the thermal noise, a multiple sensor system is developed based on the averaging theory, according to which a combination of multiple independent signals contaminated with the uncorrelated noise results in a higher signal-to-noise ratio (SNR). The sensors are implemented around a 1-DoF nanopositioner in order to produce independent measurements of the displacement in real time. The experimental results on three sensors demonstrate a 4dB SNR enhancement, which is in close agreement with the theory.

Electrothermal sensors can be operated in constant current (CC) or in constant voltage (CV) excitation modes. In the third approach, an analytic comparison of the two methods is presented. It is shown that from the SNR point of view, the benefits of operating a sensor in CC mode are only marginal. The analytical investigation is supported by experiments performed on sensors integrated into a silicon on insulator (SOI) MEMS nanopositioner with low noise read out circuits, which leads to a 0.04nm/$\sqrt{Hz}$ displacement resolution for both excitation modes.

A new 2-Degree of Freedom (2-DoF) MEMS nanopositioner with electrothermal sensors is also presented in this dissertation for the scanning stage of the atomic force microscope (AFM), which is a significant achievement towards the miniaturisation of low cost AFMs for high speed operation. The electrothermal sensors are integrated to the stage in order to provide displacement information for a feedback control loop,
which should be otherwise provided by an external sensor, such as an interferometer. The images obtained by the controlled nanopositioner demonstrate a great quality enhancement compared to the open loop nanopositioner.

Finally, a piezoresistive sensing technique is presented in this thesis as an alternative integrated small area solution. Compared to the existing piezoresistive sensors, it is fabricated in the standard SOI process without any customised fabrication step. The embedded sensor is designed and characterised for a 1-DoF nanopositioner and exploits the differential architecture to achieve higher sensitivity, linearity and common mode interference rejection.
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Publications

The research presented in this thesis has resulted in the following publications.

Journal Papers

Most of the material presented in Chapter 3 is published in [1]. Sections 1, 2, and 4 of Chapter 4 are published in [2] and accepted for publication in [4], respectively. Chapter 5 is a collaborative work which is accepted for publication in [3] (online version is available). The research results presented in section 3 of Chapter 4 and entire Chapter 6 are under review.


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Conference Papers


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1 Introduction

The positioning, manipulation and interrogation of microscopic objects demand sophisticated devices that are capable of producing repeatable, high-precision motions with nanometer resolution. The evolution of *Nanopositioning* during the last decade has led to incredibly efficient products, such as atomic force microscopes (AFM) [20] and microgrippers [21, 22], which are essential tools for the progression of biomedical science and nanotechnology. Ultra high resolution AFM equipped with integrated microheaters have been widely used in nanoscale material deposition [23], microfluidic actuation [24], localised thermal system analysis [25] and meteorology [26]. High density probe storage devices [27] and optical alignment systems [28] are amongst the other applications of nanopositioners.

Despite their fascinating advancements, nanopositioners are still expensive devices that are unable to fulfil fast operations. Faster nanopositioners can produce video rate images in microscopy compared to available products that is capable of producing single images only. The ongoing research consists of system and device level approaches to improving the speed while reducing the cost factor. The systemic approaches are about the development of new control algorithms to speed up the system performance. In the device perspective after development of mesoscale solutions, microscale components have attracted significant attention. Microelectromechanical systems (MEMS) are expected to provide high speeds at a lower cost. Figure 1.1
Figure 1.1: Conventional mesoscale AFM and a recently proposed MEMS AFM [1].

illustrates a recent comparison between the bulky scanner of a conventional AFM that needs laser alignment with a MEMS based AFM that features a miniature scanner and does not require alignment. Also a MEMS device offers fast lateral scanning with lower mechanical drift [1]. However, the performance of macro/mesoscale components in terms of noise, resolutions and linearity is higher than MEMS devices. Displacement sensors particularly play an important role in a nanopositioner performance. The research conducted in this thesis investigates different displacement sensing techniques available in MEMS.

1.1 Nanopositioning

A nanopositioner consists of a microactuator operating in a closed loop, where a feedback controller is used to improve the system dynamics. Conventional nanopositioners employ piezoelectric actuators and optical sensors, which are slow, bulky and expensive. As an alternative, MEMS have the potential to offer a miniaturised, high speed and low cost solution for nanopositioning purposes. However, achieving the performance criteria similar to that attained by macro/mesoscale solutions is yet an
open field of research. The systemic operating principle of nanpositioners follows the same mechanism employed in accelerometers and gyroscopes. The conventional open loop actuators utilised in these systems suffer from mechanical creep and drift that can be alleviated by a controller in a closed loop system, as shown in Figure 1.2. The forward path microactuator, which can be either electrostatic [8], electromagnetic [7] or electrothermal [29] has a displacement output. For example, the actuator moves a scan table in the case of AFM or an arm in the case of microgripper. The feedback network consists of a displacement sensor and readout circuitry. Lateral nanpositioners are used as the scanner table in AFM [30], which consists of an atomic sharp probe that scans the sample surface located on top of a scan table and produces a high-resolution image of the sample, as illustrated in Figure 1.2. The forces measured between the sample and the sharp tip of the AFM probe represents different properties of the sample such as topography and electric or magnetic properties of the sample. In order to operate the AFM, control of two main quantities is needed:

1. The lateral position of the sample relative to the probe tip, obtainable by using displacement sensors, is thoroughly investigated throughout the thesis.
2. The force regulation between the tip and the sample or vertical nanpositioning,

Figure 1.2: A nanpositioner closed loop control system including actuators, sensors and a controller.
which is beyond the scope of this thesis.

The closed loop control can also mitigate the limited dynamic range, slow frequency response, non-linearity and stability issues, provided that an accurate source of position information is available. A displacement sensor, mechanically coupled to the moving part of the actuator, obtains the position information. The sensing mechanisms available in MEMS include capacitive [29], piezoresistive [12], magnetoresistive [31] and electrothermal sensors [20]. Regardless of the sensing mechanism, the sensor noise is directly injected into the control loop and affects the positioning accuracy, as shown in Figure 1.2. In addition, increasing the loop bandwidth, to achieve higher scan speeds, adds to the noise power that can enter the feedback loop and escalates the random error due to position sensing inaccuracy [30]. The sensor noise in MEMS may have either mechanical or electrical sources, though the bulk micromachined devices investigated in this research are dominantly contaminated by electrical noise. The electrical noise can be generated by the device itself or the readout circuitry. Most of the low noise readout techniques are proposed to reduce the circuit noise. However, the performance of resistive sensors is drastically influenced by the device noise that mainly consists of the thermal and the flicker noise of silicon. This is addressed in Chapters 2 and 4 in more detail and two practical solutions have been proposed, implemented and tested successfully.

1.2 Microelectromechanical Systems (MEMS)

Microelectromechanical systems (MEMS) are man-made mechanical or electrical components characterised by their microscopic size [19]. MEMS fabrication processes have been developed in parallel with high pace developments related to transistors as a by-product of the evolution in the integrated circuit technology. The application
of MEMS technology can be seen in the market, significantly in products such as transducers, ink-jet printers and micromirrors.

MEMS devices feature 3D movable microstructures such as cantilever beams etc. that can handle a wide variety of signal types, such as physical (electrical, mechanical, thermal and optical etc.), chemical and biological signals. Though immovable micro-electronic devices deal with electrical signals only [32]. From a financial perspective, although the semiconductor share of the market has experienced tremendous growth, most of the MEMS products have challenges that have kept them at the research level and avoided their commercialisation.

The design process of the mechanical devices within the aforementioned micrometer dimensions is associated with some practical challenges, as well as new opportunities. In other words, the priority and dominance of the physics rules for microscale devices is not the same for macroscale ones. For instance, surface mechanical forces, such as friction, dominate the gravity that leads to a significant challenge in the design of microgrippers. Since gravity is insufficient to release an ungripped object, a separate mechanism is needed [33]. Another example is the thermal conduction through the air, which is a good thermal insulator for macroscale designs. However, the heat transfer mechanism between a hot beam and cold plate in MEMS electrothermal sensing is dominated by conduction through a micrometer width air gap, rather than through convection or radiation [34].

Among several categories for MEMS applications, sensors and actuators have attracted more attention. A significant outcome of micromachining for the sensors and actuators is their implementation in batch processes [35]. This results in low cost and identical transducers, hundreds of which can be implemented on a single wafer. The ongoing research on MEMS sensors and actuators spans fields as diverse as circuit design and device physics in electronics to mechanics and thermodynamics.
The focus of this research is most concentrated on the engineering of the electrical and electronic properties of the involved materials for special purpose displacement sensors. Additionally some electronic circuit techniques have been developed to improve the performance of the MEMS displacement sensors.

1.3 Fabrication Process

Most of the devices designed in this thesis are implemented in silicon on insulator (SOI) technology. They are designed in Coventorware and fabricated by MEMS-Cap facilities, which offers the Multi-User MEMS Processes (MUMPS) [2]. Therefore, a brief description of fabrication technology is presented here.

Silicon is introduced as the most versatile material for MEMS technology after extensive research conducted in both microfabrication techniques and material properties [36]. In addition to the magnificent electronic properties exploited in integrated circuits, it demonstrates excellent mechanical characteristics, including high elasticity, stiffness, fracture and fatigue strengths, thermal conductivity, thermal stability etc. [32]. Silicon has also been proven to be a suitable material for sensing such as in piezoresistive [37] and electrothermal [34] sensors.

As stated earlier, due to the multidisciplinary nature of MEMS, several design parameters complicate the MEMS design process. Therefore, simplicity in the fabrication process, including the duration and number of steps, is highly demanding. SOI has attracted the highest attention among the straightforward MEMS fabrication processes. SOI wafer consists of a single crystalline silicon substrate on a buried oxide (BOX) layer which is set on top of a silicon substrate, i.e. active devices are implemented in the top single crystalline silicon layer, which is isolated by an insulator from the bottom layer, which can be either single/poly crystalline silicon as
a mechanical support. The sandwiched insulator consists of silicon dioxide or, in some cases, silicon nitride films. It has been used in analogue and digital integrated circuit technology for electrical isolation of each transistor from its adjacent transistors, which otherwise should be insulated by substrate reverse-biased pn junctions [38].

The wide usage of this structure in MEMS is due to its reliable electrical insulation, excellent etching stop and sacrificial layer functions. It, therefore, increases the fabrication accuracy, process simplicity and device performance [32]. SOI has also been routinely used in optical MEMS products, such as variable optical attenuators, optical switches, high resolution spectrometers and accelerometers [39].

Multi-User MEMS Processes (MUMPS) offers simple yet robust MEMS fabrication technology in the following categories [2]:

1. Metal-MUMPS
2. SOI-MUMPS
3. Poly-MUMPS.

The SOI devices in the MUMPS technology are fabricated through a four-mask process, starting on 150mm SOI wafers, as illustrated in Figure 1.3. The first deposition layer is the 500nm gold (with minor chrome proportion) pad, which is patterned through a lift-off process and used as a wire bonding pad in the devices designed in this thesis. The second mask patterns the 25um doped silicon (red), which is the main layer used as mechanical structures, resistors and/or electrical routing. Then the silicon substrate (blue) is etched from the bottom side of the wafer. The etching process releases the moving parts of the device.
1.4 Structure of the Thesis

As stated in section 1.1, the displacement sensors determine the maximum resolution of nanopositioning devices. Therefore, the available sensing schemes in MEMS with potential applications in nanopositioning are discussed from both the device design and readout circuit perspectives in Chapter 2. After giving brief definitions of sensor specifications, such as resolution, linearity and drift etc. the operating principles of displacement sensors, including the capacitive, electrothermal, piezoresistive and magnetoresistive, are addressed. This is followed by a review on the sensor readout circuits with a focus on the frequency based measurements.

In Chapter 3, a new frequency based measurement system is introduced for electrothermal sensor in a 1-Degree of Freedom (1-DoF) nanopositioner. Chapter 4 focuses on noise reduction techniques for electrothermal sensors. Two novel techniques are introduced and implemented for reducing the noise level at the MEMS device. In Chapter 5, the electrothermal sensors are integrated to a 2-DoF nanopositioner, which can be used as a scan table for atomic force microscopy. A differential piezoresistive displacement inbuilt sensing system is designed and implemented for a 1-DoF nanopositioner in Chapter 6. Chapter 7 concludes the thesis with detailed information about the achievements in MEMS displacement sensors.
2 MEMS Displacement Sensors and Interface Circuits

A displacement sensor is a transducer that produces an electrical output, such as electrical charge, current or voltage, proportional to the position of a moving mechanism. The sensor performance is commonly assessed by its linearity, drift and stability, bandwidth, noise and resolution. Cost and form factors are also important parameters in choosing a suitable sensor for a particular application [19]. In this chapter, after a brief introduction to the sensor performance criteria, the operation principle and specifications of displacement sensing techniques including capacitive, electrothermal, piezoresistive and magnetoresistive sensors are investigated. The MEMS resistive sensors can be implemented in smaller areas compared with capacitive sensors. Therefore, electrothermal and piezoresistive sensors are elaborated on with more details. The electronic interface circuits used for resistive displacement sensors are also discussed afterwards.

2.1 Sensor Parameters

2.1.1 Linearity, Sensitivity and Offset

Transfer function is the functional relationship between the physical input, which is the displacement in position sensors, and the electrical output signal [19]. It can be
represented graphically by a characteristic curve, as shown in Figure 2.1. In general terms, \textit{linearity} is a measure of the manner in which the error varies with the range in the transducer and is expressed as the \textit{rms} error or the maximum error in the sensor because of its characteristic deviation from a straight line. \textit{Sensitivity} is the ratio of small variations in the output electrical signal to small variations in the input physical signal, i.e. the slope of the characteristic curve. More information can be obtained from the transducers’ \textit{frequency response} characteristic, rather than a single value, which shows how the output amplitude and phase changes versus frequency for the set level of the input [40]. All sensors have measurement uncertainties, such as \textit{offset} and \textit{gain error}, that must be considered in practice. This can be seen in Figure 2.1, which illustrates the inverse transfer function of a capacitive displacement sensor. The displacement is mapped to sensor voltage by a nonlinear function. The transfer function nonlinearity is the result of varying sensitivity. The error voltage is produced as the difference between the real measured and the linear expected values. In order to mitigate the nonlinearity, the sensor needs to be calibrated. The calibration process consists of finding the most accurate measurement and minimising the error by the corresponding electronic readout circuits. Although this measurement is routinely accomplished by an optical interferometer [41], however, in [42] and [43] capacitive and piezoresistive sensors are used for calibration. The calibration process involves systematic errors, such as gain, offset and linearity, and excludes random imperfections, such as noise. Let $f_a(v)$ and $f_{\text{cal}}(v)$ represent the actual measured and linear calibrated value of displacement, respectively. The mean-square error value is defined as:

$$
\epsilon = \sum (x_i - f_{\text{cal}}(y, v_i))^2,
$$

(2.1.1)

where $v_i$ and $x_i$ are the measured data points and $\epsilon$ is the optimal parameters vector for $f_{\text{cal}}(\epsilon, v)$. It is expected to be minimised after calibration. The simplest
Figure 2.1: Nonlinear characteristic curve of a capacitive displacement sensor.

calibration curve is described as:

\[ f_{\text{cal}}(v) = \varepsilon_0 + \varepsilon_1 v, \]  

(2.1.2)

where \( \varepsilon_0 \) and \( \varepsilon_1 \) are the sensor offset and sensitivity, respectively. Higher order polynomials can also be used in more complex calibrations. The real value of position can be obtained after calibration. The difference between calibration function and actual value is defined as calibration error:

\[ e_m(v) = f_a(v) - f_{\text{cal}}(e, v), \]  

(2.1.3)

To report the sensor linearity performance, and make comparisons between different products, their nonlinearity is reported as a percentage of full-scale range (FSR):

\[ \text{nonlinearity} = \frac{\max|e_m(v)|}{FSR} \]  

(2.1.4)

The linearity discussion so far has involved the sensor static behaviour, which is subject to ambient conditions, including temperature, vibration, acoustic noise level and humidity. In contrast, dynamic linearity represents the sensor agility in following
the fast inputs, and is defined by the sensor response time and the distortion in
phase and amplitude of the sensor response to fast changing inputs. As illustrated in
Figure 2.2, the symmetric and asymmetric calibration curves lead to even and odd
harmonic distortion. It represents the sensor response at frequency $w$ to excitations
at harmonics such as $nw$. Total harmonic distortion and intermodulation distortion
can be investigated to compare the dynamic linearity of sensors [40].

The *Pearson correlation coefficient* (PCC) is another parameter that quantifies the
system linearity. PCC determines the linear dependence (correlation) between two
variables by a figure between $+1$ and $-1$, with 0 for no correlation:

$$
\rho_{XY} = \frac{\text{cov}(X, Y)}{\sigma_X \sigma_Y} = \frac{E[(X - \mu_X)(Y - \mu_Y)]}{\sigma_X \sigma_Y} \tag{2.1.5}
$$

where $\sigma, \mu$ and $E$ are standard deviation, mean and expectation of the variable,
respectively.

In addition to calibration, which is a signal processing method and needs data
conversion, device level solutions have also been proposed to improve the linearity of

**Figure 2.2:** Dynamic Nonlinearity: a) odd harmonics, b) even harmonics.
sensors. For example in [11], where the electrothermal sensor is modelled as a first order system. The differential sensing is proven to enhance the linearity through the pole cancellation by a zero due to the differential structure. Also in Chapter 3, the combination of actuator and sensor readout circuit nonlinearities produces a linear input-output transfer function.

2.1.2 Noise and Resolution

The production of nano-devices that look like micromechanical sensors proves that the miniaturisation of mechanical sensors is not limited by the fabrication processes of small structures. In fact, the performance limits of these sensors are set by noise [19]. For instance, the mechanical noise power that is determined by thermal noise energy \((k_B T)\) is fixed for all devices \((k_B \text{ and } T \text{ stand for Boltzmann constant and temperature, respectively})\). Due to the fact that the smaller devices presumably produce lower signal levels, their signal to noise ratio is drastically influenced by noise.

Noise is the random fluctuation of atomistic particles, including electrons, atoms and molecules. For instance, the random motion of electrons in conductors results in voltage noise. Most of the mechanical noise sources, such as vibrations, are extrinsic. However, the thermal vibrations of atoms in a mass result in random fluctuations of the mass position, which is called Brownian noise [44]. The bulk micromachined MEMS products are influenced by electrical noise while the surface micromachined devices are prone to mechanical noise. Sensor noise is generally defined as the random fluctuations at the sensor output with no input signal. It is an inherent property of a sensor to generate noise in addition to the output signal, which is a real translation of the physical input. Sensor noise is mostly generated by the resistive elements, which contribute to Flicker and Thermal noise, and the readout circuitry. It is investigated,
with more details on the electrothermal displacement sensors, in this thesis.

The probability density function for most physical noise sources is Gaussian, i.e. despite the fact that the instantaneous deterministic value is not measurable, it falls between certain values, with certain probability, as follows:

\[ p(v_n) = \frac{1}{\sqrt{2\pi}v_{rms}} e^{-\frac{v_n^2}{2v_{rms}^2}} \]  

(2.1.6)

where \( v_n \) represents the measured noise voltage and \( v_{rms} \) is the standard deviation from the average voltage. Although the voltage noise of bulk micromachined devices is dealt with in this thesis, the rules are applicable to any random noise regardless of its origin and type. The probability that the noise voltage value falls within \( v_1 \) and \( v_2 \) is obtained by integrating the Gaussian distribution function:

\[ P = \int_{v_1}^{v_2} f(v_n)dv_n \]  

(2.1.7)

In the presence of multiple noise sources, their interaction is determined by their correlation status. Correlated noise sources are added like deterministic signals:

\[ v_{rms,tot} = v_{rms,1} + v_{rms,2} \]  

(2.1.8)

However, the total noise generated by two uncorrelated noise sources is:

\[ v_{rms,tot} = \sqrt{v_{rms,1}^2 + v_{rms,2}^2} \]  

(2.1.9)

The rms-noise represents the noise magnitude in the time domain. The frequency domain characteristic of the noise is determined by its power spectrum or power spectral density (PSD), which reflects the amount of noise energy that exists per
2.1 Sensor Parameters

unit bandwidth of frequency, on average, and is demonstrated by $\overline{v_n^2}$. In order to find the total noise energy (rms value), PSD must be integrated over the frequency bandwidth. In addition, a system response to a noise input can be obtained by the noise PSD, provided that the system transfer function is identified. The noise sources are named by corresponding colours based on their PSD. However, in instrumentation applications most of the physical noise sources are categorised as white noise and low frequency noise. The white noise PSD is constant and is independent of frequency. Though, in a measurement, the noise is shaped by the system transfer function:

$$
\overline{v_{n,\text{out}}^2(f)} = |H(f)|^2 \overline{v_n^2}
$$

(2.1.10)

where

$$
H(f) \equiv \frac{v_{\text{out}}}{v_{\text{in}}}
$$

(2.1.11)

represents the system transfer function. The low frequency noise includes all the noise sources that their energy is distributed inversely proportional with the frequency. Common low frequency noises are flicker and Brownian noises. The underlying mechanisms of these noise sources are different but they all adversely affect the low frequency signal measurements. For example, the random generation and recombination of charge carriers is introduced as the origin of the flicker noise in electronic components, while the quartz mechanical resonators are prone to flicker noise due to nonlinear forces in the quartz crystal. The electrical flicker noise is more crucial in the available MEMS devices, however in nanoscale devices mechanical flicker noise can dominate. Flicker noise can be generally described as [19]:

$$
\overline{v_{1/f}^2} = \sqrt{\overline{v_n^2} \frac{f_c}{f}}
$$

(2.1.12)

where $f_c$ denotes the corner frequency as the frequency where the $1/f$ noise amplitude
equates to the white noise amplitude $v_n^2$. The rms value of the flicker noise is obtained as:

$$v_{rms}^2 = \int_{f_L}^{f_H} v_f^2 df = v_n^2 f_L \ln(f_H/f_L)$$  \hspace{1cm} (2.1.13)

where $f_L$ and $f_H$ are the lower and upper frequency limits of measurement.

As stated earlier, resistors are one of the major noise sources in electronic systems. In particular, the contribution of the resistive sensor device to the total noise is higher because of its location at the input of the readout circuit. Due to the focus of this thesis on the resistive sensors, a review of the noise sources in resistive elements would provide a good insight.

The minimum level of the detectable signal on a resistive element is set by the thermal or Johnson-Nyquist noise power, which has a white spectral density:

$$\overline{v_n^2} = 4k_BT R$$  \hspace{1cm} (2.1.14)

where $k_B$ is the Boltzmann constant, $T$ is the temperature (in Kelvin) and $R$ is the resistance value.

An advanced expression for the resistor flicker noise voltage spectral density is described by the experimental Hooge model:

$$v_n = \sqrt{\frac{\alpha}{fN}} V,$$  \hspace{1cm} (2.1.15)

where $V$ is the bias voltage across the resistor, $N$ is the number of charge carriers, $f$ is the frequency and $\alpha$ (Hooge factor) is experimentally determined and depends on the material and geometry of the device. The larger value of $\alpha$ for resistive devices, compared to other semiconductor devices, has been a matter of controversy. Several heuristic and theoretical approaches were used for determining its value in resistive
2.1 Sensor Parameters

elements and transistors [45, 46, 47]. However, in this work we are looking at systemic methods to cope with noise rather than material and process solutions. Also, the proposed techniques in this thesis are evaluated by experimental comparisons on the same device, which makes the argument independent of the absolute value of $\alpha$. The flicker noise has been distinctly investigated for single and polycrystalline silicon devices. Monocrystalline silicon is utilised in the silicon on insulator process, which demonstrates lower flicker noise level [14, 48].

*Shot noise* is generated by random current variations across a potential barrier. For example, the random diffusion of electrons and holes through semiconductor junctions generates the current noise, whose spectral density is expressed as:

$$\bar{t}_n^2 = 2q_e I$$  \hspace{1cm} (2.1.16)

where $q_e$ is the electron charge and $I$ is the current passing the junction. It is assumed that there is no potential barrier in resistive elements except for poor contacts between metal pads and the semiconductor. Due to the fact that shot noise dominates only in ultra-high resolution optical MEMS displacement sensors [49], it is beyond the scope of this thesis to be further investigated.

*Noise floor* is an indicator of the amount of the output signal uncorrelated with the input. The noise floor shows the magnitude of the time varying output signal level without any input signal, which cannot be detected without the use of sophisticated signal processing approaches. Its value can also be calculated by integrating the noise PSD on the desired frequency bandwidth. It is often used in the thesis to compare the different sensing techniques in a certain bandwidth for evaluation purposes. Noise floor is reported electrically (V and A) or in terms of the desired physical quantity such as displacement (m).
Resolution is another performance characteristic of sensors, which is the smallest step that can be reliably resolved by the sensor. The transducer sensitivity and electrical noise floor determine its resolution. For example, resolution of the displacement sensor can be derived by dividing the sensor’s electrical output noise floor by its displacement sensitivity.

### 2.1.3 Drift and Stability

The accuracy of the sensor transfer function can be affected by changes in the transfer characteristic, in addition to the nonlinearity error argued earlier. Stability defines how constant the sensor output is in constant conditions. Drift is the sensor output tendency to vary monotonically and very slowly, compared to the times associated with sensing [3, 19, 40]. Changing temperature, humidity etc. which are called “influence quantities”, can also cause drift. There is no standard measure for drift, but it is possible to limit the variations caused by an uncertainty in the sensitivity and offset [3]:

$$f_a(v) = (1 + k_s)f^*_a(v) + k_o,$$  

(2.1.17)

where $k_s$ and $k_o$ are the sensitivity and offset variations, respectively and $f^*_a(v)$ is the nominal calibrated transfer function. Consequently, the mapping error, with the inclusion of sensitivity and offset variations, is:

$$e_d(v) = (1 + k_s)f^*_a(v) + k_o - f_{cal}(v).$$  

(2.1.18)

Assuming that the nominal mapping error is negligible ($f^*_a(v) \simeq f_{cal}(v)$), Equation 2.1.18 can be simplified to:

$$e_d(v) = k_s f_{cal}(v) + k_o$$  

(2.1.19)
Therefore, the maximum error due to drift is:

$$e_d = \pm (k_s max|f_{cal}(v)| + K_o) \quad (2.1.20)$$

### 2.1.4 Bandwidth

The bandwidth of a displacement sensor determines the maximum frequency range that it can follow the displacement variation. It translates to the frequency at which the magnitude of transfer function $v(s)/x(s)$ drops by -3dB, where $x$ and $v$ represent the displacement and sensor output voltage, respectively. The resolution of a sensor is determined by integrating the noise floor over its bandwidth, but more information is needed to predict errors caused by sensor dynamics. The position error in frequency domain can be expressed as:

$$e_{bw} (s) = x(s)(1 - p(s)), \quad (2.1.21)$$

where $p(s)$ is the sensor transfer function. Maximum error for a sinusoidal displacement, with amplitude $A$, is:

$$e_{bw} = \pm A|1 - P(s)|, \quad (2.1.22)$$

A higher order filter is expected to reduce the error due to its faster roll-off, however, the longer phase-lag associated with a higher order filter negates this assumption. Figure 2.3 demonstrates the error of a Butterworth filter response against the normalised frequency. Assuming filter poles coincide with the sensor cut-off frequency, the low frequency magnitude of $|1-P(s)|$ is approximately:

$$(1 - P(s)) \approx n \frac{f}{f_{bw}}, \quad (2.1.23)$$
where \( n \) is the filter order and \( f_{bw} \) is the bandwidth. This translates to:

\[
e_{bw} \approx \pm An \frac{f}{f_{bw}}.
\]

Consequently, the error is proportional to the signal magnitude, filter order and normalised frequency. This is particularly important considering the fact that to avoid the dynamic errors due to the sensor, its bandwidth should be significantly higher than the operating frequency.

### 2.2 Displacement Sensing Schemes

The high precision displacement sensing schemes can be classified in two main categories. The integrated sensors, which are embedded with the moving object in either standard or customised MEMS processes, and the discrete sensors, which are built separately. The emphasis of this work is on the sensors available in MEMS.
2.2 Displacement Sensing Schemes

Figure 2.4: Embedded optical encoder micromachined in glass: (a), (b) and (c) show the motion schematically, and (d) is a microscopic image [4].

Therefore, the discrete sensors are introduced briefly, and a more detailed review is conducted afterwards for integrated sensors, including capacitive, electrothermal, piezoresistive and magnetoresistive sensors.

Optical (laser) interferometers are possibly the most commonly used high precision displacement sensors. These expensive and bulky devices can achieve nanometer precision with the highest accuracy, stability and linearity compared with other sensors. However, interferometer is not comparable with low range sensors such as capacitive and electrothermal sensors, due to the fact that it is capable of measuring several meters [3]. An integrated optical displacement sensor was patented recently, which was reported in [4], based on the optical waveguide. The monolithic transducer made on a glass substrate measures the displacement using the intensity variation between a fixed and a moving waveguide array. The sensor structure and operation are shown schematically in Figure 2.4. At position (a), the waveguides are all...
aligned and the transmitted signal is maximum. As the stage moves to the right, to position (b), the middle waveguide is not aligned any more and the received signal is significantly lost. Further movement to the right realigns the waveguides. An optical microscope image of the device is illustrated in Figure 2.4(d), in which the waveguides, and moving and fixed parts are labelled. An under 100nm resolution is reported for this sensor without enough measurement elaboration.

The linear variable differential transformer (LVDT) is another widely used displacement sensor. Displacements of the core in a coil leads to variations in the induced voltage at the coil terminals. It is a rather large structure that needs to be mechanically coupled to the moving stage, which has adverse effects on the performance of microdevices. Also the measurement range is between 1mm and 50 cm, which does not fit in the desired ranges [3]. The limited resolution of LVDT in [50] has imposed special considerations in the design. Its sensitivity is 6.76μm/V, where the total travel range is around 100μm.

### 2.2.1 Capacitive Sensors

Capacitive sensors measure different physical quantities, such as acceleration, velocity, displacement, humidity etc., by detecting the current through the capacitor, which is given by:

\[
\frac{\partial}{\partial t}(CV) = C \frac{\partial V}{\partial t} + V \frac{\partial C}{\partial t},
\]

In a typical measurement, one of the two terms in Equation 2.2.2 dominates. If the sensor operates based on the first term, it is called the displacement measurement. Otherwise, if it operates based on capacitance change, it is called the velocity or rate of change measurement. In the second scenario, the sensor output current is
very small at low frequencies. Therefore, it is not used for low frequency or static measurements. To operate the sensor in displacement measurement mode, a high frequency signal drives the sensor and the output current is expressed as:

\[ i \approx C \frac{\partial V}{\partial t} = CV, \]  

(2.2.3)

where \( \dot{V} \) is the time derivative of the voltage [19].

The operation of the MEMS capacitive sensors follows the same principles as their macro scale counterpart. However, a smaller footprint necessitates advanced geometric design to achieve a practical value of capacitance. Figure 2.5 demonstrates a geometric evolution of MEMS capacitive sensors [3]. The basic comb sensor shown in Figure 2.5(a) is used most frequently for nanopositioning applications, as in [15, 29]. In [15] it is experimentally shown that the noise at the capacitive sensor output has a lower low-frequency power spectral density, compared with the electrothermal sensor. The basic comb performance is enhanced by the modification made as shown in Figure 2.5(b) [3]. Terminals 2 and 3 are driven by 180 degrees out of phase signals to ground the comb connected to terminal 1. This adds all the benefits associated with the differential sensing, such as higher sensitivity. However, the travel range of the object whose displacement is being measured is limited to finger spacing.
The configuration in Figure 2.5(c) is developed to mitigate the limited travel range issue [51, 52]. Its operation is based on the encoder principle, originally designed for optical interferometers. As comb 1 moves horizontally an alternative phase change occurs between terminals 2 and 3. The position information is obtained by decoding these voltages. Despite the increased travel range, the resolution of the incremental capacitive encoder is not as much as that which can be achieved by conventional capacitive combs. Compared with other sensing schemes, capacitive sensors occupy a great portion of the device area. The sensor bandwidth is not measured in the reported designs because the capacitive sensing bandwidth is wide enough to cover almost all instrumentation applications.

### 2.2.2 Resistive Sensors

Resistive position sensors convert the displacement to a resistance variation. Despite the differences in sensing mechanisms, the final output is a resistor variation, which offers both simplicity, from the readout circuit design perspective, and compactness, in terms of occupied area. However, power consumption is higher in these sensors and their inherent noise affects the measurements precision drastically.

#### 2.2.2.1 Electrothermal Transduction

Air is a good thermal insulator, i.e. the heat conduction between two objects in a macroscopic domain is a slow process. However, in the case of objects with micrometer distance, which are feasible using MEMS technology, the heat conduction dominates other mechanisms, including convection and radiation. It is also efficient and fast enough to be used as a function of distance for a sensing mechanism. The displacement sensing with electrothermal sensor was first proposed in [53] for high density data storage device, which was itself inspired from the usage of nanometer-
sharp tips for imaging and investigating the structure of materials in atomic force microscopy [27]. During the write process of the storage device, indentations made by the sharp tip of a silicon cantilever on a heated polymer surface represents logic bits (ones and zeros) as illustrated schematically in Figure 2.6. During the read process the same heater acts as a sensor, i.e. a constant electrical power is applied to the heater and the distance between the heater and the storage media (polymer surface) modulates the temperature of the heater. The distance between the heater and the polymer changes by 10nm while scanning bit indentations [5]. The temperature change is measured by the heater resistance changes, and depends on the heat sinking capability of the heater-surface environment. An order of magnitude higher sensitivity ($\Delta R/R = 10^{-5}/nm$) is reported here for electrothermal sensors, compared with piezoresistive strain sensing, which will be introduced next. However, the required accuracy for positioning the cantilever tip is 10nm, which is to cover a bit area, which is around 30nm. A more detailed structure of the cantilever is schematically illustrated in Figure 2.7. It consists of a heater platform with the tip on
top, the legs acting as a mechanical spring and an electrical connection to the heater. Silicon is used as the main material for its good mechanical and thermal stability. The resistivity of the heater is 15 times more than the highly doped interconnection (400Ohms) and the heater time constant is about 1μs. Further mechanical and device level information are provided in [5], [53] and [54].

Another example of a thermal conduction based transducer is the thermopile accelerometer that is reported in [6]. As illustrated in Figure 2.8, the proof mass acts as a heat sink and the polysilicon thermopile generates heat. The heat transfer rate between the heater and the suspended proof mass determines the acceleration. The sensitivity, temperature dependence, frequency response and linearity of this two-part device are comparable with a commercial accelerometer. It shows less temperature dependence than a piezoresistive accelerometer and it mitigates the parasitics issue associated with capacitive sensors.

The temperature dependence of silicon resistance is used later in [7], where again the
heat conduction rate change is exploited for displacement sensing. As illustrated in Figure 2.9, the heater consists of an elongated U-shaped cantilever, which is made of single-crystal silicon in a standard bulk micromachining fabrication process. The highly doped holding legs act as conductive leads to deliver electrical power to the middle moderately doped heater to increase its temperature. Hence, after applying power the resistance of the middle part is much larger than the legs, even with equal temperature, which is particularly important for higher sensitivity.
The usage of a heater as a displacement sensor requires the object of interest to move at a very close vicinity to the heater, so that the heat transfer mechanism can be dominated by high efficiency and fast conduction [34]. This is schematically illustrated in Figure 2.10, where two heaters are placed on top of a moving stage so that their surface is parallel to the stage in order to increase the conduction efficiency. This structure also facilitates the differential displacement sensing, whose benefits are explained later. Applying electrical power heats up the resistors and increases their electrical resistance. Thermal conduction develops through the air between the heater and the stage, which acts as a heatsink. As far as the stage overlaps equal areas of the heater in Figure 2.10(a), the thermal conduction rate is identical for both heaters. Therefore, their temperature and resistances are equal. As the stage moves toward the direction shown in the schematic, the overlapped area for $R_1$ increases, which raises the thermal conduction rate, and reduces the heater temperature, as well as its resistance. Opposite simultaneous changes take place for $R_2$. The resistance changes can be measured by the Wheatstone bridge or a transimpedance amplifier.
and converted to voltage in a straightforward procedure. The two methods are compared using analytic and experimental approaches in Chapter 5.

The conduction between the heaters and the stage is the most efficient heat transfer mechanism in the mentioned system. However, there are unwanted interferences that affect the sensor resistance. Although a single sensor is capable of achieving a very high displacement resolution, the differential architecture makes the system robust against external interferences, such as ambient temperature and ageing. Also this structure rejects power supply noise provided that there is a good matching between the resistors and also between the amplifiers. The sensitivity to any undesired motion that affects both sensors in the same way is suppressed by differential sensing. This property is particularly important in stages with more than 1-degree-of-freedom (1-DoF), where displacement in $x$ and $y$ directions, for example, need to be differentiated.

In order to characterise the sensors, an optical measurement instrument is utilised to carry out a direct displacement measurement of the stage. Voice coil actuators are used to move the stage, which can achieve long range strokes compared with electrostatic actuator. The sensitivity of these sensors can be increased by reducing the spacing between the sensor and the stage, and also by increasing the heating electrical power. However, this leads to an increase in the thermal noise level of the heater as illustrated in Figure 2.11. Therefore, the electrical power level determines the optimum point. Increasing the heating power up to 32mW results in a resolution increase, in this case, and after that downgrades the resolution. The bandwidth of this device is measured by applying a step voltage signal to one of the heaters while bypassing the other with a constant current source. The increase in bias voltage increases the temperature of the resistor, which raises the resistance in turn. The heat transfer between the heater and the surrounding ambient is modelled as a first
order system which determines the time response of the sensor. This device can achieve 0.5nm displacement resolution in a dc to 10kHz bandwidth within ±50μm stage strokes. Further detailed information about the device design and operation is provided in [55].

All previously reported works have either assembled the electrothermal actuator with off-chip actuators or utilised specialised fabrication processes to integrate the electrothermal sensor with actuators. In [8, 9], the electrothermal sensor is integrated to actuators on the same chip in a single mask monolithic process. Both works describe 1-DoF nanopositioners actuated electrostatically and electrothermally, respectively. The integration of sensor and actuator avoids misalignment, friction and hysteresis [56]. Figure 2.14 and Figure 2.13 illustrate the chip micrographs of monolithic nanopositioners in which electrothermal sensors are integrated to electrostatic and electrothermal actuators, respectively. The in-plane motions of the suspended stages are carried out by the actuators in a left-right direction. The sensor designed in [9] is characterised in a closed loop, equipped with a PI

![Graph](image)

**Figure 2.11:** Displacement sensor sensitivity and noise versus applied electrical power [7].
Figure 2.12: a) Optical micrograph of the 1-DoF nanopositioner with electrothermal sensor and electrostatic actuator [8], b) Magnified SEM image of the Sensor 1 in the chip.
controller to improve the positioning accuracy and robustness of the system. It has been experimentally shown that the positioning error of the nanopositioner is improved from 0.62 µm in open loop to 7.9 nm in closed loop setup. The measurements are carried out by applying step inputs to the actuator input and by comparing the optical observations with the sensor output voltage.

The schematic diagram in Figure 2.14 illustrates the dimensions of the electrothermal sensor and its surrounding air-gaps. An analytical model that divides the heater into several lump elements, consisting of the silicon heat capacitance coupled to its electrical resistance, is proposed in [8]. The model facilitates an optimised sensor design to achieve the highest sensitivity as a function of the doping concentration, the heater dimensions, temperature and operating mode (constant current/voltage). The sensor height is determined by the silicon layer thickness in the fabrication technology. The length and width of the heater is investigated in the model. According to the thermal power flow from the heaters towards the substrate and bond-pads, both
of which are assumed to be losses to sensitivity, an optimum length is calculated. Model based simulations also show that the lower heater width results in a higher resistance change when the stage moves from minimum to maximum overlap, i.e. higher sensitivity, which is also verified by experiments. However, the minimum width of the heater is limited by the fabrication technology as well. Doping concentration, temperature and the operating mode are investigated next, with further details in [11, 34].

Despite the simple design principle and wide variety of applications, most of the initial models investigate the underlying mechanism in quasi-static conditions [25, 26, 57]. However, some recent studies address the system dynamics, which is essential for high speed applications [34]. The numerical approaches seem rather attractive due to complications in the heater thermal and electrical responses [58, 59]. A control system approach, introduced in [10, 11, 60] to model the electrothermal sensor as
This is followed by the incorporation of the sensor noise, which bandwidth of the read sensor for the probe-storage application.

The write microheater used in thermo-mechanical probe-storage applications is used for this paper.

The transfer function approach presented in this paper extends the dissipated power separately. Note that this separation concept is not really novel [19], [22], but so far, it was mostly

The two main components of the model are the heater voltage is the electrical input to form indentations thermomechanically.

The read heater has a lateral dimension of the microheater compared with analytical or numerical thermal system. The sensing problem gets translated to the signals measured are thought of as disturbances to the microheater operator model and its experimental identification in more detail. The write microheater used in thermo-mechanical probe-storage applications is designed for write/read heaters in a prob-based data storage system but it is justication for the linearity assumption is presented in the next

one can systematically derive these voltage profiles. If the first assumption could be violated, note that, in some cases, the first assumption is estimated to be less than a few tens of degrees.

As illustrated in Figure 2.15, the model consists of two main operators: \( T_{TP} \), that translates power to temperature and, \( g(\cdot) \), that maps the temperature to electrical resistance. The system input is the bias voltage across the heater. The heat transport, including the diffusion within the heater and the conduction through air to the stage,
is assumed to be a linear function of the electrical power delivered to the heater. Another assumption is that $g(\cdot)$ is memoryless, but that it can be nonlinear. This is also explained by the extremely small thermal equilibration time of the phonon compared with electrons. A rather straightforward experimental method is used to identify $g(\cdot)$. That is, to measure the current output of the heater by varying its bias voltage. The resistance versus temperature could be obtained as shown in Figure 2.16. Assuming that $T_{TP}$ is linear, its frequency response can fully identify the characteristic. A small noise/chirp signal ridden a dc voltage is applied to the heater bias and the corresponding current output signal is measured. Consequently, the resistance fluctuation is in hand and by means of the $g(\cdot)$ operator inverse the temperature is attained:

$$\bar{T} = g^{-1}(R) - g^{-1}(R_0),$$

(2.2.4)

Then the power fluctuations is derived as:

$$\bar{P} = VI - V_0I_0,$$

(2.2.5)

where $R_0, V_0$ and $I_0$ represent the nominal values of $R$, $V$ and $I$, respectively. The frequency response of the resulting transfer function, $T_{TP}$, is illustrated in Figure 2.17, according to which a third order fit can give an appropriate description of the system. It should be noted that the mentioned model is for a write heater only. To extend it to cover the sensor (read) heater, the displacement functionality of $T_{TP}$ should be taken into account. The heater displacement over an indentation (see Figure 2.6) reduces the distance between the substrate (stage) and the heater. This change in the heater-substrate distance ($x$) decreases the heater temperature, and its resistance reduces consequently. The displacement dependence of $T_{TP}$ is entered to the model as shown in Figure 2.18, where $x$ only affects the amplitude of $T_{TP}$. Figure 2.18(b)
shows the linearised model assuming that only small signal fluctuations are entered. This model is derived by Taylor series expansion of nonlinear transfer functions around their operating point as:

\[ V = V_0 + \tilde{V}, \quad I = I_0 + \tilde{I} R = R_0 + \tilde{R}, \quad T = T_0 + \tilde{T} \quad \text{and} \quad P = P_0 + \tilde{P} \quad (2.2.6) \]

The model is ruled by the following transfer functions:

\[ \tilde{T} = T_{TPx} \left( \tilde{P} + \frac{K'(x_0)}{K(x_0)} P_0 \tilde{x} \right) \quad (2.2.7) \]

\[ \tilde{P} = -I_0^2 \tilde{R} + 2I_0 \tilde{V} \quad (2.2.8) \]

\[ \tilde{R} = g'(T_0) \tilde{T} \quad (2.2.9) \]

\[ \tilde{I} = -\frac{I_0}{R_0} \tilde{R} + \frac{\tilde{V}}{R_0} \quad (2.2.10) \]
Figure 2.18: a) The extended model for heater stage distance inclusion, b) Linearised model and c) the contribution of displacement (x) [10].
where operating the heater with constant bias voltage dictates:

\[ \tilde{V} = 0 \quad (2.2.11) \]

\( T_{I\tilde{V}_x} \) is the experimentally measurable transfer function and \( T_{TP_x} \) is related to \( T_{I\tilde{V}_x} \) as:

\[ T_{I\tilde{V}_x} = \frac{1}{R_0} \left( \frac{1 - I_0^2 g'(T_0) T_{TP_x}}{1 + I_0^2 g'(T_0) T_{TP_x}} \right) \quad (2.2.12) \]

The desired transfer function to characterise the sensor is \( T_{I\hat{X}_x} \), which describes the displacement dependence of output current. To define the displacement contribution in the model it is assumed that \( x \) affects only the amplitude of \( T_{TP_x} \), which simplifies the problem. The dynamics of \( T_{TP_x} \) are described as:

\[ T_{TP_x}(s) = K(x) \frac{b(s)}{a(s)} \quad (2.2.13) \]

where \( K(x) \) is the gain purely dependent on the heater-stage distance, \( x \), and \( b(s) \) and \( a(s) \) are the polynomials of \( s \). The following extension linearises \( T_{TP_x} \) around the operating point \( x_0 \):

\[ T_{TP_x}(s) = K(x_0) \frac{b(s)}{a(s)} + K'(x_0) \frac{b(s)}{a(s)} \tilde{x} \quad (2.2.14) \]

Hence, the small signal model is modified as illustrated in Figure 2.18(c), from which the displacement functionality of the sensor output current is derived as:

\[ T_{I\hat{X}_x} = \frac{K'(x)}{K(x)} \left( \frac{I_0 - g'(T_0) P_0 T_{TP_x}}{R_0 1 + I_0^2 g'(T_0) T_{TP_x}} \right). \quad (2.2.15) \]

One of the drawbacks of the electrothermal sensor is its noise, which is strongly dominated by the noise sources inherent to silicon. Regardless of the crystal ori-
entation, the doped silicon generates flicker and thermal noise. The power spectral
density of flicker and thermal noise were described earlier. Two distinct approaches
are proposed in Chapter 4 to mitigate the noise inherent to the sensor device. The
displacement sensor can be operated by either a constant voltage or constant current
supply [8]. The aforementioned model is modified in Chapter 4 to investigate the
displacement sensor resolution under constant voltage and constant current supply.

### 2.2.2.2 Piezoresistive Transduction

The effect of mechanical stress on electrical resistance (piezoresistance) had been
widely studied before the large piezoresistivity in silicon measured in [37]. Finding
the higher sensitivity in semiconductors (around fifty times), compared with former
metal strain gauges, was a breakthrough in 1950’s. Following developments in
semiconductor technology resulted in integrated piezoresistive sensors, i.e. instead of
using cement for connecting the sensitive element to the force collector, they were
co-fabricated on one device. The electrical resistance can be expressed in term of
physical dimensions and resistivity ($\rho$) as [12]:

$$ R = \frac{\rho l}{A} \quad (2.2.16) $$

where $l$ and $A$ are the length and the cross section area of the device, respectively.
Mechanical pressure can influence the resistance value either by changing the device
dimension or its resistivity. The cross section decrease, in response to the longitudinal
strain, follows Poisson’s ratio, $\nu$, which varies from 0.20 to 0.35 for most metals and
from 0.06 to 0.36 for anisotropic silicon. The fractional resistance change due to
geometric and resistivity variations can be expressed as:

$$ \frac{\Delta R}{R} = (1 + 2\nu)\epsilon + \frac{\Delta \rho}{\rho} \quad (2.2.17) $$
where $\epsilon$ is the mechanical strain. Although the resistance of metals and semiconductors are affected almost similarly by geometric changes, however, silicon and germanium resistivity can be 50-100 times more sensitive than metals. The silicon crystal orientation determines its directional piezoresistivity, e.g. a $<110>$ aligned piezoresistor on a (100) wafer is proved to have high equal and opposite longitudinal and transverse piezoresistive coefficients. Crystal orientations are determined during the lithography and etching processes.

The piezoresistance of silicon and germanium is expressed by the piezoresistivity coefficient ($\pi$). A comprehensive definition of $\pi$ requires tensors with four subscripts to consider the applied electric field (potential difference), current density and two stress components. The piezoresistivity in a certain voltage and current polarity ($\omega$) is expressed as [61]:

$$\frac{\Delta \rho_{\omega}}{\rho} = \sum_{\lambda=1}^{6} \pi_{\omega\lambda} \sigma_{\lambda}$$  \hspace{1cm} (2.2.18)

where $\sigma$ is the mechanical stress. Four different possibilities of the device are demonstrated in Figure 2.19. In A and C the longitudinal piezoresistance and in B and D the transverse piezoresistance of the device is illustrated through voltage drop measurement. The device is biased by a constant current source. The origin of piezoresistivity is investigated in strain impact on carrier energy levels. Therefore, the type of impurity and its concentration affect the piezoresistivity.

Another more practical coefficient used to describe the piezoresistivity is the gauge factor $GF$ which is the fractional change in resistance per unit strain.

$$GF = \frac{\Delta R/R}{\epsilon} = \frac{\Delta R/R}{\Delta L/L} = 1 + 2\nu + E\pi$$  \hspace{1cm} (2.2.19)

where $E$ represents the Young’s modulus. The term $1 + 2\nu$ is due to geometrical changes and $E\pi$ is related to material properties. The GF of several materials are
2.2 Displacement Sensing Schemes

![Diagram](image.png)

Figure 2.19: Longitudinal (A and C) versus transverse (B and D) piezoresistance [12].

Table 2.2.1: Gauge factors for selected materials [19].

<table>
<thead>
<tr>
<th>Material</th>
<th>GF</th>
</tr>
</thead>
<tbody>
<tr>
<td>Al</td>
<td>1.4</td>
</tr>
<tr>
<td>Cu</td>
<td>2.1</td>
</tr>
<tr>
<td>Ni</td>
<td>-12.62</td>
</tr>
<tr>
<td>Pt</td>
<td>2.60</td>
</tr>
<tr>
<td>Si(SC)</td>
<td>-102 to 135</td>
</tr>
<tr>
<td>Si(Poly)</td>
<td>-30 to 40</td>
</tr>
</tbody>
</table>

listed in Table 2.2.1 for comparison [19]. The small GF of the metals is explained by the fact that their resistance variation is only due to the geometrical changes. Larger GF of some metals such as Nickle is justified by their significant resistivity changes. The single-crystal silicon shows the largest GF, which makes it attractive for piezoresistivity sensing.

The silicon piezoresistivity has been utilised in atomic force microscope cantilevers [62], pressure sensors [63] and accelerometers [64]. A novel differential displacement sensing technique is proposed in Chapter 6, based on the silicon piezoresistivity, for
2.2.2.3 Magnetoresistive Transduction

An alternative discrete sensing technique is to use the magnetoresistive element, whose small size, high sensitivity and wide bandwidth make it a suitable option for nanopositioning apparatus. The giant magnetoresistive (GMR) sensors exploit the spintronic effects made by nanofabricated metallic multilayers. A basic schematic view of the GMR sensor is presented in Figure 2.20, where a thin material conductive layer is sandwiched by ferromagnetic layers. The conductivity of the middle layer can be affected by the ferromagnetic layers being magnetised. Without an external magnetic field, the spin of conduction electrons at the magnetised top and bottom films are in opposite directions, as shown in Figure 2.20(a). Hence, instead of entering the neighbouring layer they scatter off. This results in a short mean free path length for electrons, which translates to high electrical resistance. Applying a strong enough external magnetic field will cancel the existing antiferromagnetic coupling and gives identical spin to electrons in both layers, which is associated with a lower electrical resistance (See Figure 2.20(b)). The intensity of magnetic field is a function of distance. This has been exploited in GMR sensors for displacement sensing.

A hybrid solution is reported in [65], where the resonance of a MEMS cantilever is measured by a GMR sensor and a permanent magnet coupled to the cantilever. The sensitivity and noise floor are comparable with the optical measurement ($F_{res} = 262kHz$). However, the performance of GMR sensors is influenced by the flicker noise at low frequencies. A mechanical approach is taken in [66] to cancel the low frequency noise of GMR sensors. The magnetic flux gate is actuated by an electrostatically driven gate electrode at a frequency beyond the sensor $1/f$ knee fre-
Figure 2.20: The GMR sensor cross section schematic, a) no external magnetic field, b) external magnetic field dominates the magnetisation direction [13].
frequency. Consequently, the desired displacement signal modulates the high frequency oscillation of the permanent magnet. The fabrication of magnetic sensors is not fully integrated in the standard MEMS processes. As long as this thesis is mostly concentrated on the sensors that are integrated with MEMS devices in the same fabrication process, the magnetoresistive sensing is only briefly introduced here.

### 2.3 Readout Circuits

The performance of a sensing system is determined by its readout circuit as well as the device section. The circuit noise, linearity, bandwidth etc. can be as effective as the device itself. Typical readout circuits of sensing systems such as low noise amplifiers, auto zeroing and chopper amplifiers convert the capacitance, inductance or resistance of a sensor device to an electrical signal. However, indirect measurements provide some particular improvements such as accuracy, linearity or cost. The two widely used indirect approaches are the frequency detection and time of flight methods [40]. The frequency based measurement methods are introduced here and a new frequency based readout circuit is presented in Chapter 3 for the electrothermal displacement sensing.

*Ratiometric* measurement of resistors or capacitors in instrumentation applications is introduced to make the measurement independent of the resistance or the capacitance absolute values. The single-ended ratiometric resistive topography is shown in Figure 2.21(a) where:

\[
V_{\text{out}} = \frac{R_1}{R_1 + R_2} V = \frac{R_0 + \Delta R}{2R_0 + \Delta R} V \approx \frac{1}{2} V + \frac{\Delta R}{4R_0} V, \tag{2.3.1}
\]

The Wheatstone bridge circuit is a differential form of the ratiometric front-end
that removes the $\frac{1}{2} V$ large offset:

$$V_{out} = V_1 - V_2 = \frac{\Delta R}{2R_0 + \Delta R} V$$  \hspace{1cm} (2.3.2)

Increased sensitivity and linearity are also the benefits of differential sensing [19]. Consequently, a ratiometric front-end interface is employed to couple the sensor parameter variations to the following signal conditioning stage, which can be an ordinary amplifier or a frequency based circuit.

Sensor excitation methods with different input signals also impact the total system performance significantly [8, 34]. For example, constant current, voltage and power input affect the achievable signal to noise ratio in an electrothermal sensor. Direct current (dc) and alternative current (ac) input voltages are also used to shift the measurement operation to higher frequencies where the flicker noise level is lower. These are elaborated in Chapter 4 with new solutions for lower noise and higher sensitivity achievements.
2.3.1 Frequency Based Measurement

Frequency detection in sensing systems usually involves a capacitor \( C \) and an inductor \( L \) in a resonant oscillator, where output frequency is as following:

\[
\omega_n = \sqrt{\frac{1}{LC}}
\]  

(2.3.3)

This non-linear transfer function can be linearised by expansion as in:

\[
\omega_n = \sqrt{\frac{1}{L_0C_0(1 + \alpha)}} \approx \sqrt{\frac{1}{L_0C_0}} \left[ 1 - \frac{\alpha}{2L_0C_0} \right]
\]  

(2.3.4)

where \( \alpha \) is the fractional variation in the transducer property. Although it is possible to relate the oscillation frequency of an LC (resonant) oscillator to a resistive element, however, its sensitivity is not as high as capacitor and inductor. Therefore, alternative oscillators are utilised for the resistive element readout circuit. Relaxation oscillator is a low frequency and low noise oscillator which is widely used for capacitive and resistive sensors [67, 68]. A basic schematic of the relaxation oscillator is illustrated in Figure 2.22. Although different front-end configurations, such as Wheatstone bridge, have been involved following the same oscillation principle, the output frequency can be expressed as:

\[
f \propto \frac{1}{RC}
\]  

(2.3.5)

where \( R (R = R_1 = R_2 = R_3) \) and \( C \) can be any of the sensing elements.

Ring oscillators have also been used in the frequency-based readouts of capacitive and resistive sensors [69, 70, 71, 72]. In its basic form, a ring oscillator consists of an odd number of inverters cascaded in a feedback loop, as illustrated in Figure 2.23. The oscillation frequency of the resulting positive feedback is expressed as:

\[
f = \frac{1}{2\pi nRC}
\]  

(2.3.6)
where $n$ is the number of inverters and $RC$ is the time delay generated by each inverter output resistance and parasitic capacitance. The output frequency can be a function of a resistive or capacitive sensor variation if the sensor is connected to a node in the loop. The simple structure and high sensitivity of ring oscillators are rather attractive. Also their straightforward integration in microchips provides a viable signal processing mean due to the highly digital nature of ring oscillators. The high frequency output of the ring oscillator can be used for wireless data transmission. 

Ring and relaxation oscillator based sensor readout instances are listed in Table 2.3.1. They are used for both resistive and capacitive sensors. However, a fundamental difference is seen in the parameter variation range and the oscillator output frequency.
<table>
<thead>
<tr>
<th>Readout circuit</th>
<th>Sensor type</th>
<th>Parameter Variation</th>
<th>Freq. Variation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Ring Osc. [70]</td>
<td>Pressure (Capacitive)</td>
<td>2.9-3.5pF</td>
<td>1.7-1.9MHz</td>
</tr>
<tr>
<td>Relaxation Osc. [67]</td>
<td>Force (Capacitive)</td>
<td>200-500pF</td>
<td>40-90kHz</td>
</tr>
<tr>
<td>Relaxation Osc. [73]</td>
<td>Humidity (Resistive)</td>
<td>5-45kΩ</td>
<td>1.2kHz/6.8kHz</td>
</tr>
<tr>
<td>Ring Osc. [72]</td>
<td>CMOS Piezoresistive</td>
<td>NA</td>
<td>5.2-5.25MHz</td>
</tr>
</tbody>
</table>

The relaxation oscillators provide shorter frequency spectrum for a wider sensor parameter change. This translates to a higher sensitivity for the ring oscillator, whose output frequency variation range is wider for a smaller parameter variation.

A new ring oscillator based readout circuit is presented for an electrothermal sensor in Chapter 3. The half-bridge sensor interface is used to couple the resistive changes of the sensor to a ring oscillator power supply which can control the ring oscillator frequency output. In addition to the achieved high sensitivity, the linearity of the whole system, from the actuator input to the sensor output, is improved. However, the ring oscillator is the main source of the noise and deteriorates the sensor resolution despite applying the low noise design techniques.

### 2.3.2 Sensor Excitation Methods

Typically, sensing systems have an excitation input current or voltage signal whose response by the system is measured as the sensor output. In the case of resistive sensors constant current (CC), constant voltage (CV) and constant power (CP) excitation modes have been reported in the literature [8, 74, 75, 76]. In addition to circuit complexities, the resulting SNR and bandwidth of the sensor can be affected by these modes. A comparison between constant current and constant voltage modes is conducted in [8] for electrothermal displacement sensors. The circuit schematic shown in Figure 2.24(a) and (b) demonstrates the two readout circuit architectures used to bias the sensor with constant current and voltage, respectively. The sensitivity of the
measurement system is investigated in the two cases and current mode excitation is proved to have a higher sensitivity. To evaluate the resolution of the sensors two readout circuits, with identical noise properties, are proposed in Chapter 4. A detailed model based on this analysis is also provided to compare the two operation modes.

2.4 Conclusion

The most common sensor specifications, including linearity, noise, resolution and drift are introduced in this chapter. Then the existing displacement sensing tech-
niques available for high precision applications were investigated. The capacitive, and resistive sensors were discussed, with a special focus on electrothermal and piezoresistive sensors due to their smaller footprints, which is a superb advantage in the miniaturised nanopositioning applications.

Several integrated displacement sensors for nanopositioning purposes are listed in Table 2.4.1. The sensitivity comparison is not straightforward as each of them are reported in a different way. However, critical researches are conducted in [15] and [77] to make comparisons between capacitive versus electrothermal and piezoresistive versus electrothermal sensors, respectively. According to [15] capacitive sensors show a better noise performance at lower frequencies while at high frequencies the electrothermal sensor is a better candidate. In this work, the displacement of a stage is measured simultaneously by capacitive and electrothermal sensors. The capacitive sensor demonstrates a lower PSD at low frequencies while the electrothermal sensor has lower PSD at higher frequencies. According to [77] the electrothermal sensor has offered more than an order of magnitude higher resolution compared with the piezoresistive sensor. In addition the occupied area by capacitive sensors is obviously much larger than electrothermal and piezoresistive sensors. Therefore, the focus of this thesis in the following chapters is on the electrothermal and piezoresistive sensors.
Table 2.4.1: Integrated displacement sensors in MEMS nanopositioner (C, Et and Pz represent capacitive, electrothermal and piezoresistive sensors, respectively).

<table>
<thead>
<tr>
<th>Type</th>
<th>Resolution/Noise floor</th>
<th>Travel range</th>
<th>Sensitivity</th>
<th>BW</th>
<th>Size (Area)</th>
</tr>
</thead>
<tbody>
<tr>
<td>C [29]</td>
<td>0.3 nm</td>
<td>66 µm</td>
<td>0.333 fF/nm</td>
<td>NA</td>
<td>NA</td>
</tr>
<tr>
<td>C [51]</td>
<td>10 nm &amp; 25 nm</td>
<td>32 µm</td>
<td>0.5 µV/nm</td>
<td>NA</td>
<td>length ≈ 800 nm</td>
</tr>
<tr>
<td>C [78]</td>
<td>100 nm</td>
<td>3 µm</td>
<td>1 fF/nm</td>
<td>NA</td>
<td>200 × 850 µm²</td>
</tr>
<tr>
<td>C [15]</td>
<td>0.76 nm/√Hz</td>
<td>2.3 µm</td>
<td>0.137 V/V</td>
<td>NA</td>
<td>≈ 100 × 200 µm²</td>
</tr>
<tr>
<td>Pz [62]</td>
<td>10 A</td>
<td>NA</td>
<td>0.75 ppm/A°</td>
<td>100 kHz</td>
<td>100 × 5 µm²</td>
</tr>
<tr>
<td>Et [15]</td>
<td>0.14 nm/√Hz</td>
<td>2.3 µm</td>
<td>0.137 V/V</td>
<td>&lt;10 kHz</td>
<td>50 × 2 µm²</td>
</tr>
<tr>
<td>Et [77]</td>
<td>0.01 nm/√Hz</td>
<td>NA</td>
<td>10 V/nm</td>
<td>NA</td>
<td>50 × 20 µm²</td>
</tr>
<tr>
<td>Pz [77]</td>
<td>0.2 nm/√Hz</td>
<td>NA</td>
<td>1 V/nm</td>
<td>NA</td>
<td>50 × 20 µm²</td>
</tr>
</tbody>
</table>
3 A Frequency-based Readout Circuit for Resistive Sensors

The frequency modulation technique can be applied to MEMS transducers that require some form of resistive sensing, as stated in the previous chapter. A highly sensitive read out circuit is proposed in this chapter, which converts $10\Omega$ change of resistance in a $400\Omega$ electrothermal sensor to more than $200\text{kHz}$ frequency variation ($350\text{kHz}$–$550\text{kHz}$). The frequency variations are then converted to voltage values by means of a frequency demodulation. The frequency demodulation is achieved by two distinct methods: a phase locked loop and a frequency to voltage converter. In addition, the proposed technique achieves high linearity from the voltage applied to the actuator to the voltage measured at the sensor’s output, which can potentially eliminate the need for an additional linearisation if the sensor is used in a feedback loop. The proposed approach leads to high sensitivity in the MEMS electrothermal sensing since the method is not affected by amplitude variations that could arise from the readout circuit.

3.1 Introduction

The design of MEMS transducers is a multi-disciplinary area of research that spans fields as diverse as circuit design and device physics in electronics to mechanics and
Two interesting applications of MEMS to emerge in recent years are nanopositioners and microgrippers. Nanopositioners have found applications in scanning probe microscopy [79], atomic force microscopy [20] and ultra-high-density probe storage systems [80]. A typical nanopositioner consists of high precision sensors and actuators plus a feedback control loop. The displacement produced by the actuator is measured by a sensor, whose output is fed back to the actuator to increase positioning accuracy of the system. As illustrated in Figure 3.1, precise positioning is achieved through acquiring the location or displacement of the actuator by a sensor and its application via a feedback loop to the actuator input. Several sensing techniques for precise control of MEMS microactuators have been proposed, offering high sensing accuracy. For instance, in [16] a piezoresistive sensor is used to measure the displacement of an actuator. Among the different sensing and actuation schemes in MEMS, electro-static (capacitive) and electrothermal (resistive) methods have attracted much attention [35]. Capacitive sensing is widely used for displacement measurement in MEMS transducers. One of the issues with this method of sensing is the relatively large footprint of the resulting sensor. Electrothermal sensing has been proposed as an alternative solution [7]. These sensors can be

![Microactuator feedback loop block diagram.](image)

Figure 3.1: Microactuator feedback loop block diagram.
designed to be of a very small form factor, freeing up space that, otherwise, can be used for other purposes, e.g. actuation. As explained in the previous chapter the sensor performance depends on the readout circuitry designed for its output measurement as well as the efficiency of its mechanical structure. Regarding the readout circuitry, conventional amplifiers may not be the best option for use as a sensor front-end. Common low frequency signal measurement techniques include low noise amplifiers, auto-zeroing, and chopper modulation [81]. The latter has emerged as a popular sensing front-end for capacitive MEMS transducers since it avoids the 1/f noise associated with operational amplifiers by using an amplitude modulation technique. The principle of chopper modulation is to up-convert the signal to higher frequencies where amplifier flicker noise is less than the thermal noise floor. Then, the up-converted signal is amplified and finally down-converted to obtain the original low-frequency signal. The clock feed-through and jitter are the main drawbacks of this method. A chopper amplifier has been proposed for electrothermal sensing in [82]. However, only simulation results are presented and most of the practical issues are ignored.

A frequency modulation (FM) technique is introduced in this chapter to realise a sensor readout circuit for electrothermal sensors. With this method, the signal obtained from the sensor is measured in the frequency domain rather than the amplitude domain. The frequency modulated signals are less prone to amplitude noise. Typical distortions on the signal, arising from the electronic circuits, will not normally affect the performance of an FM signal as they are based on frequency, not amplitude. The proposed method features a linear input-output relationship, by combining the non-linear frequency modulation with the inherent non-linearity of the actuator. The remainder of this chapter continues as follows. In section 3.2, the operation principles of the microactuator used in the experiments are addressed.
The frequency based front end for electrothermal sensing, including the circuit and the system, is proposed in section 3.3. This is followed by implementation details, experimental results and conclusions.

3.2 Electrothermal Microactuators

Electrothermal displacement sensors and actuators are commonly fabricated from doped silicon and/or poly-silicon. Figure 3.2 shows a 1-Degree of Freedom (1-DoF) MEMS nanopositioner, which works based on the concepts of electrothermal actuation and electrothermal sensing [9]. An input voltage is applied to the actuator component of the device via ports 1 and 2. This results in a proportional current passing through

Figure 3.2: Thermal Sensor integrated to thermal actuator in 1-DoF nanopositioner.
the silicon beams of the electrothermal actuator. The actuator’s end point is attached to a stage (S), whose movements are to be controlled. The stage functions as a heat absorbing mass, which moves relative to a pair of resistive position sensors (RMEMS1, RMEMS2). The resistance of each sensor is proportional to its temperature. The heat generated by the resistors is transferred to the mass, which is at a much cooler temperature. The structure of the MEMS device is designed such that the movement of the heat absorber in either direction (up or down in Figure 3.2) has opposite effects on the heat transfer rate from the two sensors. Therefore, any displacement would cause a differential variation in the resistances $R_{MEMS1}$ and $R_{MEMS2}$. Doped silicon resistors suffer from flicker noise as well as thermal (Johnson) noise [14, 48, 83], as in:

$$V_j = \sqrt{4k_BTR} \quad [V/\sqrt{Hz}]$$

$$V_f = \sqrt{(KR^2(I^a/f^b))} \quad [V/\sqrt{Hz}]$$

where $V_j$, $V_f$ stand for thermal and flicker noise, respectively. $T$, $R$ and $k_B$ in Equation 3.2.1 represent the sensor temperature, the resistance value and Boltzmann constant. The flicker noise power is proportional to the applied current $I$, and inversely proportional to the applied signal frequency $f$ and cross-sectional area of the resistor $R$. $K$ is a constant for the device, $a$ and $b$ are constant values in Equation 3.2.2. Commonly, the flicker noise power overwhelms the thermal noise at low frequencies, as illustrated in Figure 3.3 [14]. In particular, sensing resistors should be heated up in order to reach adequate sensitivity. A significant current in the order of $mA$ must pass through the resistor to heat it up. Thus, the low frequency noise associated with the resistor becomes significant. The electrothermal resistance changes should be measured by the interface readout circuit. Most readout circuits, as proposed in the literature, utilise low noise instrumentation amplifiers [9]. In addition to the intrinsic flicker (i.e. $1/f$) and thermal noise of the electrothermal
sensors, the amplifier noise also adds to the output noise, and thus will degrade the 
SNR performance of the system. Consequently, a higher frequency measurement is 
likely to reduce the amplifier noise effect on the final result.

The relationship between the applied input voltage and the achieved displacement in 
a typical MEMS electrothermal actuator is non-linear. For example, the measurement 
results for the MEMS actuator in [15] illustrates a quadratic transfer function from 
the actuation input voltage to displacement. In the case of using the available readout 
circuitries which offer a linear transfer characteristic, the feedback network should 
compensate the non-linearity, which increases the system complexity. The readout 
technique proposed here leads to a linear input-output relationship in addition to 
the sensing at high frequencies to reduce the effect of noise sources.

### 3.3 A Frequency Modulation Sensing Technique

This section first describes the proposed approach for sensing the resistivity changes
corresponding to the electrothermal sensors in the system level. Then, the circuit design procedure, including the ring oscillator and front-end circuitry, are discussed.

3.3.1 Theory of the Frequency-based Measurement

Significant efforts to develop frequency based measurement methods for resistive sensing have been reported in the literature [71, 84, 85, 86]. The non-linear relationship between the input voltage and the output frequency is a common issue with previous works. An electrothermal sensor measurement with a high frequency output is reported in [86]. An oscillator based signal conditioning is also introduced in [71]. Neither of these approaches returns the information in voltage amplitude form, as needed in a feedback-controlled system. The time based measurement method, introduced in [85] and [84], that is designed for gas sensors, converts large resistance variations in the range of 1kΩ to 1GΩ to a measured signal. However, this method is not applicable to MEMS resistive sensors that typically undergo much smaller changes, e.g. less than 10Ω resistance change in a 400Ω sensor, as is the case in the systems considered here. The proposed frequency-based measurement in this thesis offers high sensitivity and avoids the noise contribution of the amplifiers in conventional readout circuits. Additionally, the circuit configuration is designed to provide a linear relationship between the actuator input and the sensor output voltages. The electrothermal resistive changes, which contain the displacement information associated with the stage (See Figure 3.2), demonstrate a non-linear behaviour, as shown in Figure 3.4(a). The front-end circuit translates the resistive changes to voltage values in a linear way (Figure 3.4(b)). Then, the voltage changes result in the frequency variations in a ring oscillator configuration, which is inherently non-linear (Figure 3.4(c)). The voltage-frequency characteristic of the ring oscillator is designed to be the inverse of the former non-linearity. The final output is either a dc, or a slowly varying voltage, which indicates the resistance of the sensor. This
voltage is obtained by demodulating the frequency modulated signal (Figure 3.4(d)). The measured non-linearity of the thermal actuator transfer function is depicted in Figure 3.4. Proposed method mitigates the non-linearity from actuation input to the sensor output.

Figure 3.4: Proposed method mitigates the non-linearity from actuation input to the sensor output.

Figure 3.5. In addition the displacement to resistive changes transfer function adds to this non-linearity in the same manner. This has been proved through thermodynamic and electrical analysis. The system’s functionality is depicted in Figure 3.6. The changes in the sensing resistor translate to frequency variations ($F_{IN}$) that could be demodulated to a voltage (Figure 3.6(a)). As shown, in more details, in Figure 3.6(b), electrothermal resistive changes modulate the output frequency of the oscillator VCO1 (voltage controlled oscillator). That is, the output frequency of VCO1 is modulated by changes in the resistance of the electrothermal sensor. The FM signal is then demodulated back to a DC voltage (i.e. sensing voltage).

The frequency variations can be detected using a phase locked loop (PLL) circuit,
Figure 3.5: The measured non-linearity of the thermal actuator transfer function.

Figure 3.6: Proposed frequency based system: a) system architecture b) frequency demodulation based on PLL.

which is a widely used frequency demodulation technique. In Figure 3.6(b) the output frequency of the reference oscillator (VCO1) is the PLL input frequency ($F_{IN}$). The phase-frequency detector (PFD) in the PLL compares the two inputs and
locks the VCO2’s output frequency to the PLL input frequency \( F_{IN} \). Consequently, the loop follows the VCO1’s frequency variations. Until the loop stays in the lock condition, very slight frequency variations appear as phase difference between PFD inputs. The phase difference is extracted by the PFD and transformed to a slowly varying (dc) voltage through a low pass filter (LPF) at the VCO2 input. The frequency-demodulated signal is obtained at the LPF output. PLL FM demodulators have superior linear performance in comparison with the other types of demodulators. While the LPF determines the bandwidth of the demodulated output, it can be utilised to control the low frequency noise. In telecommunication applications where PLLs are used as frequency demodulators, the signal bandwidth is often much wider than what is pursued in this work.

Another popular demodulation approach circuit for frequency modulated signals is frequency to voltage conversion. As illustrated in Figure 3.7, the input frequency \( F_{IN} \) triggers the one-shot circuit which subsequently steers a constant current source \( \alpha \) to either the output or the summing node of the integrator. This produces a dc voltage proportional to the ON time of the one-shot circuit and the input frequency that is the duty cycle of the current pulse applied to the integrator input. Therefore, the current signal at the integrator input will be a pulse wave with a frequency equivalent to \( F_{IN} \) and on-time equal to the one-shot time constant \( t_{OS} \). The integrated output voltage is as follows:

\[
V_{OUT} = t_{OS} \alpha R_{INT} F_{IN}
\]  

A comparison of the two aforementioned frequency demodulation schemes is provided later, in this section.
Figure 3.7: Alternative frequency demodulation approach: a frequency to voltage converter (FVC).

3.3.2 The Ring Oscillator

A ring oscillator circuit is utilised to translate the resistive changes to frequency variations (VCO1 in Figure 3.6(a)). A ring oscillator can be implemented with an odd number of single ended inverters connected in a positive feedback configuration, as illustrated in Figure 3.8. The oscillation frequency of the system is determined as follows [87]:

$$f_{osc} = \frac{1}{2\pi n \tau}, \quad \tau = RC = \frac{C}{g_m}$$

$$f_{osc} = \frac{g_m}{2\pi n C}$$

where, $n$ is the number of stages, and $\tau$ is the delay generated by each inverter stage. $R$ represents the inverse of the transistor’s transconductance ($g_m$) and $C$ represents the node’s capacitance, both of which are varying by the supply voltage. Oscillation frequency can be tuned by manipulating the number of stages, loading, drive strength and supply voltage of the inverters. Furthermore, a proper combination
of the mentioned methods can result in a wider tuning range [88]. However, changing the supply voltage is the most convenient and the only practical approach available to us. It also leads to a conveniently wide tuning range. The only disadvantage associated with this method is the presence of a higher level of jitter at low frequencies. Supply induced jitter reduction techniques are discussed in the next section.

External large capacitors are added to the inverter input/output nodes to cancel out the voltage variable parasitic capacitances associated with the inverter transistors. Hence, the only viable option to control the oscillation frequency of the system is to manipulate the $g_m$ with the supply voltage. A transistor’s transconductance is not a linear function of the supplied voltage. This leads to a non-linear relationship between the supplied voltage and the oscillation frequency. The non-linearity of voltage to frequency transfer characteristic of a VCO with differential inverters is explained in [89]. The input control voltage drives the inverter transistors in the cut-off, saturation and linear regions. The analytical relationships for all of the operation regions are derived. An inverter based ring oscillator has been simulated here to illustrate the non-linearity of voltage to frequency transfer characteristic of a VCO. This is established by simulating the operation of a 3-stage inverter-based

![Figure 3.8: Ring oscillator using digital inverter chain.](image)
ring oscillator, consisting of TSMC 0.18\textmu m CMOS transistors. Simulations were performed on a Spice simulator and the results are illustrated in Figure 3.9.

![Figure 3.9: Simulation of ring oscillator frequency versus supply voltage transfer characteristics.](image)

The linearity of the voltage to frequency (V/F) conversion of ring oscillators is a matter of concern in conventional applications, and several linearisation techniques have been proposed in the literature, e.g. see [90, 91]. In the following it is proven that the inherent non-linearity can be utilised to mitigate the non-linear response of electrothermal actuators. As mentioned previously, displacement of an electrothermally actuated nanopositioner is a non-linear function of the actuating voltage [9, 92]. Such quadratic non-linearities are commonly observed in MEMS electrothermal actuators of various types, as well as in MEMS capacitive actuators.
This non-linearity is very similar to the inverse of the V/F characteristic of a VCO. Thus, by combining the two, one may expect to obtain a relatively linear relationship between the actuating voltage and the measured displacement. This is achievable by selecting a proper value for the node capacitance of the ring oscillator. As illustrated in Figure 3.9, the different non-linear relationships are achievable by using different parasitic capacitors \( C_p \) at the inverters output node. The curves in Figure 3.9 are obtained by simulating the same ring oscillator with different node capacitances. The mathematical relationship is fitted to the curves in MATLAB. For \( C_p = 1pF \), the added parasitic is in the order of parasitic capacitances of transistors. Therefore, the voltage (supply voltage in this case) controls the oscillation frequency through both transconductance and internal capacitances of the transistors. However, for \( C_p = 7pF, 15pF \) the added capacitance dominates the node capacitance value so the denominator of Equation 3.3.3 is fixed with respect to control voltage. Accordingly, large capacitances are used in the experiments to control the achievable non-linearity.

Electronic oscillators used in telecommunication systems generally suffer from phase noise. Phase noise in the frequency domain translates to jitter in the time domain. The phase noise phenomenon in ring oscillators has been discussed in [93]. The main contributor to the phase noise close to the centre frequency is the flicker noise of the inverters’ tail current that up-converts into the close vicinity of the oscillation frequency. Insertion of a large capacitor between the supplying tail current and the ground, as illustrated in Figure 3.10, can reduce this noise [94]. The large capacitance, when combined with the output impedance of the current source acts like a low-pass filter and suppresses the flicker noise at the baseband, avoiding its up-conversion. The available simple digital gates, at the inverter chain, are utilised to explain the operation of ring oscillators. However, digital gate inverters degrade phase noise performance of ring oscillators in comparison with the differential and
current steering inverters [95]. Further improvement can be achieved by replacing voltage supplies with current sources. This has been shown to lead to 4 to 6dB phase noise reduction at the same oscillation frequency with similar design parameters [96].

![Figure 3.10: Circuit designed for VCO1 in the proposed system.](image)

### 3.3.3 Front-End Circuit Design

Applying the aforementioned phase noise reduction techniques, the circuit depicted in Figure 3.10 is designed to replace the reference oscillator VCO1 in Figure 3.6(b). In order to convey the resistive changes to the oscillator control voltage, an interface circuit including transistors $M_1$, $M_2$ and reference resistors is designed, as shown in Figure 3.10. Since $M_1$ and $M_2$ are operating in saturation the input-output transfer characteristic is linear. This circuit transforms the variations in the electrothermal...
sensor’s resistance to frequency variations. As a result, the oscillation frequency of the ring oscillator is a function of the MEMS resistor and, consequently, the actuator displacement. The reference resistor \( R_{ref} \) is a discrete (lumped) resistor selected to be almost the same value as the MEMS electrothermal sensor resistance. This is clarified through the following explanation:

Assuming \( R_{MEMS} = x \) and \( R_{ref} = y \), the \( M_1 \) gate voltage is as follows:

\[
V_g(x, y) = V_{dd} \frac{y}{x + y}
\] (3.3.4)

The sensitivity of \( V_g \) to \( x \) is therefore defined as:

\[
S^V_x = \frac{\partial V_g}{\partial x} = V_{dd} \frac{-y}{(x + y)^2}
\] (3.3.5)

In order to maximise the sensitivity of the circuit to MEMS electrothermal resistor changes, the maximum of \( S_x \) with respect to \( y \) should be found. A differential equation is obtained by taking the partial derivative of Equation 3.3.5 with respect to \( y \),

\[
\frac{\partial S^V_x}{\partial y} = \frac{\partial^2 V_g}{\partial x \partial y} = V_{dd} \frac{y - x}{(x + y)^3}
\] (3.3.6)

\[
\frac{\partial S^V_x}{\partial y} = 0 \rightarrow x = y
\] (3.3.7)

From Equation 3.3.7 it is inferred that the maximum sensitivity of the circuit to \( R_{MEMS} \) variations will be achieved if \( R_{ref} = R_{MEMS} \). Fortunately, this coincides with the optimum bias point of \( M_1 \) to operate in saturation mode. \( R_{ref} \) is composed of standard 220Ω lumped resistors in series with 100Ω potentiometers to

The power supply induced jitter requires attention since the noise from the MEMS devices reaches the supply line of the ring oscillator by going through \( M_1 \) and \( M_2 \). This noise, as indicated earlier, has a low pass characteristic. According to [95], a
digital gate has a low-pass transfer function from the power supply to output of
the ring oscillator, while the differential pair and the current steering gates have
a high-pass nature. Obviously, better noise performance can be expected for the
system if the high-pass transfer functions of those inverters filter out the supply
induced noise.

Here, the oscillator was realised from off-the-shelf digital gates. This is expected
to affect the overall noise performance of the proposed scheme since they pass the
supply noise, which is affected by the noise of MEMS devices. Thus, in order to
suppress the noise over the lowest possible frequency ranges, a large noise cancelling
capacity (C in Figure 3.10) is connected to the common drain of \( M_1 \) and \( M_2 \), both
of which are operating as tail current sources. Each of these techniques significantly
improves the implemented circuit performance in terms of phase noise and jitter.

### 3.3.4 Experiments and Discussion

The sensing technique explained in section 3.3 has been applied to the MEMS
device reported in [9]. The device is utilised as an electrothermal displacement
sensor. Electrothermal position sensor resistors (\( R_{MEMS1} \), \( R_{MEMS2} \) in Figure 3.2,
Figure 3.10) are connected in series with the reference resistors of the same values to
a 5v dc supply (\( V_{dd} \) in Figure 3.10). The actuation mechanism of this device is also
electrothermal. By applying the variable actuation voltage (\( 0-5V \)) to ports 1 and
2 in Figure 3.2, the stage (which functions as the heat absorber for electrothermal
sensors) moves relative to the hot sensor resistors. Since the stage is cooler than the
sensors, it absorbs the heat and changes the sensor resistance.

The sensing circuit is implemented using off-the-shelf integrated circuits (ICs) as
follows. \( M_1 \) and \( M_2 \) in Figure 3.10 are implemented using CD4007 [97], which
contains three NMOS and three PMOS transistors. In order to adjust the quiescent
points of \( M_1 \) and \( M_2 \) (as it was impossible to change the aspect ratio of the transistors) two NMOS and three PMOS transistors were connected in parallel. The ring oscillator is realised from five 74HC04 inverters [98]. First a PLL based demodulation technique is used to recover the sensing signals. A phase/frequency detector (PFD) and VCO2 in Figure 3.6(b) are integrated in 74HC4046, which is a PLL IC [99]. The low pass filter is externally connected to the PLL.

As illustrated in Figure 3.11, the experimental results show that VCO1 and VCO2 track each other while the PLL is in phase lock condition. The demodulated output is represented with Ch1. The oscillation frequency of VCO1 increases from 390kHz to 550kHz for the actuation input voltage varying from 1V to 6V. The demodulated dc output varies almost linearly from 3.4V to 4.8V for the same actuation voltages as shown in Figure 3.11. Therefore, the output voltage which is to be used as a feedback signal is a linear function of input actuation voltage. This is an alternative to the previous work, in which a lookup table was designed in the feedback path to mitigate the non-linearity of forward path [9].

As indicated in section 3.3, the linearity of the transfer function from the input actuation voltage to the last sensor output in microactuators is highly desirable. A ring oscillator generates a non-linear transfer function from the input voltage to the output frequency. By placing the ring oscillator after the front-end of the readout circuitry an almost linear relationship between the signal applied to the actuator (which is a non-linear device) and the sensed output voltage is achieved, as shown in Figure 3.12. Figure 3.13 is also obtained by an Agilent spectrum analyser (N9020A) tracing the output spectrum of the oscillator while the actuation voltage was being increased.

The power spectral density of the measured low frequency signal is shown in Figure 3.14. The data was acquired using an HP 35670A dynamic signal analyser,
which can repeat the measurement up to 32 times in the averaging process. The low frequency noise is shown at various offset frequencies. In particular, -72dBv rms output noise power is achieved at 2Hz offset frequency. In order to demonstrate the MEMS resistor’s noise contribution it was replaced with a lumped resistor, of the same resistance. About 10dB noise power difference can be observed in the 2-20Hz frequency range. This implies that, in addition to the flicker noise discussed in section 3.2, the sensing system is prone to thermal noise arising from the heated MEMS resistors. During these experiments the lumped resistor was kept at the room temperature.

The most common approach to frequency demodulation involves using a PLL. However, in order to understand the overall frequency modulation technique clearly, such as the relationship between the actuation voltage and the demodulated output voltage and the noise performance, a frequency to voltage converter (FVC) circuit
was also implemented, as shown in Figure 3.7 [100]. The ring oscillator output VCO1 was applied to the frequency input of the FVC. The output spectrum is illustrated in Figure 3.15. In order to determine the contributions of VCO1 and the front-end to the resulting low frequency noise, a signal generator was directly applied to the FVC. The result is about 10 dB lower noise power at 1Hz. The difference is partially due to the MEMS device and partially due to the oscillator circuit. The former could be improved by some modifications in the mechanical design of the device, while the latter requires an improvement of the utilised circuit design techniques and components.

Since the two demodulators are implemented with different off-the-shelf devices, a fair comparison of their performances may be rather difficult. However, the second approach requires fewer components and involves a simpler architecture to realise the requisite demodulation circuit. Thus, it has the potential to result in less noise at
the output. Furthermore, in the first approach, the input signal bandwidth follows that of the PLL’s, which is directly proportional to its lock range. Moreover, due to the trade-off between the PLL bandwidth and the phase noise passed through the PLL, a narrower bandwidth is desired. This constraint does not apply to the second approach. Instead, the settling accuracy of the circuit shown in Figure 3.7 is determined by the integrator’s R and C. Besides, the dc output voltage generated by F/V converter depends on the power supply and could be wide enough. A comparison of the proposed system with similar circuits proposed for resistive sensing in the literature is summarised in Table 3.3.1. They differ in the resistance variation range and its temperature. In the application pursued here, the resistors are heated up to increase the sensitivity of device versus temperature. On the other hand the output noise level increases as a heating consequence. Also the non-linear circuit designed

**Figure 3.13:** Ring oscillator frequency versus actuation voltage.
in this work combines with the non-linearity of the actuator and produces a linear input-output relationship. However the interface circuits designed in [84, 94, 101] are linear readout front-ends which are not suitable for this application.

**Figure 3.14:** Noise power spectrum of demodulated output.

**Table 3.3.1:** Comparison between resistivity sensors.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>[94]</th>
<th>[84]</th>
<th>[101]</th>
<th>This work</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resistor variation</td>
<td>Not reported</td>
<td>1MΩ-10GΩ</td>
<td>4kΩ Active</td>
<td>400-410Ω</td>
</tr>
<tr>
<td>Sensitivity</td>
<td>1000V/V</td>
<td>Not reported</td>
<td>1V for ∆R/R = %1</td>
<td>2V for ∆R/R = %2.5</td>
</tr>
<tr>
<td>Linearity</td>
<td>Linear</td>
<td>Linear</td>
<td>Linear</td>
<td>non-linear</td>
</tr>
<tr>
<td>Implementation</td>
<td>Simulation</td>
<td>Off-the-shelf</td>
<td>0.35μm CMOS</td>
<td>Off-the-shelf</td>
</tr>
<tr>
<td>Application</td>
<td>Thermopile</td>
<td>Gas Sensor</td>
<td>MEMS Strain gauge</td>
<td>MEMS Disp. Sensor</td>
</tr>
<tr>
<td>Circuit Architecture</td>
<td>Chopper Amplifier</td>
<td>Relaxation Osc.</td>
<td>Wheatstone bridge</td>
<td>Frequency Modulation</td>
</tr>
</tbody>
</table>
It should be noted that the test setups are operated in the room conditions (temperature, pressure and etc.) as it should be in practice. The dependency of the sensor system on ambient conditions such as temperature can be due to either the MEMS device or readout circuit parts. The ambient temperature cannot have considerable influence on the sensor performance because:

1. The differential structure desensitizes the sensor to ambient variations such as temperature.

2. The operating temperature of the sensor is around 400°C - 600°C, which is much higher than the room temperature. Therefore, changes in room temperature are negligible.
3.4 Conclusion

This chapter presents the design, analysis and implementation of a highly sensitive sensor readout circuit. The readout circuit uses a frequency modulation and demodulation technique to sense the changes of a MEMS-based electrothermal sensor’s resistance. A linear relationship between the input actuation voltage and the sensor output signal has been achieved, while maintaining the same noise performance as other high sensing measurement techniques. After actuation, the variation in the sensing element is frequency modulated using a ring oscillator, oscillating in the frequency range of 350–550kHz. Thus, the sensing is done at a much higher frequency than the low frequency ranges where there are noise sources. A practical demodulation circuit is implemented to recover the modulated sensing signal. Experimental results demonstrate about -72dBv rms output noise floor at the 2Hz offset frequency. Noise contribution due to the doped silicon is also characterised and discussed. Furthermore, the overall transfer characteristic from the input actuation voltage to the sensor output is almost linear.
4 Resolution Enhancement

Techniques for MEMS Electrothermal Sensors

The resolution of a microactuator is highly dependent on the noise and the sensitivity of its sensor, as stated earlier in Chapter 2. The electrothermal displacement sensor noise is strongly dominated by the heater noise profile, particularly the flicker and thermal noise inherent to the doped silicon. In this chapter, two techniques have been proposed to tackle the flicker noise as well as the thermal noise. First, it is shown that the flicker noise in a MEMS electrothermal displacement sensor can be reduced by driving the silicon heaters with a high frequency voltage. The proposed technique has been applied to a MEMS electrothermal sensor fabricated in the standard silicon on insulator (SOI) process. Experimental results demonstrate an 8dB improvement in noise level compared to the conventional measurement technique. The achieved noise floor is less than -100dBVrms around the 20Hz measured signal.

In the second approach a multiple sensor displacement measurement system is designed to alleviate the thermal noise. The sensitivity of an electrothermal displacement sensor increases with its temperature, whereas a higher temperature range leads to higher thermal noise level, which imposes a trade-off on the sensor’s achievable
resolution. The proposed technique, which is implemented on a 1-Degree-of-Freedom (1-DoF) SOI MEMS nanopositioner, mitigates this trade-off. By combining three identical sensors a 4dB improvement in SNR is achieved, which is in close agreement with the averaging theory. Experiments show that the displacement resolution is improved from \(0.3 \text{nm/} \sqrt{\text{Hz}}\) to \(0.15 \text{nm/} \sqrt{\text{Hz}}\) in the prototype nanopositioner.

In addition, a critical analysis is conducted in this chapter to compare the sensor excitation methods. The sensor resolution is dependent on the excitation method, such as constant current (CC) source or constant voltage (CV) source. The CV mode is more commonly used. However, there have been reports that the CC excitation mode may lead to a larger measured signal, and thus it may be a better choice than the CV mode [8]. An analytic comparison of the two methods is proposed, and it is shown that from a signal-to-noise ratio (SNR) point of view, benefits of operating a sensor in CC mode are only marginal. The analytical investigation is supported by experiments performed on sensors integrated in a SOI-MEMS nanopositioner with low noise read out circuits.

### 4.1 Electrothermal Sensor Noise in a Microactuator

A nanopositioner features a microactuator operating in closed loop, where a feedback controller is used to improve the system dynamics, as discussed in Chapter 1. Such a feedback control loop depends on the availability of a displacement sensor that supplies the position information, as illustrated in Figure 4.1. The displacement sensor specifications have a significant influence on the system performance. In particular, sensor noise adds to the reference signal in the feedback loop after being measured by the readout circuit. Therefore, the displacement precision is highly dependent on the sensor resolution.
The operating principle of electrothermal sensors can be explained using Figure 4.1. Applying electrical power to doped silicon resistors (sensors) increases their temperature. The heat conduction through a thin air gap between the stage and sensors is proportional to the overlapped area. The stage movements change the area overlapped and consequently the sensor’s temperature. The resistance of sensors varies as a function of temperature. Hence, the stage displacement translates to resistance variations. Sensors are also designed to operate in a differential configuration in order to improve the transducer linearity and reject the common mode signals. However, resolution of the electrothermal sensors is significantly limited by the noise associated with their silicon heaters [15].

The noise in MEMS devices can be of mechanical or electrical origins [44]. Mechanical noise (Brownian noise) is more crucial in surface micromachined devices. In bulk micromachined devices the electrical noise is dominant [19]. The electrical noise in MEMS consists of the thermal noise (also known as Johnson noise) and flicker noise. Johnson noise is the result of thermal agitation of charge carriers. The flicker noise is believed to be a result of fluctuations in conductivity in a semiconductor device.
The power spectral density (PSD) of the thermal noise is described as:

\[ V_j = \sqrt{4k_BTR} \] (4.1.1)

where \( k_B \) is the Boltzmann constant, \( R \) is the resistance and \( T \) is the temperature of the device. Power spectral density of flicker noise is known to be:

\[ V_f = \sqrt{(KR^2(I^a/f^b))} \] (4.1.2)

where \( K \) is a constant for the device, \( I \) is the current, \( f \) is frequency and \( a \) and \( b \) are constants [14].

### 4.2 Flicker Noise Reduction Technique

The concept of electrothermal sensing is discussed in [8], [15], [34] and [7]. The 1-DoF nanopositioner illustrated in Figure 4.2 is fabricated using the SOI MUMPs standard technology [15]. The objective is to measure the displacement of the moving stage that is actuated electrostatically by comb drives. \( R_1 \) and \( R_2 \) are two sets of doped silicon resistors whose resistance is a function of temperature. They are heated up by passing a dc current through them. The moving stage is acting as a heat sink and its movements change the sensor temperature and, thereby, its resistance. A readout circuitry measures the resistance changes and maps them to displacements. The resolution of this system is drastically influenced by the noise associated with the two sensing resistors. In this section an excitation method for reducing the flicker noise component of the sensor noise is proposed. The method is implemented and verified by experiments on the 1-DoF nanopositioner in Figure 4.2.

Conventional low noise readout techniques for low frequency measurements, including
chopper modulation and auto-zeroing, are designed to cope with the amplifier imperfections [81], [82]. These techniques up-convert the signal to higher frequencies, so that the amplifier is operated over its low noise range (away from low frequencies where 1/f noise is dominant). The signal is then down-converted back to the original

Figure 4.2: Differential electrothermal displacement sensor in a 1-DoF electrostatic microactuator, implemented in standard SOI MUMPS technology [15].
frequency. As a consequence, the noise originating from the electronics is reduced, however the noise arising from sensors will persist.

It can be inferred from Equation 4.1.1 and Equation 4.1.2 that the power spectrum of the thermal noise is white and that the flicker noise reduces with increasing frequency of the applied voltage. Electrothermal sensors investigated in this work are doped silicon monocrystalline devices. It is known that they suffer from both flicker noise and thermal noise \[102\], \[14\] and \[48\]. Nanopositioners are often operated at low frequencies. Hence, in a MEMS nanopositioner that uses electrothermal sensing, flicker noise is dominant.

### 4.2.1 Proposed High Frequency Excitation Technique

The temperature of resistors rises when conducting the electrical power \[34\]. The power is proportional to the RMS value of the applied voltage. To the author’s best knowledge, in all previous research reported in the literature a dc voltage is used to heat up the electrothermal sensors \[8\], \[15\], \[7\], and \[102\]. An ac (alternative current) voltage source can achieve the same temperature provided that its root mean square (RMS) value is identical to the applied dc voltage. As stated in (Equation 4.1.2), the flicker noise of doped silicon is inversely proportional to the frequency \[14\]. As a result, the noise power generated by the doped silicon, due to the high frequency input (i.e. frequency of interest), is lower. Hence, a lower noise floor can be expected by increasing the frequency of the heating voltage. Thus, a high frequency voltage source is used here to heat up the sensors. The noise power depends on the frequency of the applied electrical signal. Its PSD is higher in low frequencies where flicker noise dominates, though decreases at higher frequencies where it meets the thermal noise level. The intercept frequency at which these two noise spectrums meet is a selection criterion for the ac excitation signal frequency. The supply frequency can be
chosen at any point higher than the intercept point. However, the instrumentation amplifier of the readout circuit has a limited bandwidth, which sets an upper limit on the supply voltage frequency.

The proposed readout technique is illustrated in Figure 4.3(a), which converts the resistance \( R_{MEMS} \) variation to the voltage output \( V_{\text{out}} \). The high frequency voltage source \( V_h \) is used to heat up the resistors. The voltage \( V_a \) actuates the stage. \( V_a \) is typically a low frequency voltage source. The stage displacement changes the temperature of the resistors via thermal conduction through a thin air gap. As illustrated in Figure 4.3(b), the MEMS sensor model consists of a constant resistor in series with a signal source, \( V_i \), which represents the resistive variations due to the stage displacement. \( V_i \) has the same frequency as \( V_a \), but a smaller amplitude.

The signal at the transimpedance (TI) amplifier input in Figure 4.3, is a high frequency and high amplitude voltage. This voltage is added by \( V_i \), which is a low frequency and low amplitude signal. The resistance variations are differential, while the heating voltage is common for the differential amplifier inputs. Therefore, the output voltages of the TI amplifiers are as follows:

\[
V_m = A(V_i \pm V_h)
\]  

(4.2.1)

where \( A \) is the gain of the TI amplifier. The differential amplifier, also a gain stage, amplifies the resistance variation as a desired signal and attenuates the heating signal with regard to its common mode rejection ratio (CMRR). The simulation results of the model are illustrated in Figure 4.4. \( V_m \) is the output of the TI amplifier. The high frequency high amplitude component of this signal \( V_h \) is attenuated by the differential amplifier, while the low frequency and low amplitude component \( V_i \) is amplified at the output \( V_{\text{out}} \).
Figure 4.3: Proposed high frequency excitation technique for flicker noise reduction, a) circuit schematic, b) MEMS resistive sensor model.

4.2.2 Experiments

The proposed method was tested on the 1-DoF nanopositioner shown in Figure 4.2. A 12V (peak-to-peak), 5kHz sinusoidal voltage ($V_h$) was applied to the sensor resistors (heaters). The resistors were heated up to the same level as a 4.3V dc voltage. A 20Hz actuation voltage ($V_a$) was applied to the electrostatic actuator to move the stage. The $V_{out}$ signal in Figure 4.3(a) and (b) is a measure of stage displacement. The differential amplifier, which is the main gain stage, is a low noise instrumentation amplifier (INA128) which offers a gain of 100 at 200kHz bandwidth. Power spectral density of the output signal is plotted in Figure 4.5. $V_a$ is the actuator input signal, which is supplied to the comb drives. $V_{out,AC}$ is the sensor output attained by applying $V_h$ to the heaters. It should be noted that the dominant noise contribution in the
proposed technique is determined by the high frequency signal source attributes. For example, the -100dBVrms/$\sqrt{Hz}$ is the minimum noise floor attained by an Agilent/HP210A signal generator, as illustrated in Figure 4.5.

In another experiment, with the same setup, a 4.3V dc voltage is applied to the heaters and the sensor output is recorded as $V_{outDC}$ in Figure 4.5. This experiment shows that the noise floor around the 20Hz signal is about 8dB less than what is measured with the DC heating signal. In terms of noise floor, it is improved from 0.2nm/$\sqrt{Hz}$ to about 0.1nm/$\sqrt{Hz}$. The 50Hz noise and the 20Hz harmonics can be removed by applying a low-pass filter to the output signal. The PSDs were recorded by a HP35670A spectrum analyser.

The bandwidth of the sensor has been determined by measuring its step response. This is found to be 10kHz. The resonance frequency of the stage is 1.6kHz. The heating signal’s frequency (5kHz) is chosen so that it is above the bandwidth of the mechanical system, but low enough to avoid excessive phase noise due to the signal
generator which increases with the drive frequency.

One further step is required to extract the low-frequency content of the sensor output, which is proportional to the stage displacement. As stated earlier, the differential amplifier output signal consists of the small variations of the low-frequency displacement signal added to the peak of a high-frequency large heating signal. Therefore a peak detector circuit, as illustrated in Figure 4.6, is designed to remove the high frequency signal and extract the displacement signal. A Wheatstone bridge is used here as the sensor interface instead of the transimpedance amplifiers. A thorough comparison of the aforementioned techniques is conducted through circuit analysis in Chapter 5. $R_{MEMS1}$, $R_{MEMS2}$ and $R_F$ in Figure 4.6(a) represent the MEMS sensor resistances and the reference resistor that form the Wheatstone bridge. As long as $V_h$ is a common mode signal for differential amplifier, $V_{out}$ is expected to
Figure 4.6: The peak detector circuit designed for extraction of the displacement signal.
be the $V_i$ amplified. However, the differential amplifier has a limited CMRR and $V_h$ is a large signal. Therefore the differential amplifier does not reject the common mode signal completely and the output of differential amplifier is still dominated by $V_h$, i.e., the amplified differential signal is added on the peak of the attenuated $V_h$. Thus, a peak detector circuit is designed, as illustrated in Figure 4.6(b) to extract the low frequency content of the $V_{out}$. $V_{PD}$ represents the final output, which is proportional to the stage displacement. The high precision peak detector works only for negative cycle of its input signal. Since the heating signal is symmetric at the positive and negative peaks, one cycle peak detection is enough to extract the displacement data.

The high frequency excitation and peak detection method was tested on the 1-DOF nanopositioner by applying a 9V (peak-to-peak), 5KHz sinusoidal voltage ($V_h$) to the sensor resistors (heaters). The resistors were heated up to the same level as a 3.2V dc voltage. A 20Hz actuation voltage ($V_a$) was applied to the electrostatic actuator to move the stage. Power spectral density of the output signal is plotted in Figure 4.7. $V_{acpd}$ is the sensor peak detector output attained by applying $V_h$ to the heaters. In another experiment, with the same setup, a 3.2V dc voltage is applied to the heaters and the sensor differential amplifier output is recorded as $V_{dc}$ in Figure 4.7. Similar results is attained in terms of noise floor reduction. Displacement noise floor improvement from $0.2nm/\sqrt{Hz}$ to about $0.1nm/\sqrt{Hz}$ is achieved here as well. As long as the sensitivity is equalized for both circuits and measurements are carried out over the same frequency bandwidth, equal peak values of the 20Hz signal directly translates to two times better displacement resolution for high frequency excitation. The flicker noise due to doped silicon sensor can be reduced down to its thermal noise floor. The flicker noise in Figure 4.7 is due to the high frequency heating signal close-in phase noise (close to the centre frequency), which is down-converted to low
Figure 4.7: Signal power spectrum at the peak detector output.
frequencies and the operation and instrumentation amplifiers flicker noise.

4.3 Thermal Noise Mitigation

As stated earlier, MEMS offer a high-speed and low-cost solution compared to existing mesoscale approaches. In addition, a smaller footprint and higher sensitivity make the electrothermal sensing a viable method for displacement measurement. The sensitivity of an electrothermal sensor increases proportional to its temperature, according to the models proposed in [34]. However, the thermal noise floor also increases with temperature, which forces the user to strike a compromise between the thermal noise and sensitivity.

Thermal noise reduction by averaging repetitive measurements, known as serial averaging, is routinely used in instrumentation by processing the recorded data, however, real-time parallel averaging has been reported only in a few cases [103]. A general purpose averaging system was first patented in [104], where a digital circuit is utilised to examine the polarity of multiple analogue inputs and inputs with similar polarity are amplified by a variable gain amplifier. In [105], sixteen piezoresistive elements are combined to reduce the thermal noise of a tactile sensor but no quantitative improved figure was reported. Furthermore, parallel dummy memory cells were combined in high density memories to recover degraded signals and generate accurate reference signals in [106]. Due to the Monte Carlo simulation and measurements, it was shown that parallel averaging has greater stochastic results in noise reduction than the serial approach. A new architecture for MEMS electrothermal displacement sensing is proposed based on the averaging theory, in this section. The scanning electron microscope (SEM) image of a prototype nanopositioner implemented in the standard SOI technology for this purpose is
4.3 Thermal Noise Mitigation

![Diagram of noise sources](image)

**Figure 4.8:** Uncorrelated noise sources \( (n_1, \ldots, n_m) \) added to the desired displacement signal \( (d) \).

shown in Figure 4.9. A description of the sensor noise phenomena and theoretical expectations of the proposed technique are followed by the experimental details and measurement results.

### 4.3.1 Proposed Multi-Sensor Architecture

The electrical noise of a MEMS resistor made of doped silicon consists of thermal noise and flicker noise. A high frequency excitation method to mitigate the flicker noise in electrothermal displacement sensors is described in section 4.2. The thermal (Johnson) noise is the result of the thermal agitation of charge carriers, and its white power spectrum is described in Equation 4.1.1. The resolution of an electrothermal displacement sensor at higher frequencies is limited by its thermal noise. In particular, it is a crucial factor in high speed nanopositioning. A trade-off between the thermal noise and the sensitivity sets the resolution limit, i.e. to achieve a higher displacement resolution it is essential to improve the SNR of the sensor.

Signal averaging provides a means of estimating the shape of a repetitive response buried in non-coherent interference. The SNR can be improved by using the redundant
**Figure 4.9:** Proposed multiple-sensor displacement measurement system on a 1-DoF SOI MEMS nanopositioner (one electrothermal sensor magnified at top).
information inherent in multiple measurements [103]. The averaging approach to
sensor white noise reduction is depicted in Figure 4.8, where the desired displacement
signal, sensor noise and output voltage are represented by $d$, $n$ and $v$, respectively,
and $m$ is the number of measurements. Let $\sigma$ be the rms value of the random variable
$n$. Then the total noise ($N$) and signal ($S$) power at the output are:

$$N = \sigma \sqrt{m}$$  \hspace{1cm} (4.3.1)  

$$S = md$$  \hspace{1cm} (4.3.2)  

Thus, the resulting SNR is:

$$SNR_{avg} = \frac{mS}{\sqrt{mN}}$$  \hspace{1cm} (4.3.3)  

which is $\sqrt{m}$ times larger than what it would be by each sensor alone. This technique
is routinely implemented by sampling the signals at certain time intervals and storing
the data in a memory (serial averaging). In comparison, parallel averaging is based
on combining independent measurements of displacement using multiple sensors
[105, 106]. In nanopositioning applications, the delay generated by serial averaging
may adversely affect the performance of the closed-loop system. In addition, it may
add quantisation noise to the measured quantity, which further deteriorates the SNR.
The parallel approach is immune to quantisation noise and time delays that may
arise in serial methods.

A real-time analogue averaging method that avoids the complications arising from the
digitisation and storage of data is proposed in this chapter. The 1-DoF nanopositioner
designed for this purpose is illustrated in Figure 4.9. The operating principle of
each electrothermal sensor is similar to that explained earlier. Applying electrical
power to doped silicon resistors (see $S_1$ magnified) increases their temperature. The
heat conduction through a thin air gap (2$\mu$m) between the stage and the sensors is
proportional to the overlapped area. The stage movements change the area overlapped and, consequently, the sensors temperature. The resistance of sensors varies as a function of temperature. Hence, the stage displacement translates to resistance variations. Four pairs of sensors are on the two opposing sides of the moving stage. Each sensor is independently measuring the same displacement. Hence, they translate the same amount of movement to resistance variations. However, imperfections in the MEMS fabrication process lead to different resistance values. Consequently, even with the same excitation voltage or current the sensors may heat up to different temperatures leading to different sensitivities and unequal output signals. It is known that the sensor temperature is determined by the amount of injected electrical power. Therefore, in order to have sensors with identical sensitivities, the applied electrical power is kept constant in the measurements. This is achieved by operating the sensors in constant voltage mode and monitoring the sensors bias current and voltage simultaneously.

The proposed readout circuit consists of two transimpedance amplifiers (TA) that add the sensor outputs and a differential amplifier to provide the required gain, as shown in Figure 4.10. $R_{1p}$ and $R_{1n}$ are nominal resistances of a differential sensor and $V_h$ and $V_l$ represents the heating voltage and resistance changes due to the stage displacement, respectively.

### 4.3.2 Experimental Results

The 1-DoF nanopositioner shown in Figure 4.9 is fabricated in a SOI MEMS process. The sensors length, width and thickness are $50\mu\text{m}$, $2\mu\text{m}$ and $25\mu\text{m}$, respectively, and their spacing from the stage is $2\mu\text{m}$. A 3V dc voltage is applied to heat up the sensors. Larger voltages are avoided due to the current limit of the following amplifiers. The stage is moved by the electrostatic forces generated by the comb
drives. A 110Hz actuation signal is applied to the electrostatic actuator in order to operate the device sufficiently away from the low frequency noise and the power line 50Hz noise. Despite monitoring the electrical power delivered to sensors to ensure equal sensitivity, $S_4$ displayed a different characteristic, which is most likely due to fabrication tolerances. Therefore, three sensors were utilised in the experiments. The power spectrum of each sensor ($S_1$, $S_2$ and $S_3$), and their combination ($S_{123}$), were measured. These are depicted in Figure 4.11. The signal level is increased by about 9dB. In order to measure the noise level a separate noise floor (nf) measurement was carried out by removing the actuation signal. The integrated noise floor over a 10Hz bandwidth shows that the noise floor of the combined output is increased by about 5dB. Therefore, the SNR is improved by approximately 4dB, which is very close to the theoretical prediction of $20\log(\sqrt{3})$.

The bandwidth of the sensors is about 6.6kHz in both cases, which is obtained by

Figure 4.10: Proposed readout circuit designed for multiple sensor outputs.
applying a step input to the sensor bias and measuring the settling time, assuming that it resembles a first order transfer function \[7\]. The sensor output voltage was measured versus the stage displacement by an optical microsystem analyser (MSA-050-3D) for a single sensor \(S_1\) and combination of three sensors \(S_{123}\). This was achieved by applying a slowly varying triangular signal to the actuator and recording the stage displacement and the sensor output voltage simultaneously. The transfer characteristics are shown in Figure 4.12. \(S_{123}\) is scaled down to match its sensitivity with \(S_1\). In order to compare the linearity, the maximum residual of a linear fit is divided by the full scale of the sensor output which leads to 2.32\% and 3.12\% for \(S_1\) and \(S_{123}\), respectively. The three sensors are operated independently, therefore the increased non-linearity can be attributed to the limited output current of transimpedance amplifiers. Three sensors draw higher currents, which drives the amplifier closer to its non-linear region. The sensor dynamic range (\(\frac{\text{maximum displacement input}}{\text{sensor resolution}}\)) is limited by the actuator displacement to 40.5dB. As

\[\text{Figure 4.11: Measured power spectral density for three sensors and their combination.}\]
4.4 Sensor Excitation Effects on Resolution: A Comparison

The electrothermal sensor shown in Figure 4.13 is comprised of a pair of resistors that detect the displacement of a moving stage. The sensing mechanism is based on the heat conduction between the resistors and the stage, as discussed in Chapter 2. The sensitivity of the sensor is dominated by the temperature gradient, i.e. a higher sensor temperature translates to a higher displacement to resistance variation gain. The thermal noise of the sensor increases with the temperature as well. Hence, the trade-off between sensitivity and low thermal noise determines the signal to noise ratio that translates to displacement resolution.

In order to heat up the resistors, electrical power is delivered either by a constant current (CC) [8] or a constant voltage (CV) source [11]. The CV mode is widely

![Figure 4.12: Sensor transfer characteristic.](image-url)
reported in the literature. However, the study reported in [8] indicated that the CC excitation leads to a higher sensitivity. Achieving higher resolution is critical for MEMS displacement sensors. In order to properly analyse this phenomenon, a new readout circuit that guarantees equal operating conditions for both CV and CC excitation modes is reported in this chapter. A dynamic model method of electrothermal sensors is presented to analytically compare the two approaches. Experimental results support the simulations that the achievable SNR is almost identical for both excitation modes.

### 4.4.1 Current Excitation Model

The electrical power delivered to the sensor with a constant voltage source \( V \) and a constant current source \( I \) is described by Equation 4.4.1 and Equation 4.4.2, respectively:

\[
P = \frac{V^2}{R} \tag{4.4.1}
\]

\[
P = R I^2 \tag{4.4.2}
\]
4.4 Sensor Excitation Effects on Resolution: A Comparison

Any displacement that results in a resistor temperature increase adds to its resistance value. According to Equation 4.4.1, in the CV mode, this leads to a reduction in the electrical power dissipation, which results in a decrease in temperature. In comparison, Equation 4.4.2 implies that any increase in the resistor temperature in CC mode leads to a higher electrical power dissipation that translates to a higher resistor temperature, and this cycle makes the sensor more sensitive in CC mode. A comprehensive model that addresses the dynamic behaviour of MEMS electrothermal systems [11], was discussed in Chapter 2. It explains the displacement \( x \) functionality of the sensor resistance in the CV mode, as depicted in Figure 4.14. Here, \( T_T \) and \( g(\cdot) \) are the two main operators that relate the electrical power to temperature and the temperature to the sensor resistance, respectively. The operating point of the sensor is described by \( V_0, I_0, R_0, P_0, T_0 \) coefficients, in terms of its voltage, current, resistance, electrical power and temperature, respectively. The displacement perturbation \( \tilde{x} \), as the system input, results in temperature changes \( \tilde{T} \). This, in turn, leads to resistance fluctuations \( \tilde{R} \) that are measured by the output current variation \( \tilde{I} \). For small signal analysis, the Taylor series expansion of Equation 4.4.1 can be used to linearise the non-linear blocks around the operating point of the system:

\[
P = \frac{V_0^2}{R_0} + \frac{2V_0\tilde{V}R_0 - V_0^2\tilde{R}}{R_0^2}
\]  \hspace{1cm} (4.4.3)

The resulting perturbation in electrical power \( \tilde{P} \) can be extracted from Equation 4.4.3 by removing the constant term \( (P_0=V_0^2/R_0) \):

\[
\tilde{P} = 2I_0\tilde{V} - \tilde{R}I_0^2
\]  \hspace{1cm} (4.4.4)
In order to adapt the model to the CC excitation mode, Equation 4.4.2 is expanded as:

\[ P = R_0 I_0^2 + \tilde{R} I_0^2 + 2R_0 I_0 \tilde{I} \]  

(4.4.5)

which leads to

\[ \tilde{P} = \tilde{R} I_0^2 + 2R_0 I_0 \tilde{I}. \]  

(4.4.6)

Thus, the CC mode model can be obtained by applying the electrical power signal in positive feedback to the system, as illustrated in Figure 4.15. Note that constant coefficients that match with the CC mode equations are shown in the highlighted box.

Positive feedback typically leads to a larger gain. It often has a similar effect on noise. The signal transfer functions \( T_{I_x} \) and \( T_{V_x} \) and the noise transfer functions \( T_{I_n} \) and \( T_{V_n} \) for CV and CC modes, are obtained as:

\[ T_{I_{\bar{x},x_0}} = \frac{K'(x_0)}{K(x_0)} \left( -\frac{I_0}{R_0} \frac{g'(T_0)P_0 T_{TP_{x_0}}}{1 + I_0^2 g'(T_0) T_{TP_{x_0}}} \right) \]  

(4.4.7)
Figure 4.15: Modified Model for the Electrothermal Sensor in CC Mode

\[
T_{\tilde{V}_{n_{x_0}}} = -\frac{I_0/R_0}{1 + I_0^2 g'(T_0)T_{TPx_0}} n_R
\]

\[
T_{\tilde{R}} = -\frac{I_0/R_0}{1 - I_0^2 g'(T_0)T_{TPx_0}} n_R
\]

where \( K(x) \) determines the \( T_{TP} \) gain as a function of displacement \( x \), and \( n_R \) is the sensor noise. The power spectral density of the net resistance noise is the sum of the thermal noise \( (n_{R_T}) \) and the flicker noise \( (n_{R_f}) \) as:

\[
S_{nR}(f) = S_{nR_T}(f) + S_{nR_f}(f) = 4k_BT_0R_0^2 f_{N_{carr}}
\]

where \( k_B \) and \( f \) represent the Boltzmann constant and frequency, respectively and \( \alpha \) (the Hooge factor) and \( N_{carr} \) are experimentally determined. The SNR for CV and CC modes is defined by Equation 4.4.11 and Equation 4.4.12 over a desired frequency bandwidth.

\[
SNR_{CV} = 20 \log \left( \frac{\sqrt{\int |T_{\tilde{I}_{n_{x_0}}}(f)|^2 df}}{\sqrt{\int |T_{\tilde{I}_{n_{R_{x_0}}}}(f)|^2 S_{nR}(f) df}} \right)
\]

(4.4.11)
**Figure 4.16:** The signal to noise ratio obtained by the analytical approach.

\[
SNR_{CC} = 20 \log \left( \frac{\sqrt{\int |T_{\tilde{V}\tilde{x}_0}(f)|^2 df}}{\sqrt{\int |T_{\tilde{V}n_{R,x_0}}(f)|^2 S_{n_R}(f) df}} \right)
\]

(4.4.12)

The SNR values corresponding to Equation 4.4.11 and Equation 4.4.12 were evaluated and plotted in Figure 4.16. It can be observed that the CC excitation mode leads to a slightly higher SNR (about 2dB). The 5kHz simulation frequency bandwidth covers the typical electrothermal sensor bandwidth [11]. Further simulations also showed that the empirically defined values, such as \(N_{carr}\) and \(\alpha\), affect the absolute noise values but the differences are the same.
4.4.2 Experimental Verification

Experiments were performed on a SOI-MEMS nanopositioner with integrated electrothermal sensors, as shown in Figure 4.13. The width, length and height of each sensor are 2μm, 50μm and 25μm, respectively, which translates to around 300Ω heated electrical resistance ($R_0$) for a typical bias current of ($I_0=16mA$). $G(\cdot)$ and $T_{TP}$ transfer functions were identified based on the procedure proposed in [11]. A readout circuit was designed and tested to verify the analytic model and simulations. In the conventional CV mode (Figure 4.17(a)) the first pair of operational amplifiers acts as a transimpedance amplifier that converts the current variations arising from differential resistive changes to a voltage output. $V_h$, $R_M$ and $v_{nT}$ represent the heating bias voltage, the MEMS sensor resistor and the sensor noise, respectively. The subsequent differential amplifier produces a voltage output ($V_{out}$), which is a linear function of the resistive changes. In order to guarantee equal readout circuit noise contributions, a similar configuration is designed for the CC mode excitation by only swapping the feedback resistor ($R_L$) with the MEMS resistor, as depicted in
During experiments, the heat-sink (stage) shown in Figure 4.13 was actuated by an electrostatic comb drive. The sensor output voltage is measured against the stage displacement by an optical microsystem analyser (MSA-050-3D), as illustrated in Figure 4.18. The sensitivity is clearly higher in CC mode (1.3V/µm) compared with the CV mode (0.38V/µm). However, the achievable displacement resolution is determined by the sensitivity and output voltage noise floor. Since the amplifiers have identical noise contributions, the output power spectrum can be used to compare the achievable SNRs corresponding to the two modes. The stage is actuated by a 30Hz sinusoidal signal and the sensor output power spectrum was measured, as illustrated in Figure 4.19. The CC excitation results in approximately 10dB higher signal level. It is observed that the noise floor (NF) has also risen by about the same amount, which is in close agreement with the analysis. The measured noise floor and sensitivity translates to $0.04\text{nm}/\sqrt{\text{Hz}}$ displacement resolution for both cases. The small difference between the simulations and experimental results is most likely due to the
approximations used in the analytical modelling, particularly for the identification of the physical attributes of the silicon.

### 4.5 Conclusion

The positioning precision of a MEMS nanopositioner is strongly dependent on its sensor specifications. In this chapter the resolution of electrothermal displacement sensors has been investigated through a circuit and device level approach. An alternative sensing procedure has been introduced to reduce the flicker noise. Driving the electrothermal sensors with an ac source instead of the commonly used dc source led to a lower noise level in the low frequency range. The effectiveness of the method is verified experimentally with electrothermal sensors on a 1-DoF nanopositioner. The flicker noise floor level has been reduced by 8dB around the
20Hz signal, which translates to a resolution improvement from 0.2nm/√Hz to approximately 0.1nm/√Hz.

In addition, a multiple sensor based 1-DoF MEMS nanopositioner has been designed to cope with the thermal noise of the sensors. A 4dB SNR enhancement is achieved successfully, which is close to the theoretical improvement attainable by the averaging technique, and leads to a resolution enhancement from 0.3nm/√Hz to 0.15nm/√Hz. The number of sensors can be increased to achieve a higher resolution. However, the limited space available around the stage and higher power consumption must be considered in the design process.

Furthermore, the noise performance of a MEMS electrothermal displacement sensor excited by a constant current (CC) and a constant voltage (CV) source has also been investigated. The analytical model shows that the achievable SNR in the CC mode is slightly higher (about 2dB). The experiments have been carried out on electrothermal sensors integrated in a 2-DoF SOI-MEMS nanopositioner. The measurements show that although the CC mode leads to a higher output signal level, it increases the noise floor as well. Hence, the maximum achievable SNR and displacement resolution by the two modes is almost identical, which is in close agreement with the analytical result.
A 2-degree-of-freedom (2-DoF) nanopositioner for on-chip atomic force microscope (AFM) is designed and implemented as a microelectromechanical system (MEMS). The device is fabricated using a silicon-on-insulator-based process in order to function as the scanning stage of a miniaturised AFM. The lateral displacements of the scan table are measured using a pair of electrothermal position sensors. An innovative structure is proposed to integrated the sensors in the nanopositioner device with minimal modifications. The integration of the electrothermal sensors in the microactuator facilitates the closed loop control of the electrostatically actuated stage. The electrothermal sensor bandwidth is larger than the stage lateral resonance frequency, which is at approximately 850Hz. The incorporated electrostatic actuators achieve a travel range of 16µm in each direction. These sensors are used, together with a positive position feedback (PPF) controller, in a feedback loop, to damp the highly resonant dynamics of the stage. The sensitivity of the sensor readout circuits are also investigated through both circuit analysis and experimentation. The feedback controlled nanopositioner is used, successfully, to generate high-quality AFM images at scan rates as fast as 100Hz.


5.1 Introduction

Nanopositioners are high precision microactuators that are used to produce repeatable motions with nanometer resolution. The precisely controlled motions are particularly demanding in diverse fields of micro and nanotechnology, where positioning, manipulation and interrogation of samples is the main objective. The applications span from molecular biology to nanoassembly and optical alignment systems [107, 108, 109, 110]. As stated in the previous chapters, recent developments of novel nanopositioning devices based on MEMS fabrication processes have attracted significant attention [111, 112, 113, 114]. Compared to their macro/meso scale counterparts, the miniaturised nanopositioners demonstrate substantial improvements in terms of the operating bandwidth, fabrication costs and the packaged size of the system [111, 115]. For example the high density probe storage device implemented in MEMS, has successfully achieved these attributes as reported in [116, 117, 118].

Atomic force microscope (AFM) is one of the most influential tools in the development of science and engineering in recent times [20]. The AFM features a sharp probe, of a few nanometers width, that scans the sample surface, which is located on a stage (See Figure 1.2). The lateral motions, in order to scan the sample, can be achieved either by moving the stage or the cantilever itself. The MEMS-based nanopositioner can be used as the scanning stage of an AFM in order to reduce the size of the main components of the system. The design of such a MEMS-based AFM scanner was addressed in [119], where a 2-degree-of-freedom (2-DoF) MEMS nanopositioner was designed to operate as the scanning stage for an off-the-shelf AFM.

The recently reported servomechanisms for MEMS nanopositioners apply a feedback control to the actuator input by means of the integrated sensor output, which enhances the dynamic and static performance of the mechanical system. As shown in [17, 120], a closed-loop control system can be used to achieve a higher positioning
accuracy and alleviate imperfections inherent in the microactuator such as drift and vibration.

In this chapter, the design and implementation of a controlled scanner stage is presented, which consists of a 2-DoF MEMS nanopositioner equipped with electrothermal sensors, as illustrated schematically in Figure 5.1. An external displacement sensing system, which is typically an optical interferometer, was required in order to control the previous MEMS nanopositioner [119]. The proposed nanopositioner is upgraded by integrating electrothermal displacement sensors with the stage in the same MEMS process in order to measure the lateral motions of the stage along each of its two axes. This introduces a major step for implementing the entire AFM on a MEMS chip. In addition to the lower cost and size factors, integration of sensors in the same standard fabrication process mitigates the required sensor calibration and enhances
the positioning accuracy. It should be noted that the gap between the electrothermal sensors and the 1-DoF stage introduced in Chapter 2 was assumed to be constant. However, this assumption is not satisfied in the 2-DoF nanopositioner schematic diagram. An innovative mechanical structure is designed in this work to operate the electrothermal sensors based on the same principles.

The remainder of this chapter continues as follows. The design of the MEMS scanner and its modal analysis result are described in section 5.2. The electrothermal sensor design and the readout circuit are presented in section 5.3. The characterisation of the device, which includes sensor sensitivity and frequency response measurements, as well as the static non-linearity linearisation of the nanopositioner, is discussed in section 5.4. The design and implementation of controllers to suppress resonant modes of the device and to facilitate good tracking of reference signals is presented in section 5.5. Finally, the performance of the MEMS nanopositioner for AFM imaging is evaluated in section 5.6.

## 5.2 2-DoF MEMS Scanner

The main components of a scanner consist of actuators, suspension beams and the stage. The design and implementation of a MEMS scanner stage with similar features is reported earlier in [119]. In order to add sensors a second version of the device is designed and fabricated in this work using a commercial SOI MEMS process. As stated earlier, real-time measurements of the stage lateral displacements, along the x and y directions, can be facilitated by embedded displacement sensors, which are integrated with the nanopositioner in the same fabrication process. Figure 5.2 shows the scanning electron microscopy (SEM) image of the proposed nanopositioner, equipped with the electrothermal sensors. The magnified images of the sensor part
and reference gold features on the stage surface are illustrated at the top, which will be explained in the following sections.

The parallel-kinematic configuration is used to design the actuation system, which moves the scanner stage with two planar mechanical degrees of freedom. Two electrostatic actuators pull the stage in x and y directions by the force generated between the engaged comb fingers. The stage area is $3 \text{mm} \times 3 \text{mm}$. The stage is suspended by means of two beam sets designed at the opposite sides of the
actuators, which decouple the x and y forces as well. Hence, it is ideally expected that the motions in each direction are solely controlled by the corresponding actuator. However, it has been shown that absolutely independent x and y motions are hardly achievable. The maximum displacement of the stage is limited by the mechanical stoppers to 20μm.

The minimum feature size available in the MEMS fabrication technology determines the size of comb fingers and the gap between engaged comb fingers, which sets a limit on the maximum force and displacement range that can be achieved by electrostatic actuators. The width of each finger in this design is 2μm and the gap between the fingers is 2μm. A higher force is needed for a high resonance frequency in the mass-spring system of the nanopositioner, which is essential for a high speed scanning. The thickness of the moving parts including the suspension beams are limited by the fabrication technology to 25μm. The height to width ratio of the beam is designed to be 7:1 in order to ensure higher stiffness for the stage in z direction compared with in-plane directions [121].

The initial design including the desired displacement and frequency response is followed by a simulation in a MEMS design software package (Coventor Ware) for analysing the resonant modes of the structure. The simulation result is shown in Figure 5.3 where the in-plane resonant frequency of the nanopositioner is located at approximately 903Hz. The symmetrical design of the device guarantees similar resonant behaviour for both of the in-plane motions. The nanopositioner is fully fabricated in a commercial SOI MEMS process owned by MEMSCAP [2].
5.3 Position Sensors for the 2-DoF Nanopositioner

5.3.1 The Electrothermal Displacement Sensor

The displacement sensing technologies available in MEMS devices are based on the capacitive [122], piezoresistive [16], and electrothermal [7] effects. Fabrication of the capacitive sensors is highly compatible with standard MEMS fabrication processes. However, piezoresistive and electrothermal sensors can be fabricated with much smaller footprints, which makes them rather attractive in applications where space is at a premium. These sensors map the displacements of a stage to variations in the resistivity of a transducer, typically made of doped silicon. Piezoresistive sensors can be operated at lower power levels compared to electrothermal sensors. However, the resistance of the doped silicon is more sensitive to temperature changes than variations in mechanical stress [12]. The small footprint and high sensitivity of electrothermal sensors were the motivations to utilise this technology to measure
displacements of the nanopositioning stage.

An early implementation of the electrothermal displacement sensor was reported in [7], and it was subsequently used in a probe-based data storage device reported in [116]. The operating principle of the sensor is based on utilising the temperature sensitivity of doped silicon resistance. The conduction of electrical power through the doped silicon increases its temperature. The transfer of heat between a moving heat sink (the nanopositioner stage, in this case) and a stationary hot resistor (sensor) changes the sensor resistance, as depicted in the 1-DoF nanopositioner schematic in Figure 4.1, where the stage is actuated electrostatically. The narrow spacing between the heat sink and the sensor results in the heat transfer mechanism being dominated by conduction [8]. As the stage moves, it absorbs heat from the resistor. When the resistor temperature changes its resistivity will change. Therefore, voltages derived from the change of the sensor’s resistivity are highly correlated with displacement, and can be used to represent the displacement of the stage. The nanopositioner’s electrothermal sensors are arranged in a differential topology, in order to improve the sensor linearity and reject common mode signals. In previously reported electrothermal sensing implementations, the heaters have typically featured a uniform cross-section [7, 8, 115]. Here, the integrated sensors are designed to have a nonuniform profile. As discussed in [123], electrothermal heaters with this shape display a flatter spatial temperature distribution that leads to higher sensor linearity and sensitivity. Further analysis of the sensor design and characteristics, including drift and bandwidth, are also presented in [123].

### 5.3.2 Proposed Structure for the Integrated Sensor

The use of electrothermal sensors in 1-DoF nanopositioners has been reported in recent literature, e.g. see [8, 15]. In such designs the distance between the sensor
and the stage must be kept fixed in order to make precise measurements of the lateral movements of the stage. It is rather difficult to achieve this objective in a 2-DoF planar design as illustrated in Figure 5.1. Here, the unique structure of the 2-DoF nanopositioner, as described in section 5.2, is used in order to incorporate the electrothermal displacement sensors for both axes. A scanning electron (SEM) micrograph of the proposed 2-DoF nanopositioner, with integrated electrothermal displacement sensors, is shown in Figure 5.2. The frames have elongated ends, compared to the previous design reported in [119], in order to accommodate the addition of the electrothermal sensors. The decoupled mechanical design facilitates independent motions of the frames along the two lateral axes. Thus, the frame that tracks the stage displacement in one direction is fixed in the orthogonal direction. Hence, the lateral motions of the stage are followed by the frames independently, which leads to a fixed gap between each sensor and its corresponding frame. The length of each sensor is 50µm, which is sufficient for measuring the full travel range of the stage. The sensors were designed to have resistance values of approximately 200Ω. In the fabricated nanopositioner, however, the sensor resistances vary from 160Ω to 300Ω. The offset voltages generated by these differences can be corrected through the use of potentiometers in the readout circuitry. The sensor resistance variations may still result in varying sensitivities between the sensors. However, these variations are addressed through the use of lookup tables, as articulated in section 5.4.

5.3.3 Readout Circuitry

As stated earlier, the desired sensor output is the heater resistivity changes, which are highly correlated with the movements of the stage. Common readout circuits for resistive measurements are based on techniques involving RC-decay, oscillator
frequency, resistance-to-current conversion and resistance-to-voltage conversion [124]. In the first approach, a voltage pulse is applied to the RC circuit, and the time it takes for the output voltage to reach a certain threshold can be measured from the variation in resistance or capacitance [125]. This method is effective for resistive sensors that have a large dynamic range. The resistivity change in an electrothermal sensor, however, is typically less than 10% of its nominal value, which does not lead to a significant RC variation. The effectiveness of the second approach is impeded by the phase noise of the oscillator. Particularly, if a ring oscillator is used, it is known that the phase noise contributes substantially to the total measurement noise and reduces the achievable SNR, as discussed in Chapter 3.

To avoid these complications, the readout circuit is designed based on the concept of resistance-to-voltage conversion, which uses a Wheatstone bridge, and resistance-to-current conversion, which uses transimpedance amplifiers (TA). Both methods are implemented and tested on the MEMS nanopositioner. The Wheatstone bridge is commonly used in resistive readout circuits. However, the conclusion were that for identical resistive changes, a TA-based readout circuit offers a higher sensitivity.

The readout circuits corresponding to the two methods are schematically illustrated in Figure 5.4. The actuation voltage, represented by $V_a$, drives the electrostatic actuator. $V_h$ is the dc heating voltage, $R_{MEMS1,2}$ are the sensor heated resistors and $R_{r1,2}, R_{F1,2}$ are the bridge reference and the TA feedback resistors, respectively. The sensor resistance variations can be described as:

$$R_{MEMS1,2} = R \pm \delta R,$$  \hspace{1cm} (5.3.1)

where $R$ is the MEMS resistor value with the stage at the middle (no displacement) and $\delta R$ is the resistance changes associated with the stage displacement. Assuming
that $R_{MEMS1}=R_{MEMS2}=R$, the bridge output voltage is:

$$V_{outB} = A \frac{2R \delta R}{(R + R_{ref})^2} V_h,$$

(5.3.2)

where $A$ is the differential amplifier voltage gain and $V_h$ is the bias voltage, which heats the sensors. The balanced bridge conditions lead to $R=R_{ref1,2}$. Therefore, the total gain can be obtained as:

$$\frac{V_{outB}}{\delta R} = \frac{AV_h}{2R}.$$

(5.3.3)
Similarly, the output voltage for the TA circuit is given by:

\[ V_{outT} = A \frac{2R_F \delta R}{(R^2 - \delta R^2)} V_h \] (5.3.4)

As long as no voltage gain is expected from the TA, we may assume \( R = R_{F1,2} \). Therefore, neglecting \( \delta R^2 \), the total gain can be approximated as:

\[ \frac{V_{outT}}{\delta R} = \frac{2AV_h}{R} \] (5.3.5)

Comparing Equation 5.3.3 and (Equation 5.3.5), it can be inferred that the achievable sensitivity with the TA readout circuit is four times higher than that with the bridge circuit. This is supported by experimental results, illustrated in Figure 5.5, where for the same actuation voltage, a larger output voltage is obtained with the TA circuit. In particular, it should be noted from this figure that the slope of the TA readout circuit transfer characteristic, i.e. its sensitivity, is much larger than that of the bridge circuit. Additionally, in the TA topology the heating voltage across the resistor is kept fixed, which leads to a constant voltage mode operation. In contrast, the bridge topology does not guarantee a constant voltage across the resistor. The curves shown in Figure 5.6(a) are four iterations of the same measurement, using the bridge circuit on one sensor, which indicate that a repeatable measurement cannot be ensured. The constant voltage mode operation achieved by the TA leads to similar outputs for the four iterations, as demonstrated in Figure 5.6(b)
5.4 Device Characterisation

5.4.1 Sensor Sensitivity Measurement

A Polytec PMA-400 planar motion analyser is used to measure the lateral displacements of the stage, which are called x and y directions. This instrument measures the displacement based on the principle of stroboscopic video microscopy, which is carried out by the inbuilt image analysis software of the PMA. The stage moves in the x and y directions by the electrostatic force generated by applying the actuation voltages, from 0V to 45V, to the comb drives. The corresponding displacements (measured by the PMA) and sensor output voltages are recorded simultaneously. The results are plotted in Figure 5.7. The non-linearities shown in these plots are due to the non-linear relationship between the generated force and the voltage applied to the comb actuators. With a maximum actuation voltage of 45V, the measured displacement is 16µm for both x and y axes. From the measured displacement versus sensor output plots shown in Figure 5.7(c), the mean sensitivity of the x and y sensors are estimated from the slopes of linear best-fit curves fitted to the data. The

![Figure 5.5: Comparison of bridge and TA sensitivity.](image)
mean sensitivity of the x and y sensors are approximately 2.29µm/V and 3.42µm/V, respectively.

The power spectral density (PSD) of the sensor output is measured by a HP35670A signal analyser in order to determine the sensor resolution. The noise floor for the x-axis sensor (worst case) is -90dBVrms/√Hz. In the open-loop actuator-sensor system this typically translates to a 1 nm displacement resolution over a 100Hz frequency bandwidth.

Figure 5.6: Sensor sensitivity curves achieved by the same setup over four iterations using (a) the Wheatstone bridge and (b) the TA circuit.
5.4 Device Characterisation

5.4.2 Linearisation of the Static Nonlinearity

The non-linearity between the input actuation voltage and the actuator displacement can be estimated by a high-order polynomial function. To linearise this, a lookup table is implemented in dSPACE and placed in series with the plant, as shown in Figure 5.8. The lookup table stores the displacement data and actuation voltages of the MEMS nanopositioner. These stored values are obtained from the data plotted in Figure 5.7(a). For every predefined input displacement $u$ (in $\mu$m), the lookup table
generates an output voltage $V_a$ by interpolating/extrapolating among the stored values. The generated signal $V_a$ is fed to a voltage amplifier which has a gain of 20. The output voltage of the amplifier is in turn used to drive the nanopositioner.

The sensors exhibit minor non-linearities, which may be attributed to device imperfections due to tolerances in the MEMS fabrication process. These non-linearities were linearised using a second lookup table (see Figure 5.8) which stores the data plotted in Figure 5.7(c). Similar to the first lookup table, for every sensor voltage $S$ (in Volt) fed in to the lookup table, the lookup table generates an output displacement $d$ (in $\mu$m). By cascading the lookup tables with the plant, the input-output relationship from $u$ to $d$ is linearised.

### 5.4.3 Frequency Response Measurement

The x and y axes frequency response functions (FRFs) of the MEMS nanopositioner are obtained using the signal analyser. The displacements of both the axes are biased to their mid-range, which is at $8\mu$m. The FRFs are recorded from the inputs applied to the cascaded plant $u_x$ and $u_y$ in Figure 5.8 to the measured displacement $d_x$ and $d_y$, respectively.

Figure 5.9 and Figure 5.10 plot the experimentally determined open and closed loop FRFs for both the x and y axes. The first resonance frequency of the x axis is located
5.5 Control Design and Implementation

![Diagram of frequency responses](image)

**Figure 5.9:** Measured open and closed loop frequency responses of the x axes.

at 860Hz. For the y axis, the resonant peak appears at 850Hz. These measured resonances are close to the ConventorWare simulated values. The dynamic range of the x and y resonant peaks are 36.6dB and 37.5dB, respectively.

### 5.5 Control Design and Implementation

The measured motion coupling from x to y, and y to x, are -40.7dB (0.92%) and -42.5dB (0.75%), respectively, which are remarkably low. Thus, the MEMS nanopositioner can, effectively, be considered as two single-input single-output (SISO) systems. A second-order model was fitted to the measured frequency response of each axis using the frequency-domain subspace algorithm [126]. The following are
Next, a positive position feedback (PPF) controller is designed and implemented on the fast axis (x axis) of the nanopositioner in order to suppress its first resonance frequency. PPF, proposed by Caughey and coauthors [127, 128], is known to be an effective controller capable of providing substantial damping to collocated structures [129]. The transfer function of a PPF controller rolls-off at 40dB/decade at high
5.5 Control Design and Implementation

Inversion-based Feedforward

 Offline

\[ G_{cl}^{-1}(s) \]

Real-time

\[ G_{cl}(s) \]

\[ K_{ix} \]

\[ + \]

\[ \frac{K_{ix}}{s} \]

\[ u_x \]

\[ C_{PPF} \]

\[ G_{dxu_x} \]

\[ d_x \]

\[ r_x \]

\[ \hat{x}_x \]

\[ u_x \]

\[ d_x \]

\[ C_{notch} \]

\[ u_y \]

\[ d_y \]

\[ r_y \]

\[ + \]

\[ \frac{K_{iy}}{s} \]

\[ G_{dyu_y} \]

\[ d_y \]

**Figure 5.11:** Block diagram of control structures for the x and y axes.

frequencies, which is a desirable property to avoid the excitation of the high frequency dynamics of the nanopositioner. An integral tracking controller was implemented in the outer loop to improve low frequency tracking (see Figure 5.11). The combined control strategy is known to reduce crosstalk between the two lateral axes [130]. In a typical raster application, the y axis is used to track a slow ramp set-point. Hence, the closed-loop bandwidth requirement on this axis is less demanding. A notch filter combined with an integral tracking controller [131] is designed and implemented on the y axis, as shown in Figure 5.11. The transfer functions of the two controllers are
described in Equation 5.5.3 and Equation 5.5.4 below:

\[
C_{PPF} = \frac{5.5 \times 10^6}{s^2 + 4161s + 3.533 \times 10^7} \tag{5.5.3}
\]

\[
C_{\text{notch}} = \frac{s^2 + 289.9s + 2.574 \times 10^7}{s^2 + 2.899 \times 10^4s + 2.574 \times 10^7} \tag{5.5.4}
\]

A dSPACE-1103 rapid prototyping system, working at a sampling rate of 80 kHz, is used to implement the controller. The closed-loop frequency responses of the two axes are plotted in Figure 5.9. The PPF controller in the damping loop reduces the resonant peak by 22.1dB. Together with the integral controller with a gain of \(K_{ix} = 700\), the achievable closed-loop bandwidth is 165Hz. The gain and phase margins are 7.24dB and 80.2°, respectively. For comparison, in a closed-loop system without the damping loop, but with an integrator, the highest closed-loop bandwidth obtainable is only 16Hz. The tracking bandwidth is increased tenfold with the implementation of the proposed PPF controller. For the slow axis, where a smaller closed-loop bandwidth is acceptable, the measured bandwidth is 18Hz with gain and phase margins of 27.2dB and 87.1°, respectively. This bandwidth is sufficient for tracking a slow ramp reference input.

### 5.5.1 Inversion Feedforward

Without feedback control, the highest scan rate achievable on a nanopositioner with a resonance frequency of 860Hz is 8.6Hz [107]. Here, the objective is to obtain AFM images with raster scan rates up to 100 Hz. The obtained closed-loop tracking bandwidth of 165Hz is insufficient to track a fast 100-Hz triangular waveform. Therefore, the inversion-based feedforward technique is used to further increase the tracking bandwidth. A model-based feedforward technique may not be able to compensate for plant uncertainties [132]. However, plant uncertainties can be
minimised by means of feedback control [121, 133, 134], i.e. by implementing the two feedback control loops as presented in section 5.5. A model $G_{cl}(s)$ is fitted to the closed-loop frequency response data of the x axis for inversion, see Figure 5.11(a). In order to obtain an accurate inverse model $G_{cl}^{-1}(s)$, the frequency range of the model was restricted to 1.5 kHz to reduce modelling errors due to uncertainties in the measured closed-loop data. The inversion feedforward inputs $\hat{r}_x$ were obtained by using all odd harmonics of the triangular waveform that lie within the bandwidth of 1.5kHz, where the amplitude of the harmonics is scaled by $|G_{cl}^{-1}(i\omega)|$, and the phase of the harmonics was shifted by $\angle G_{cl}^{-1}(i\omega)$. Note that $\hat{r}_x$ are obtained offline as shown in Figure 5.11(a). By implementing the inversion feedforward technique, the tracking bandwidth of the system was increased from 165 Hz to 1.5kHz.
5.6 AFM Imaging Performance

This section reports on the AFM images obtained with the feedback controlled MEMS nanopositioner. The experimental setup, consisting of the MEMS scanner mounted on a printed circuit board, together with the readout circuitry and a Nanosurf EasyScan 2 AFM, is illustrated in Figure 5.12. The experiments are performed in the scan-by-sample mode where the scan table, which is deposited with calibration features (see Figure 5.2), is moved in relation to the static probe. The z scanner of the commercial AFM and its in-built vertical feedback controller are activated during the “landing” process to regulate the probe-sample force. After successfully “landing” the probe, the vertical feedback controller is turned off. The MEMS nanopositioner is driven in a raster pattern during the scans. This scanning mode is known as the constant-height contact-mode.

A cantilever probe with a resonance frequency of 13kHz and a stiffness of 0.2N/µm is used during the scans. The height of the features is approximately 550nm. The remaining dimensions can be found in Figure 5.2. An image area of 12.7µm×12.7µm is scanned at 10Hz, 50Hz and 100Hz in open-loop, closed-loop and closed-loop with inversion. Figure 5.13 plots the 3-dimensional topography images, the fast x axis displacements and the tracking errors of the nanopositioner.

At 10Hz, vibrations are not noticeable in any of the scans. While vibration-induced artefacts appear in the open-loop scans at 50Hz and 100Hz, these vibrations are

<table>
<thead>
<tr>
<th>Scan freq. (Hz)</th>
<th>OL Max</th>
<th>OL RMS</th>
<th>CL Max</th>
<th>CL RMS</th>
<th>CL+Inv Max</th>
<th>CL+Inv RMS</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>119.6</td>
<td>87.6</td>
<td>328.4</td>
<td>308.0</td>
<td>215.2</td>
<td>108.4</td>
</tr>
<tr>
<td>50</td>
<td>766.7</td>
<td>313.8</td>
<td>541.8</td>
<td>463.8</td>
<td>366.4</td>
<td>166.5</td>
</tr>
<tr>
<td>100</td>
<td>1435.2</td>
<td>965.4</td>
<td>1379.4</td>
<td>1018.8</td>
<td>500.4</td>
<td>167.5</td>
</tr>
</tbody>
</table>
suppressed in closed-loop. However, the associated tracking errors are relatively large, which can be seen from the time signal plots in Figure 5.13. This is due to the inadequate closed-loop tracking bandwidth of the fast axis, which is only 165Hz. Image artefacts due to poor tracking can be observed in the 100Hz closed-loop scans where the round features are smudged and elongated. With the implementation of the inversion feedforward technique, these artefacts are eliminated in the 100Hz scan, which improves the image quality substantially. Table 5.6 shows the maximum and RMS tracking errors for 90% of the scan range. With the inversion feedforward technique and the feedback control loops combined, the tracking errors of the 50Hz and 100Hz triangular references are significantly reduced by a factor of 1.8 and 8.2, respectively, compared to their open-loop counterparts.

5.7 Conclusion

The electrothermal displacement sensors are integrated in a 2-DoF electrostatically-actuated nanopositioner on a SOI MEMS chip. The stage, which functions as the scan table, has an area of $3\text{mm} \times 3\text{mm}$, and a maximum displacement of $16\mu\text{m}$ in both the x and y directions. The electrothermal sensor outputs are measured by a transimpedance amplifier circuit, with this output being utilised as a measurement by a controller in a feedback loop. The frequency response and static non-linearity of the device are characterised in order to implement the controller. The comparison of the scan results obtained by the AFM in open-loop, closed-loop and closed-loop with inversion configurations demonstrate the effectiveness of MEMS-based sensor integration in achieving high-quality AFM scans at high scan speeds.

The MEMS scanner stage equipped with electrothermal sensors is used as a feedback controlled system for generating images of the gold samples on the stage. This is a
major step toward producing an AFM on a single chip. The remaining challenge is to integrate the $z$-scanner in the same system that can work as a true on-chip AFM.
Figure 5.13: AFM scan results obtained at 10Hz, 50Hz and 100Hz, in open-loop, closed-loop and closed-loop with inversion feedforward. 3-D topography of the sample is plotted. For the time signal plots, the left-hand scale is for the fast-axis reference (blue) and displacement (red) signals of the nanopositioner. The right-hand scale is for the tracking errors (green). These are plotted in ($\mu$m Vs. msec).
6 Embedded Piezoresistive Displacement Sensor

This chapter reports the design and characterisation of a novel piezoresistive sensing technique for a 1-Degree-of-Freedom (1-DoF) MEMS nanopositioner. The MEMS device fabrication consists of a standard silicon on insulator (SOI) bulk micromachining process. The proposed technique exploits the piezoresistivity of the doped silicon used in suspension beams of the nanopositioner. Therefore, neither an external sensing structure is added to the system, nor is a customised step needed in the fabrication process, which reduces the design cost and complexity. The beams are designed to be connected to the stage with an angle, which results in opposite axial forces regardless of the direction of stage motion. The opposite piezoresistive variations are used for the differential sensing architecture. The differential architecture of the sensor beams improves the measurement resolution from 319nm to 1.5nm (3σ resolution), as well as linearity from 80 to 1 in terms of Pearson Correlation Coefficient (PCC), compared with a non-differential sensor. The frequency response of the 1-DoF nanopositioner is measured by the piezoresistive sensor as well as an optical microsystem analyser. The piezoresistive sensor follows the nanopositioner displacement up to 7kHz.
6.1 Introduction

The piezoresistivity of silicon, as discussed in Chapter 2, is a major sensing scheme that has been widely used for displacement, velocity, acceleration, force, pressure and deflection measurements. The physical concept and applications of silicon piezoresistivity are thoroughly investigated and reviewed [12]. The piezoresistivity of cantilever beam was first used for atomic force microscope (AFM) in [135]. Compared with other sensing schemes, piezoresistive transduction better suits AFM as it integrates with the AFM structure so that distinct external elements such as optical leverage systems and position sensitive detectors (PSD) are not needed. It is also needless of precise alignments and its readout circuitry is simple. In a piezoresistive based AFM the cantilever deflections resulting from the contacts between its sharp tip and a sample surface apply a stress that leads to changes in the beam resistance. Consequently, the sample topography can be extracted by the beam scanning the sample surface. The piezoresistivity of silicon is determined by local implantation of the impurity concentration in the cantilever. As illustrated in Figure 6.1, the beam is capable of detecting both the lateral and vertical deflections required for a sample topography formation. The lateral detection is achieved by sidewall ion implantation in the lateral deflection sensor. Other impurity doping methods used to create piezoresistive zones include diffusion and epitaxial growth. However, fewer added steps to the fabrication process is preferable, both in terms of production costs and possible damages to the device [12].

The piezoresistive cantilever beam sensitivity and noise have been extensively studied in [136], and an analytical model is proposed which specifically takes the non-uniform doping into consideration. An optimised design procedure is introduced for maximum sensitivity and minimum noise as a function of the bias voltage, diffusion length and cantilever dimensions [137]. Flexible polysilicon beams are also used in [16] to
Figure 6.1: The piezoresistive sensor implanted in an AFM cantilever beam (a) schematic representation (b) SEM image [12].
measure the displacement of a thermally actuated shuttle. The resistance of the beams that are separated from the actuation beams by their bonding pads, (See Figure 6.2), changes when they bend. On-chip force and position sensors based on piezoresistivity have also been used for the closed loop control of the Z-shaped thermal actuators, as illustrated in Figure 6.3 [17], where the sensor resolution is calculated based on the sensitivity, which is reported as a function of the electrical current and noise. Effects of the stray electromagnetic fields present inside the scanning electron microscope (SEM) chamber, stage vibration and electron-beam-induced heating are included in the sensor noise. It should be noted that conducting the experiment in a vacuum setup with fixed temperature rather than air does not cancel the thermal and flicker noise of the sensor resistors.
6.2 Inbuilt Piezoresistive Sensor for a 1-DoF Nanopositioner

A novel piezoresistive sensor is designed for a 1-DoF nanopositioner, which can also be used for similar mechanically suspended structures, such as microtweezers. This microactuator is implemented through a SOI MUMPs\(^1\) micromachining process [2]. The sensor is designed based on differential piezoresistive effects of suspension beams. Figure 6.4(a) illustrates the SEM image of the proposed device, in which the motion of the electrostatically actuated suspended stage is measured by the variations in the resistances of suspension beams. \(R_{p1}\) and \(R_{p2}\) represent the resistances of the straight suspension beams, which are also used as piezoresistors in the sensor. The schematic view of the proposed device is shown in Figure 6.4(b). The stage displacement along the \(x\) direction, caused by the electrostatic force of the comb drives, applies dominant longitudinal stresses along the piezoresistive beams. The beams are slightly angled with respect to the \(y\)-axis (about 0.86°) in order to develop opposite axial forces in the beams. Consequently, a differential sensing scheme is achieved, which has not been reported before. As the stage is actuated, one of the beams experiences compression, while the other is stretched. The two lower suspension beams are coated with a gold layer, which is part of the standard micromachining process. This electrically connects the piezoresistive sensors to the ground of the interface circuit. As illustrated in Figure 6.5, another pair of identical beams, the same size as the suspension beams but without any mechanical stress, are fabricated on the same chip in order to complete the Wheatstone bridge configuration. The sensor and actuator circuit schematic are shown in Figure 6.6, where \(V_b\), \(v_D\), \(V_a\) and \(v_a\) represent the sensor bias, sensor output, actuator bias and differential actuation

\(^1\)Multi-User MEMS Processes
Figure 6.4: Proposed 1-DoF nanopositioner with electrostatic actuator and embedded piezoresistive sensors: a) SEM image, b) schematic diagram.
voltages, respectively. The resistance values of the sensors and dummy resistors are reported in Table 6.2.1. Off-chip 100kΩ potentiometers are used to compensate for the resistance mismatches. The tolerances of the MEMS fabrication process is an instinct property of all devices fabricated in these processes and in this case it is elaborated on in the SOI-MUMPS design rules [2]. The nanopositioner stage is equipped with two sets of comb drive actuators to make the bidirectional motions along the x-axis. It was also shown in [18] that biasing both of the stage actuators
with a dc voltage avoids the inherent quadratic non-linearity of the comb drives. In order to prove this, assume that the electrostatic energy stored in a unipolar comb actuator, as illustrated in Figure 6.7(a), is expressed as:

$$ E = \frac{1}{2} (c_{11}v_1^2 + c_{22}v_2^2 + 2c_{12}v_1v_2) $$ \hspace{1cm} (6.2.1)

where $c_{ij}, (i, j = 1, 2, 3)$ represent the capacitance between conductors, i.e. combs i and j. The force generated by such a capacitor is derived as:

$$ F = \frac{\partial E}{\partial z} = c_n(v_1^2 + v_2^2 + 2v_1v_2), \hspace{1cm} (6.2.2) $$

$$ c_{ij} = c_n(z_0 + z), \hspace{0.5cm} c_n = \frac{N_n\varepsilon_0\varepsilon_r l_n z_0}{a} $$

where $\varepsilon_0$, $\varepsilon_r$, $l_n$, $z_0$, $a$ and $N_n$ are the permeability of vacuum, relative permeability of air, length, initial engagement of the comb fingers, gap between the comb fingers and number of the comb fingers, respectively; while $z$ and $v_i$ represent the displacement and electric potential of the comb i, respectively. As illustrated in Figure 6.7(b), for a bipolar comb actuator, given that:

$$ c_{11} = 2c_l x_0, c_{22} = c_{12} = c_l(x_0 - x), c_{33} = c_{13} = c_1(x_0 + x), c_{23} \approx 0, c_l = \frac{N_l\varepsilon_0\varepsilon_r l_t}{a}. \hspace{1cm} (6.2.3) $$

and assuming that the stored energy can be represented as:

$$ E = \frac{1}{2} (c_{11}v_1^2 + c_{22}v_2^2 + c_{33}v_3^2 + 2c_{12}v_1v_2 + 2c_{13}v_1v_3 + 2c_{23}v_2v_3) \hspace{1cm} (6.2.4) $$

<table>
<thead>
<tr>
<th>Sensor Resistance Value</th>
<th>$R_{p1}=535 , \Omega$</th>
<th>$R_{p2}=553 , \Omega$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dummy Resistor Value</td>
<td>$R_{pA1}=551 , \Omega$</td>
<td>$R_{pA2}=537 , \Omega$</td>
</tr>
</tbody>
</table>

Table 6.2.1: Resistance values of the MEMS device fabricated in SOI technology
the generated force is derived as:

\[
F = \frac{\partial E}{\partial z} = \frac{1}{2} \varepsilon_0 (v_2^2 + v_3^2 - 2v_1v_2 + 2v_1v_3)
\]  

Choosing a fixed value for \(v_1\) and equal magnitude but opposite polarity for \(v_2\) and \(v_3\), cancels the second order terms in Equation 6.2.5 as:

\[
F = 2\varepsilon_0 v_1 v_2
\]  

which represents the electrostatic force as a linear function of applied voltage.

The sensor consists of a Wheatstone bridge and a differential amplifier. Two elements of the bridge are active (mechanically stressed beams) and the other two are dummy resistors with similar passive resistance values. A fixed voltage is switched at the amplifier input in order to characterise the device single-ended (non-differential) operation and compare it with the differential mode. Due to the mismatch of the resistors fabricated in MEMS processes, potentiometers with small values are used for initial zero adjustment.
Figure 6.8: The piezoresistive sensor static transfer function for differential and single-ended (non-differential) operation modes.

### 6.3 Sensor Characterisation

The stage displacements are measured by an optical microsystem analyser (MSA) in order to characterise the integrated piezoresistive sensor. For static characterisation, a slowly varying (0.5Hz) triangular signal \( v_a \), biased with \( V_q = 30V \), is applied to the actuator and the sensor output voltage is measured. The amplitude of the triangular signal is set to move the stage up to the actuator maximum limits, which are around ±7\( \mu \)m in this case. Applying a 4V bias voltage to the Wheatstone bridge, the sensor output voltage versus stage displacement (measured by the MSA) transfer characteristics are obtained for rising and falling half-periods of triangular actuation input, as shown in Figure 6.8.

The transfer curves demonstrate a linearly varying output voltage for the differential
sensor, with respect to the stage displacements in either positive or negative directions, without any hysteresis. To compare the differential and non-differential sensor performances, a fixed voltage is applied to one of the amplifier inputs to null the corresponding sensor variations. Hence, the resistive variation of only one of the sensors is amplified. Switching power supplies are replaced with batteries at this stage to avoid different switching noises at the amplifier inputs, which otherwise can significantly corrupt the desired signal at the sensor output. The transfer curves are redrawn for the differential and non-differential sensors and $V_b = 6.3V$ (available battery cells). The results show that the differential sensors are more sensitive than the non-differential ones. The linearity of the two cases is quantified by the Pearson correlation coefficient, as defined in Chapter 2:

$$NL = 10^4(1 - |\rho|)$$  \hspace{1cm} (6.3.1)

where magnitude of the Pearson product-moment correlation coefficient $\rho$ varies between 0 and 1, with 0 for the highest non-linearity and 1 for a straight line. According to the measured values reported in Figure 6.8, the differential sensing clearly improves the linearity of the sensor transfer characteristic.

The sensor frequency response to the actuation input is obtained and compared with the displacement frequency response measured by the optical MSA in order to characterise the system dynamics. As illustrated in Figure 6.9, the sensor follows the stage displacement as the frequency increases up to 7kHz, which is well above the stage mechanical bandwidth that is below its resonance frequency (3kHz). The necessity of this specification was discussed in Chapter 2. At higher frequencies the sensor response starts to deviate from the stage displacement. This is possibly due to the common issue known as feed-through in resonators with piezoresistive sensors [138]. The non-zero resistance of the gold-coated suspension beams and lower
Figure 6.9: The sensor and stage displacement frequency response to actuation input.

The impedance of the capacitive comb drives are amongst the main reasons. To prove this experimentally, the bias voltage of the actuators are switched off to stop their motion. The sensor frequency response for a zero displacement is recorded, as shown in Figure 6.9. Therefore, the sensor response beyond its bandwidth is mainly due to electrical coupling from the actuator to the sensor. The sensor output power spectral density is measured in three different conditions as illustrated in Figure 6.10. It is clearly shown that the differential sensing removes the common mode interferences such as ambient temperature variations and regulated power supply effects, which are appeared as noise in the non-differential sensor output. The switching power supply induces noise in the differential sensor output, which is significantly reduced by using battery cells. However, unequal supply voltages, that leads to different noise levels generated at the differential sensor output, avoid an appropriate comparison. It should be noted that as the available battery voltages of 6.3V is more than the 4V supply voltage, even more improvement is expected for the differential sensor using 4V battery cells.
In addition, the time domain recordings of the sensor outputs in differential and non-differential modes are measured for zero actuation input in order to characterise the sensor noise performance (see Figure 6.11). A switching power supply is used to generate the data in Figure 6.11. Clearly, the differential sensing improves the undesired common mode interferences but, theoretically, the thermal and flicker noises are random processes and they are not included in this improvement. The noise performance and resolution of the sensor are summarised in Table 6.3.1 for both modes. The sensor resolution is measured by different sensor bias voltage only while the actuation voltage is switched off, on identical frequency bandwidth. The differential sensing has improved the resolution significantly.
Figure 6.11: Sensor time-domain outputs for differential and nondifferential modes.

Table 6.3.1: Sensitivity and resolution of the proposed sensor.

<table>
<thead>
<tr>
<th>Sensor Settings</th>
<th>Sensitivity (mV/µm)</th>
<th>3σ-Resolution (nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Differential-6.3V battery</td>
<td>172</td>
<td>1.5</td>
</tr>
<tr>
<td>Nondifferential-6.3V battery</td>
<td>89</td>
<td>319</td>
</tr>
<tr>
<td>Differential-4V Switching Supply</td>
<td>88</td>
<td>3.9</td>
</tr>
</tbody>
</table>

6.4 Conclusion

An embedded displacement sensing system was presented for nanopositioning purposes. The proposed sensor is implemented for a 1-DoF MEMS nanopositioner in a standard SOI process without any customised fabrication step. This is in comparison with similar piezoresistive sensing systems such as [12, 139], which use selectively doped regions in silicon to implement piezoresistor. This work exploits the piezoresistivity of the suspension beams in a differential architecture in order to measure the stage displacements. Another unique feature of the proposed sensor is that it is designed based on axial forces orthogonal to the direction of displacement, allowing larger travel range for the stage, compared to a few similar works whose displacement is limited by using axial forces along the displacement direction. The sensor bandwidth is limited to 7kHz by the electrical coupling from the actuation input, which is well above the mechanical bandwidth of the stage. Substantial improvements in sensor resolution and linearity are achieved by the differential sensing.
This thesis aimed to make new contributions in the design and implementation of displacement sensors in microelectromechanical systems (MEMS). The miniaturised nanopositioners, such as those in the scanning stages of the atomic force microscope (AFM), are among the most prominent applications for these sensors. Therefore, a conceptual introduction to the nanopositioning systems is presented in Chapter 1. The importance of miniaturisation, as well as the role of the sensor in the nanopositioner servomechanism, is described afterwards. The MEMS devices presented in this thesis are all fabricated in the silicon on insulator (SOI) process. Hence, a brief introduction to the corresponding SOI process is also included in Chapter 1.

The research objective is to design a high precision and high accuracy sensing scheme for nanopositioning applications. The displacement resolution of these devices is strongly dependent on the sensor specifications. The performance criteria of the displacement sensors, such as noise, sensitivity, resolution, linearity and bandwidth, are elaborated on with a special focus on noise in Chapter 2 of the thesis. The sensor performance is dependent on the device sensing mechanism, as well as the readout circuitry. Hence, the displacement sensing techniques for high precision applications are reviewed first.

The feasibility of integrating piezoresistive and electrothermal sensors in miniaturised nanopositioners is higher due to their smaller footprint on the MEMS chip. The
electrothermal sensors are investigated with higher priority, throughout the work, due to their relatively simple structure and the higher reliability of these sensors, as well as their higher sensitivity and dynamic ranges. In addition, the readout circuits are discussed and alternative methods, such as frequency based interfaces, are introduced. Ring and relaxation oscillators are used as the interface circuits for sensing applications such as in capacitive and resistive sensors. However, the relaxation oscillator is most commonly used for resistive sensors that offer a wide resistivity variation range, which is not the case in electrothermal sensors. The electrothermal sensor excitation methods, such as constant current and constant voltage, are also discussed in Chapter 2. A review of the thesis achievements is presented in the following sections.

7.1 Design and Implementation of a Frequency-Based Readout Circuit

Frequency-based measurements can offer a higher sensitivity compared with the conventional methods. Chapter 3 includes the design and implementation of a new frequency based readout circuit for an electrothermal sensor. In the presented circuit, the sensor’s resistive variation translates to the frequency variation of a ring oscillator, which is coupled to the resistive sensor by means of a ratiometric interface circuit. The 10Ω resistive variation is converted to the ring oscillator output frequency variation range of 350–550kHz. A phase locked loop (PLL) and a frequency to voltage converter are implemented as the two distinct frequency demodulation techniques in order to produce a voltage signal suitable for the closed-loop control. The experiments are carried out on a 1-Degree-of-Freedom (1-DoF) nanopositioner equipped with electrothermal sensors and actuator. The inherent non-linearity
of the ring oscillator voltage to frequency conversion characteristic mitigates the
quadratic nonlinearity of the actuator. Consequently, a linear input-output transfer
characteristic is achieved, which can potentially eliminate the need for an additional
linearisation if the sensor is used in a feedback loop.

7.2 Electrothermal Sensor Resolution Enhancement

The high sensitivity of the electrothermal sensor is degraded by the noise profile of
the silicon heaters, which consists of flicker noise and thermal noise. Three different
circuit and device level approaches are proposed to investigate and improve the
sensors noise performance in Chapter 4. Firstly, a high frequency excitation method
is introduced to reduce the flicker noise. In this method an ac voltage source is used
to drive the sensor instead of the commonly used dc sources. The experiments on
a 1-DoF MEMS nanopositioner confirm the proposed method. A 20dB noise floor
reduction is measured in the low frequency range, which translates to a resolution
improvement from 0.2nm/√Hz to approximately 0.1nm/√Hz.

The sensitivity of the thermal sensor increases as a function of its temperature.
However, higher temperature escalates the thermal noise level. In the second approach
a multiple sensor 1-DoF MEMS nanopositioner is designed and implemented to
alleviate the thermal noise issue. This idea is the realisation of the averaging theory,
according to which, combining multiple signal sources that are contaminated with
uncorrelated noise improves the signal to noise ratio (SNR). A 4dB SNR enhancement,
which is equal to a resolution enhancement from 0.3nm/√Hz to 0.15nm/√Hz, is
achieved by exploiting the averaging theory for three sensors that are contaminated
with uncorrelated thermal noise. During the experiments the electrical power injected
to each sensor was kept constant to guarantee equal output signal.
Furthermore, the impact of the excitation methods of the electrothermal sensors, including constant current (CC) and constant voltage (CV), are investigated by analytic models and experimental approaches. The model is a modified version of a previously proposed work. A new circuit for CC mode is designed to guarantee equal noise contribution from the circuit elements. The measurements show that although the CC mode leads to a higher output signal level, it increases the noise floor as well. Hence, the maximum achievable SNR and displacement resolution by the two modes is almost identical, which is in close agreement with the analytical result.

## 7.3 Integration of the Electrothermal Sensors to a 2-DoF Nanopositioner

The integration of electrothermal sensors to the electrostatic comb actuators of a 2-DoF nanopositioner stage is reported in Chapter 5. The device is implemented using a standard SOI MEMS fabrication process and successfully tested for operating as the scanner stage of the AFM. An external cantilever of a commercially available AFM is used for scanning the sample surface. Maximum displacement of the $3 \times 3 \text{mm}^2$ stage is $16\mu\text{m}$ in the $x$ and $y$ directions. The electrothermal sensor outputs are measured by a transimpedance amplifier circuit, with this output being utilised as a measurement by a controller in a feedback loop. The sensitivity of the transimpedance amplifier and Wheatstone bridge circuits are analysed, and the former is proven to be more sensitive by circuit analysis and experiments. The sensor frequency response and linearity are characterised for the purpose of controller design. The feedback controlled nanopositioner is tested successfully by producing images of the gold features on the 2-DoF MEMS scanner. The image quality enhancement due to the feedback controller is obviously recognisable.
7.4 Integration of the Piezoresistive Sensors in a 1-DoF Nanopositioner

A new displacement sensing architecture, based on the piezoresistivity of silicon, is embedded in a 1-DoF nanopositioner. The suspension beams are designed to be connected to the stage with an angle such that stage movements produce opposite axial forces applied to the beam. This results in differential piezoresistivity variations in the beams. The proposed differential architecture adds no extra hardware by exploiting the piezoresistivity of the suspension beams. It has been fabricated in a standard SOI MEMS fabrication process, without any customised fabrication step, which can reduce the fabrication costs and complexity. The sensor bandwidth is expected to be higher than what has been measured on this device, with the possible reason being the signal feed-through by the actuator.

7.5 Future Work

Despite the high sensitivity of the electrothermal sensor in the displacement measurements achieved by the ring oscillator in Chapter 3, the noise floor of the ring oscillator has adverse impacts on the final resolution. An analytical model that considers the noise contribution of each unit of the sensor in the final displacement resolution can give a clearer perspective of the ring oscillator benefits and drawbacks. The high frequency excitation technique and multiple sensor displacement measurement are implemented distinctly and are addressed in Chapter 4. The combination of the two methods on a system, in order to achieve higher specifications, requires alternative circuit design considerations. For example, a high current power supply is needed to drive multiple sensors with a high frequency excitation signal.
The number of sensors can be increased to achieve a higher resolution in the averaging technique proposed for the thermal noise mitigation. However, there is a practical limit, which is set by the added noise sources, such as the amplifier noise, in addition to the available space and power consumption. An analytical modelling approach is helpful to optimise the number of sensors in a way that leads to the minimum noise level.

The ultimate goal of the 2-DoF nanopositioner stage is to integrate all of the AFM components on a chip. One further step is to embed the vertical scanner in the same MEMS chip, which is impossible with the currently available SOI processes. However, this might be feasible with two chips aligned on top of each other.

The piezoresistive sensors, proposed in Chapter 6, can be integrated in a 2-DoF nanopositioner. The cross-coupling of the x and y actuators should be mitigated on the sensor. Also the wide bandwidth of the piezoresistive sensor in nanopositioning application is adversely affected by the feed-through from the electrostatic comb drives. Differential actuation and sensing, that has been proposed for other resonators, can reduce the feed-through in nanopositioners.
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