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Ultra-Low-Power Interface Circuits for Gas Sensors with a High Resistance Range

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Abstract

As technology has developed, an ever increasing number of gases are being used in industrial processes, transportation and household applications. This trend, has brought about the need for precise measurements of gas concentrations to control processes, detect leaks and monitor the environment. In contrast with mainstream gas sensing technologies, novel chemiresistive gas sensors allow for the integration of tenths of devices in the same package with low power consumption and submillimetre size. Using well-known microfabrication techniques, like lithography and inkjet printing, the electrodes can be miniaturized and the sensing layer, selected from a wide range of chemical compounds, printed on top of them. This new paradigm enables applications, like remote environmental sensing systems or disposable gas sensors, where previous technologies were unable to meet specifications in terms of power consumption, size and economic cost. During the last three decades, research on new gas sensing materials has taken place with remarkable results in two areas. Firstly, conductive polymers like polypyrrole (PPy), polyaniline (PANI), polyphenylene sulfide and poly(3,4-ethylenedioxythiophene), and secondly, in graphene and its derivatives including graphene oxide (rGO). Researchers have experimented with different formulations, dopings and morphologies and implemented sensors for a wide range of gases, including CO, NO₂, CO₂, NH₃, O₂, N₂ and volatile organic compounds (VOCs).

Even though the gas sensing properties of these materials are well understood, a widespread effort to investigate its electrical characteristics for the integration in miniature gas sensing systems is still lacking. We begin this dissertation with an in-depth study of the current-to-voltage curves, the noise power spectral density (PSD) and the response time of PPy, PANI

and graphene sensors. We found that these types of sensors have a non-linear current-to-voltage relation and flicker noise is predominant. Additionally the level of noise increases with the bias voltage, which leads to a modified flicker-noise mathematical model that has been validated experimentally.

After the sensor characterization we review the state-of-the-art of read-out circuits (ROICs) for resistive sensors. Given the multi-sensor approach, flexibility is paramount, which is best served by relaxation-oscillator-based ROICs. After this review, we designed and fabricated a proof-of-concept ROIC in a bulk 180 nm CMOS process. It can measure sensors from 1 k Ω up to 33 M Ω , with a minimum SNR of 57 dB. It also offers a wide range of input configurations in terms of resistance and bias voltages to select the optimal operating conditions for each sensor, as well as to accommodate sensors with dissimilar baseline resistances. The circuit was tested in a real application for the measurement of NH_3 and NO_2 using two different types of sensors validating the design objectives and the capability of the read-out circuit for system-on-chip integration.

Once the ROIC architecture was defined and validated with the proof-of-concept circuit, we studied how to improve the power consumption profile of relaxation-oscillator-based ROICs, as the energy required to quantify the sensor's resistance increases proportionally with the resistance value. This presents a dilemma, as increased miniaturization will lead to smaller sensors with higher resistances. Two adaptive biasing techniques for comparators are presented to achieve ultra-low-power and high dynamic range in ROICs for resistive gas sensors. The first technique is based on an on-demand biasing that generates a bias current of up to 3 μA when the differential input voltage is close to zero, while minimizing the bias current when a comparison is not imminent. The second technique takes advantage of the correlation between sensor current and delay requirements to bias the comparator with the appropriate current for each situation. These techniques were applied to two readout circuits, demonstrating a reduction of the energy per measurement to less than a fifth for the largest input resistance in comparison to fixed biasing without affecting the SNR. The input range is above five decades in both cases, from 2.4 k Ω up to 1 G Ω and from 1 k Ω up to 270 M Ω respectively. The signal-to-noise ratio ranges from 55 dB up to 70 dB which provides enough margin to interface a wide range of gas sensors. Lastly, one of the designs was successfully employed to measure varying concentrations of ammonia with a polypyrrole gas sensor.

After reviewing the ROICs, it was seen that to optimize performance, not only the ROIC must be adapted to the sensors characteristics but the performance and capabilities of the interface microcontroller must also be taken into account to assess how the digitization process must be done. In our analysis, we compared how the digital implementation in a 180 nm SOI process performs against a modern low-power microcontroller (STM32U575 from ST Microelectronics) and found that it is best to reuse as many digital blocks (counter, PLL, oscillator) from the microcontroller as possible. This is for two reasons, firstly, it is very energy-expensive to toggle the input/output pads of the integrated circuits, therefore, high-frequency signals should be kept inside the integrated circuit. Secondly, the microcontroller comes equipped with these blocks as peripherals, and some, like the oscillator are already running when a measurement is taking place. Furthermore, the studied microcontroller is fabricated in a 40 nm CMOS process, which is more efficient for digitally-intensive processing. In all, it was seen that the energy advantage of integrating the digital blocks in the same ASIC as the ROIC is at best marginal. The microcontroller timer peripheral consumes 270 nW/MHz, while bare a counter implemented in the ASIC with a 5 V digital library (at 1.8 V) requires 253.3 nW/MHz without any of the functionalities available in the microcontroller's timer.

To close this dissertation, we do a comparison of the performance of the proposed ROICs to the state-of-the-art and offer some perspective and ideas on how this work should proceed.

Publications

Directly related to the thesis

Journal Papers

[JP1] R. Puyol, S. Pétré, Y. Danlée, T. Walewyns, L. A. Francis and D. Flandre, "An Ultra-Low-Power Read-Out Circuit for Interfacing Novel Gas Sensors Matrices," in IEEE Sensors Journal, vol. 22, no. 10, pp. 9521-9533, 15 May. 15, 2022.

[JP2] R. Puyol, T. Walewyns, L. A. Francis and D. Flandre, "Design of Ultralow-Power Sensor Readout Circuits Through Adaptive Biasing for Gas Monitoring With Chemiresistive Sensors," in IEEE Sensors Journal, vol. 23, no. 22, pp. 27468-27477, 15 Nov.15, 2023

Conference Papers

[CF1] R. Puyol, S. Pétré, Y. Danlée, T. Walewyns, L. A. Francis, D. Flandre, " Design Considerations of Ultra-Low-Power Polymer Gas Microsensors Based on Noise Analysis", nanoFIS, Graz, Austria, 2020, 56, 19.

[CF2] R. Puyol, Y. Molle, S. Pétré, T. Walewyns, L. A. Francis and D. Flandre, "A Practical Approach for the Evaluation of Noise in Oscillator-Based Resistive Sensor Interfaces," 2020 IEEE SENSORS, Rotterdam, Netherlands, 2020, pp. 1-4.

Unrelated to the thesis

[UR1] A. Arnaud, R. Puyol, D. Hardy, M. Miguez and J. Gak, "Bulk Linearization Techniques," 2019 IEEE International Symposium on Circuits and Systems (ISCAS), Sapporo, Japan, 2019, pp. 1-5

[UR2] A. Arnaud, M. Miguez, J. Gak, R. Puyol, R. Garcia-Ramirez, E. Solera-Bolanos, R. Castro-Gonzalez, R. Molina-Robles, A. Chacon-Rodriguez, R. Rimolo-Donadio, "A RISC-V Based Medical Implantable SoC for High Voltage and Current Tissue Stimulus," 2020 IEEE 11th Latin American Symposium on Circuits and Systems (LASCAS), San Jose, Costa Rica, 2020, pp. 1-4.

List of Acronyms

ADC	Analog to Digital Converter
AFE	Analog Front-End
ASIC	Application-Specific Integrated Circuit
BDC	Bridge-to-Digital Converter
CMOS	Complementary Metal-Oxide-Semiconductor
CCO	Current-Controlled Oscillator
CPU	Central Processing Unit
DAC	Digital to Analog Converter
DR	Dynamic Range
ENOB	Effective Number of Bits
EPC	Energy per Conversion
FOM	Figure of Merit
FPGA	Field-Programmable Gate Array
IDAC	Current Digital to Analog Converter
IDE	Interdigitated Electrodes
LoD	Limit of Detection
LSB	Least Significant Bit

MCU	Microcontroller Unit
MEMS	Micro-Electromechanical System
MIM	Metal-Insulator-Metal
MOx	Metal-Oxide Gas Sensor
MSB	Most Significant Bit
OTA	Operational Transconductance Amplifier
PANI	Polyaniline
PCB	Printed Circuit Board
PFD	Phase Frequency Detector
PFM	Pulse-Frequency Modulation
PLL	Phase-Locked Loop
PPy	Polypyrrole
PSD	Power Spectral Density
PVT	Process-Voltage-Temperature
PWM	Pulse-Width Modulation
RDC	Resistor-to-Digital Converter
rGO	Reduced Graphene Oxide
ROIC	Read-Out Integrated Circuit
SEM	Scanning Electron Microscope
SNR	Signal-to-Noise Ratio
UV	Ultraviolet Light
VCO	Voltage-Controlled Oscillator
VOC	Volatile Organic Compound
ZXB	Zero-Crossing Bias Current Generator

1

Introduction

GAS detection techniques date back to the late 19th century when animals like canaries or mice were used for the detection of carbon monoxide in coal mines [2]. By the turn of the 20th century, the need for detection and recognition of gases grew even more with the expansion of the oil and gas industry. Since then, their applications have broadened to include environmental control, health monitoring and machinery health management, among others. This development was also accompanied by progress in the sensor's technology. Starting with catalytic sensors in the 1920s, joined by metal-oxide (MOx) in the 1960s [3], and recently by more sophisticated devices like compact nondispersive infrared sensors (NDIR) and chemiresistive sensors made with conductive polymers or graphene [4], [5], [6], Fig. 1.1.

Each technology has been subsequently developed to address, at least in part, some of the shortcomings of previous generations of gas sensors. For this reason, several technologies coexist and the choice is determined by the application's requirements. To better understand each technology, next, we briefly describe the principle of operation of each type of sensor, and a qualitative comparison is presented in Table 1.1, where we list the technologies that have achieved a commercial readiness level as reviewed from publications like [6] and the main manufacturers product range (SGX

Sensortech S.A. , Figaro Engineering Inc., Honeywell International Inc., Draegerwerk AG & Co. KGaA and Kemet Electronics Corporation).

Electrochemical (EC): They are electrochemical fuel cell devices of 2, 3 or 4 electrodes, in which oxidation-reduction reactions take place on the surface of the working electrode that generates a small current proportional to the exposed gas concentration. Target gases: CO, Ethylene, Cl₂, H₂S, NO, NO₂, O₂, O₃, SO₂, Ethylene Oxide, NH₃, ClO₂, H₂, Hydrogen Cyanide, Phosphine and Acrylonitrile.

Metal-Oxide (MOx): The electrical resistance of a piece of heated semiconductor material increases as oxygen from the atmosphere is adsorbed on the surface. When a reducing gas is present, the density of oxygen atoms in the sensor surface is reduced and the resistance decreases. Typically made with semiconductors like tin oxide (SnO), zinc oxide (ZnO), or titanium oxide (TiO). These materials are most sensitive at several hundred degrees celsius, so a heater is used to increase the temperature of the sensing element. Target gases: VOCs, CO, NO₂, NH₃, H₂, Ethanol, CH₄, C₃H₈, C₄H₁₀, Isobutane and Refrigerant gases (R-32, R-1234yf, R-1234ze, R-454B, R-404a, R-290, R-410a, R-22, R-134a).

Non-Dispersive Infrared (NDIR): They contain three elements: an emitter, a filter and a detector. The gas under inspection is illuminated with the filtered infrared light and the absorbed spectrum is measured by the detector. Then, a digital signal processor quantifies the absorbed wavelength and the concentration of the gas. Target gases: CO₂, CH₄, C₃H₈, C₄H₁₀, Acetylene and Refrigerant gases (R-32).

Catalytic (CA), Beads and MEMS: An electrical resistance difference is detected by comparing the resistance of two heating elements (e.g. platinum coils), one exposed to pure air and the other to the ambient. In the presence of flammable gases, the heat generated by its combustion increases the resistance of the coil exposed to the gas. Target gases: Combustible gases (H₂, C₄H₁₀, CH₄, C₃H₈), NH₃ and N-Methyl-2-pyrrolidone.

Thermal Conductivity (TC): They output a signal correlated to the heat transfer rate of the sensed gas which is measured by generating a temperature difference in a chamber. As with catalytic sensors, a sample of a reference gas is needed to compare to the measured value. Their selectivity is very low for a single temperature point. Taking advantage of the uncorrelated response of thermal conductivity to temperature of different gases, some devices allow to make two or more measurements at points several degrees Celsius apart to increase selectivity. Target gases: Flammable gases, N₂, O₂, Ar, CO₂, CH₄, C₂H₆, C₃H₈, CO, Ethylene, Acetylene, H₂, He

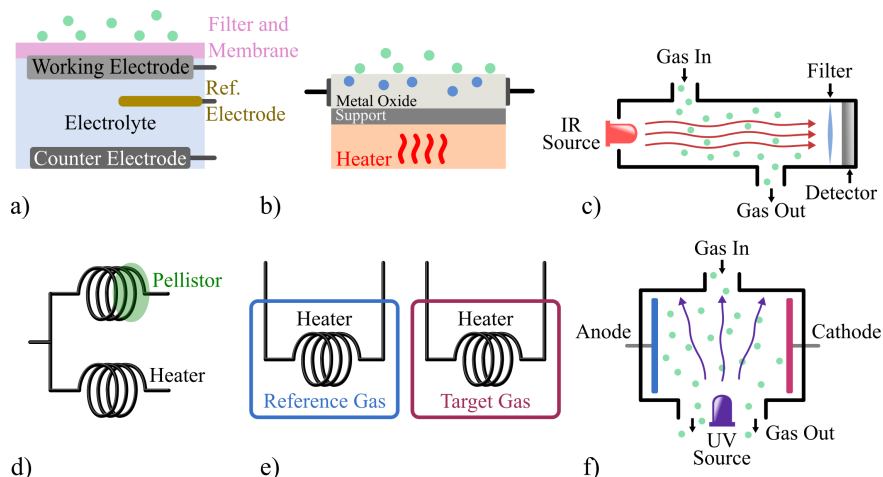


Fig. 1.1 Mature gas sensing technologies: a) Electrochemical, b) MOx, c) NDIR, d) Catalytic, e) Thermal Conductivity, f) Photoionization.

and NH_3 .

Photoionization (PI): They are used for the detection of VOCs by ionizing the target gas in an ion chamber with UV light. The ions created in the chamber are accelerated by an electric field to generate a current that is proportional to the gas concentration. Even though they are capable of detecting VOCs in the ppb range, they make a very small portion of the gas sensors market due to their high cost. The generated current is very small (at most nA [6]), therefore ultra-low-noise amplification and signal processing are required. Target gases: VOCs.

1.0.1 Requirements of State-of-the-Art Sensing Systems

Transistors in integrated circuits have been made continuously smaller since they were first introduced, and for several decades the complexity of integrated circuits has followed an exponential trend. The reduction in transistor size coupled with improvements in materials and production processes, has delivered devices with higher operating frequency and lower switching energy [7]. Such devices have enabled ultra-low-power microcontrollers as well as radio-frequency transceivers with ever-improving noise figures and modulation techniques like chirp spread spectrum or ultra narrowband. These modulation technologies, have

Table 1.1 Sensor Technologies Overview.

Technology	Power	Advantages	Disadvantages
EC	nW-uW	High sensitivity Stability	Short lifespan Slow response
MOx	mW	High sensitivity Fast response time Long lifespan Low cost	Heating High power Low selectivity
IR	uW-mW	Good antitoxicity High sensitivity Long lifespan	Complex structure High cost
CA	mW	High accuracy Fast response Long lifespan	Only flammable gases Low selectivity Poisoning risk
TC	mW	Wide range of gases	Resistance variations
PI	mW	Fast response High sensitivity	High cost

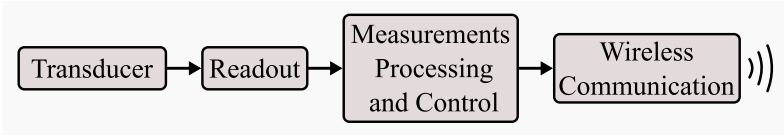


Fig. 1.2 Signal path in a conceptual sensing node.

been employed in low-power wireless communications technologies like LoRa, Sigfox and NB-IoT, which can be used to communicate with remote sensors tenths of kilometres away [8]. The power efficiency of LoRa and Sigfox has been compared to other wireless communication protocol and found to require a similar energy per transmitted useful bit to: Bluetooth low energy, TSCH, IEEE 802.15.4 PSM or IEEE 802.11ah (Wi-Fi HaLow) when the application throughput is low, e.g. 100 Bytes per day but can reach much longer distances, kilometres instead of hundred of meters, at most with the other technologies [9].

These technological developments have been applied to sensing systems for physical quantities that can operate autonomously for several years, with low power, low cost and low (or none) maintenance. The basic components of such systems are depicted in Fig. 1.2. Part of the control

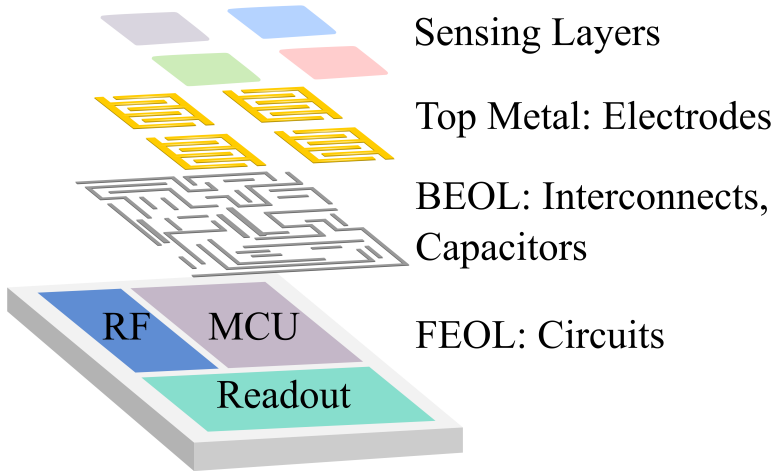


Fig. 1.3 Concept of an integrated gas sensing system including: sensors, ROIC, microcontroller and communications.

and communications block are standard across different sensing systems as they are based on commercially available technologies. The differences are mostly related to the firmware configuration depending on the network architecture, the payload size and the transfer rate. On the other hand, the sensors and the readout circuit (ROIC) are specific for each application and to maximize the life of the battery it is necessary to keep their power consumption low in comparison to the microcontroller and communication circuits.

None of the traditional gas sensors technologies has been able to offer at the same time good selectivity, long lifespan at low cost while having a power consumption compatible with ultra-low-power wireless sensing systems, which has prevented the extensive deployment of remote gas sensing nodes. The demand of which, is expected to grow as new applications arise, the most notable are:

- Environmental monitoring, indoor and outdoor,
- Combustion engines exhaust and cabin air quality monitoring in the automotive and aerospace industry,
- Gases leakage detection in batteries,

- Automotive and residential hydrogen storage,
- HVAC and refrigeration,
- Enhanced industrial safety and process control.

Photoionization and NDIR sensors require complex electronics to illuminate the gas with UV or IR light and then process the measured signal. Electrochemical sensors, on the other hand, work by exploiting redox reactions, therefore the electrodes wear-out with time requiring frequent calibrations and having a short lifespan of no more than two years for commercial sensors. MOx sensors are poorly selective and the sensing mechanism is temperature dependent. They are most sensitive when the semiconductor is at several hundred degrees, requiring a power-hungry heating element, although MEMS microheaters have shown great potential to reduce the power consumption [10]. Catalytic and thermal conductivity sensors are also power consuming, as the sensing gas has to be heated, again to several hundred degrees Celsius. With regards to the sensor composition, noble metals are frequently used for the electrodes in electrochemical sensors and for the heater in MOx and catalytic sensors [11], resulting in costly devices. In terms of volume, NDIR and MOx have achieved the smallest form factor as most of the sensor is built around traditional CMOS or MEMS processes with the addition of the infrared emitter, receptor or the semiconductor oxide element.

Several efforts have been carried out to obtain smaller sensors with lower power consumption and cost [12], [13], [14]. Chemiresistive sensors made with conductive polymers or graphene have shown promising results in this aspect due to the ease of fabrication and the abundance of sensing layer formulations that have been developed in the last decades [15], [16], [4], [17]. To ease the integration with traditional CMOS processes, the use of planar interdigitated electrodes (IDE) is preferred as they can be built using only the top metal layer of an integrated circuit and the remaining metallization can be used for the interconnection of other circuits of the system, like the ROIC, Fig. 1.3. For the top metallization, pure aluminum or alloyed with small quantities of copper (e.g. 0.5 %) is sometimes used, as in the case of X-FAB XT018 and UMC RFCMOS 180 nm processes. Some foundries also offer gold or platinum plating of the top layers, which have been used for making electrodes in multiple applications due to their resistance to corrosion as they are noble metals.

Inspired by the human nose which contains nearly 400 different olfactory receptors that simultaneously contribute to produce different smells, researchers have tried to improve the performance of gas sensors by using arrays of dissimilar sensors in parallel [18]. For this approach, chemiresistive sensors are good candidates as they are easy to miniaturize, suitable for many types of sensing layer materials and capable of integrating the readout electronics in the same chip. The chemistry and dimensions can be tuned to achieve very low area and power consumption; hence tens of sensors can be integrated in a single chip with an area of a few square millimeters as demonstrated by Le Maout et al. [19], Ge et al. [20] and Simonenko et al. [21]. On the other hand, such sensors present a challenge for the readout integrated circuit (ROIC) as the baseline resistance of each sensor can be greatly different, spanning several decades [4], [5].

1.1 Research Objectives of The Thesis

While chemiresistive sensors have the potential to address some of the limitations of traditional gas sensors and their chemistry has been widely researched, there is still a lack of knowledge about their electrical performance and characteristics. More specifically, countless formulations of conductive polymers and doping's exist, but the available literature focuses mainly on the sensitivity and selectivity of the sensors to gases, whereas electrical models to be used in circuit design are scarce. At the same time, the characterization of current-to-voltage (I-V) curves, noise, SNR and power consumption can guide the evaluation and selection of the appropriate ROIC architecture and optimize the design parameters. Therefore, the electrical modelling of chemiresistive gas sensors is one of the main objectives of our work. On top of this, the ROIC needs to be flexible enough to interface different sensor, in a multi-sensor system without a cumbersome reconfiguration. Resistive sensors are ubiquitous, therefore several different ROIC designs exist. On the other hand, none of them has been developed specifically for chemiresistive sensor arrays therefore further research is needed in this aspect. With this in mind, this thesis will try to answer the following questions:

1.1.1 What are the electrical characteristics of chemiresistive gas sensors of relevance to electronic interfaces?

The sensor is the core of a sensing system, therefore its electrical characteristic must be known to design a system that is most efficient and provides the suitable operating conditions to achieve the lowest measurement uncertainty. In chapter 2, we will study the I-V curves, the intrinsic electronic noise, the SNR and the sensitivity for several examples of gas sensors. Additionally, we will provide a simple model for the noise PSD. The studied chemistries are: Polypyrrole (PPy), Polyaniline (PANI) and graphene.

1.1.2 What ROIC architecture is most suitable for a multi-sensor system?

Resistive sensors are available for other physical quantities besides gases, like: pressure, light, temperature and weight. Therefore, many architectures for ROICs have been proposed which can be repurposed for gas sensing with chemiresistive sensors. In chapter 3, we will review the most relevant publications and with the trade-offs weighted, explain why relaxation-oscillator-based ROICs are the best candidates in terms of input range while keeping power consumption low. A proof-of-concept application-specific integrated circuit (ASIC) will be presented.

1.1.3 What are the implications in terms of power consumption and SNR for a relaxation-oscillator-based ROICs interfacing chemiresistive gas sensors?

Relaxations-oscillator-based ROICs are time based, therefore the measurement time is highly dependent on the sensor resistance. Consequently, this ROICs have a unique power consumption profile which can span several decades. In chapter 4, we model the power consumption and propose two adaptive biasing techniques to improve the efficiency. In chapter 5, we introduce some additional considerations regarding the power consumption of auxiliary circuits and some metrics from a system point of view. Additionally, in chapters 2 and 3 we study how the SNR of chemiresistive sensors and the ROIC changes with the bias voltages.

This thesis is structured as indicated in Fig. 1.4. Chapter 2 deals with the sensor characterization, followed by the ROIC study and optimization in chapters 3 and 4. In chapter 5 the period quantization methods are studied and compared. Finally in chapter 6 we offer some conclusions and perspectives. There are two annexes, in the first, the challenges encountered when fabricating chemiresistive gas sensors with aluminium electrodes are

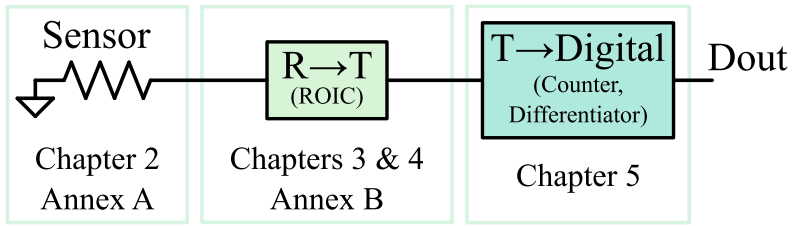


Fig. 1.4 Thesis structure and chapters in relation to the studied components of the gas sensing system .

explained and in the second we include additional measurements of the linearity of the ROICs presented in chapter 4.

2

Characterization and Modeling of Chemiresistive Gas Sensors

Abstract: This chapter explores the electrical characteristics of chemiresistive gas sensors made with polypyrrole, polyaniline and graphene in terms of: I-V curves, noise, signal-to-noise ratio and bias response time. We propose a three-parameter model of the noise PSD derived from measurements which includes the bias voltage dependency. The I-V curves show a non-ohmic behaviour throughout the studied bias voltage: from 20 mV up to 1.1 V. With respect to noise, flicker noise dominates from 0.1 Hz up to 100 kHz and the exponent coefficient is close to 1. The SNR is between 23 dB and 79 dB for all types of sensors.

2.1 Construction

Chemiresistive gas sensors are constructed with two interdigitated electrodes (IDE), made of (or coated with) noble metals, where the sensing layer is applied on top of them. The electrodes are preferably planar, to take

2 Characterization and Modeling of Chemiresistive Gas Sensors

advantage of widespread microfabrication techniques, like photolithography, chemical vapour deposition and etching. Sensing layers made of conductive polymers and graphene differentiate from other chemiresistive sensors like MOx in that their sensing mechanism takes place at ambient temperature and no heating element is needed, reducing complexity and power consumption. Additionally, the sensing layer can be prepared in a solution which facilitates the application by drop casting, inkjet printing or electropolymerization. The first two techniques refer to the application of the sensing layer in liquid form, followed by the evaporation of the solvent and the hardening of the polymer. Electropolymerization is the formation of a polymer from the oxidation or reduction of a monomer by an electric current. For example, pyrrole can oxidize and form a polymer chain of polypyrrole [22].

As a side note, even though the sensors used in the experiments were made with gold electrodes, we also experimented aluminium electrodes as explained in annex A. None of these sensors were functional, as no conductivity could be achieved due to the native aluminium oxide layer.

2.2 Electrical Characterization of Chemiresistive Gas Sensors

2.2.1 Baseline Resistance and Dynamic Range

There are four degrees of freedom for the sizing of the electrodes: the width (W), the spacing (S), the length (L) and the number of fingers (N), as shown in Fig. 2.1 (a). The height of the electrodes, named Z, is taken to be constant as it is determined by the fabrication process. Assuming a perfect coverage of the fingers by the sensing material, the resistance of the sensor can be approximated with equation 2.1 derived from the resistance of two parallel half-cylinders separated by a distance S, where δ is the conductivity of the sensing material. This approximation is valid for a non-polarizable sensing layer that extends infinitely upwards. In practice, a layer with a thickness of 2Z already contains 90 percent of the current, for electrodes separated less than 10 μm .

$$R = \frac{2 \ln \left(\frac{s}{z} + 1 \right)}{\delta N L \pi} \quad (2.1)$$

In Fig. 2.1, the total width of the sensor is $L + 3W$ and the length $(2N + 1)W$, the maximum area per sensor is limited by the size of the chip and the

2.2 Electrical Characterization of Chemiresistive Gas Sensors

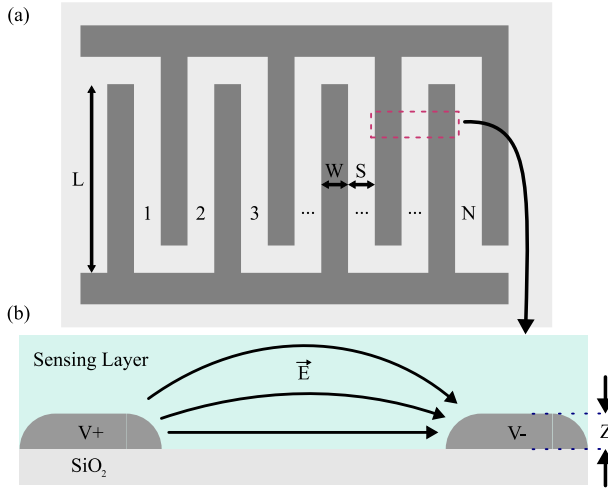


Fig. 2.1 Structure and dimensions of interdigitated electrodes for chemiresistive sensors.

number of sensors employed. The conductivities for a selection of sensing materials are listed in Table 2.1. Even though, the electrodes dimensions can help equalize the final sensor resistance, the conductivity range is considerably larger than what dimensions can compensate for, resulting in sensors with baseline resistance spanning several decades.

During our research, we experimented with sensors with an area up to 1 mm^2 with W and S in the range of $2 \text{ }\mu\text{m}$ up to $4 \text{ }\mu\text{m}$, Z fixed to $2 \text{ }\mu\text{m}$. The lower limit in sensor area is not set by the possibility of making such small electrodes but by the difficulty of accurately and reliably depositing the sensing layer without automated equipment, like inkjet printing. Our smallest sensors made by drop casting are $200 \text{ }\mu\text{m}$ by $200 \text{ }\mu\text{m}$. Considering these values, for the same conductivity, we obtain a ratio of 400 between the largest and smallest resistances by only changing electrodes dimensions. Fig. 2.2 shows the resulting dc resistance for sensors made with the constraints imposed by the electrode area and the conductivities of table 2.1. Many sensing layer formulations with different properties fall in between the extremes, for this reason it is very important, to make a read-out circuit that can accommodate very dissimilar sensors. Some formulations present a very high conductivity, resulting in sensors in the Ω 's range and below. Such devices present a challenge in terms of biasing,

2 Characterization and Modeling of Chemiresistive Gas Sensors

Table 2.1 Conductivities of several materials used as sensing layers in chemiresistive gas sensors.

Material	Doping/Treatment	$\sigma (Scm^{-1})$	Ref.
PPy	None	9.33×10^{-9}	[23]
PPy	None	3.2×10^{-3}	~*
PPy	ClO_4	10.95	[24]
PPy	ClO_4	65.34	[24]
PANI	None	4.4×10^{-11}	[25]
PANI	HNO_3	1.63×10^{-5}	[26]
PANI	-	2.56×10^{-4}	~*
PANI	CSA	7.99	[26]
rGO	None	1.18×10^{-4}	[27]
graphene	None (Pristine)	5×10^{-3}	~*
rGO	1100°C	210	[28]

* Measured in the laboratory.

because very low bias voltages will still produce large currents and compromise power consumption. These materials are better addressed either by including a resistance in series and consequently reducing the signal-to-noise ratio or by changing the chemical formulation to increase the resistance.

For our experiments, due to practical limitations when sourcing the sensing layer materials, we used one of each type: PANI, PPy and graphene. We used sensors which were assembled and tested in the institute previously as well as new sensors which were made by drop casting with 30 gauge needles (159 μm inner diameter) placed on a fixed rig. The sensing layers come from three sources: PPy was prepared in the institute, PANI was provided by Dr. Carmen Moldovan from IMT Bucharest and graphene was provided by Tiziana Polichetti from ENEA Portici Research Center. VOCsSens (Mont-Saint-Guibert, Belgium) also provided some of the samples.

2.2.2 Current-to-Voltage Curves

Even though chemiresistive sensors resemble a theoretical two-terminal resistor, the materials employed may not necessarily be ohmic, meaning that the slope of the current to voltage curve may not be constant. The current-to-voltage (I-V) curves for several sensors were measured and plotted in Fig. 2.3 (a). In Fig. 2.3 (b), the percentage change of the sensor's resistance

2.2 Electrical Characterization of Chemiresistive Gas Sensors

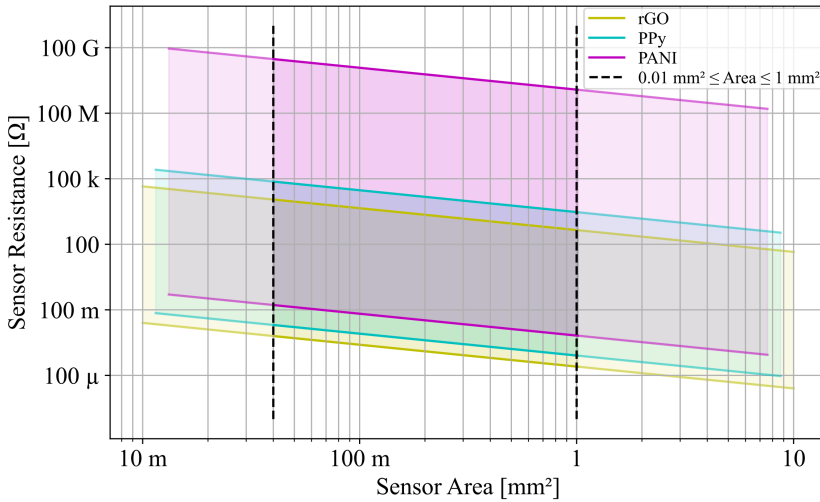


Fig. 2.2 Range of resistances for different formulations of sensing materials as a function of area. Finger width and spacing equal to 3 μm .

from the reference value at 1 V bias is calculated. PPy sensors (A-6, N01-5, N02-5, N03-4, N04-4, N04-7, 17-7) exhibit a nominal resistance at 1 V between 810.4 Ω and 5.93 k Ω . The resistance for PANI and graphene (E3) sensors are 348.4 k Ω and 9.07 M Ω respectively. PPy sensors consistently showed a decrease of resistance with bias voltage, of up to 40 % when increasing the bias from 20 mV up to 1 V. For the case of graphene, the result is similar but the measurement are affected by noise, the current is 110.3 nA at 1 V, or 90.7 M Ω . For the case of PANI the opposite behaviour is observed becoming more resistive as the bias voltage increases, phenomena like self-heating or humidity desorption could not be accounted for in the results. Although not a common figure in sensor publications, some authors have reported the I-V curves. For example, in [29], an almost perfectly ohmic characteristic is reported for PANI-based sulphide gas sensors, the same was reported in [30] for PANI films and in [29] and [25] for PANI emeraldine salt.

2 Characterization and Modeling of Chemiresistive Gas Sensors

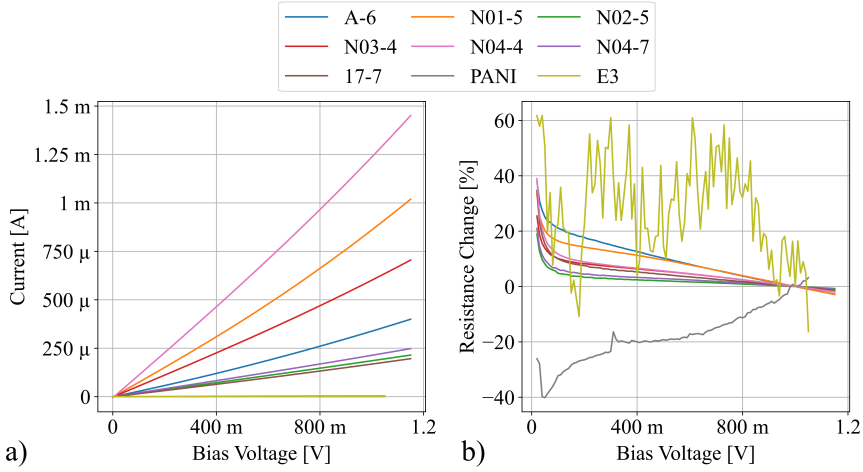


Fig. 2.3 a) I-V curves for several sensors. b) Change of the resistance with respect to 1 V bias.

2.2.3 Noise

The conduction mechanism in conductive polymers, which are organic materials, and graphene are different from that of metals and semiconductors, therefore the available models for noise in resistors and transistors may not be valid for chemiresistive gas sensors. In the literature this has not been significantly studied and more research is needed, the most relevant work are that of: Toušek et al. [31], Bruschi et al. [32], [33], Robinson et al. [34], Liang et al. [35] and Cui et. al [36]. In these publications flicker noise was found for PPy, PANI and rGO devices, although a model of the noise, that could aid circuit design, was not proposed in either of the publications.

To study the noise profile, we used two different methods: direct power spectral density measurements with an Advanced Low Frequency Noise Analyser (A-LFNA, Keysight E4727B) and secondly, transient recordings of the sensor current with the transimpedance amplifier of Fig. 2.4 which also provides the biasing for the sensor. In the latter, a PicoScope 5243D oscilloscope with 16 bit of vertical resolution and a buffer of 100 million samples was used for recording. The component values are: $R1 = [27 \text{ k}\Omega, 56 \text{ k}\Omega]$, $R2 = 100 \text{ }\Omega$, $R3 = 1.2 \text{ k}\Omega$, $R4 = 1.2 \text{ k}\Omega$, $R5 = [33 \text{ k}\Omega, 56 \text{ k}\Omega, 220 \text{ k}\Omega]$, $C = 158 \text{ nF}$, $f_0 = 19.913 \text{ kHz}$. The values of $R1$ and $R5$ were chosen depend-

2.2 Electrical Characterization of Chemiresistive Gas Sensors

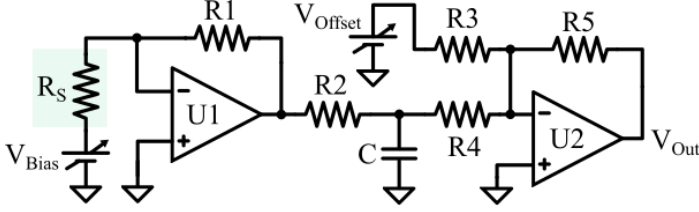


Fig. 2.4 Circuit used for the measurement of transient current noise.

ing on the sensor's resistance to achieve high gain and avoid saturation. The operational amplifiers are low-noise, low-offset LTC6241 from Linear Technologies. Equations 2.2, 2.3, 2.4, are the output voltage, gain and pole frequency respectively. In equation 2.2, $i_n(t)$ is the sensor's instantaneous current noise, having an RMS value equal to the square root of the integral of the noise PSD (measured in A^2/Hz) in the bandwidth of interest.

Table 2.2 Sensors used for noise characterization.

Sensor Code	Material
17-7	PPy
A-6	PPy
N01-5	PPy
N02-5	PPy
N03-4	PPy
N04-4	PPy
N04-7	PPy
E3-1	rGO
P1	PANI

$$V_{Out} = R4 \left(\frac{V_{Bias} R1}{R_S (R2 + R4 + R2 R4 C s)} - \frac{V_{Offset}}{R3} \right) + \frac{i_n(t) R5 R1}{R2 + R4 + R2 R4 C s} \quad (2.2)$$

$$G_0 = \frac{R5 R1}{R2 + R4} \quad (2.3)$$

$$f_0 = \frac{R2 + R4}{2\pi R2 R4 C} \quad (2.4)$$

2 Characterization and Modeling of Chemiresistive Gas Sensors

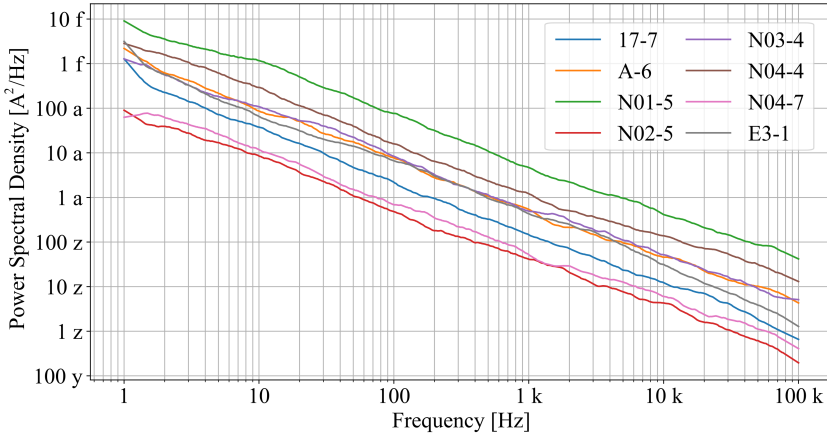


Fig. 2.5 Current noise PSD of several sensors biased with 1 V, measured with the Keysight A-LFNA.

The sensors used for the tests are listed in table 2.2. The PSDs measured with the A-LFNA are plotted in Fig. 2.5. At the same time, the noise of three of these sensors and a reference resistor (100 k Ω) were measured with the transimpedance amplifier and plotted in Fig. 2.6 b). Given that the transimpedance amplifier was assembled specifically for these measurements, the 100 k Ω resistor was used to evaluate the noise level of the circuit. The noise measured with the resistor is orders of magnitude lower than that of the sensors, verifying its adequacy. Still, the theoretical thermal noise of a 100 k Ω resistor is $0.166 \times 10^{-24} \text{ A}^2/\text{Hz}$ while the measured value at 1 kHz is $14.5 \times 10^{-24} \text{ A}^2/\text{Hz}$. This difference is explained by the noise of the other devices and the quantization noise of the oscilloscope.

From the noise measurements of the sensors (with both methods), we see that flicker noise is predominant in the band of interest (0.1 Hz - 100 Hz), therefore it is convenient to model the noise, by a multiplicative factor α and an exponent of the inverse of the frequency, named β , resulting in the model of equation 2.5. These parameters were extracted for the measured curves which were taken at different bias voltages, the results are presented in Fig. 2.7 for the A-LFNA and in Fig. 2.8 for time measurements. The normalized (to the sensor current) multiplicative parameter is plotted in Fig. 2.9.

2.2 Electrical Characterization of Chemiresistive Gas Sensors

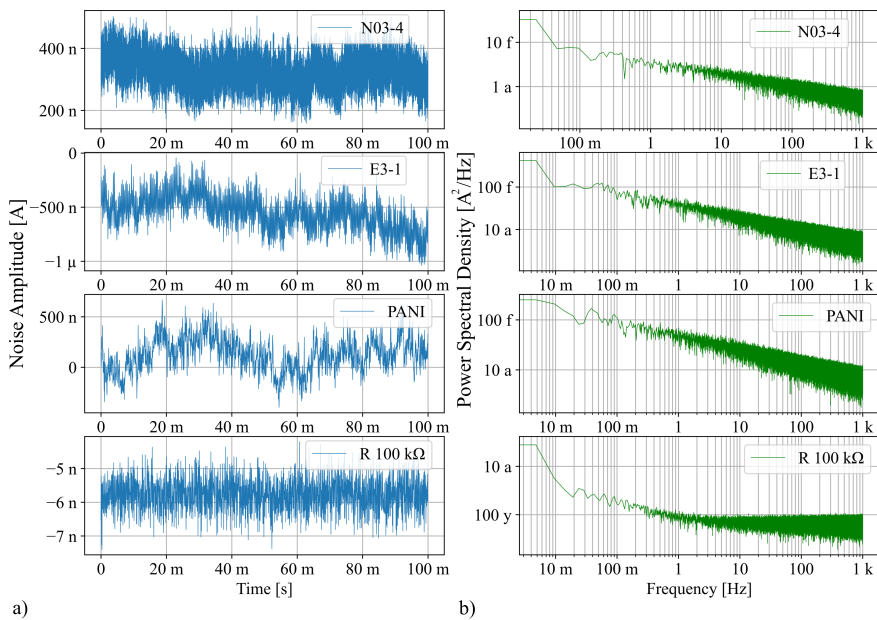


Fig. 2.6 a) Time records of sensor current noise for three different sensors and a resistor as a reference (last row). b) PSD of the time records.

2 Characterization and Modeling of Chemiresistive Gas Sensors

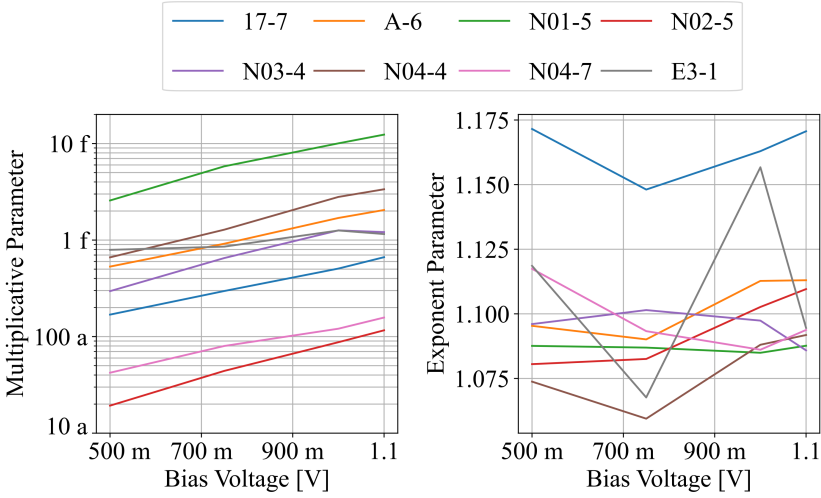


Fig. 2.7 Extracted α and β parameters of the noise model presented in equation 2.5, from the measurements presented in Fig. 2.5.

$$S_n(f) = \frac{\alpha}{f^\beta} \quad (2.5)$$

The exponent parameter β does not follow a clear pattern and its small variations seem to be related to the fitting algorithm used. On the other hand, α is causally related to the bias voltage (and hence to the DC current). A simple model that fits this correlation is presented in equation 2.6 where a new parameter λ is introduced as an exponent of the bias voltage. The values of λ in table 2.3, were extracted by fitting equation 2.6 to the previously extracted values of α .

$$\alpha(V_{Bias}) = \sigma V_{Bias}^\lambda \quad (2.6)$$

In Fig. 2.10 the model with the best-fit parameters is plotted (in dashed) together with the initially extracted values of α . For both PPy and graphene the model predicts reasonably well the dependence of noise with the bias voltage but for PANI the error is greater. The complete noise model, with the three parameters is presented in equation 2.7. This model is similar to that proposed by F.N. Hooge in [37] for flicker noise in semiconduc-

2.2 Electrical Characterization of Chemiresistive Gas Sensors

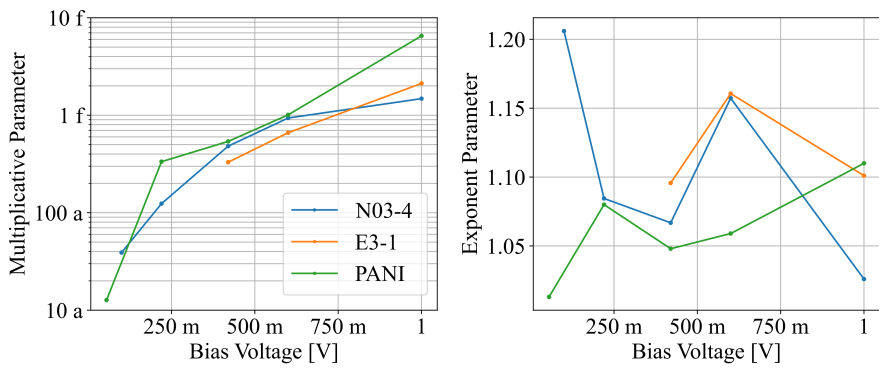


Fig. 2.8 Extracted α and β parameters of the noise model presented in equation 2.5, from the measurements presented in Fig. 2.6.

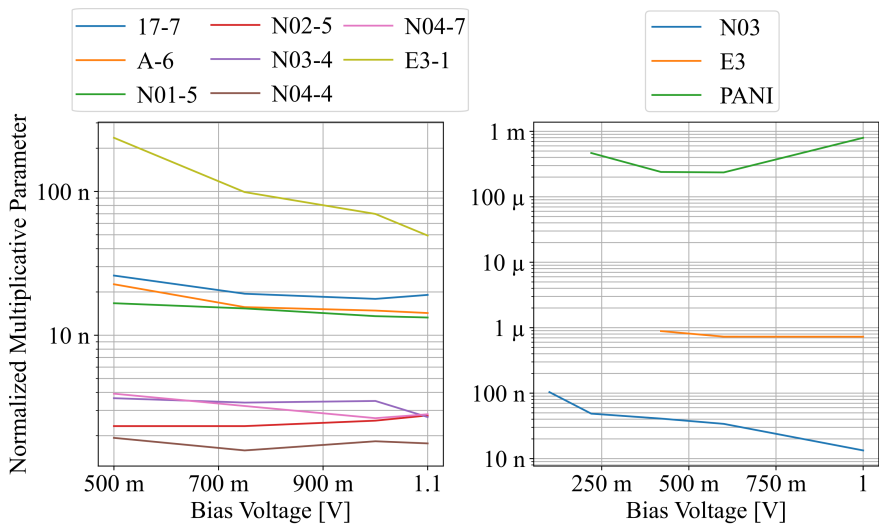


Fig. 2.9 Normalized multiplicative parameter for A-LFNA and transient noise measurements.

2 Characterization and Modeling of Chemiresistive Gas Sensors

Table 2.3 Estimated λ for A-LFNA and time noise records.

Sensor	$\lambda(\text{A-LFNA})$	$\lambda(\text{time})$
17-7	1.90	-
A-6	1.89	-
N01-5	1.98	-
N02-5	2.45	-
N03-4	1.74	1.66
N04-4	2.30	-
N04-7	1.70	-
E3-1	0.61	2.15
P1	-	2.00

tors, equation 2.8, where ϵ , γ and σ are constants and N_c is the number of carriers. Hooge's model was developed for homogeneous samples and the noise level is inversely proportional to the number of mobile charge. Nevertheless, this model was not developed for conductive polymers and therefore, an in-depth study of the conduction mechanisms is needed to confirm a similar origin of flicker noise.

$$S_n(f) = \frac{\sigma V_{Bias}^\lambda}{f^\beta} \quad (2.7)$$

$$S_v(f) = \gamma \frac{V_{DC}^{2+\sigma}}{N_c f^\epsilon} \quad (2.8)$$

The PSD obtained from transient measurements and the one predicted by the model of 2.7 are plotted in Figs.: 2.11, 2.12, 2.13 for PPy, PANI and graphene sensors respectively and several bias voltages. For PPy and graphene the fit in magnitude is close to measurements but there is a small difference in the slope of the plot, set by the frequency exponent. This is in part explained by the fact that not all the curves have the same slope, and even for a single measurement the slope is not steady throughout the whole frequency range. For the case of PANI, as noted before, the modelling of the effect of bias voltage on noise is not accurately achieved. Given that only two PANI samples were tested it is important to confirm or reject this results with further experiments in the future.

2.2 Electrical Characterization of Chemiresistive Gas Sensors

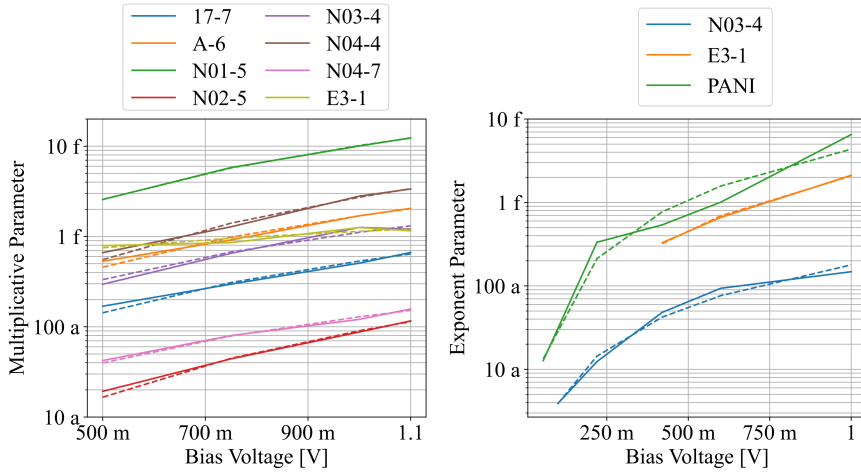


Fig. 2.10 a) α as calculated from measurements in solid lines and its approximated value in dashed as calculated with equation 2.6. Measurements taken with the A-LFNA. b) Same as a) but for measurements made with the circuit of Fig. 2.4.

Signal-to-Noise Ratio

The process of measuring a physical magnitude is ruled by the statistics which describe how likely is the measurand to be found in a certain region given the characteristics of the measurement equipment. The noise in electronic circuits is mainly represented by Gaussian processes (according to the central limit theorem), as such noise is caused by the aggregation of the random movement of many electrons whose statistics need not be known.

¹ Therefore, the noise of the interfacing and processing circuits (including the sensor's noise) increase the uncertainty of the measurement and may obscure what the real value of the physical variable is.

In Fig. 2.14, R_{x1} , R_{x2} and R_{x3} are all possible outcomes of the measurement of R_0 , with different degrees of likelihood. When the noise of the sensing system is reduced the standard deviation is smaller, and values like R_{x1} and R_{x3} become less probable. As a consequence, when a measurement is now performed, the new measured value is likely to be closer

¹ Random telegraph noise (RTN), which is a non-gaussian telegraph process, has been observed in MOS transistors, especially in nanometer-scale devices and could alter the statistics if its contribution is significant.

2 Characterization and Modeling of Chemiresistive Gas Sensors

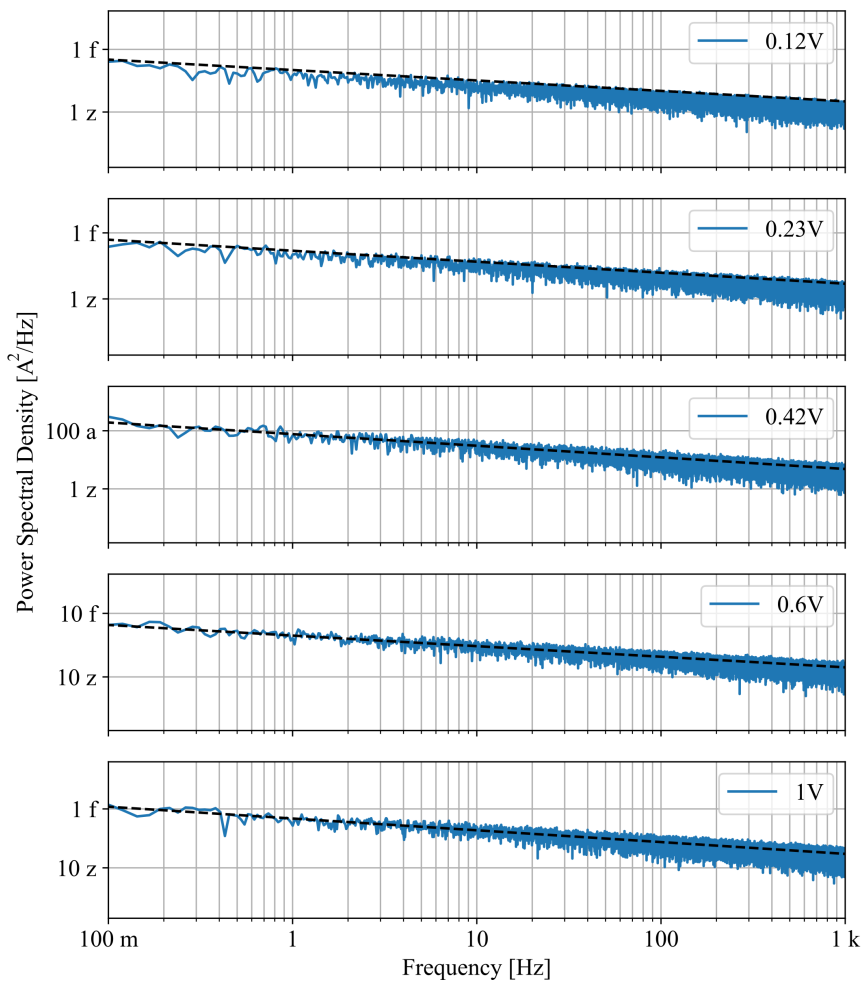


Fig. 2.11 PSD of current noise in a PPy sensor for five bias voltages: 0.12 V, 0.23 V, 0.42 V, 0.6 V, 1 V in blue. The predicted noise by the proposed model is plotted in black dashed lines.

2.2 Electrical Characterization of Chemiresistive Gas Sensors

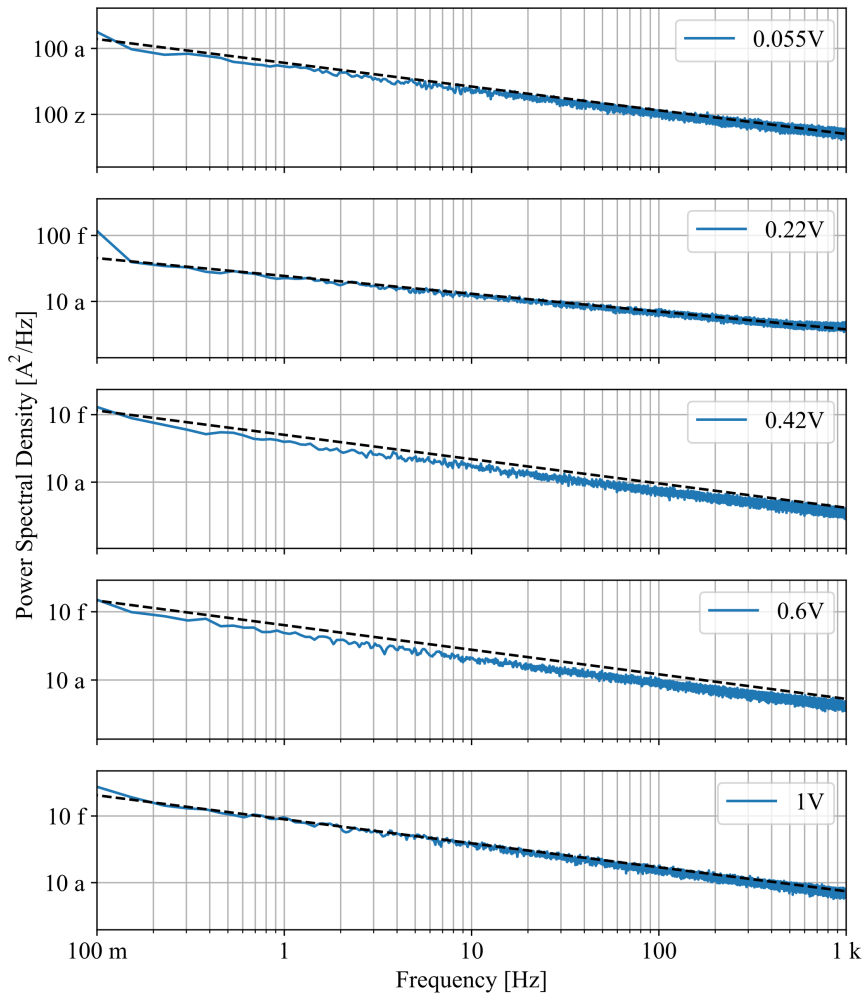


Fig. 2.12 PSD of current noise in a PANI sensor for five bias voltages: 0.055 V, 0.22 V, 0.42 V, 0.6 V, 1 V in blue. The predicted noise by the proposed model is plotted in black dashed lines.

2 Characterization and Modeling of Chemiresistive Gas Sensors

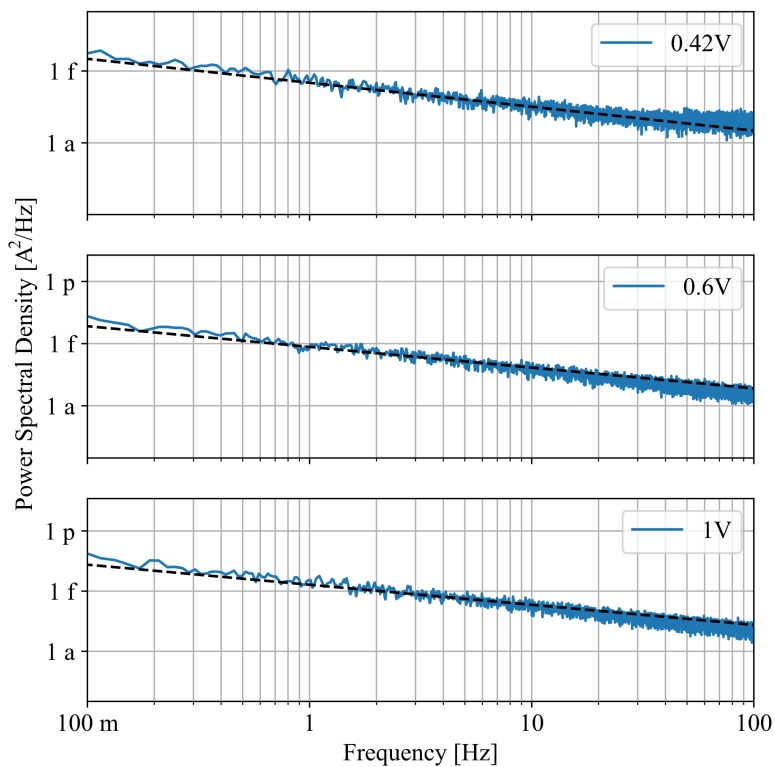


Fig. 2.13 PSD of current noise in a graphene sensor for three bias voltages: 0.42 V, 0.6 V, 1 V in blue. The predicted noise by the proposed model is plotted in black dashed lines.

2.2 Electrical Characterization of Chemiresistive Gas Sensors

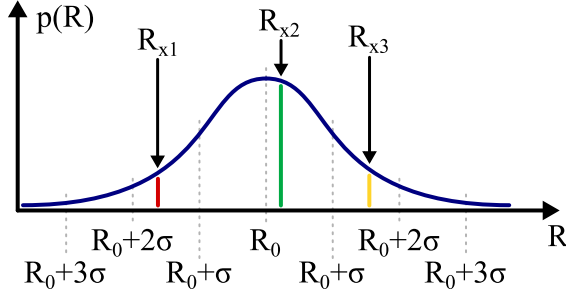


Fig. 2.14 Measurement uncertainty (normal distribution) and resistance estimation.

to the true value of the magnitude than before, assuming no bias exists [38].

The SNR is defined as ten times the logarithm of the ratio between the signal power and the noise power, equation 2.9. Therefore, both must be defined in relation to the magnitude of interest. The sensitivity of gas sensors is determined by the chemical composition of the sensor and is set once the sensor is fabricated. From a circuit design point of view, the sensitivity cannot be modified so the interface circuit should be able to discern resistance changes comparable (or smaller) than the uncertainty in the sensor resistance caused by the sensor's intrinsic noise.

$$SNR = 10 \log_{10} \left(\frac{P_{Signal}}{P_{Noise}} \right) \quad (2.9)$$

Given the above, we define the SNR as the ratio between the sensor's DC current power I_S^2 and the integrated PSD of the noise $i_n(t)^2$ (expressed in A^2), as calculated in equations: 2.10, 2.11, 2.12 and Fig. 2.15. Where $i_n(t)^2$ is the square of the instantaneous noise and S_n the sensor's current noise PSD (in A^2/Hz). Also, $i_n(t)$ equals the RMS amplitude of the noise current in the time domain, or one standard deviation of the Gaussian distribution.

$$i_n^2(t) = \int_{f_0}^{f_1} S_n(f) df \quad (2.10)$$

$$I_S(t) = \frac{V_{Bias}}{R_S} \quad (2.11)$$

2 Characterization and Modeling of Chemiresistive Gas Sensors

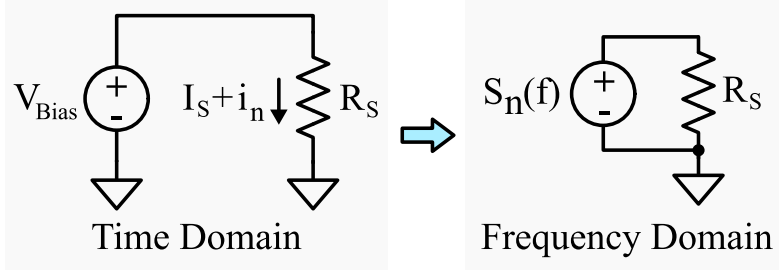


Fig. 2.15 Noise models in time and frequency domains.

$$SNR = 10 \log_{10} \left(\frac{I_S^2}{i_n^2} \right) = 20 \log_{10} \left(\frac{V_{Bias}}{R_S i_n} \right) \quad (2.12)$$

Fig. 2.16 presents the SNR of the studied sensors for different bias voltages, where the bandwidth considered for the integration is from 0.1 Hz up to 100 Hz. Even though, the noise was measured up to 100 kHz, above 100 Hz there are no signals of interest due to the slow reaction time of the sensors. A capacitor, placed in parallel with the sensor, can help reduce the high frequency noise if it impacts the SNR considerably.

2.2.4 Sensitivity

The sensitivity of each sensing layer formulation is set by the chemistry and its affinity with certain gases. The standard figure of merit used in the literature to quantify it is the fractional change from its baseline value [30], [39], expressed as a percentage, equation 2.13. Where R_a is the resistance of the sensor when exposed to pure air and R_g is the resistance when exposed to a given gas concentration.

$$S = 100 \left(\frac{R_a - R_g}{R_a} \right) \quad (2.13)$$

The minimum detectable concentration, also known as limit of detection (LOD) is usually defined as the value for which a gas can be detected with a given confidence, e.g. 99 % or 3 standard deviations. To obtain the LOD for chemiresistive sensors, first we have to obtain the uncertainty in the estimation of the resistance of the sensor, which can be done with equation 2.14. Next, we calculate the LOD, with equation 2.15, where G is the

2.2 Electrical Characterization of Chemiresistive Gas Sensors

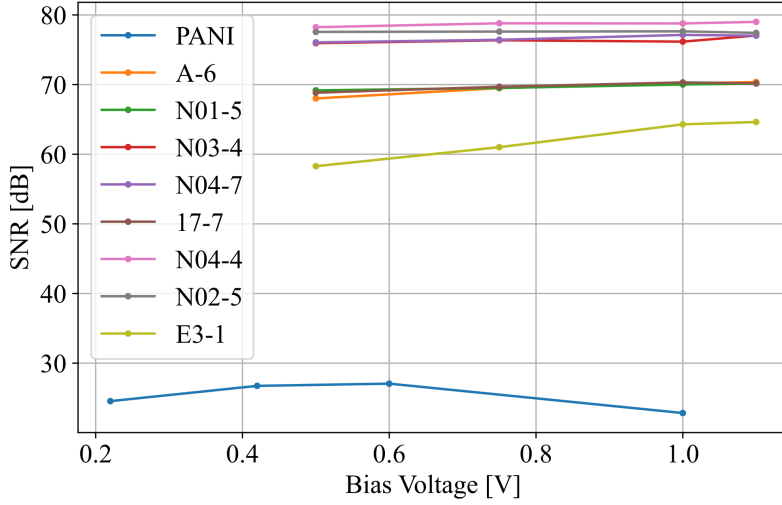


Fig. 2.16 SNR for the studied sensors as calculated with equation 2.12.

sensitivity of the sensor (expressed in ppm/Ω).

$$R_S + r_n = \frac{V_{Bias}}{I_S - i_n} \rightarrow r_n = \frac{V_{Bias}}{I_S - i_n} - \frac{V_{Bias}}{I_S} \approx \frac{V_{Bias} \cdot i_n}{I_S^2} \quad (2.14)$$

$$LOD = 3 \cdot \sigma_R \cdot G = 3 \cdot r_n \cdot G = \frac{3 \cdot V_{Bias} \cdot i_n \cdot G}{I_S^2} \quad (2.15)$$

The purpose of the ROIC is to quantify the resistance of the sensor, and given the sensor noise level, a minimum discernible resistance will be found, as explained in the SNR subsection. Therefore, even though the sensitivity (together with noise), sets the limit-of-detection, in general it does not impose restrictions in the design of the interface, as the ROIC estimates a resistance which is afterwards converted to a gas concentration, with a resistance-to-concentration lookup table. The only exception is when the sensitivity is very large, like the methane sensor reported in [40] with a sensitivity of $72.6 \times 10^6 \text{ } \%/ppm$ (or $890 \text{ k}\Omega/ppm$). In this case, a change in the concentration of the gas will produce a significant change in the resistance and consequently in the point of operation of the ROIC.

In Fig. 2.17 we present the measured sensitivity for one of our PPY

2 Characterization and Modeling of Chemiresistive Gas Sensors

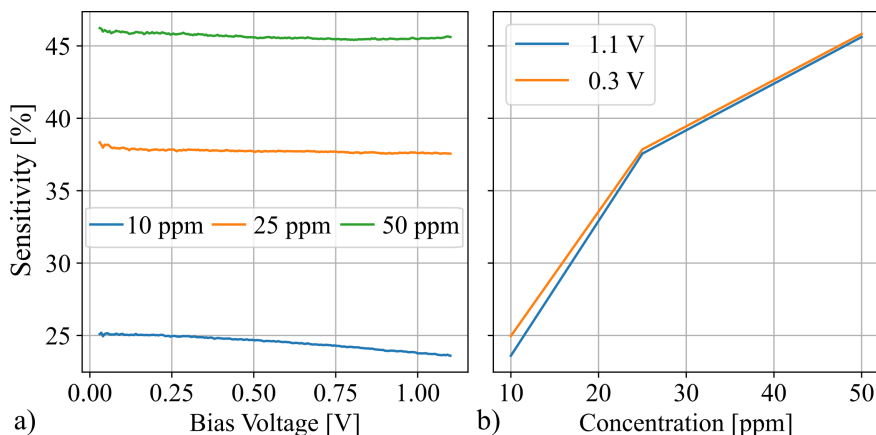


Fig. 2.17 a) Sensitivity of a PPy sensor to different concentrations of ammonia and a range of bias voltages, expressed with the FoM of 2.13. b) Sensitivity expressed as a function of the ammonia concentration.

sensors for different ammonia concentration and bias voltages, the LOD of this sensor is 0.044 ppm when biased at 1 V.

Outside this work, the characterization and figures of merit related to the noise of gas sensors are not unique. In the literature, as it was noted before, most of the published papers do not study the noise of the sensor, but some do provide the LOD, we include a few examples in table 2.4 for PPy. What is most usually found is the detection range [15], [16], [41]. There is not an unambiguous definition and some papers give a detection range starting from 0 ppm. This means that the sensor noise has not been studied and not much can be said about the LOD. In other papers, when the detection range does not begin in 0 ppm there is not an explicit explanation for this. It is not clear if the noise was measured and the smallest range coincides with the LOD, or it could mean that the characterization was performed in that concentration range or that the LOD is smaller but an accurate value is not provided.

When it comes to commercial sensors, we see in the datasheets of Figaro Engineering Inc. sensors, that the detection range start from 0 ppm for some of their sensors and at higher values for others, leaving unanswered if this figure contemplates the noise or not. On the other hand, Honeywell International Inc. indicates the resolution for a few of their sensors, meaning that some sort uncertainty characterization was performed.

Table 2.4 LOD of published PPy sensors.

Sensor Type	LOD (ppm)	Reference
PPy nanoribbons	0.5	[42]
PPy nanoparticles	5	[43]
Multidimensional PPy nanotubes	0.01	[44]
PPy with SnO_2 or ZnO	10	[45]

2.2.5 Response time

The response time of chemiresistive sensors involves two different phenomena, on one side the stabilization time that is required for the bias to settle when no gas is present and on the other hand, the time it takes for the sensing mechanism to take place once the sensor is exposed to a gas. The former is of most importance from ROIC design point of view as it sets the biasing strategy for each sensor, the latter impacts mainly the use case, and in turn this will determine the acceptable delay, for example in applications like alarm systems speed is paramount, where in other cases like in soil monitoring, a sensor with a lower response time may be sufficient. The response time to changes in the concentration of gases is the most studied in the literature and is widely published in sensors review papers, [4], [5], [15], [16]. There are only a handful of examples where the response time is less than 1 seconds, for this reason it is sufficient to have a ROIC with a bandwidth of only a few Hertz.

Regarding the time response to the bias voltage, in Fig. 2.18 a square wave of 200 mV peak-to-peak amplitude is applied to PPy sensor, responding instantly to the voltage step.

2.3 Conclusions

In this chapter the electrical characteristics of PPy, PANI and graphene gas sensors were studied. The most remarkable results are the wide diversity of formulations of sensing layers which yielded a wide range of baseline resistances. Secondly, we observed a non-ohmic behaviour for all of the tested sensors, where the resistance decreases with bias voltage, except for PANI where the opposite was observed. Thirdly, noise was studied, and it was found that flicker noise is predominant for all sensors in the bandwidth where gas sensing operates. On top of this, the amplitude of the

2 Characterization and Modeling of Chemiresistive Gas Sensors

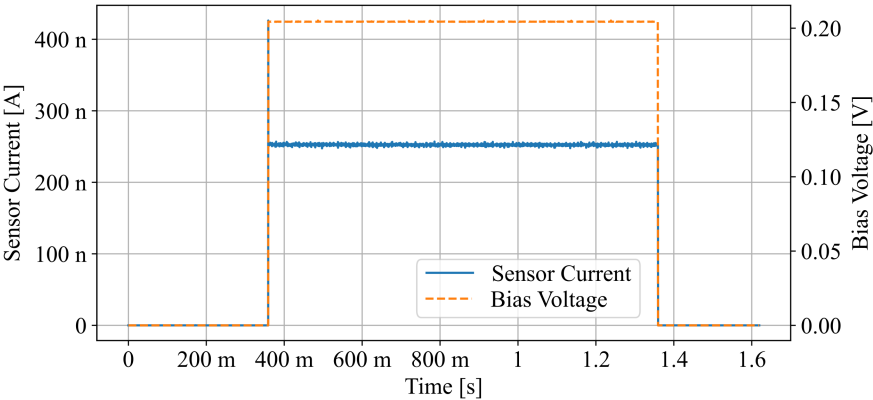


Fig. 2.18 Current of a PPy sensor when biased with a 200 mV square wave.

noise is also related to bias voltage, increasing for larger values. The SNR is between 23 dB and 79 dB, being almost independent of the bias voltage. The sensitivity of PPy is unaffected by the bias voltage, so the LoD is constant while the power increases exponentially with it. These constraints are carried on to the next chapter where the types of ROICs used to interface chemiresistive gas sensors will be discussed.

3

Readout Circuits for Resistive Sensors

Abstract:

As seen in Chapter 2, new gas sensing materials, like conductive polymers and nanomaterials-based composites, together with integrated circuit advancements have enabled a new paradigm in gas sensing where an array of different types of sensors is used to improve selectivity and sensitivity in gas sensing. Here, we begin with a discussion of the different types of resistive sensor ROICs, followed by a review of the state of the art, after which we present a highly flexible ROIC for acquiring the DC resistance of chemiresistive sensors. It can measure sensors from 1 k Ω up to 33 M Ω , with a minimum SNR of 57 dB. It also offers a wide range of input configuration in terms of resistance and bias voltages to select the optimal bias point for each sensor and to accommodate a large range of sensor types. It achieves very low power consumption with a maximum current consumption of 194 μ A ($V_{DD} = 1.8$ V) and an energy per conversion ranging from 1.21 nJ up to 188 nJ. The circuit was fabricated in a 180 nm bulk CMOS process from United Microelectronics Corporation (UMC) and a complete characterization of the circuit is presented including: current consumption, SNR, and transfer function. Finally, the circuit was tested in a real application for the measurement of NH_3 and NO_2 using two different types of

3 Readout Circuits for Resistive Sensors

sensors validating the design objectives and the capability of the ROIC for system-on-chip integrations.

3.1 Requirements of a ROIC to Interface An Array of Chemiresistive Gas Sensors

The objective of a sensor readout circuit is to provide a digital representation of the measured quantity with an acceptable resolution and uncertainty for the application. As explained in Chapter 2, chemiresistive gas sensors have some unique features (large resistance range, flicker noise and nonlinearity) that sets them apart from other well-known resistive sensors, like temperature detectors, whose behaviour closely resembles that of a theoretical resistor. For a multi-sensor system, the most critical aspect is the capacity to interface a wide range of sensors while maintaining an acceptable SNR in each case. This, extends the range of possible sensor to be interfaced (mainly differentiated by the sensing layer material), enhancing the capacity to cross-correlate the measurements, hence improving selectivity and reducing the LoD. Having flexibility in the bias voltage is also desirable, as it was seen that noise and power depend on it, and can be used as a knob to optimize the operating point of the sensor.

Sensor ROICs are analog-to-digital converters themselves in the sense that they produce a digital word from an analog magnitude, but with a very specific application considered in the design from the start. Given this, they may not necessarily follow the traditional voltage ADC topologies and figures of merit: sensor biasing may be part of the ROIC, the sensor model dictates the input characteristics and no generic usage is foreseen like in off-the-shelf ADCs. ROICs used to estimate the sensor's resistance can be broadly classified into two categories: those that produce a voltage amplitude and those that produce a time signal whose period, frequency or duration (non-oscillatory ROICs) is proportional to the measurand.

With respect to amplitude readout circuits, they measure the voltage directly with a traditional analog-to-digital converter (ADC). Frequently, Wheatstone bridges or current digital-to-analog converters (IDAC) are used to bias the sensor and extend the input range, as is the case of [46]. They offer good linearity and integrability. On the other hand, achieving simultaneously high dynamic range and low power could prove challenging, due to the large number of bits needed both in the ADC and the IDAC

to achieve a large input dynamic range (e.g. 150 dB)[47], [48], [49].

Time-based ROICs can be further sub-classified into two categories: open or closed-loop. Open-loop designs are simpler and offer the benefit of separating the biasing of the sensor from the rest of the circuit, imposing no restrictions on the sensor bias point. On the other hand, closed-loop designs use the output frequency to adjust the sensor bias current and make the oscillator, a voltage-controlled oscillator (VCO) or a current-controlled oscillator (CCO), operate around a fixed input voltage or current, avoiding the usual nonlinearity in the voltage-to-frequency conversion found for larger input swings, where the phenomena that sets the frequency deviates from a small-signal linear behaviour. All time-based approaches require an additional circuit to digitize the period.

To a lesser extent, other architectures like triggered ROICs have also been investigated, producing a measurement in the time domain but without oscillating [50].

In the next sections, after reviewing the figures of merit (FOM), we describe in detail each of the main categories of ROICs and the state of the art, citing the most relevant publications for each type.

3.2 Figures of Merit

3.2.1 Energy per Conversion

Together with standard FOMs, like static current or power consumption, one that has received significant attention is the energy per conversion (EPC), equation 3.1. The EPC is the energy needed to produce one estimation of the measurand, represented as a digital number in steady state. It is similar to Walden's classic ADC FOM [51] (equation 3.2), that is sometimes used in conjunction or alternatively, but does not account for the effective number of bits (ENOB).

$$EPC = Power \cdot t_{conversion} \quad (3.1)$$

$$FOM_{Walden} = \frac{Power \cdot t_{conversion}}{2^{ENOB}} \quad (3.2)$$

In open-loop readout circuits, the bandwidth and output jitter evolve throughout the input range giving a different SNR, and thus an ENOB, for each operating point. For this reason, it is sometimes preferred to study the energy profile separated from that of noise as they can be uncorrelated.

3 Readout Circuits for Resistive Sensors

Furthermore, ideally the noise caused by readout circuits should not be larger than that caused by the sensor, so it is the sensor that will ultimately determine the SNR and thus EPC can serve as a comparison metric for the complete system. Like Walden's FOM, the EPC does not account for the stabilization time that is needed after power-up, nor sensor specific variables like bias voltage, but it still captures the intrinsic efficiency of the circuit.

3.2.2 Signal-to-Noise Ratio

The SNR must be consistent with the figure used for the sensor to enable a direct comparison between the sensor and the ROIC. For both, amplitude and oscillatory ROICs we define the SNR as the input current power over the input resistance uncertainty power. Where I_{in} is the sensor's current, and i_{noise} the instantaneous RMS noise obtained from the integration of the PSD.

$$SNR = 10\log_{10} \left(\frac{I_{in}^2}{i_{noise}(t)^2} \right) = 10\log_{10} \left(\frac{V_{bias}^2}{R_S^2 i_{noise}(t)^2} \right) \quad (3.3)$$

3.2.3 Dynamic Range

For the dynamic range, the definition used is as described by equation (3.4), where SNR_{min} is the minimum SNR in the considered resistance range and R_{MAX} and R_{MIN} are the maximum and minimum input resistance, respectively. This convention, which assumes the worst case, is used for our circuits, but is not always available for other published works in same test conditions (bias voltage, linearity). Additionally, we also use the ratio of R_{MAX} and R_{MIN} .

$$DR = SNR_{min} + 20\log_{10} \left(\frac{R_{MAX}}{R_{MIN}} \right) \quad (3.4)$$

3.3 Analog-to-Digital Converter ROICs

Voltage ADCs have been the backbone of signal acquisition systems from the analog to the digital world. Although current ADCs do exist, they have been considerably less adopted than their voltage counterparts. Throughout the evolution of CMOS technologies, uncountable architectures and

3.3 Analog-to-Digital Converter ROICs

designs of voltage ADCs have been proposed to best meet the requirements of a wide range of applications, the main categories are:

- **Flash:** The input signal is compared to a set of equidistant analog voltages producing a digital conversion in one clock cycle. Consequently, they are the fastest type but lag in number of bits as 2^n analog voltages have to be generated simultaneously for a converter with n bits. As n increases, also does the area and the power consumption, as 2^n comparators are required.
- **Successive-Approximations (SAR):** This type of converters is widely used in microcontrollers because they offer a good balance between number of bits, conversion speed and area. When a conversion begins, a register with n bits (for an n -bit converter) is cleared. This parallel register is connected to an n -bit DAC that produces an analog voltage that is compared to the input voltage with a comparator. Starting from the MSB each bit is sequentially set to one to see if the output of the comparator toggles. If it does, it means that the given bit is zero, if it does not toggle the bit is one. This is done for the n bits until the LSB is reached. At the end of the conversion the value of the register is the digital approximation of the input signal.
- **Sigma-Delta ($\Sigma\Delta$):** Typically used for instrumentation, they provide the largest effective number of bits (ENOB), with commercial integrated circuits reaching precisions above 24 bits. They are highly digital and use a $\Sigma\Delta$ modulator to push quantization noise to higher frequencies which is then low-pass filtered. They trade lower noise for conversion speed, still with modern manufacturing processes their frequency limitation has been to a certain extent overcome. For example, the LTC2500 from Linear Technologies has an ENOB of 17 bits at 1 MSPS.
- **Pipeline (subranging quantizer):** The input voltage is compared to half the reference voltage to find the MSB. If the MSB is one, half the reference voltage is subtracted from the input signal and no subtraction is performed if the MSB is zero. Then, this new voltage is amplified with a gain of two and compared to half the reference voltage as was done for the first bit, to find the MSB-1 bit. This process is repeated n times for an n -bit converter, from the MSB to the LSB in a sequential architecture. In this way, n simultaneous conversion

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can be carried out in a n -bit converter, producing one conversion per clock cycle.

- **Folding:** Two ADCs work in parallel, one that performs the MSBs conversion and the other the LSBs. The latter also includes a folding circuit which transform the input voltage range into a sequence of smaller voltage ranges, in a sawtooth fashion. The MSB ADC finds the appropriate input folding range that is used in the LSB ADC to set the correct position within the folded curve. It is similar to a flash ADC but requires less comparators. Finally, a digital circuit combines the results of both ADCs.
- **VCO-Based (or CCO-based):** In these oscillators, the output frequency changes proportionally to the input voltage around a center frequency f_0 , according to equation 3.5, where ρ is the gain. The frequency is subsequently quantized, to obtain a digital value proportional to the input voltage by subtracting the center frequency. This is sometimes used as the backbone of oscillator-based ROICs.

$$f_{OUT} = f_0 + \rho V_{in} \quad (3.5)$$

In Fig. 3.1, the SNR is plotted as a function of the energy-per-conversion step for a selection of ADCs published from 1997 until 2023 and compiled by B. Murmann in [1].

From the survey, we observe that $\Sigma\Delta$, SAR and VCO are viable candidates for implementing a ROIC, while flash and folding types fall short on SNR, due to the lower ENOB. Those ADCs that achieve simultaneously a high SNR and low power consumption present the lowest sampling rates, which is not an issue for the ROIC, given the low bandwidth requirement of gas sensing. The sensor bias is set externally but it has to be consistent with the input range of the ADC, therefore, it can be set directly as a function of the digital output to avoid overrange situations.

ADC ROICs can be open or closed loop as in Fig. 3.2. For closed loop circuits, a current DAC is typically employed to bias the sensor Fig. 3.2 (b). The maximum sensor current is limited by the reference voltage of the ADC (V_{REF}) and the sensor resistance (R_S). As shown in the diagram of Fig. 3.3, these ROICs have the property that the larger the bias current the smaller the relative resistance change corresponding to one LSB, as a larger voltage drop is produced for the same resistance change. Consequently, for a higher SNR, larger bias currents are needed. If the sensor's

3.3 Analog-to-Digital Converter ROICs

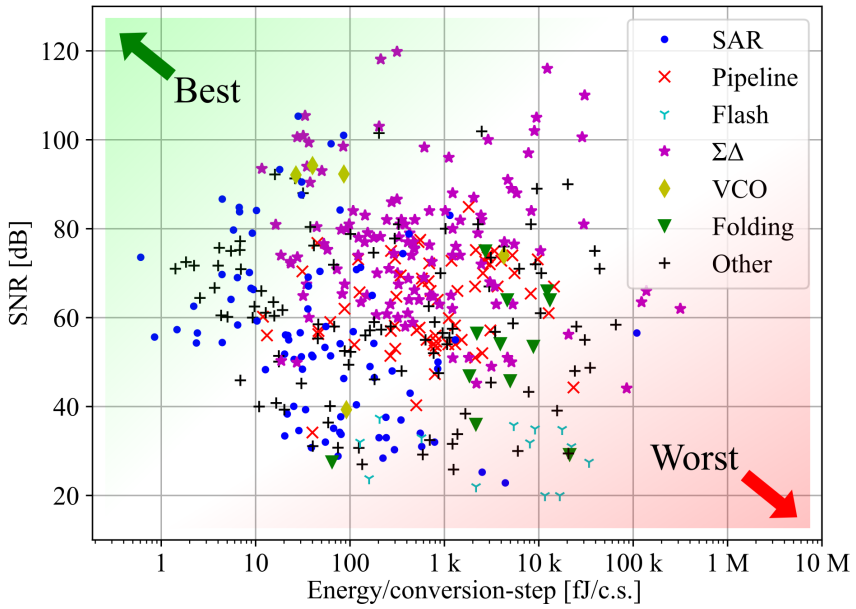


Fig. 3.1 Relation of SNR to energy-per-conversion step for a selection of publications taken from the ADC survey of [1].

3 Readout Circuits for Resistive Sensors

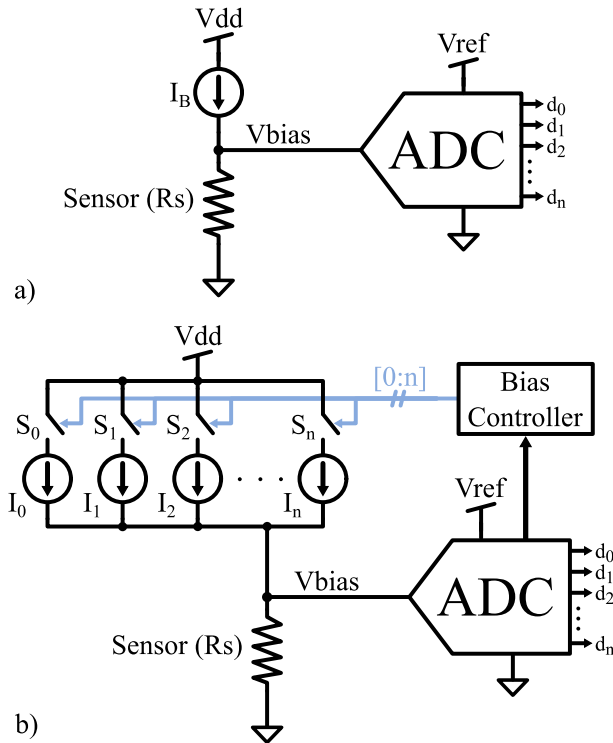


Fig. 3.2 a) ROIC with fixed-biased sensor and an ADC in open-loop configuration. b) ROIC with an ADC and an IDAC in a closed-loop configuration.

3.3 Analog-to-Digital Converter ROICs

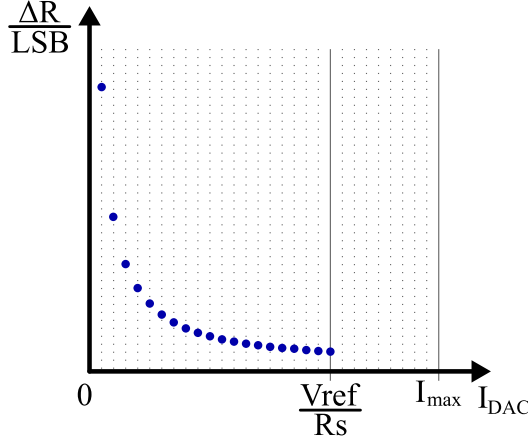


Fig. 3.3 Resistance change to LSB ratio versus bias current for a given sensor resistance.

resistance is too low, the current capacity of the IDAC sets the SNR upper limit in terms of quantization noise. The conversion time of ADCs is independent of the input voltage, therefore, the conversion time of the ROIC is also mostly unaffected by the changes in the sensor's resistance. With this assumption, the EPC of an ADC-based ROIC can be estimated with equation 3.6, where T_{Conv} is the conversion time, V_{DD} the supply voltage, I_{Bias} the current through the sensor and I_{ADC} the average current consumed by the ADC in one conversion.

$$EPC = T_{Conv} V_{DD} (I_{Bias} + I_{ADC}) \quad (3.6)$$

Besides gas sensing, ADC-based ROICs have been developed for other resistive sensors like RTDs, thermistors, strain gauges, magnetoresistive sensors, pressure sensors and photoresistors. For example, Xin et al. [46] presented a four-terminal sensing ROIC for resistive sensors in implantable or wearable medical devices, that can measure resistances from almost 0Ω up to $30 \text{ k}\Omega$. This 65 nm design, uses a 5-bit IDAC to bias the sensor with a current between $1 \mu\text{A}$ and $31 \mu\text{A}$. The ADC is a 10 bit SAR with a sampling rate between 0.1 S/s and 12.5 kS/s . Correlated double sampling is used to reduce offset and flicker noise. To compensate for temperature variations, a system-level ratiometric measurement is done with a reference resistor. The lowest power consumption is 0.32 nW at 0.1 S/s . Due to such low sam-

3 Readout Circuits for Resistive Sensors

pling rate, the EPC is high, at 3.2 nJ. When the sampling rate is increased to 100 Hz or 25 kHz, the EPC drops to 26 pJ or 22.3 pJ respectively. The SNR reaches 49.1 dB with the highest bias current (31 μ A) and the energy-per-conversion step ranges from 98 to 552 fJ/conv-step. The silicon area is 0.076 mm².

A similar ROIC for gas sensors was proposed in [52]. Here, a 10-bit SAR ADC with correlated double sampling is used but the structure of the biasing circuit is slightly different. Two 6-bit IDACs are used to bias the sensor (R_S) and an 8-bit adjustable resistor (R_A) respectively. Each of them is connected to one of the inputs of the correlated-double sampler. The ROIC operates in two phases: during the first phase, the trimming value that equals the adjustable resistor to the sensor is found. This is done by connecting both: R_S and R_A in series and the remaining terminals one to ground and one to a reference voltage, forming a resistor divider. The output of this divider is compared to a reference voltage with a comparator. The trimming value of the adjustable resistor is increased, until the comparator toggles. In the second phase, the sensor and the adjustable resistor are disconnected and each of them connected in series with an IDAC. The IDACs are turned on, and the digital conversion is performed after amplifying the differential input voltage (voltage difference between the sensor and the resistor). The LSB of the IDAC and the adjustable resistor are 190.25 nA and 8 k Ω respectively. The input range is from 8 k Ω up to 2 M Ω and the ENOB 14.1 bits. In terms of power consumption, it lies between 2.51 mW and 2.61mW depending on the operating conditions. The energy per conversion-step is 1.43-1.49 pJ/step. A 180 nm process was used for the fabrication, requiring 1.17 mm² (including another oscillatory ROIC).

An innovative ROIC with a 10-bit ADC is presented in [53]. The sensor is biased with an IDAC with 4 coarse subranges and 14 fine subranges. The input resistance range is split into subranges that increase exponentially. To get the appropriate subrange, the logarithm of the resistance must be calculated, for this, a comparator operating near its metastable point is used. The starting sensor current is 300 nA, then the logarithmic subrange detector sets the optimal coarse and fine subranges in 1 cycle. The ASIC was fabricated in a 180 nm process achieving a power consumption of 1.7 μ W with an EPC of 1.42 nJ. The resistance range is from 10 k Ω up to 10 M Ω . The maximum conversion rate is 1.2 kS/s. Throughout the whole input range, the measurement error is below ± 0.2 %.

A multi-sensor ROIC is described in [49] where current or resistive sensors can be interfaced. For current sensors, a dual-mode current mir-

3.3 Analog-to-Digital Converter ROICs

ror with two selectable copy ratios is used to improve the performance when small currents are being measured. This is followed by a coarse self-cancellation scheme with an 8-bit IDAC and an 8-bit binary search algorithm with a SAR control loop. Finally, the remainder current is converted by a $\Delta\Sigma$ current detector. For low resistance sensors, a second ROIC uses a correlated-double-sampler to reduce offset and noise, where two 8-bit IDACs are embedded in a SAR loop. The sensor is placed in the feedback loop of a transimpedance instrumentation amplifier whose output is converted with an 8-12 bit dual-mode two-step incremental SAR. For high resistances, the sensor is reconnected to a potentiostat, that is already available for electrochemical sensors, where the working electrode is used for biasing, then the signal follows the same path as current sensors. The ROIC has been tested with resistances in the range of 7 k Ω up to 50 M Ω , from 7 k Ω up to 120 k Ω the CDS+ADC ROIC is used, then from 120 k Ω up to 1 M Ω the $\Delta\Sigma$ -detector ROIC is employed in high-current mode and then from 1 M Ω up to 50 M Ω in low-current mode. The power consumption is 0.425 mW for the $\Delta\Sigma$ -detector and 0.5 mW for the CDS+ADC. The SNR and DR are 91.9 dB (@ 7 k Ω) and 116.7 dB for the ADC+CDS and 108.6 dB (@ 120 k Ω) and 160.9 dB $\Delta\Sigma$ -detector.

Magnetoresistive sensors are capable of detecting magnetic nanoparticles and can be applied for immunoassay as presented in [54], where an interface for these type resistive sensors is developed. The low-noise ROIC contains a fixed-current bias block, a capacitively coupled instrumentation amplifier (CCIA), and a zoom ADC. The CCIA is differential, where the non-inverting input is connected to the sensor and the inverting to a reference voltage. A servo loop cancels the baseline resistance of the sensor and a ripple rejection loop is used to diminish the chopping ripple. The zoom ADC can operate with a 6-bit SAR or with a 13-bit $\Delta\Sigma$ modulator for coarse or fine quantizations respectively. As the magnetoresistive sensor has a predictable resistance, the ROIC has been designed to work only at 1.3 k Ω . The input referred noise is 87.3 $\mu\Omega_{RMS}$, resulting in an ENOB of 23.8 bits. The power consumption is 4.32 mW and the EPC 3.04 mJ. Each channel requires 1.92 mm² in a 180 nm process.

A 20-bit ROIC with 0.04 % gain error for bridge transducers is described in [55]. Their circuit uses a current-feedback instrumentation amplifier to amplify the Wheatstone bridge differential output voltage, followed by a switched-capacitor incremental $\Delta\Sigma$ ADC. Chopping is used to reduce the flicker noise and offset of the instrumentation amplifier and dynamic element matching to adjust the gain of the transconductors. The ROIC is

3 Readout Circuits for Resistive Sensors

designed to interface general bridge transducers, like strain gauges or thermistors. The current consumption is $270\text{ }\mu\text{A}$, at 5 V , with an EPC of $229.5\text{ }\mu\text{J}$.

Some of the published ROICs, like that present in [56], propose to implement only the analog part in the AFE and use an external off-the-shelf ADC (or the ones available in microcontrollers) to improve efficiency and reduce design complexity. Their design uses the well-proven amplifier with chopper scheme and an IDAC to bias the sensor in a resistance range tracker configuration. The power consumption is between 90.2 nW and $4.616\text{ }\mu\text{W}$, plus 9 nW for a 6-bit ADC. Following the same approach, Moayer et al. [57] designed a ROIC for resistive and capacitive sensors, the input range from $3.7\text{ k}\Omega$ up to $5.1\text{ M}\Omega$ is adequate for chemiresistive sensors, and the power consumption is low at $3.56\text{ }\mu\text{W}$ (maximum), the ADC is again not included in the ASIC.

Spectroscopy ROICs are capable of interfacing resistive sensors if they can produce DC excitation currents, but are in general not favoured due to their higher power consumption. A ROIC for electrochemical sensors which can perform spectroscopy up to 200 kHz is presented in [58]. This ROIC has also a potentiostat to measure DC currents. The impedance measurement consumes 20.6 mW while the potentiostat draws $8\text{ }\mu\text{A}$. In [59] another spectroscopy ROIC is presented for biological tissue. Two IDACs ($1\text{ }\mu\text{A}$ to $100\text{ }\mu\text{A}$) bias the sample and the voltage drop is measured with an instrumentation amplifier followed by a 12-bit SAR ADC. The power consumption is $155\text{ }\mu\text{W}$ at 250 kSps .

Other publications of lesser relevance are that of [60], [61], where similar AFEs for measuring impedances were proposed including current sources to bias the DUT, the power consumption is $152.3\text{ }\mu\text{W}$ and $81.6\text{ }\mu\text{W}$ respectively. An AFE without a specific application is described by Badami et al. [62], with a low power consumption at only $1.8\text{ }\mu\text{W}$. In [63] a single slope ADC is used to interface chemiresistive sensors. The sensor is biased with a series resistor connected to the supply voltage, giving a single operating point. In [64] an ultra-low-noise bridge-to-digital ROIC with digital compensation is implemented. In this work, a TIA with chopper and a continuous-time $\Delta\Sigma$ modulator are used. The offset is corrected by means of capacitive DAC. Another bridge-to-digital ROIC was published in [48]. Even though the general architecture follows other designs, the power consumption is reduced by sampling the bridge output voltage for a short time.

Lastly, there are ROICs still being designed with discrete off-the-shelf

components soldered on a PCB. We will not dive into them as the power consumption is several orders of magnitude above that of ASIC designs. As an example, a ROIC for an array of chemiresistive sensors is described in [65] consuming 150 mW at 5 V.

3.4 Time-based ROICs

Time-based ROICs use an oscillator or a triggered circuit to produce a signal with a period correlated to the sensor resistance. The design requirements can be tuned to match the sensor performance without having to design a general-purpose ADC. As mixed-signal circuits, some of the building blocks, like counters, differentiators and filters can be implemented in the digital domain making the design more portable between CMOS processes and integrated in the digital design flow.

The building block of most oscillatory ROICs is a VCO or a CCO, in which the output frequency is proportional to the analog input voltage with an offset (called the quiescent frequency). Among all implementations, a notable subclass are the relaxation-oscillator-based ROICs, because they allow for a wide-range adjustment of the oscillation frequency. In all cases, the output is digital in amplitude but with an analog phase like in pulse-width modulation (PWM) or pulse-frequency-modulation (PFM).

3.4.1 Oscillatory ROICs

In open-loop oscillatory ROICs, Fig. 3.4 (a), the CCO is driven by an input current generated from the biasing, filtering and amplification of the sensor. The CCO phase is quantified by a counter followed by a differentiator, running at a sampling frequency f_s , lower than that of the CCO. For closed-loop ROICs, the structure is the same, but an integrator is added at the output which in turn drives a DAC to produce an offset of the sensor signal, Fig. 3.4 (b). In this way, the input range of the CCO is reduced.

The CCO behaves like an integrator in the frequency domain, so with the differentiator the system resembles a $\Sigma\Delta$ modulator [66]. Hence, first-order (quantization) noise shaping is obtained. The signal and noise transfer functions for open and closed-loop ROICs are: 3.7, 3.8, 3.9, 3.10 as calculated in [66]. Where, k_{CCO} and f_0 are the gain and the free running frequency of the CCO respectively, and G_{DAC} is the gain of the IDAC. In this chapter we concentrate on the sensor biasing and operation of the CCO, leaving the quantization of the period to chapter 5 where the implications

3 Readout Circuits for Resistive Sensors

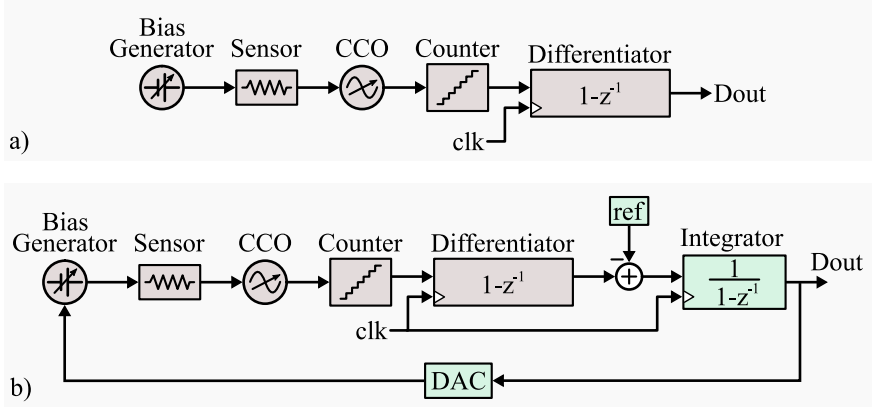


Fig. 3.4 a) Open-loop oscillator ROIC structure. b) Closed-loop oscillator ROIC.

of the digital interfacing are studied.

$$STF(z)_{OL} = k_{CCO}2\pi f_0 \quad (3.7)$$

$$NTF(z)_{OL} = 1 - z^{-1} \quad (3.8)$$

$$STF(z)_{CL} = \frac{k_{CCO}2\pi f_0}{1 - z^{-1} + G_{DAC}k_{CCO}2\pi f_0 z^{-1}} \quad (3.9)$$

$$NTF(z)_{CL} = \frac{1 - z^{-1}}{G_{DAC}k_{CCO}2\pi f_0} \quad (3.10)$$

Two of the most relevant VCO-based ROICs are those presented by Sacco et al. in [67] and [68]. The former employs a VCO, followed by a PFD, whose outputs drives a digital-controlled oscillator. The output of the second oscillator is used by a digital core to perform count and differentiation operations. The sensor is biased with a 4-bit IDAC that is driven by the output value of the ROIC. This ROIC is capable of interfacing two resistive sensor simultaneously, e.g. a temperature sensor and a pressure sensor. The resistance input range is from 5 k Ω up to 20 k Ω . The core power consumption is 109 μ W and the EPC 2.73 nJ or 54.5 nJ depending on the resolution selected. The SNR is 83.6 dB and 96.9 dB for each case.

The ROIC of the second publication [68], closely resembles the architecture depicted in Fig. 3.4 (b) but employs a second loop in sturdy multi-stage configuration to produce second-order noise shaping. In this case, the input range is from $20\ \Omega$ up to $20\ \text{k}\Omega$. The core power consumption is $109\ \mu\text{W}$ and the EPC $34.2\ \text{nJ}$ at a 89 dB resolution. Both ROICs were fabricated in a 180 nm processes taking $0.1\ \text{mm}^2$ and $0.064\ \text{mm}^2$ of silicon area respectively.

Lee et al. [69], presented a ROIC for the detection of benzene, toluene, ethyl-benzene, and xylene with a resistive MEMS sensor. The aforementioned sensor is biased with a differential voltage regulation loop implemented with symmetrical NMOS transistors driven by voltage-follower amplifiers that track a reference voltage. One of these circuits is used for each of the two sensor terminals, only differentiated by the value of the reference voltage (the sensor is connected to the sources of the two branches). A digitally tunable resistor (RDAC) is also connected between each terminal of the sensor and ground. The input voltage of the VCO is the differential voltage between the drain terminals. The RDAC is controlled by a feedback loop from the output to keep the drain voltage difference within a given range without affecting the sensor current. The VCO contains two 15-stages current starved ring oscillators which are driven by differential-input-output transconductors. The output phases of the VCOs are sampled with a 16-level phase quantizer. In terms of performance, the power consumption is $95\ \mu\text{W}$, with an EPC of $47.5\ \text{nJ}$. The input resistance range is from $20\ \text{k}\Omega$ up to $500\ \text{k}\Omega$ and the resolution 18.8 bits. It was implemented in a 110 nm process, taking $0.093\ \text{mm}^2$.

In [70], two parallel VCOs are again used with a binary phase-detector. In this paper, the sensor is placed in one of the branches of a Wheatstone bridge. The other branch is implemented with a RDAC that is driven by the output signal of the ROIC passing through a proportional-integral controller. The goal of this RDAC is to minimize the voltage difference between the two branches of the Wheatstone, because the bridge is ideally unaffected by supply variations or noise and hence has an infinite PSRR at this point. The ASIC, implemented in a 130 nm process, has a power consumption of $124.5\ \mu\text{W}$, an EPC of $6.23\ \text{pJ}$. The circuit was designed for temperature sensing, hence the sensor resistance is fixed at $10\ \text{k}\Omega$.

A high-dynamic-range VCO ROIC is described in [71]. Their use case includes electrode-sensing, bio-potentials and motion artifacts therefore they do not include a sensor biasing circuit. The differential input is chopped and amplified, followed by two VCOs with a phase detector for

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each of them. The phase difference is fed to a gain error correction logic. Differential pulse-code modulation is applied to reduce the amplitude of the input signal to the VCO's quantizer, resulting in a very low THD of -112 dB. It was fabricated in a 65 nm process taking just 0.08 mm². Dynamic element matching was employed to reduce the impact of mismatch. The total power consumption is 3.2 μ W, the SNR 89.2 dB and DR 94.2 dB.

As noted in the beginning of this section, relaxation-oscillators-based ROICs share the same architecture as other time-based ROICs, but the oscillator is of relaxation type. The operation is explained in Fig. 3.5 (a). On the left side, a voltage regulator, formed by a transconductance amplifier U1 (or an operational amplifier) and a MOS transistor Q, sets the bias voltage of the sensor equal to an external reference voltage, V_{REF} . The current flowing through Q is copied and divided by a factor δ to feed the current-controlled relaxation oscillator (CCRO). The CCRO is made of a comparator U2 with hysteresis voltage V_{HIST} , a logic inverter U3, an integration capacitor C and two control switches: S1 and S2 which transfer the current to or out of C. When the comparator output (V_x) is low, S1 is open and S2 closed, thus the capacitor is discharging until its voltage reaches the lower threshold of the comparator's hysteresis. At this point, the output of U2 becomes high, S1 is closed, S2 opened, and the capacitor begins charging. This repeats as long as the circuit is operated, Fig. 3.5 (b). The period, equation 3.11, can be calculated from the equation of the capacitor and the circuit parameters, where $V_{HIST} = V_H - V_L$. Lastly, an integrated or external counter (e.g. from a microcontroller) transforms the period length into a digital value.

$$T = \frac{2\delta RC(V_H - V_L)}{V_{REF}} = \psi R \quad (3.11)$$

This kind of oscillatory ROICs has been employed mainly for high-dynamic-range applications like gas sensing. For example, in [72] a 128 dB DR ROIC for MOX gas sensors is presented and tested. The sensor is biased with a classical voltage regulator where an operational amplifier drives the gate of a NMOS transistor. The output current of the voltage regulator, that is equal to the reference voltage divided by the sensor resistance, is divided and feed to a relaxation oscillator. One P and one N cascoded-regulated current mirrors alternate to either charge or discharge the integration capacitor. The capacitor is part of an opamp integrator to keep the drain voltage of the current mirrors constant at a given common-mode voltage. The triangular output levels are set by the reference voltage

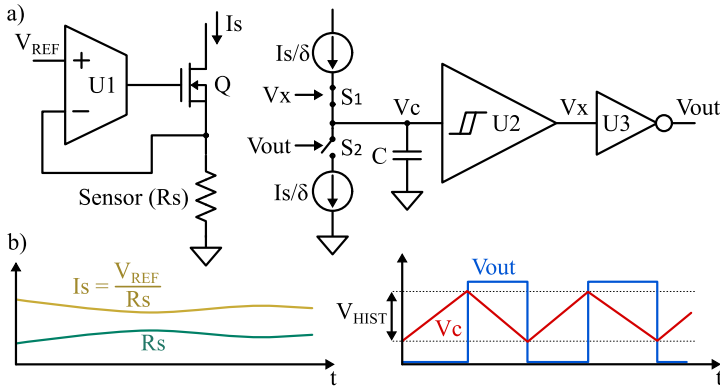


Fig. 3.5 a) Open-loop relaxation-oscillator-based ROIC core components.
b) Current and voltages at different nodes.

of two comparators that switch the active current mirror and consequently the direction of charge flowing into the integration capacitor when each threshold is reached (V_H and V_L). A SR latch holds the current configuration of the system. Finally, the period of the modulated digital signal is quantified (externally) with a counter. The resistance input range is from $100\ \Omega$ up $1\ \text{M}\Omega$. The power consumption is $68\ \text{mW}$ and the EPC is $150\ \text{pJ}$ for $100\ \Omega$ and $675\ \text{nJ}$ for $1\ \text{M}\Omega$. A $130\ \text{nm}$ CMOS process was used for the fabrication requiring $0.125\ \text{mm}^2$.

A similar architecture was presented in [73] for chemiresistive gas sensors. In this ROIC, there is also an operational amplifier integrator but the generation of the sensor current is different. The capacitor is connected between the output and the inverting input of the operational amplifier, while the non-inverting is kept at $V_{DD}/2$. The sensor is connected with a limiting resistor in series, to the non-inverting input as well. The other terminal of the sensor is either connected to V_{DD} or to ground, for charging or discharging the capacitor. The benefit of this approach is that the current through the sensor is not wasted to ground. The resistance input range is from 100 Ω up to 4.7 M Ω . The power consumption is 0.37 mW and the maximum EPC 296 nJ for a 4.7 M Ω sensor. A 350 nm CMOS process was used for the fabrication requiring 0.24 mm².

Two ROICs have been combined in [52] to operate the system in two modes: a) monitoring (with a relaxation-oscillator) or b) high precision

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(with an ADC and an IDAC). In monitoring mode, it presents an EPC of 89.5 pJ but the SNR is restricted to 46.3 dB. The input range is from 250 k Ω up to 2 M Ω .

3.4.2 Other Time-Based ROICs

Some researchers have proposed ROICs which are still time-based, but do not follow the oscillator-based architectures described above. Here we describe some relevant publications that fall within this category. Woo and Yang [50] proposed a discharge time ROIC with the sensor embedded in a RC discharge circuit with an integrated capacitor connected to a comparator. The capacitor is charged to a known voltage, a timer is cleared when the discharge begins and the stopped when the voltage in the tank capacitor reaches that of a given reference voltage, hence the comparator toggles. The initial discharge voltage is fixed at 0.3 V and the counter frequency at 5 MHz. The input range is from 1 k Ω up to 100 k Ω with an SNR of 65.6 dB. The footprint is 2,500 μm^2 . A 65 nm CMOS process was used for fabrication.

A similar design to that of Woo and Yang is presented in [74]. In this publication the sensor is biased with a constant current and is placed in series with a diode-connected PMOS to copy the current through the sensor. This current is then integrated in a capacitor and is compared to a reference voltage. An identical circuit is constructed but using a resistive DAC instead of the sensor. The output of the two comparators are connected to a flip flop and a SAR Logic. The SAR Logic compares the time delay between the circuits and finds the value of the RDAC that minimizes this delay. The benefit of using this approach is that it requires a counter clock frequency N times higher than that of sampling frequency to resolve N bits, unlike [50], where 2^N are needed. The power consumption is 15 μW with a sampling rate of 5.5 kS/s, giving an EPC of 2.73 nJ with an ENOB of 7.6 bits. The ASIC was fabricated in a 180 nm process, taking 0.057 mm².

Up to this point, all the time-based ROICs studied, require a counter (or timer) to produce a digital representation of the time event (oscillation period or discharge time). For the case of oscillatory ROICs, a differentiator is sometimes added at the output to generate an estimation of the frequency from the phase (output of the non-resetting counter) or alternatively a timer module with built-in auto reset function can be used. The implementation of these circuits is purely digital and will be discussed in more detail in chapter 5.

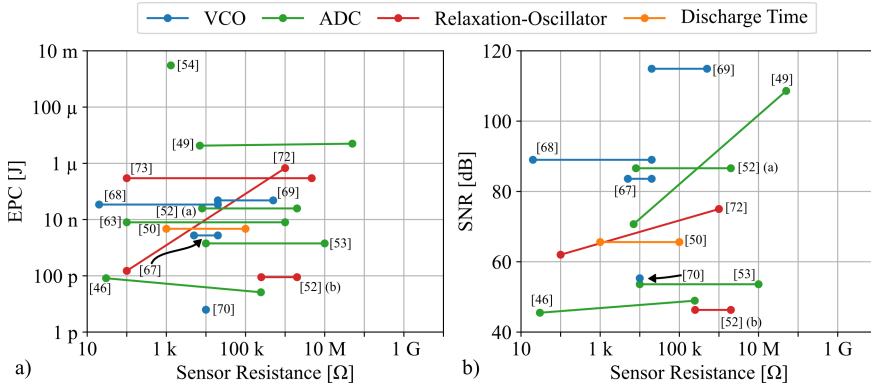


Fig. 3.6 a) EPC for several published ROICs separated by types. b) SNR for several published ROICs separated by types.

3.5 ROIC Discussion

Even though there are ample proposals for resistive sensor interface circuits, only a small portion of them are designed specifically for gas sensing applications, as seen in the reviewed publications.

Combining the results of the analyzed ROICs, Fig. 3.6 presents the plots of the EPC and SNR as a function of the resistance input range. With respect to ADC-based readout circuits, they offer good linearity but struggle to achieve simultaneously high dynamic range and low power consumption, largely explained by the large number of bits needed both in the ADC and IDAC, those where the input range is large enough and the EPC low, the SNR is just above 40 dB.

Closed-loop oscillatory ROICs have a low gain error as they use the output frequency to adjust the sensor bias current and make the VCO/CCO operate at a fixed input voltage, avoiding the usual nonlinearity in the voltage-to-frequency conversion found for larger input swings. On the other hand, achieving a low EPC is a challenge, due to the usage of digital signal processing. In [68] where a VCO-based ROIC with a second loop in sturdy multi-stage noise shaping configuration is described, the power consumption of the ROIC is $171 \mu\text{W}$ with a conversion time of 1 ms. The sensor biasing takes at most $40.32 \mu\text{W}$, therefore the ROIC consumes $130.68 \mu\text{W}$ which corresponds to an EPC of 130.68 nJ . Relaxation-oscillator-based ROICs are simpler and offer the benefit of separating the biasing of the sen-

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sor from the rest of the circuit, imposing no restriction on the sensor bias point. As seen in Fig. 3.6, they can also achieve a low power consumption.

Most of the reviewed designs were fabricated in CMOS process nodes between 130 and 350 nm, with some exceptions in the ADC group where a 65 nm process was used. This is important to consider because, ADC and VCO ROICs may reduce the power consumptions by employing a smaller CMOS process, if the reduction of the supply voltage to 1.5 V or 1.2 V is acceptable.

Other architectures, like the triggered ROICs of Woo and Yang [50] achieved a low power consumption, between 461.5 and 543.5 nW but their design is limited to a maximum sensor resistance of 100 k Ω and the sensor bias voltage is not kept constant. The latter is especially inconvenient because conductive polymers are usually not ohmic materials [4], therefore, the sensors resistance value changes during discharge and the estimation of the gas concentration from the discharge time is not obtained from the RC equations.

Some authors, like Park et al. [52], have not relied on a single approach. In their paper, they describe a multi-sensor system employing one relaxation-oscillator ROIC and one ADC ROIC, to achieve low power or high resolution as needed, even though the relaxation oscillator achieves a power consumption of just 2.95 μ W (low resolution mode), it sharply increases to 2.51 mW when the high-performance ROIC is used.

Owing to the large variability of baseline resistance in sensors made with PPy, PANI or graphene but also in legacy technologies like MOx [75], we propose to use open-loop relaxation-oscillators for chemiresistive gas sensors as they offer at the same time a wide input range, bias flexibility to interface multiple sensors sequentially [73], [72], [52] and a low EPC [76]. The main obstacle they present, is the large variability of the measurement time, depending on the sensor's resistance. For example, in [77] a readout for sensors with resistances between 100 Ω and 10 M Ω is presented, with an oscillation period ranging from 100 ns to 10 ms respectively (all other parameters remaining equal). Such wide range results in a very distinctive power consumption profile where these ROICs are highly efficient for short periods but become inefficient as the period increases and static power consumption becomes dominant.

In the rest of this chapter, we will present a proof of concept (POC) of a relaxation-oscillator-based ROIC to validate the architecture and the conclusions drawn from the literature review. In the next chapter, we will analyse the efficiency degradation of open-loop relaxation-oscillator ROICs for

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

high resistance sensors and propose two adaptive biasing techniques to improve power consumption that do not require external supervision and can retain the desirable input characteristics needed for autonomous multi-sensor monitoring systems.

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

3.6.1 ROIC Description

To validate the capacity of relaxation-oscillator-based ROICs to interface chemiresistive gas sensors, we designed and fabricated a ROIC, named hereafter POC (proof of concept), with the goal of achieving a high dynamic range and a tuneable bias voltage to allow for the experimentation with different sensors. The schematic of the circuit is presented in Fig. 3.7. It builds on the same general principles as other relaxation oscillators [72], [77], but its components are designed and optimized for a multi-sensor application, with the following requirements:

- Input resistance between 1 k Ω and 33 M Ω , which allows to interface at least some types of sensors made with PPy, PANI or graphene with the same ROIC.
- 1.8 V supply voltage ($\pm 5\%$).
- Maximum oscillation frequency of 1 MHz, to ease period quantization.
- Sensor biasing voltage (V_{ref}), between 50 mV and 1 V (chosen so that $R \cdot I_{Sensor}$ remains below 60 μA to limit power consumption). Each sensor bias voltage is set according to the application requirements with beforehand knowledge of sensor's characteristics.

As noted in chapter 2, sensors biased with larger currents require short measurement times, e.g. hundreds of nanoseconds. This means that transistors in the comparators and operational amplifiers must be biased in strong inversion where they are least efficient, therefore degrading the overall EPC. Consequently, the lowest sensor resistances are limited in the bias voltage that can be used to safeguard the EPC elsewhere.

V_{ref} is generated by a voltage reference circuit made with an NMOS transistor, Q_{1A} , driven by an operational transconductance amplifier (OTA).

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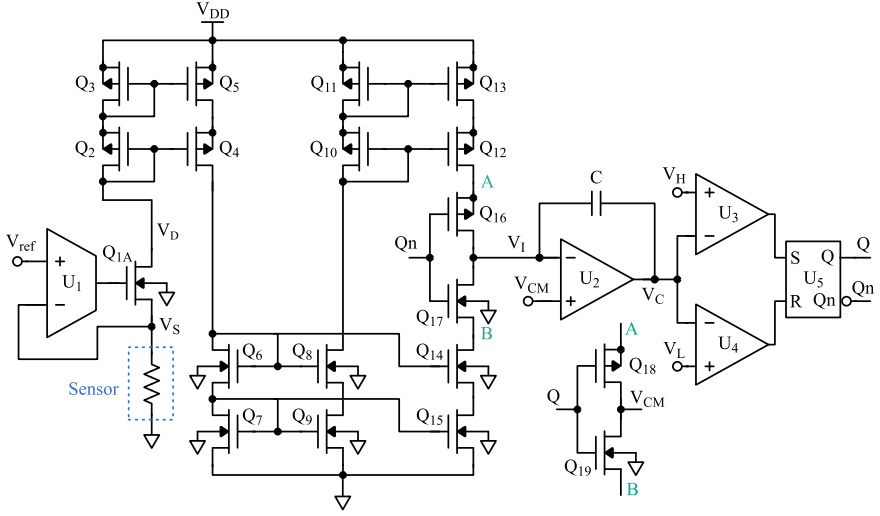


Fig. 3.7 Schematic of the proposed relaxation-oscillator based ROIC.

The current flowing through the sensor and Q_{1A} is copied and divided by a factor $\gamma = 10$ by the asymmetrical cascode current mirror $Q_{2-3-4-5}$. This current is then copied by NMOS and PMOS cascoded current mirrors to feed the integration capacitor C . $Q_{2-3-4-5}$ are all low V_T transistors to reduce the voltage drop in the mirror and increase the maximum sensor current achievable.

The complementary outputs Q_n and Q of a set-reset (SR) latch, U_5 , control the switches $Q_{16} - Q_{17}$ and $Q_{18} - Q_{19}$ that respectively connect the current mirrors to the input of the integrator, or to a constant voltage V_{CM} to diminish charge injection in the capacitor due to charging and discharging of the drain-referred parasitic capacitances. When Q_n is high, the NMOS cascode output is connected to the integrator and the PMOS is connected to V_{CM} , when Q_n is low the connections are permuted.

Assuming the sensor resistance is almost constant during each measurement cycle, the resulting current is DC, thus the voltage at the output of the integrator raises or decreases with a constant slope until it reaches V_H or V_L . At this point, the comparators U_3 or U_4 toggle as also does the latch U_5 , changing the direction of the current that is integrated in C . As a consequence, the waveform at the output of U_2 is a triangle wave where the period T is determined by 3.11. The value of C is 3.03 pF.

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

The biasing circuit is designed to provide up to 60 μA to the sensor, higher currents would result in higher oscillation frequency which would be harder to digitalize with a counter or timer. Finally, the SR latch stores the current state of the switches and generates the digital output (with sufficient drive capability) for the output pad. For the measurements of the period, we used an external timer.

3.6.2 CMOS Process Selection

For the implementation of the integrated circuit, we considered three different technologies: 250 nm and 65 nm from TSMC and 180 nm from UMC. The core voltages are: 2.5 V, 1.2 and 1.8 V respectively. For the current mirrors and the OTA, the benefit of using a smaller node is marginal, the performance of same size transistors in the three technologies is similar. There is an improvement in the Pelgrom's constants, positively affecting mismatch and linearity, for same size transistors (if a smaller process is used). In simulations, a DC gate leakage current of up to 2pA was observed for the 65 nm node for a $W = 10\text{ }\mu\text{m}$, $L = 10\text{ }\mu\text{m}$ NMOS device [78]. In the case of 180 nm and 250 nm transistors, the DC gate leakage was not modelled (virtually 0 A) and no reference to it was found in the documentation.

Other components of the ROIC, like operational amplifiers and comparators could benefit from a smaller node as they could achieve higher frequencies at the same power level, and this results in a lower energy per conversion (EPC). From simulations, to obtain an f_t of 1 GHz with a minimum length transistor, the 250 nm node requires 150 nA of bias, the 180 nm node 100 nA and the 65 nm node 35 nA approximately. Given that there is a need for a bias voltage of at least 1 V for the sensors, we decided to use the UMC RF-CMOS 180 nm technology for the implementation which provides the best trade-off between power, maximum oscillation frequency and cost. Additionally, using a supply voltage of 1.2 V is challenging for the sensor biasing at 1 V. Low-threshold voltage transistor have a V_T of over 0.3 V, complicating the implementation of current mirrors.

3.6.3 Sensor Bias

The sensor biasing is implemented with the OTA of Fig. 3.8, driving an NMOS transistor connected in a feedback loop, Q_{1A} in Fig. 3.7. The goal of this stage is to cover a wide range of sensor resistance and reference voltage combinations to accommodate dissimilar sensors, or to optimize

3 Readout Circuits for Resistive Sensors

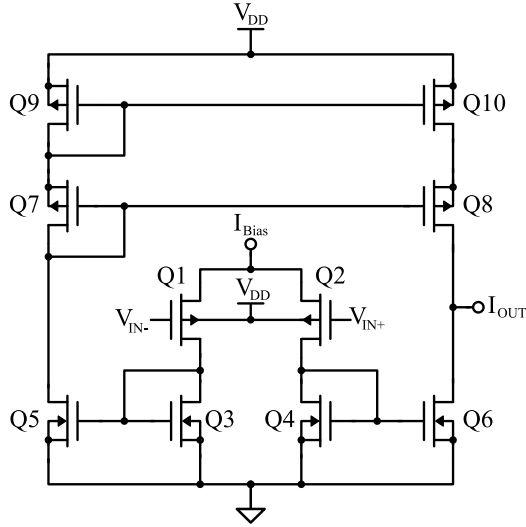


Fig. 3.8 Schematic of the transconductor used for the generation of the bias voltage, named U1 in Fig. 3.7.

the operating point of a particular sensor. Low threshold voltage PMOS and NMOS transistors were used, except for Q_{1A} to decrease leakage and decrease the output swing required in the operational amplifier.

To be able to drive the gate of Q_{1A} at low voltages (for small sensor currents), the current mirrors $Q_5 - Q_3$ and $Q_4 - Q_6$ are made wide ($W = 10 \mu\text{m}$, $L = 10 \mu\text{m}$) and the bias current, I_{Bias} is kept low ($5 \mu\text{A}$), to provide adequate transconductance in the input differential pair but to require the least overdrive voltage in the current mirrors. The differential pair transistors size is $W = 10 \mu\text{m}/L = 1 \mu\text{m}$ and they are biased in weak inversion to increase the gain efficiency, the large width also reduces the effects of mismatch and consequently offset. The highest voltage required at the gate of Q_{1A} , for the smallest R_{Sensor} , is considerably lower than V_{DD} , this allowed the usage of a PMOS cascode current mirror to reduce flicker noise and increase the output impedance of the OTA. The circuit is stable across the whole input range, having a phase margin of at least 45 degrees, as obtained from simulations.

As discussed in chapter 2, the bandwidth of interest for gas sensing is very low, below 100 Hz, thus it is essential to reduce flicker noise. Firstly, by means of using large transistors in the signal path, Q_{1A} is $W = 30 \mu\text{m}$, L

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

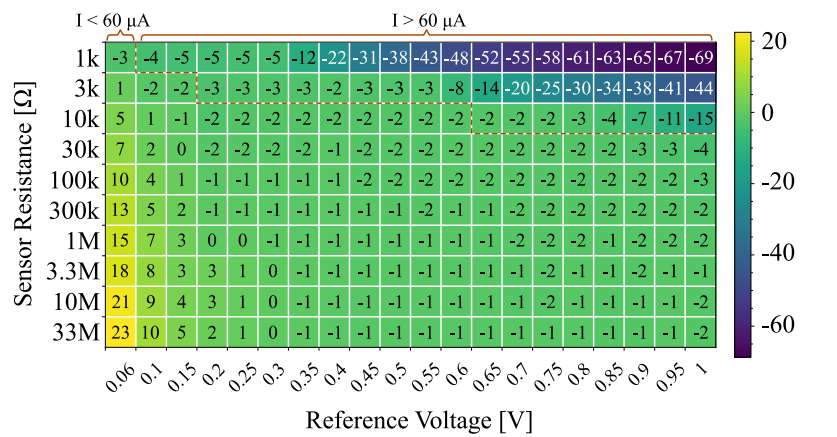


Fig. 3.9 Heatmap of the mismatch between the sensor and reference voltage.

= 1 μm and the PMOS cascode current mirror ($Q_2 - Q_3$) in Fig. 3.8 are $W = 800 \mu\text{m}$, $L = 15 \mu\text{m}$ on the left branch and $W = 800 \mu\text{m}$, $L = 15 \mu\text{m}$ in the right branch. Secondly the low bias current of the OTA helps to reduce noise even more, by decreasing the bandwidth of the feedback loop, attenuating undesired high frequency noise.

The difference between the sensor voltage to the reference voltage was measured for the complete sensor resistance and reference voltage range and is plotted in Fig. 3.9 achieving good results within the design range ($I < 60 \mu\text{A}$). Two lateral regions showed significant error. The first, outside the design range, is where the sensor resistance is low and the bias voltage high resulting in a large current that demands a large overdrive voltage in the PMOS current mirror $Q_{2-3-4-5}$ of Fig. 3.8 that cannot be achieved. The second is where the sensor resistance is highest and the bias voltage lowest, in this region leakage currents play an important role and affect the sensor current significantly. Low voltages are also the most affected by offset. The offset contributes to the nonlinearity in the overall resistance-to-period transfer function.

3.6.4 Integrator and Comparators

To keep the voltage at the output of the current mirrors constant and decrease the output impedance for driving the comparators, the integration

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capacitor is connected in the feedback loop of an operational amplifier U_2 , Fig. 3.7. The amplifier follows the traditional architecture: an OTA connected to a PMOS common source output stage, Fig. 3.10. This amplifier provides sufficient output and input voltage swing. For a more compact design and lower power consumption, a Miller operational amplifier can be used, if the input detection voltage where the comparators will operate is compatible with the output swing. To guarantee stability, a Miller capacitor of 1.1 pF is added between the output of the OTA and that of the common source. The bias current of 10 μ A, is chosen to provide the necessary frequency for the smallest sensor resistance, nominally 641 kHz. As the signal at the output of the comparator is ideally triangular, a large bandwidth overhead was needed to avoid smoothing of the discontinuities in the waveform, the achieved values for all corners are included in Table 3.1. The differential pair ($W = 8 \mu\text{m}$, $L = 4 \mu\text{m}$) size was chosen to balance mismatch while not decreasing f_t excessively. The input common mode range is small as both inputs are close to V_{CM} which is equal to $V_{DD}/2$, (0.9 V). The output range is from 0.3 V up to 1.4 V. For the highest frequency, the opamp must also have sufficient slew rate to generate the triangular waveform. The slew rate was simulated for all corners and input steps with different amplitudes, Fig. 3.11. The largest required slew rate of 1.3 V/ μ s generates a difference between the inputs of 60 mV in the typical case and 70 mV in the worst case.

Table 3.1 Operational Amplifier Transition Frequency (MHz).

Typical	Slow, Slow	Fast, Fast	Slow N, Fast P	Fast N, Slow P
2.80	2.46	3.28	2.76	2.86

Comparators U_3 and U_4 have three stages. The first one is an active-load-NMOS differential pair, which is followed by two cascaded minimum-sized inverter stages, providing minimum delay, Fig. 3.12. The bias current for the differential pair is 20 μ A. V_H and V_L are 0.339 V and 1.16 V respectively. Mismatch in the differential pair or the current mirror is not critical as it will appear as a constant offset, or uncertainty in the threshold voltage, but it will affect all frequencies proportionally.

Large decoupling capacitors were placed in the supply and threshold voltages to reduce supply ripple, which could have created interference. Especially in the current mirrors where it could couple to the signal by V_{DS} variations.

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

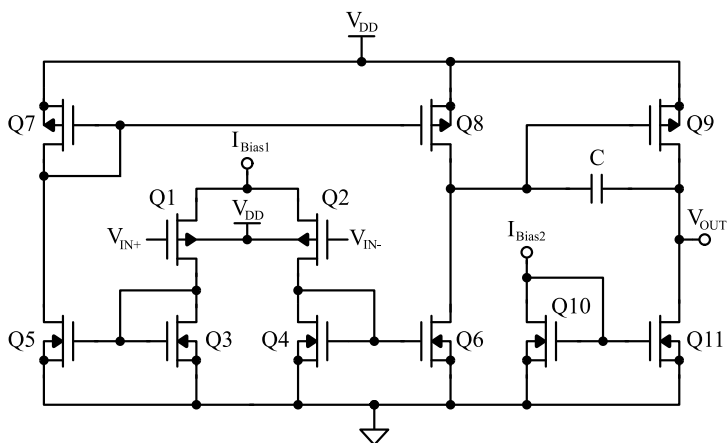


Fig. 3.10 Schematic of the operational amplifier used to charge or discharge the integration capacitor, named U2 in Fig. 3.7.

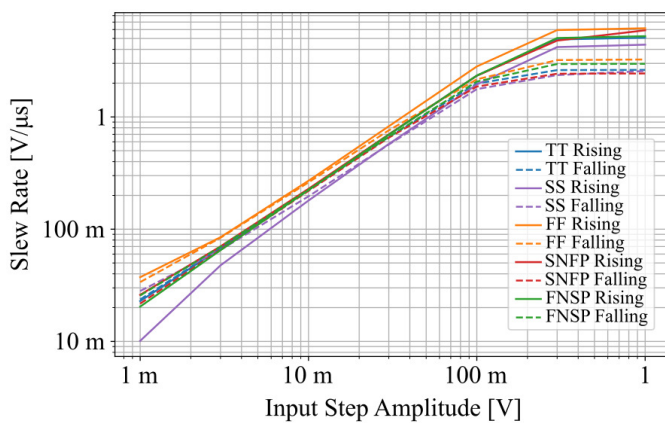


Fig. 3.11 Operational amplifier slew rate as a function of input differential voltage for all corners.

3 Readout Circuits for Resistive Sensors

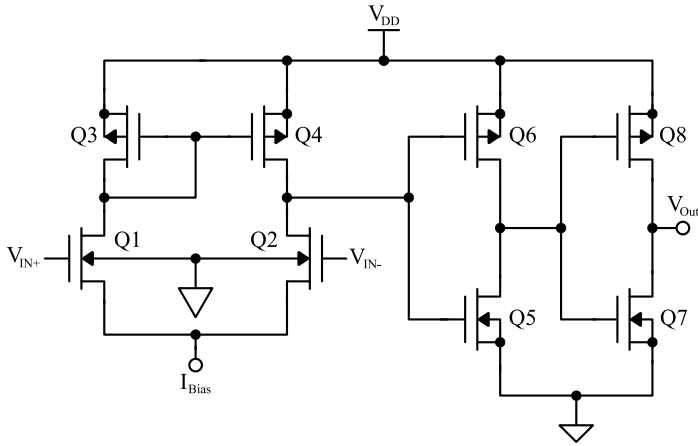


Fig. 3.12 Schematic of the comparator that detects the thresholds to toggle the integration direction, named U3 in Fig. 3.7.

The delay of each comparator, U_3 corresponding to the higher threshold voltage and U_4 to the lower, are depicted in Fig. 3.13 as a function of the input slew rate. In Fig. 3.14, the delay is expressed as fraction of the period. The delay of U_4 is intentionally larger than that of U_3 , this is done to avoid a race condition when the triangle signal (output of the integrator) reaches V_L , then the output of U_4 toggles to high level. Consequently, the SR latch begins the transition to the reset state, and the switches $Q_{16-17-18-19}$ toggle, changing the direction of current in the integration capacitor. As the capacitor is discharged, the comparator reverses to the low state but the SR latch has not completed the transition to the other state yet, therefore it can bounce back to the set state resulting in missed transitions and increased jitter or malfunctioning.

3.6.5 Reference Voltage Generation

Three voltage references are needed for the ROIC, one adjustable for the reference input, which is used to set the sensors biasing point, and two fixed for the comparators thresholds of 0.339 V and 1.16 V.

For the reference voltage we decided to use a simple digital-to-analog converter based on pulse width modulation (PWM). It provides a sufficiently small voltage step and due to the fact that the voltage is applied to a MOS gate (that behaves like a DC open circuit), the generated volt-

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

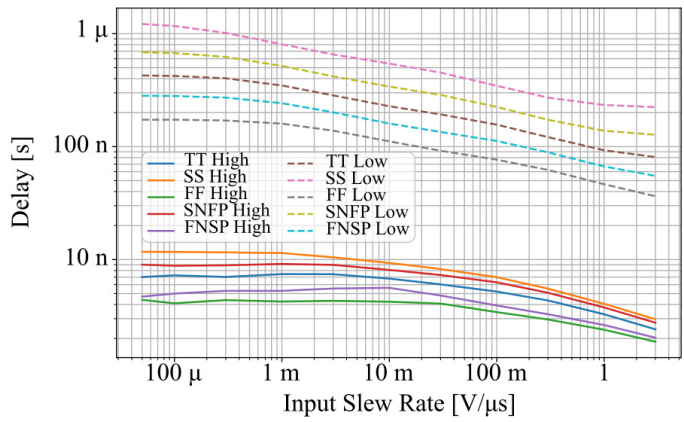


Fig. 3.13 Comparator delay as a function of slew rate for all corners.

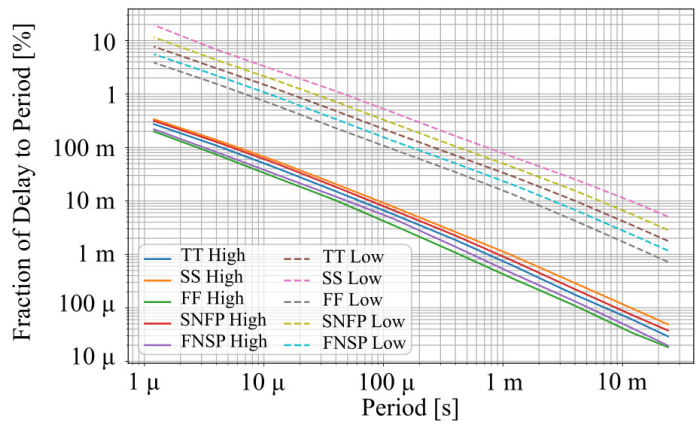


Fig. 3.14 Fraction of delay to oscillation period for the entire oscillation range.

3 Readout Circuits for Resistive Sensors

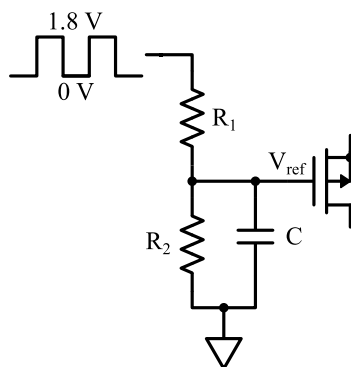


Fig. 3.15 Reference voltage generator.

age is not affected by the load. The schematic is presented in Fig. 3.15, it consists of a resistor divider connected to a capacitor and to the reference input, which is the gate of a PMOS transistor. The generated voltage by the divider is unaffected by temperature variations as both resistors have the same temperature coefficient. The input of the resistor divider is connected to a microcontroller. Most modern microcontrollers, even those for lower power applications, include an output compare or PWM peripheral. This peripheral can create a periodic signal whose duty-cycle can be configured by a setting a value in a register. A crystal is generally used as a clock source for the microcontroller, so the generated digital signal has very consistent frequency.

In our example of Fig. 3.15, we chose a 10 kHz frequency, thus if the microcontroller runs at least at 1 MHz 100 voltage steps can be defined (the peripheral frequency is usually limited to the maximum frequency of the core). The resistors are $R_1 = 6.5 \text{ M}\Omega$ and $R_2 = 10 \text{ M}\Omega$ resulting in a maximum current consumption of 100 nA. A capacitor of 22 nF, was chosen to keep the ripple amplitude well below that of the generated DC reference. For the divider, a 5 % tolerance (or lower) is standard for discrete carbon or film resistors (through hole or SMD). In the case of integrated polysilicon resistors, a very low deviation from the desired resistance ratio can be achieved, with the help of common centroid layout techniques. In Fig. 3.16, the uncertainty caused by a 5 percent tolerance of the resistors is represented by the grey shaded area. The time constant of the circuit is given by its RC value, here 87 ms.

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

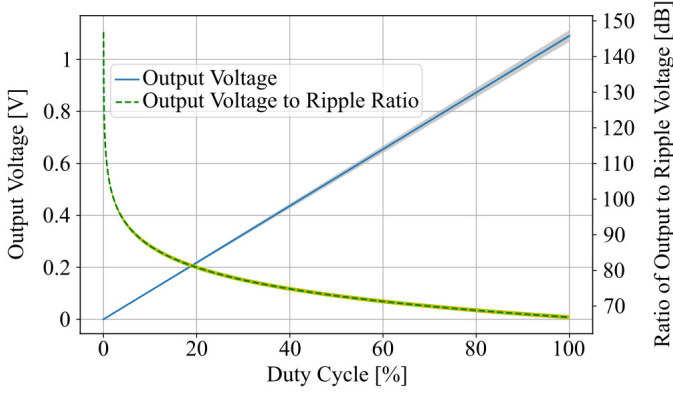


Fig. 3.16 Output voltage as a function of the duty cycle in blue and output voltage to ripple ratio in green.

For the fixed references we used high poly resistors (values: 299.7 k Ω and 1.067 M Ω), to achieve a very high resistance, thus a very low current. A decoupling capacitor is added again to mitigate supply instability and to reduce resistor noise.

3.6.6 SNR Estimation

The noise of the ROIC can be estimated by identifying the main noise sources that are added to the sensor signal. Most of the noise comes from the transistors in the sensor current path: the sensor bias generator and the current mirrors. The resistance added by metal traces, vias and contacts in the layout is not taken into account. For each transistor we estimated the noise using the classical models for thermal and flicker noise of equations 3.12 and 3.13 respectively, where k is the Boltzmann constant, T the absolute temperature, K_F a flicker noise constant obtained from noise measurements or reported by the foundry, W the width of the transistor, L the length, C_{ox} the gate capacitance per unit area and f the frequency. These models give the PSD of noise in the drain as current expressed in A^2/Hz .

$$S_{th} = \frac{8}{3}kTg_m \quad (3.12)$$

$$S_f(f) = \frac{g_m^2 K_F}{C_{ox}WLf} \quad (3.13)$$

3 Readout Circuits for Resistive Sensors

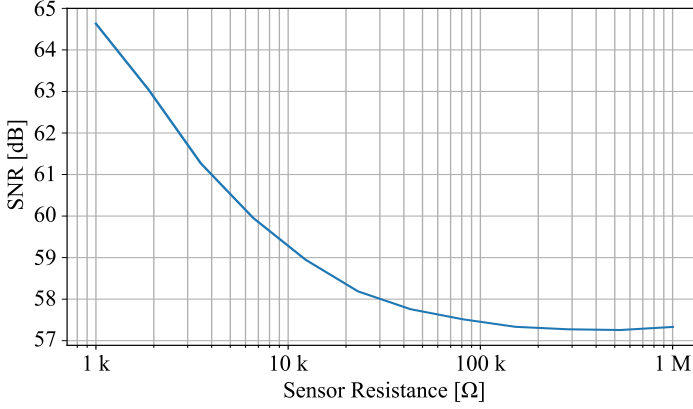


Fig. 3.17 Estimated SNR from the noise model.

To find the SNR, the signal transfer function of each noise source to the integration capacitor must be found. Each transistor is affected differently given its placement in the circuit but all experience a zero-order hold filtering due to the oscillation of the ROIC, eq. 3.14. The zero-order hold model is an approximation because the noise infinitesimally changes the frequency of oscillation and hence the value of f_s , but it is sufficient for the purpose of estimating the SNR. The simulated SNR for a 50 mV bias voltage and input resistances from 1 kΩ and 1 MΩ is plotted in Fig. 3.17.

$$ZOH(f) = \frac{\sin(\frac{\pi f}{f_s})}{\frac{\pi f}{f_s}} \quad (3.14)$$

The SNR of the ROIC must be referred to the input to enable the comparison with that of the sensor. From a practical point of view, it is easier to measure the ratio of the period to the jitter and then calculate its input counterpart: the ratio between the sensor's resistance and its uncertainty, with equation 3.15 (where δR is the uncertainty of the resistance value).

$$SNR_{ROIC} = 20 \log \left(\frac{T}{Jitter_{rms}} \right) = 20 \log \left(\frac{R}{\delta R_{rms}} \right) \quad (3.15)$$

The signal-to-noise and distortion ratio (SNDR) is ubiquitous in the ADC literature and could also be applied here. Still distortion in the transfer function is not itself a problem as long as it is predictable and that it

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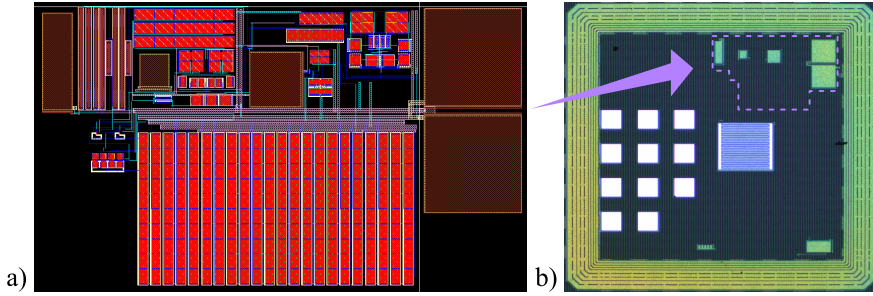


Fig. 3.18 a) Layout view of the ROIC. b) Photograph of the fabricated integrated circuit, the designed ROIC is indicated by the dashed line and covered by metals dummy fill. Bonding pads not shown.

does not diminish the gain ψ from 3.11 significantly, that in turn would lead to an increase in the minimum detectable resistance deviation.

3.6.7 Implementation

The designed layout is shown in Fig. 3.18 (a). Its width and height are $497 \mu\text{m}$ and $288 \mu\text{m}$, respectively (0.143 mm^2). The brown areas correspond to capacitors, the two on the rightmost are for supply decoupling. The cas-coded PMOS current mirror, $Q_{2-3-4-5}$ in Fig. 3.8 (a), requires the most area, covering all the bottom red region. Fig. 3.8 (b) presents a photograph of the fabricated chip. The devices are not visible because dummy metal filling has been added to comply with metal densities requirements, the capacitors are still visible. The size of the chip is $1520 \mu\text{m}$ by $1520 \mu\text{m}$, Fig. 3.18 (b). Process, voltage and temperature simulations were carried out prior to fabrication, unfortunately Monte Carlo models were not available for low- V_T transistors in this technology, mismatch was evaluated and reduced through calculations and good layout practices respectively.

3.6.8 Measurements

The measured ROIC transfer function is plotted in Fig. 3.19 for a set of 9 chips, it was measured in the time domain using a Tektronix MSO56 oscilloscope with a 125 M samples buffer and post-processed with scripts written on Python 3.7. It has a linear region in the middle but at the extremes, several non-linear phenomena become noticeable. For the lowest frequencies, we identified some non-linearity coming from the PMOS and NMOS

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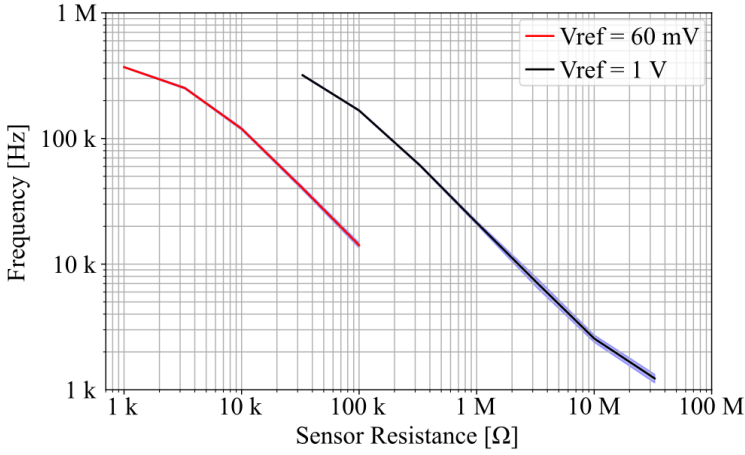


Fig. 3.19 Output frequency for two bias voltages and a range of input resistances. The standard deviation is represented by the blue area.

cascode current mirrors connected to the switches. As the drain-source voltage of Q_{13} and Q_{15} becomes smaller, the current mirrors become less effective in copying the current. Furthermore, for low sensor currents, below 50 nA, the pads leakage current becomes noticeable, measured values of up to 300 pA were observed. The nonlinearity at high frequencies is partly attributed to the integrator and partly to the comparators, the later having the largest impact, although this has not been backed by simulations as shown in their respective section.

Given the measured nonlinearity (Fig. 3.19), it is worth considering whether U2 is necessary or if it is better to place the integration capacitor between the cascodes output and ground. Related to this, the symmetrical two-stage OTA opamp does not seem to offer a better performance than a Miller opamp, only improving the output voltage range moderately.

To assess the noise characteristics of the proposed design we plot in Fig. 3.20 the average signal-to-noise ratio and the minimum fraction of resistance variation that can be measured. The SNR is at least 57 dB for the complete R_{Sensor} range, which is sufficient to interface many graphene and polymer gas sensors [36]. The measured values are slightly above those obtained in simulations, one possible explanation is the overestimation of the noise bandwidth in the calculations. Because the noise processes present

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

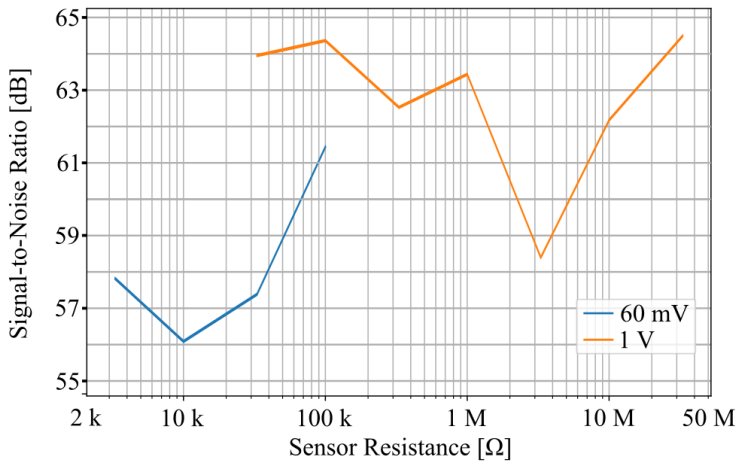


Fig. 3.20 Signal-to-noise ratio for two bias voltages.

in the ROIC are a combination of white and flicker noises, averaging is effective at increasing the SNR. For example, with a 10 samples average, the minimum SNR increases to 65 dB and with 100 to 72 dB if such need arises, while, in terms of sampling time, still maintaining the capability of interfacing 10 different sensors per second, no matter the resistance of each of them.

To keep quantization noise below the SNR of the ROIC, the timestep of the counter in the microcontroller must be small enough to provide 9.6 bits (corresponding to a SNR_{min} of 57 dB) of resolution at the highest output frequency of the interface, which translate to a timestep of 3.33 ns or a frequency of 301 MHz. In chapter 5, we will evaluate how the digitization of the period is most efficiently achieved from a system-level point of view.

Using equation 3.4 and the measured SNR we now calculate the dynamic range. The baseline resistance range spans 60 dB for the 60 mV range and 60 dB for the 1 V range (90.4 dB if both ranges are combined). Therefore, the dynamic range is 117 dB for each subrange, and 148 dB if the whole range is considered. Thus, the external counter must be at least 25 bits wide.

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3.6.9 Power Consumption

The supply current was measured for 9 chips, for a set of resistors (from 1 k Ω up to 33 M Ω) and two reference voltages: 60 mV and 1 V, Fig. 3.21. Towards the left of the chart, the sensor's resistance is lower and the current flowing through it is larger, as its value increases the static current of active components dominates. The EPC follows the expected behaviour. The energy dissipated in the sensor is independent of the period, so the longer the period (larger R) more energy is dissipated by the active components and greater is the total EPC. The changes in the trend at both extremes can be explained by the nonlinearity observed in the transfer function. The theoretical EPC is obtained with equations: 3.16, 3.17 and 3.18 (where Q_5 is the charge consumed from the supply voltage by the digital cells during one oscillation). The current subindices are referenced to the components of Fig. 3.7.

From Fig. 3.21 and 3.11 the EPC for the smallest resistance (1 k Ω) and smallest bias voltage (60 mV) is 1.21 nJ and for the largest resistance (33 M Ω) and largest bias voltage (1 V) is 188 nJ.

$$EPC = E_{sensor} + E_{ROIC} \quad (3.16)$$

$$E_{sensor} = 1.3 \frac{V_{DD}^2}{R} T = \frac{2.6 V_{DD}^2 C \gamma (V_H - V_L)}{V_{REF}} \quad (3.17)$$

$$E_{ROIC} = V_{DD} T (I_{U1} + I_{U2} + I_{U3} + I_{U4}) + Q_{U5} V_{DD} \quad (3.18)$$

3.6.10 Temperature Response

A climate chamber (Espec SH-241) with controlled temperature and fixed humidity was used to test the variations of the output frequencies for the expected temperature range in typical applications: from -20 °C up to 60 °C, a small deviation of 2 percent was seen from the room temperature value for the extremes, Fig. 3.22. The cause of the variation could be attributed to the degradation of slew rate in the operational amplifier with temperature. As the threshold voltages are generated with a resistor divider, the voltages are not greatly affected by temperature changes.

3.6 A Relaxation Oscillator-Based ROIC for Chemiresistive Gas Sensors

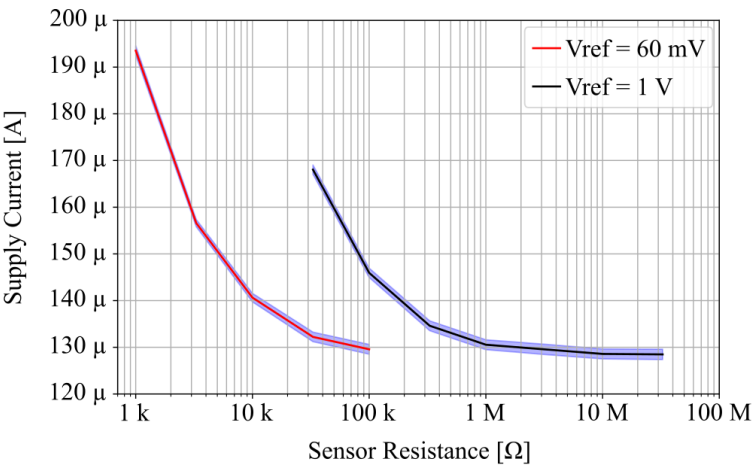


Fig. 3.21 Current consumed by the interface and sensor for two bias voltages. Average of 9 chips, the standard deviation is represented by the shaded blue area.

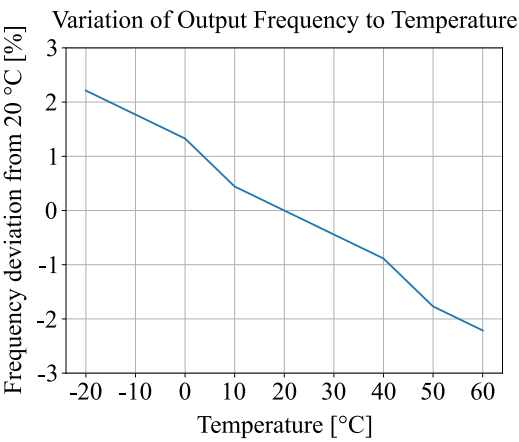


Fig. 3.22 Dependence of the oscillation frequency with temperature for $R_{\text{sensor}} = 330 \text{ k}\Omega$, $V_{\text{ref}} = 1 \text{ V}$.

3 Readout Circuits for Resistive Sensors

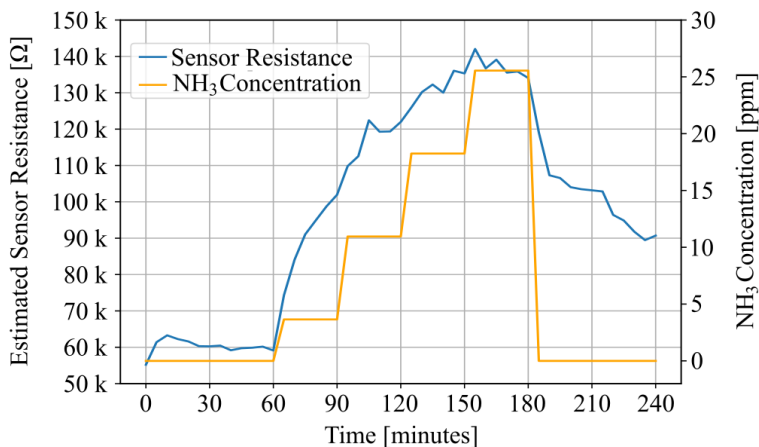


Fig. 3.23 Estimated resistance, from the output frequency for a sequence of ammonia levels applied to a PANi sensor.

3.6.11 Gas Measurements

The read-out circuit was tested in a real case scenario using two types of experimental chemiresistive sensors, serving as a means of validating the sensing system, while not being representative of state-of-the-art chemiresistive sensors. The first sensor was built by drop casting of PANI over a set of 1 mm long by 1 mm wide gold IDE's ($W = 4 \mu\text{m}$, $S = 4 \mu\text{m}$, $N = 125$, $Z = 1 \mu\text{m}$ in Fig. 2.1). This sensor has a baseline resistance of 60 kΩ and was biased with 100 mV. The sensor connected to read-out circuit powered by batteries was placed in a 2.9-liter plastic chamber where a sequence of NH_3 diluted in clean air at different concentrations (nominally 5, 15, 25 and 35 ppm) was applied at a constant flow rate of 2 liters per minute. Humidity and temperature were kept constant. As the thermal mass flow meter (Brooks GF-80 Series) readings are referenced to N_2 , a correction factor of 0.73 had to be applied to obtain the real concentration inside the chamber. The results are plotted in Fig. 3.23. A higher sensitivity is observed for the first step of NH_3 , almost saturating for values above 18 ppm.

For the second experiment a pristine graphene sensor was used. It was also built by drop casting using identical electrodes to the first case. The baseline resistance was 6.1 kΩ biased with 60 mV. The sensor was tested with a sequence of NO_2 concentrations (0.2, 0.8 and 1.4 ppm) again diluted

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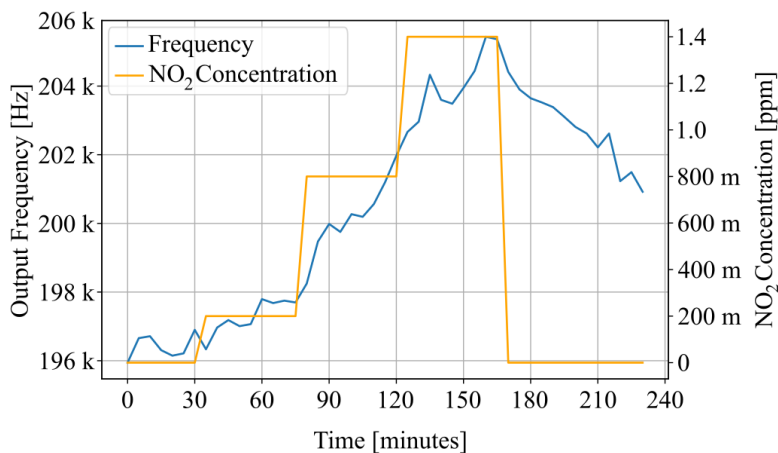


Fig. 3.24 Output frequency for a sequence of NO_2 concentration for a pristine graphene sensor.

in clean air, the response is plotted in Fig. 3.24. In this case a correction factor of 0.74 for the gas flow measurements was used. The graphene sensor, unlike the PANI, has a negative coefficient, so its resistance decreases for increasing concentration.

In both plots a significant amount of noise is observed, this is not caused by the read-out but by the sensors themselves, as explained in chapter 2. This highlights the importance of the knowledge of each sensor behaviour from a signal point of view to improve concentrations estimates.

In terms of the LOD, we have measured PPy sensors for ammonia and found a sensitivity S (equation 2.13) of 4.8 %/ppm for low concentrations (1.85 ppm NH_3), yielding a limit of detection of 426 ppb with a requirement of detecting variations of 0.37 percent from the baseline resistance, which is within the read-out range. One application case of this system is for example for breath analysis in patients with renal disease. Breath ammonia levels can be around 2000 ppb and 200 ppb before and after haemodialysis respectively [79].

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3.7 Conclusions

After reviewing the available ROIC architectures for resistive sensors we see that ADC-based and time-based designs are both viable options and the choice comes down to the specifications that are prioritized in the application case. Given the flexibility needed for interfacing dissimilar chemiresistive sensor we argued that relaxation-oscillator-based ROICs are the best candidates. After implementing and testing our proof-of-concept design, we confirmed the functionality with two chemiresistive gas sensors in a real case scenario. Most of the design objectives, like: input range, bias voltage and circuit size were met. The read-out circuit successfully provided estimations of the sensor resistance and values in the ppb range can be measured for both NH_3 and NO_2 . We think there is room for improvement in the EPC, for high resistance sensors and in the SNR to be above that of chemiresistive sensors.

4

Adaptive Biasing For Relaxation Oscillator-Based ROICs

Abstract:

In this chapter, we present two adaptive biasing techniques for comparators to achieve ultra-low-power and high dynamic range in readout circuits for resistive gas sensors. The first technique is based on an on-demand biasing that generates a bias current of up to $3\ \mu\text{A}$ when the differential input voltage is close to zero, while minimizing the bias current when a comparison is not imminent. The second technique takes advantage of the correlation between sensor current and delay requirements to bias the comparator with the appropriate current for each resistance. These techniques were applied to two readout circuits, ULP1 and ULP2, demonstrating a reduction of the energy per measurement to less than a fifth for the largest input resistance in comparison to fixed biasing without affecting the SNR. The input range is above 5 decades in both cases, from $2.4\ \text{k}\Omega$ up to $1\ \text{G}\Omega$ and from $1\ \text{k}\Omega$ up to $270\ \text{M}\Omega$ respectively. The SNR ranges from 55 dB up to 70 dB which provides enough margin to interface a wide range of gas sensors. An ASIC in a 180 nm SOI process was fabricated.

4 Adaptive Biasing For Relaxation Oscillator-Based ROICs

Lastly, one of the designs was successfully employed to measure varying concentrations of ammonia with a PPy gas sensor.

4.1 Adaptive Biasing

The power consumption of relaxation-oscillators-based ROICs is distributed among its components and is highly dependent on the sensor's resistance and the bias voltage. A part of the energy spent in a conversion is dissipated in the sensor and the rest in the readout as calculated in equations: 4.1, 4.2, 4.3. The EPC results from multiplying the power associated with each functional part of the circuit by the period of oscillation T and adding the results, 4.1. The power consumed by the sensor is that of a resistor: $V_{DD}(V_{REF}/R_s)$, whereas that of the ROIC is the product of V_{DD} and ΣI , where ΣI denotes the sum of the bias currents of each circuit block: U1, U2 and U3, Fig. 3.5. As the period is proportional to the sensor resistance 3.11, the energy dissipated in the sensor is constant for all R_s (assuming a perfectly linear relationship between R_s and the period), 4.2. On the other hand, the term arising from the bias currents, 4.3, is proportional to the period with the addition of the energy used by the digital devices $E_{DIGITAL}$, namely U3 and the counter. Finally equation 4.4, results from the substitution of 4.2 and 4.3 in 4.1.

$$EPC = E_{sensor} + E_{ROIC} = T(P_{sensor} + P_{ROIC}) \quad (4.1)$$

$$E_{sensor} = 2\delta CV_{HIST}V_{DD} \quad (4.2)$$

$$E_{ROIC} = \left(\frac{2\delta R_s CV_{HIST}}{V_{REF}} \right) V_{DD} \Sigma I + E_{DIGITAL} \quad (4.3)$$

$$EPC = 2\delta CV_{HIST}V_{DD} \left(1 + \frac{R_s \Sigma I}{V_{REF}} \right) + E_{DIGITAL} \quad (4.4)$$

4.1.1 Biasing Current Sizing

The maximum attainable frequency by the ROIC is set by the comparator delay, therefore it is important to size it properly in relation to the oscillation period but without overreaching, as this would require a larger bias current and efficiency would be degraded. In this subsection, we explain how to size the bias current using the simplified model of Fig. 4.1.

4.1 Adaptive Biasing

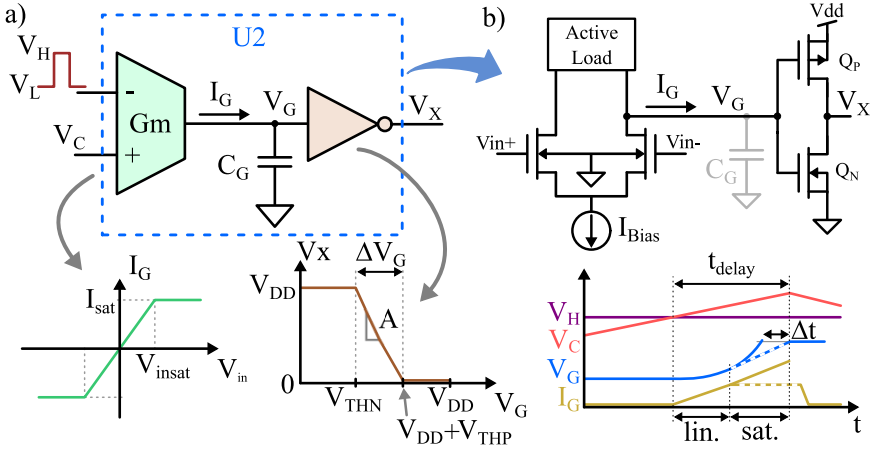


Fig. 4.1 a) Block diagram of the comparator with the transfer function of each device. V_{THN} and V_{THP} are the threshold voltages of NMOS and PMOS transistors respectively. b) Schematic of a generic comparator and time-domain signals at different nodes.

Comparators have been extensively researched and uncountable implementations have been proposed [80], [81], [82]. To keep the analysis as general as possible, we consider a basic comparator architecture: a differential transconductor, G_m , drives a high-gain amplifier, like a digital inverter whose output toggles between V_{DD} and ground upon a small change of its input voltage, Fig. 4.1(a). G_m is modelled as a MOSFET differential pair with an active load where one input is connected to the integration capacitor C and the other toggles between two voltages V_H and V_L : $V_{HIST} = V_H - V_L$ in sync with the output. The transconductance is determined by the transistor's geometry and bias current, which also sets the maximum sink and source output current drive, I_{sat} .

The high-gain amplifier (e.g. a digital inverter), can be modelled by a gain A and a capacitor C_G at its input, that mimics the pole in that node, Fig. 4.1(a). To toggle the output of the amplifier from one digital value to the other, G_m must charge or discharge C_G , by $\Delta V_G = V_{DD} + V_{THP} - V_{THN}$. The process of charging V_G is depicted in Fig. 4.1 (b). After the differential input of the comparator crosses zero, I_G increases and so does V_G . Here, two different situations may occur depending on I_{sat} and the slew rate of V_C . In one scenario (dashed line in Fig. 4.1 (b)), G_m saturates at I_{sat} before

4 Adaptive Biasing For Relaxation Oscillator-Based ROICs

C_G is fully charged, increasing the delay by Δt , in the other situation (solid line), G_m does not saturate, and V_G has a parabolic shape achieving a lower delay. These delays are calculated in equations 4.5 and 4.6 for each case respectively using the equations of the devices, Fig. 4.1 (a).

$$\Delta t_L = \sqrt{\frac{2\Delta V_G C_G R C \delta}{g_m V_{REF}}} \quad (4.5)$$

$$\Delta t_S = \frac{2g_m V_{REF} C_G \Delta V_G + I_{sat}^2 R \delta C}{2I_{sat} g_m V_{REF}} \quad (4.6)$$

$$\left(\frac{t_0}{T}\right)_L = \sqrt{\frac{2V_{sat} C_G V_{REF}}{g_m R \delta C}} \frac{1}{V_{HIST}} \quad (4.7)$$

$$\left(\frac{t_0}{T}\right)_S = \frac{V_{sat} C_G V_{REF}}{I_{sat} R C \delta V_{HIST}} + \frac{I_{sat}}{2g_m V_{HIST}} \quad (4.8)$$

With 3.11, 4.5 and 4.6, the delay-to-period ratio takes the form of 4.7 in linear region and is aggravated when G_m saturates 4.8. In Fig. 4.2 (a), curves 4.7 and 4.8 as well as the EPC are plotted taking the technology parameters from a commercial 180 nm SOI process. The dashed parts, Fig. 4.2 (a), indicate saturation of G_m and in turn a faster increase of the delay. In terms of EPC, Fig. 4.2 (c), a bathtub-like shape appears for a 100 nA bias because the period does not decrease as fast as the resistance does, due to the delay taking a constantly increasing fraction of the period, which hinders operation at higher frequencies. Consequently, nonlinearity is aggravated, and SNR reduced as T is less sensitive to R .

Chemiresistive gas sensors are generally nonlinear devices [83] which require the knowledge of its response curve to make precise gas measurements from their estimated resistance. Taking advantage of this, the ROIC nonlinearity can be accounted for in the sensor calibration curve. In this manner, nonlinearity requirements are less stringent, and power consumption can be kept low. We chose to limit nonlinearity to 10 % to safeguard the $\Delta T / \Delta R$ gain factor and the SNR.

To keep quantization noise close to SNR at maximum frequency, the ROICs presented in the next sections of this chapter have been designed to oscillate at frequencies up to 1 MHz, limited by the practical impediments of quantifying higher frequencies in an ultra-low-power system. In Fig. 4.2 (b), I_{sat} , which is equal to the comparator bias, is taken as the independent variable and the period-delay ratio plotted again, for $I_{Sensor} =$

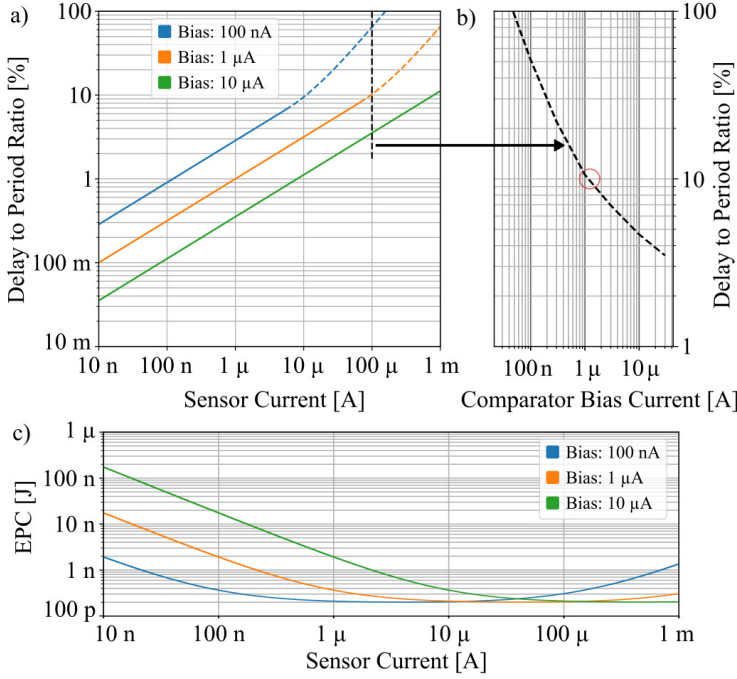


Fig. 4.2 a) Comparator delay for three bias currents. Simulation parameters: $V_{Ref} = 1$ V, $C = 8$ pF, $\delta = 10$, $V_{DD} = 1.8$ V, $V_{HIST} = 0.6$ V. Technology: X-FAB 180 nm. b) Fractional delay vs comparator bias for: $I_{Sensor} = 100$ μ A ($V_{Ref} = 1$ V and $R_s = 10$ k Ω) with G_m : $W = 1$ μ m, $L = 500$ nm, Inverter: $W = 500$ nm, $L = 500$ nm. $C_{inv} = 29.5$ fF, g_m from technology's g_m/I_d curves (27.3 μ S at 1.5 μ A). c) EPC with the same parameters as in a).

4 Adaptive Biasing For Relaxation Oscillator-Based ROICs

100 μA , which is the maximum sensor current by design. From this chart, we determine that a bias current close to 1.5 μA is required to achieve an adequate delay, below 10 % of the oscillation period. This value provides a good initial estimation. In practice, full-circuit simulations have shown that a value closer to 3 μA is more adequate. With this value of I_{bias} , the EPC has the aforementioned characteristic, spanning slightly more than two decades from its lowest value, Fig. 4.2(c).

Due to the need to integrate a large number of sensors in the same package, miniaturization is paramount. However, this results in an increase of the resistance of each device, reducing the sensor current and operating the ROIC where it is least efficient. To tackle this problem, in the next section, we propose two adaptive biasing techniques that help diminish the EPC when the sensor resistance is highest, without affecting the transfer characteristics.

4.2 Zero-Crossing Bias Current Generator

The first method is based on the observation that the comparator needs to be biased only when a switching event is imminent. To achieve this, we designed a zero-crossing bias current generator (ZXB), Fig. 4.3, where the current sourced by Q_9 to bias the comparator is maximum when the comparator inputs V_+ and V_- are equal. The currents and voltages are plotted in Fig. 4.4.

The differential stage is biased with an 80 nA dc current. When the voltages at the gates are dissimilar, the current flows either through Q_4 or Q_6 , both connected to Q_1 , so there is no current flowing through Q_2 . When the gates of Q_7 and Q_8 are at the same voltage the current flows equally through Q_3 and Q_5 or Q_4 and Q_6 . This results in a spike-shaped current in Q_2 with a peak amplitude of 40 nA. To produce a larger current, Q_9 is enlarged with respect to Q_2 , resulting in a peak current of over 3 μA , Fig. 4.4. The quiescent current is 80 nA.

The overall benefit of using the ZXB is shown in Fig. 4.5, where the ratio of peak to average consumed current is plotted. For low slew-rates this circuit produces a peak bias current more than three times the average current consumption.

Compared to other adaptive biasing techniques, this circuit offers excellent transient response, and it can operate with input slew rates above 1 V/ μs . On the contrary, in [84] a comparator with an adaptive bias current generator was proposed and tested in a 350 nm technology, reaching

4.2 Zero-Crossing Bias Current Generator

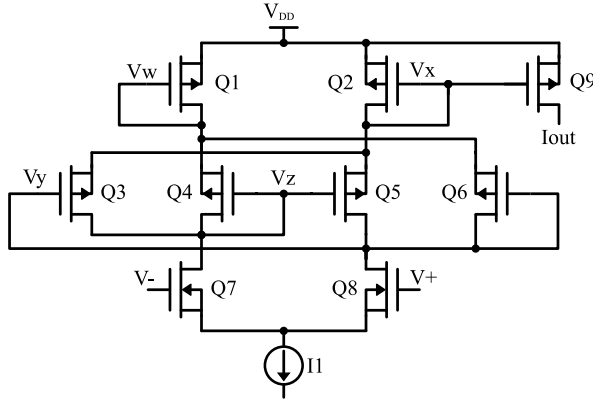


Fig. 4.3 ZXB (zero-crossing bias) current generator schematic. Sizes (W/L) $Q_1 = Q_2 = 500 \text{ nm} / 500 \text{ nm}$, $Q_3 = Q_4 = Q_5 = Q_6 = 700 \text{ nm} / 700 \text{ nm}$, $Q_7 = Q_8 = 220 \text{ nm} / 10 \text{ }\mu\text{m}$, $Q_9 = 3.75 \text{ }\mu\text{m} / 220 \text{ nm}$.

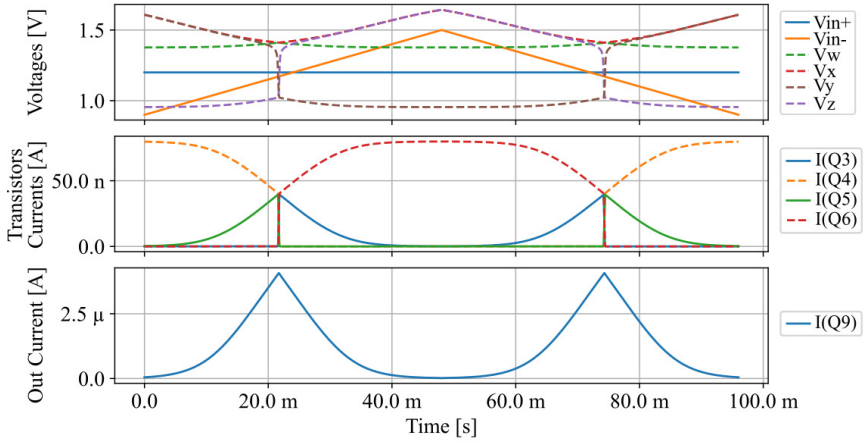


Fig. 4.4 Time-domain simulated current and voltages in the ZXB of Fig. 4.3 for one period of 10 Hz triangle waveform.

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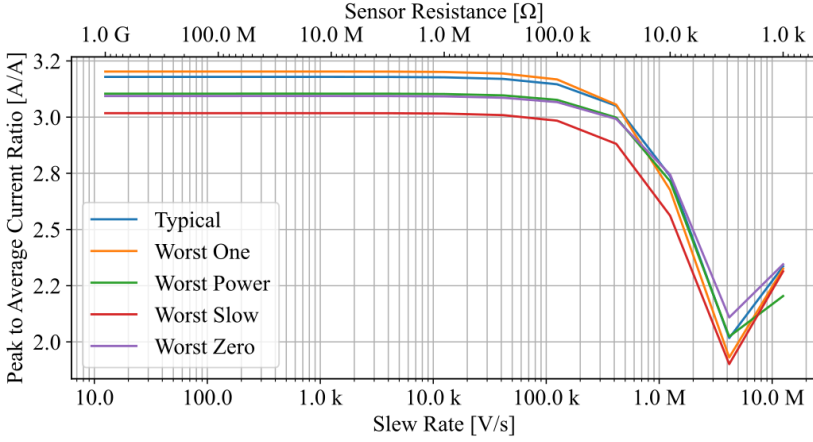


Fig. 4.5 Ratio of improvement of using the ZXB compared to a fixed bias current, simulated for all corners and the corresponding sensor resistance when $V_{ref} = 1$ V.

a maximum frequency of 130 kHz with a quiescent current of 33 nA. In the cited paper, the adaptive bias is generated using current subtractors that copy and amplify the currents flowing through diode-connected transistors that are part of the hysteresis mechanism of the comparator. The main drawback is that the current in these transistors increases with a delay due to the hysteresis and consequently the adaptive bias generation is also delayed. In [85] an OTA with adaptive biasing was developed to reduce quiescent current. In their implementation, the peak of the bias current is reached when the differential input voltage is 0.5 V. The bias current peaks at approximately $45 \mu A$, the slew rate of the OTA is above $17 V/\mu s$ at this point.

4.3 Sensor Current Proportional Biasing

The second adaptive biasing technique we propose takes a different approach, noting that the delay requirement of the comparator changes with the sensor's resistance as it sets the slew rate of the V_C , Fig. 4.1. For low values of R_S , the oscillation frequency is highest and so the bandwidth requirement of the comparator, the opposite happens for highest resistances. With this in mind, we propose to bias the comparators with the sum of two

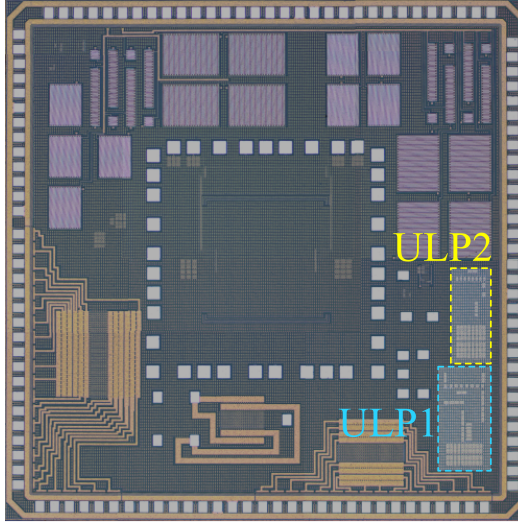


Fig. 4.6 Photograph of the fabricated integrated circuit containing both ASICs.

sources: a 100 nA fixed source and a reduced copy of the sensor current. Consequently, a larger current will be generated when a larger bandwidth in the comparator is needed. To obtain a maximum bias current close to 3 μA , a reduction factor of 30 was selected since the maximum sensor current is designed to be 100 μA ($V_{ref} = 100 \text{ mV}$, $R_s = 1 \text{ k}\Omega$). The fixed current is a scaled copy of the externally-supplied reference current, which is already used in the ROIC to bias all the active devices by a set of current mirrors.

4.4 Measurements

Two readout circuits for resistive gas sensors, named ULP1 and ULP2, were implemented in a 180 nm SOI process, XT018 from X-FAB (X-FAB Semiconductor Foundries GmbH, Erfurt, Germany), Fig. 4.6. The area of these ROICs are: 0.183 mm² and 0.119 mm² respectively. They were designed to handle sensor currents, V_{ref}/R_s , from 1 nA up to 100 μA with V_{ref} ranging from 50 mV to 1 V. The principle of operation is that of Fig. 3.5 but structurally ULP1 and ULP2 have some differences. Instead of using a Schmitt trigger comparator, two regular comparators with NMOS and PMOS input differential pairs respectively were used to allow for bet-

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ter control of the threshold voltages. Additionally, a SR latch was added at the output to avoid false toggles generated by noise that could induce jitter. To study the robustness in the copy of the sensor current used to charge and discharge C , ULP1 uses cascode regulated current mirrors while ULP2 uses standard cascodes, the former being more linear for currents in the nA-range. Both circuits share the same design parameters: $C = 8$ pF, $\delta = 10$, $V_H = 1.2$ V, $V_L = 0.6$ V. The reference voltages for the comparators V_H and V_L were generated with on-chip M Ω resistors, consuming 55 nA, V_{ref} is supplied externally. The capacitors are all metal-insulator-metal (MIM) type. For the circuit characterization, discrete resistors ranging from 1 k Ω up to 1 G Ω were used.

ULP1, Fig. 4.7, uses active current mirrors and the ZXB technique in both complementary comparators. The transconductors used in the regulated current mirrors (U_5 , U_6 , U_7) are biased with 10 nA each, having little impact in the overall power consumption. ULP2, Fig. 4.8, relies on biasing the comparators with a reduced copy of the sensor current plus a 100 nA dc source (not present in ULP1), as explained earlier. In this way, the comparators are biased with at most $3.3 \mu\text{A} + 0.1 \mu\text{A}$. When the current in the sensor is minimum, 1 nA, essentially only the fixed 100 nA remains.

The measured transfer functions as well as the simulated corners and the ideal response are plotted in Fig. 4.9 for both circuits, 10 chip samples each, biased with 50 mV and 1 V. The plots present the average values and standard deviation, the latter not being noticeable. The maximum non-linearity in the full characterization range is -32 percent for ULP1 and 41 percent for ULP2. If the nonlinearity is restricted to ± 10 % (shaded area in Fig. 4.9), the input range of ULP1 extends from 2.4 k Ω to 1 G Ω and from 1 k Ω to 270 M Ω for ULP2. For largest resistances, the benefit of using active current mirrors results in a more precise copy of the sensor current and in turn better linearity, at the cost of larger die area due to the additional 9600 μm^2 needed for MIM capacitors (22.7 pF) to guarantee stability. At such low currents, the V_{DS} in the diode-connected devices is only some tenths of mV so, a significantly larger V_{DS} in the passive copying transistor results in large differences in the current through the drain. The active current mirrors help diminish the difference between both V_{DS} to just the offset voltage of the transconductor. The nonlinearity at high currents is mostly explained by the trade-off presented in the previous section regarding comparators bias and delay. Notably, ULP2 performs better, the reason for this is that at 1 k Ω the ZXB is operating with an input slew rate of 0.42 V/ μs , where its efficacy declines, Fig. 4.5. Additional considerations re-

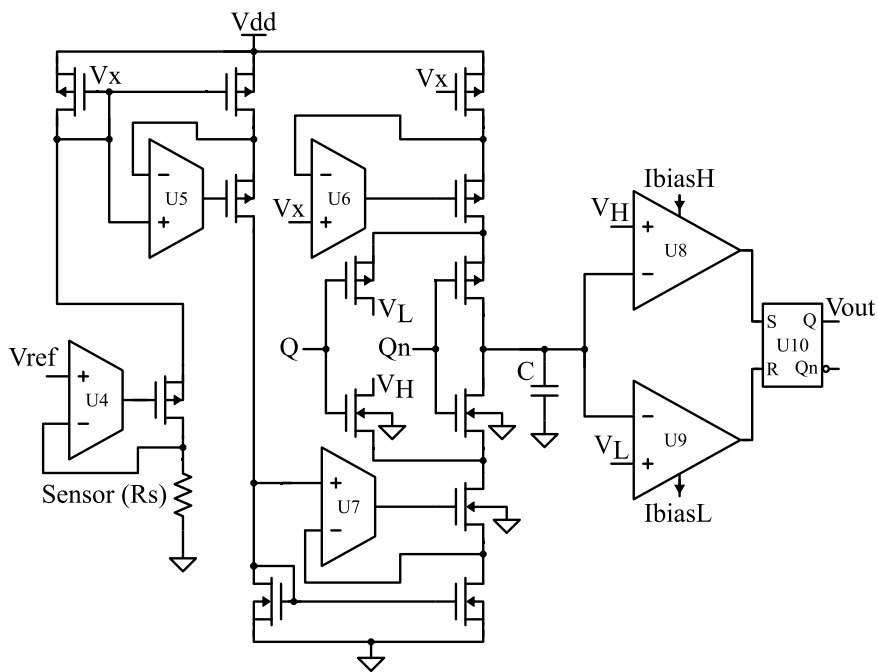


Fig. 4.7 Schematic of readout ULP1, the comparators, U8 and U9, are biased with ZXBs.

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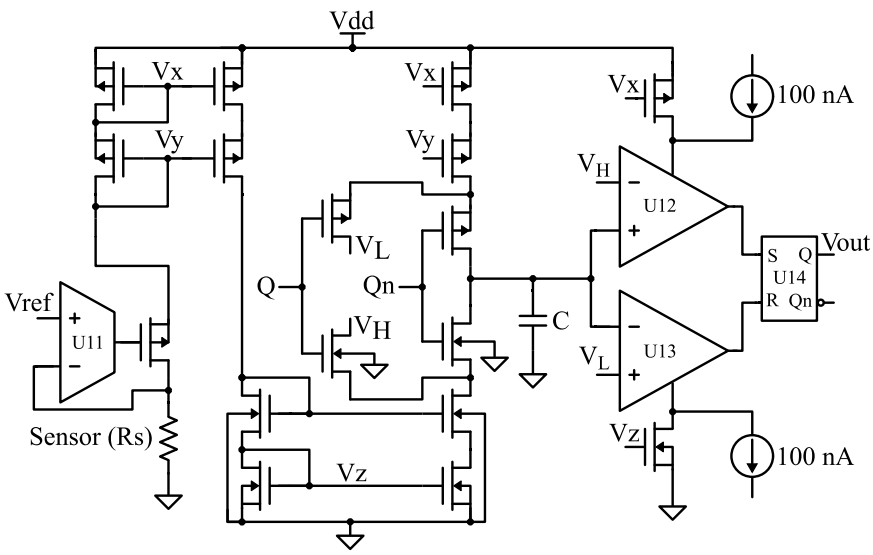


Fig. 4.8 Schematic of readout ULP2, the comparators, U12 and U13, are biased with a fixed 100 nA plus a reduced copy of the current through the sensor.

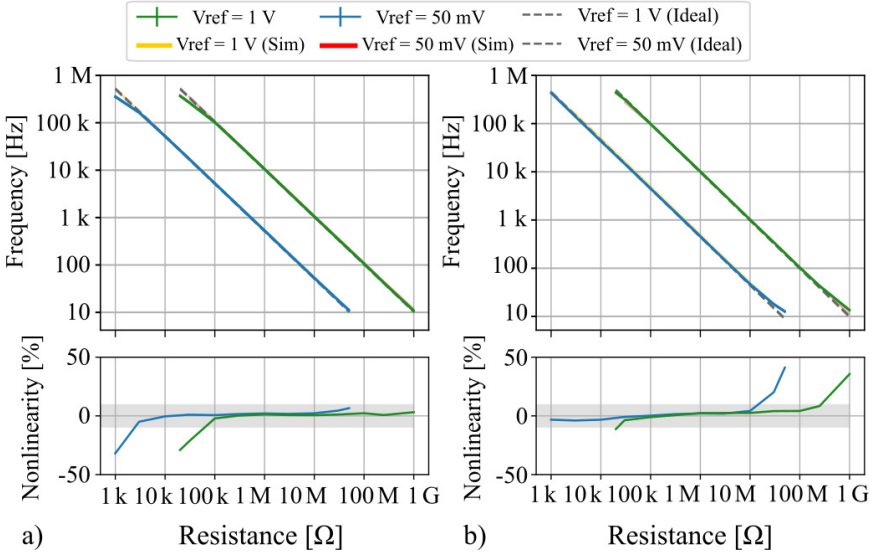


Fig. 4.9 Measured, simulated and ideal transfer functions for ULP1 (a) and ULP2 (b), the measured transferred functions are compared to the ideal response in the nonlinearity plots. Ten chips were measured and averaged. The error bars, although small, are also plotted. The shaded are indicates nonlinearity within 10 %.

garding the nonlinearity and complementary measurements are included in appendix B.

The total average supply current for ULP1 and ULP2 are plotted in Fig. 4.10, due to the dominance of the sensor current for lowest sensor's resistances, the curves are almost identical in this region. A larger variation with process corners is predicted by simulations in ULP1 when the sensor current vanishes after the corner point, because small size transistors are used for the ZXB to obtain the best frequency response, making it more vulnerable to process variations.

The supply current difference between ULP1 and ULP2 above 100 MΩ is explained by the biasing current in the comparators: in ULP2 both comparators are biased with a 100 nA fixed source each (the current copied from the sensor is negligible), while in ULP1 the ZXB still consumes on average 1 μA. The rest of the devices: U_1 , U_2 , U_3 , U_4 consume 10 nA, not affecting the analysis significantly. Time records of a few oscillation cycles

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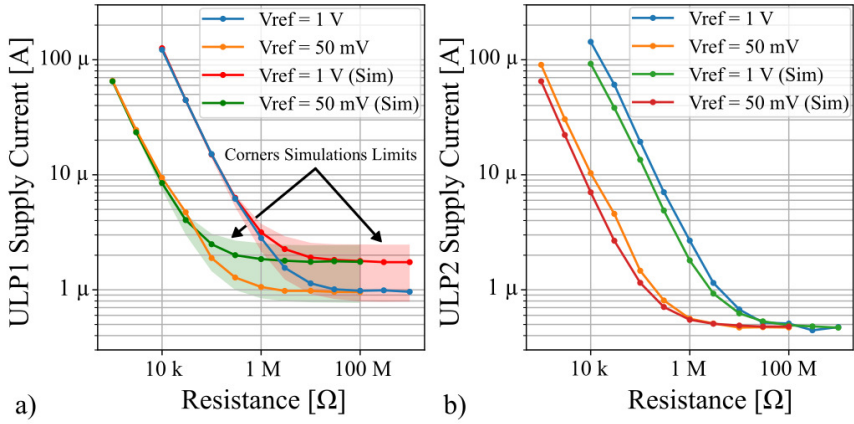


Fig. 4.10 Measured and simulated supply current for ULP1 (a) and ULP2 (b). Corners bounds are represented by the shaded area.

of the current drawn from V_{DD} are plotted in Fig. 4.11, for $V_{ref} = 1$ V. The operation of the ZXB is clearly visible for resistances larger than 3 MΩ. For ULP2 short spikes are seen which can be attributed to the charge/discharge cycles of the embedded decoupling capacitors and switching of the SR latch.

The EPC is calculated from the current and frequency measurements, ($V_{ref} = 50$ mV and 1 V) for both designs, Fig. 4.12. As the current generated by either of the two bias generation mechanisms is close to 3 μ A at its maximum point, we compare the EPC of ULP1 and ULP2 to an ideal design whose comparators are biased with a constant 3 μ A dc source. The results show that when the sensor resistance is low the EPC of the three circuits are similar but great gains are observed for the largest sensor resistance, where the EPC drops almost one decade for ULP1 and more than one decade for ULP2, i.e. 85 % and 94 % lower with respect to dc biasing respectively. A shift to the right of the corner point in 4.4 is also seen. Due to the different onset of action of each adaptive biasing mechanism, the shape of EPC is slightly different, having a similar but not equal slope in the rising part. For $V_{ref} = 50$ mV, and $R_s = 100$ MΩ, there is a decrease in the slope of ULP2 explained by the nonlinearity in the transfer function. The oscillation frequency is larger than the ideal, so the integration time of the consumed current is smaller. For the highest frequency, ULP2 offers no

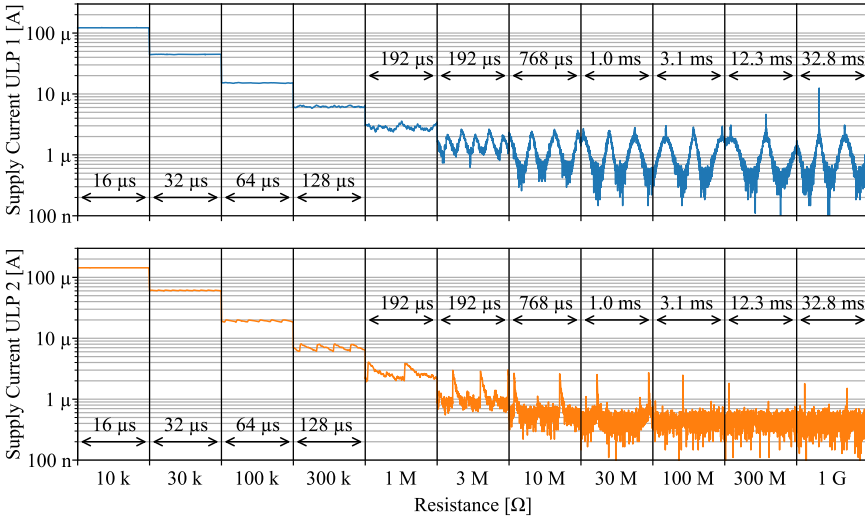


Fig. 4.11 Time fragments of the current drawn from the main supply for a range of resistances and $V_{Ref} = 1$ V.

benefit in comparison to dc biasing, on the other hand the ZXB in ULP1 is still operating and leads to a slightly lower EPC than the former.

To calculate the SNR, 4096 consecutive period measurements were done for each point of operation and the SNR calculated from the jitter, Fig. 4.13. Due to the measurement setup, the noise of the test resistor could not be removed. Sensors made with conductive polymers and graphene tend to be moderately noisy devices, for example in our measurements of Fig. 4.13, the SNR of the PPy sensor is 56 dB, for PANI sensors we have measured an SNR of 27 dB. In [72] a minimum SNR of 46 dB was reported for MOx sensors also. These values are below ULP1 and ULP2 intrinsic noise, confirming the suitability of these circuits to interface gas sensors. The calculated dynamic ranges are listed in table 4.1, exceeding 128 dB [72] in all cases.

ULP1 was used in combination with an experimental PPy sensor to estimate ammonia at three different concentrations by calculating the sensor resistance from the measured output frequency, Fig. 4.14. Four different bias voltages were employed demanding different EPCs for the same sensor, according to Fig. 4.12. For this sensor, the response time, noise and

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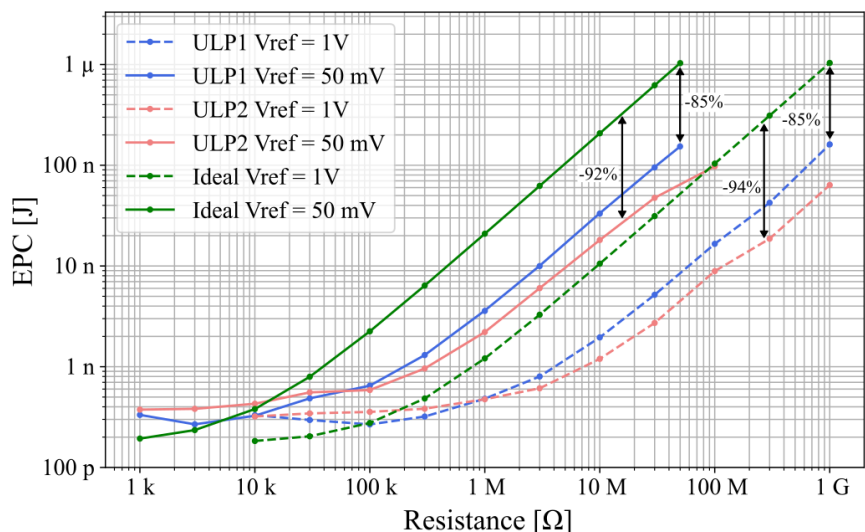


Fig. 4.12 Measured EPC for ULP1 and ULP2 compared to an ideal read-out interface whose comparators are biased with 3 μ A dc each.

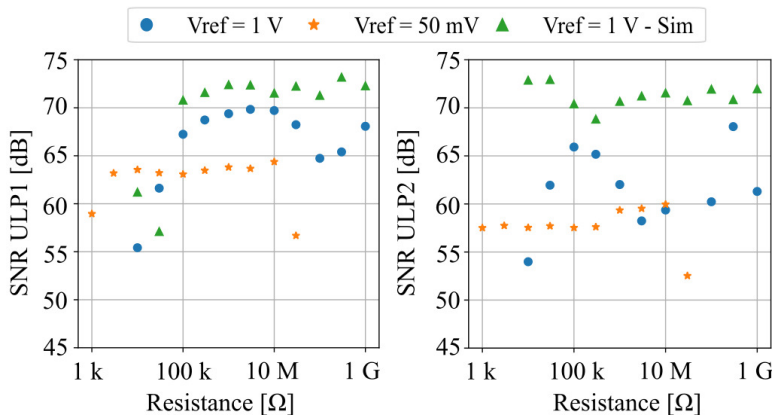


Fig. 4.13 Measured and simulated SNR for ULP1 and ULP2 for discrete resistors: 1 k Ω , 3 k Ω , 10 k Ω , 30 k Ω , 100 k Ω , 300 k Ω , 1 M Ω , 3 M Ω , 10 M Ω , 30 M Ω , 100 M Ω , 300 M Ω and 1 G Ω .

Table 4.1 Dynamic Range of POC, ULP1 and ULP2

Sensor Bias Voltage (mV)	POC (dB)	ULP1 (dB)	ULP2 (dB)
50	-	142	136
60	96	-	-
1000	119.3	139	135

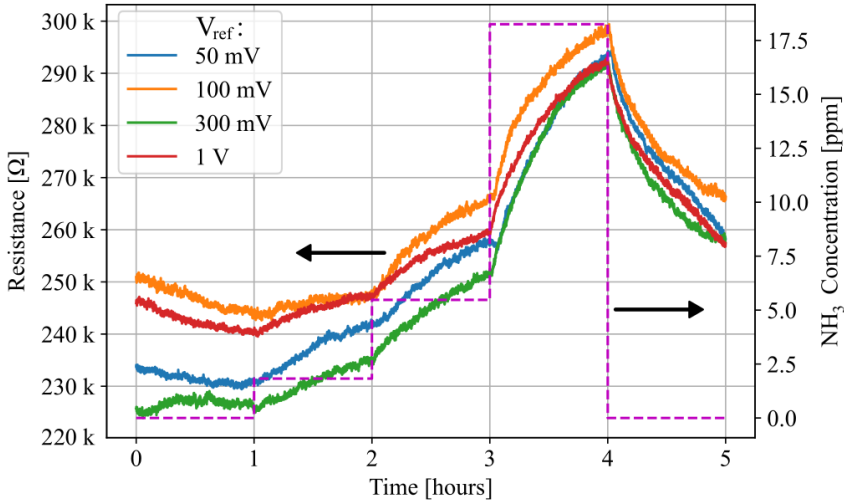


Fig. 4.14 Ammonia detection test with a ULP1 and a PPy sensor. Four bias voltage and three ammonia concentration were employed.

sensitivity are not affected considerably by the bias voltage, so the largest V_{ref} produces the lowest EPC. The counter used for these measurements as well as the SNR is from an STM32 STM32L4R5ZI microcontroller.

4.5 Conclusions

In this chapter we studied and addressed the power and efficiency challenges presented by relaxation-oscillator based ROICs. The two biasing techniques we described, improve efficiency for large resistance sensors in a similar scale, reducing energy per conversion down to a tenth versus standard dc biasing. The sensor-current-proportional biasing (ULP2) yielded the largest improvement on EPC and was least affected by process variations. ZXB offers a significant improvement throughout the whole

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frequency range, at the expense of larger PVT variations. In all, these techniques are not mutually exclusive and a combination of both could be feasible. The effect of active and passive current mirrors was also researched, and we can conclude that active current mirrors can diminish nonlinearity for nA sensor currents with a small penalty in the EPC of 9 nJ (in a total of 161 nJ for 1 G Ω). Finally, ULP1 was successfully tested with a PPy sensor to measure different concentrations of ammonia at different bias voltages, confirming the performance from a practical standpoint

5

Period Digitization and System-Level Considerations

Abstract: In this chapter we analyze several configurations to implement the counter and the clock generation circuits needed to digitize the period of time-based ROICs. We study and discuss their impact in power consumption and conclude that digitally-intensive tasks are found to be most efficiently handled by a microcontroller while the ROIC is solely dedicated to the sensor interface and analog signal processing. Two digitization methods are considered: input capture and cycle counting, where the former is found to be the most efficient.

5.1 Period Digitization

As described in previous chapters, time-based ROICs need a timer to transform the analog period of the oscillating signal into a digital word. In the literature, several approaches have been proposed, for example, microcontrollers in commercial development boards like Arduino Due and Arduino Nano have been used in [72] and in [73] respectively. In [74] and [50] the counter was integrated in the ASIC whose output is a parallel digital word. Some authors like Aragonés et al. [86] implemented the counter for their ROIC in a field-programmable gate array

5 Period Digitization and System-Level Considerations

(FPGA). Even though FPGAs are excellent for the exploration of different topologies in the prototyping phase, their real-world application is limited due to the power consumption and cost. For this reason, we do not see them as a preferred option for an autonomous sensing systems, and hence are not considered in the study.

Three main building blocks are needed to implement this circuit: an oscillator (with a time source e.g. a crystal or an RC network), a PLL if frequency synthesis is needed, and a counter. These components can be either integrated in the ASIC or reuse the ones available in the microcontroller that controls the ROIC. In Fig. 5.1 we show three possible distributions: a) all circuits are embedded in the ASIC and the digital value is transmitted to the microcontroller through a set of parallel general-purpose input/outputs (GPIOs), b) the PLL and the counter in the ASIC and the other blocks in the microcontroller and c) no block is implemented in the ASIC (besides the ROIC) and the microcontroller peripherals are used for the period digitization. A fourth case in which the counter is integrated in the ASIC and a PLL from the microcontroller is used has not been considered because it becomes very energy-demanding when the counter clock is above a few megahertz, due to the energy needed to charge-discharge the pad's stray capacitances in both the microcontroller and the ASIC.

5.1.1 Frequency Generation

We begin by studying the trade-offs of using an integrated oscillator in the ASIC, or otherwise, to rely on the microcontroller's timing sources. To evaluate the power consumption of integrated oscillators we consider two type of circuits that are found in many system-on-chip nowadays: ring oscillators and Pierce oscillators. Ring oscillators are the simplest oscillator that can be integrated without external components, requiring only three cascaded logical inverters, Fig. 5.2. In case the natural frequency of oscillation is too high for the application, more stages or starvation mechanisms can be added to delay the transition from one inverter stage to the next, increasing the power consumption per cycle.

The energy consumption per cycle of the three-stage ring oscillator can be calculated with equation 5.1. Where C_P and C_N are the equivalent capacitance seen from gate to ground of the transistors that form each inverter and V_{DD} the supply voltage. This first term corresponds to the energy required to induce or remove the conduction channel in the MOS transistors by charging or discharging the gate. In the second term, $< I_{ON} >$

5.1 Period Digitization

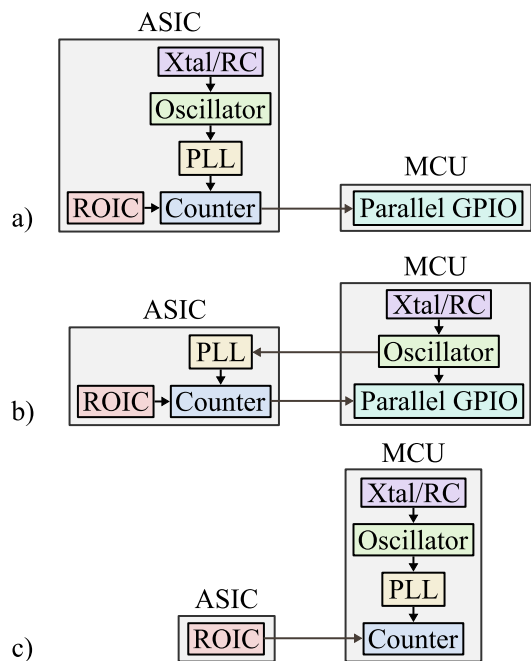


Fig. 5.1 a) Oscillator, PLL and counter in the ASIC, b) PLL and counter in the ASIC, oscillator in the microcontroller, c) Oscillator, PLL and counter in the microcontroller.

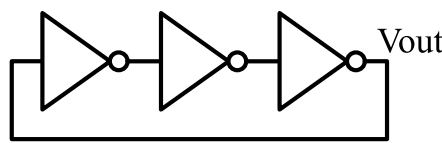


Fig. 5.2 Three-stage ring oscillator.

5 Period Digitization and System-Level Considerations

is the time average of the short-circuit current flowing through the drains of both transistors (while they are simultaneously are turned on) during a transition. T_{cycle} is the period of oscillation. In table 5.1 the power consumption and frequency of a three-stage ring oscillator has been simulated in X-FAB XT018, using standard V_T , NMOS and PMOS transistors ($W = 1 \mu m$, $L = 1 \mu m$), operated at 1.8 V, from -20 °C up to 80 °C, for typical devices.

$$E_{cycle} = 3 \left(\frac{(C_P + C_N)V_{DD}^2}{4} + \langle I_{ON} \rangle V_{DD} T_{cycle} \right) \quad (5.1)$$

Table 5.1 Three-Stage Ring Oscillator Frequency and Average Current Consumption.

Temperature [°C]	Frequency [MHz]	Current Consump- tion [μA]	Power per MHz [nW/MHz]
-20	237.0	23.7	180.0
0	226.8	23.1	183.3
20	219.3	22.6	185.5
40	212.3	22.1	187.4
60	206.2	21.6	188.6
80	200.8	21.1	189.1

To reduce the short-circuit current in each stage and to achieve low frequencies with a small footprint, Funke et al. [87] designed a thyristor-based oscillator based on the delay element presented in [88]. This oscillator runs at 20 Hz, while consuming 150 nW/MHz with a 200 mV supply and 11.6 μW /MHz with 1.8 V supply. The frequency is set by the leakage current of MOS transistors and the size of two capacitors which are charged with these leakage currents. The ASIC was fabricated in a 180 nm CMOS process.

The main drawback of ring oscillators is their vulnerability to temperature, noise and supply variations. Their frequency of oscillation is set, in part, by the transistors conductance which is significantly affected by temperature fluctuations as seen in simulations. In terms of jitter, Geraedts et al. [89] and Navid et al. [90] have studied the minimum phase noise of RC oscillators, including ring oscillators. They observed $1/f^3$, $1/f^2$ and white noise. In their 12 MHz oscillator, the phase noise is -109dBc/Hz at 100 kHz, which corresponds to the center of the $1/f^2$ region. Given that the largest SNR measured with the ROIC is expected to be 80 dB, this theoretical jitter

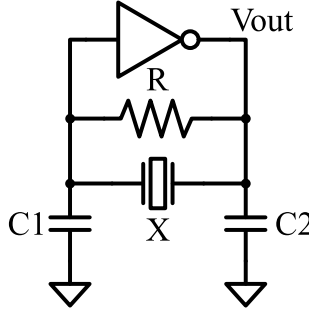


Fig. 5.3 Pierce oscillator.

figure is below the noise of the sensing system.

The Pierce oscillator has been widely adopted in embedded systems, due to its simplicity and frequency stability as it uses a quartz crystals to set the oscillation frequency, Fig. 5.3. Quartz crystals are cut with very high precision, resulting in a frequency tolerance and stability in the range of tenths of ppm [91]. Their resonant frequency is well-defined and set by the dimensions and geometry of the crystal cut. The Barkhausen criterion is met at this predefined frequency, which is stable with temperature (also in the order of tenths of ppm), therefore the frequency of the oscillator is not significantly affected by PVT variations.

The crystal is chosen so that the load capacitance seen in its terminals is equal to that obtained with equation 5.2 where C_{IN} and C_{OUT} are the input and output capacitances of the logical inverter, of Fig. 5.3.

$$C_L = \frac{(C_{IN} + C_1)(C_2 + C_{OUT})}{C_{IN} + C_1 + C_2 + C_{OUT}} \quad (5.2)$$

The load capacitance of commercial quartz crystals are in the range of 5 to 50 picofarads. This means that C_1 and C_2 are also in this range, and are not in general integrated in commercial microcontrollers and microprocessors oscillators because they depend on the selected crystal. The power consumption of Pierce oscillators depends, in part, on the chosen capacitances, the drive level of the crystal and the intended frequency stability as influenced by the crystals pullability.

Several works have proposed variations of the classical Pierce oscillator, for example, Peng et al. [92] presented a temperature-compensated Pierce oscillator, which runs at 16 MHz while consuming $150 \mu A / MHz$.

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The ASIC was fabricated in a 180 nm bulk process and uses a 1.2 V supply voltage. Another Pierce crystal oscillator was presented in [93]. It was also fabricated in a 180 nm process and consumes $3.75 \mu\text{W} / \text{MHz}$ while running at 48 MHz with a 1 V power supply. Lastly Siniscalchi et al. [94] introduced two ultra-low-voltage crystal oscillators that consume $70.63 \text{ nW} / \text{MHz}$ and $468.8 \text{ nW} / \text{MHz}$ while running at 32.768 kHz. Such low power consumption is obtained with a novel structure operating in the subthreshold region with a very low supply voltage, between 60 mV and 100 mV.

Besides, ring oscillators and crystal oscillators, LC oscillators are widely used in CMOS circuits. They employ a tank resonant circuit formed by an inductor and a capacitor to transfer energy between them continuously. The oscillation frequency is that of equation 5.3. This kind of oscillators are particularly used in RF applications for system operating in the spectrum bands of hundreds of MHz to GHz (UHF and SHF), because it is possible to implement the oscillators with integrated inductors and capacitors [95], but it is harder to achieve lower oscillation frequencies, e.g. 10 MHz as the values of the inductor or capacitors are not practical for the integration in an ASIC. Another problem with this type of oscillator in sensor applications, is that the top metal layer of the CMOS process which is used for the sensors's interdigitated electrodes is also the one used to implement the inductor, thus increasing the die area.

$$f = \frac{1}{2\pi\sqrt{LC}} \quad (5.3)$$

In modern microcontrollers, like the STM32U575ZIT6Q (STMicroelectronics N.V.), it is standard to find a Pierce oscillators with a quartz crystal element, as the backbone of a complex clock distribution tree, where PLLs, dividers and multiplexers generate the appropriate frequency for each peripheral. This ultra-low-power microcontroller can be clocked up to 160 MHz with internal PLLs and powered with a minimum supply voltage of 1.71 V. Given the complexity of the system, it is not possible to isolate the power consumption of the oscillator from that of the core. Additionally, the oscillator provides timing to the CPU, memory and other peripherals, therefore the impact in power consumption must be weighted among these blocks.

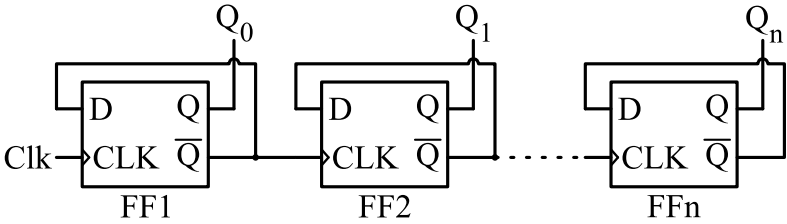


Fig. 5.4 Schematic of the counter used for simulations.

5.1.2 Counter

Once a stable timing reference is obtained, a counter is needed to quantify the period of the oscillatory ROIC. We begin by comparing a low-power timer (LPTIM1) from the STM32U575ZIT6Q microcontroller with several counters made with different standard cells in the XT018 process:

- STV: Standard cells for 1.8 V.
- 5V: Standard cells for 2.5 V / 3.3 V / 4.0 V / 5.0 V.
- HD: High density cells for 1.2 V / 1.8 V.
- HDLL: Low leakage, low power cells for 1.2 V / 1.8 V.

The counters were implemented by cascading D flip flops, Fig. 5.4 and the simulated power consumption was calculated by averaging the total current consumption, with the supply voltage kept constant at 1.8 V for all cases. The results for each type of standard cells and corner are presented in table 5.2, simulated at 23 °C.

Table 5.2 Counter Average Current Consumption for Several Standard Cells Libraries and Process Corners.

	12-bit Counter Power Consumption [<i>nW/MHz</i>]				
	Corners				
Cell Family	Typical	Worst Power	Worst Slow	Worst One	Worst Zero
STV	628.8	816	504.4	649.9	644.3
5V	253.3	253.8	270.6	261.9	249.7
HD	440.8	606.9	350.2	520	414.3
HDLL	384.3	496.2	326	438.2	365

5 Period Digitization and System-Level Considerations

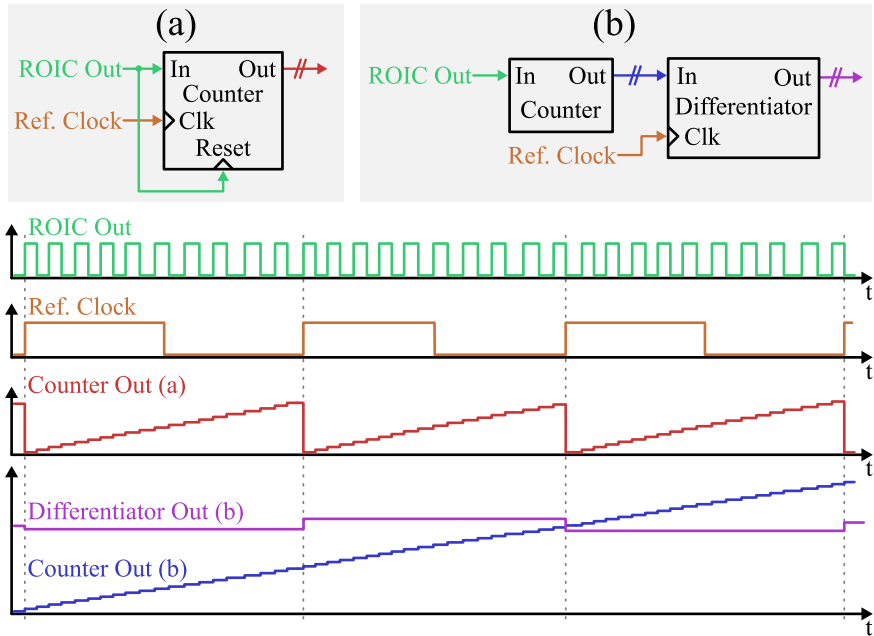


Fig. 5.5 a) Counter with reset, b) Counter and differentiator.

Even though the 5V library is rated from 2.5 V to 5 V by X-FAB, operation with 1.8 V was simulated up to 200 MHz in all corners without issues.

Some papers, like [72], use a counter with reset where the count is restarted every time an edge is detected, Fig. 5.5 a). Other papers, like [66], use a counter with sufficient bits so it does not have to be reset during operation, or at least rarely, Fig. 5.5 b). In this way, the output of the counter is proportional to the phase of the signal. In the second case, a differentiator with a fixed sampling rate is placed after the counter, providing an approximation of the frequency of the signal (derivative of the phase).

The STM32U575ZIT6Q includes low power timers, with a power consumption of 324 nW/MHz at 1.8 V. This counter includes prescalers, several clock sources, input capture, among other functionalities. This microcontroller is manufactured in a 40 nm CMOS process. Reverter and Gasulla [96] studied the power consumption of two other commercial microcontrollers: MSP430FR5969 and ATtiny2313. Even though this microcontrollers are designed for low power applications, their power consumption is larger being over 2.04 $\mu\text{W}/\text{MHz}$ for the MSP430FR5969 and 7.3

$\mu\text{W}/\text{MHz}$ for the ATtiny2313.

5.1.3 Discussion

With regards to the oscillator, if frequency stability with supply voltage and temperature variations is needed, a Pierce oscillator is the best candidate as the frequency of oscillation is set by the crystal properties. For most automotive-grade commercial crystals, the frequency stability from $-40\text{ }^{\circ}\text{C}$ to $150\text{ }^{\circ}\text{C}$ (AEC-Q100 grade 0) is less than 200 ppm over the whole temperature range. On the other hand, fully integrated oscillators, like ring oscillators can consume less energy per oscillation at the expense of more variability and noise. Given that chemiresistive gas sensors have an SNR below 80 dB, either type of time sources can achieve a sufficiently low jitter, with Pierce oscillators being the most stable. Temperature-compensated ring oscillators do exist [97], [98], [99], and their integration in the ASIC may be justified to safeguard the SNR if the timing or jitter of the microcontroller clock cannot be specified or bounded.

On top of this, most microcontrollers include an RC oscillator (fully integrated) and a Pierce oscillator with an external crystal, which are used alternatively. Therefore it is possible to reuse the active oscillator of the microcontroller as a time source for the ROIC without increasing power consumption or adding another external crystal which would increase complexity, footprint and costs.

With respect to the counter, we see that at first glance, the lowest energy is achieved with an integrated counter in the ASIC, when using the 5V cells (outside of specification). This figure however, is for a minimalistic counter, without any additional functionality. For this reason, we think that it is better to use the counter from the microcontroller which at the same time offers more functionalities and is fabricated in a much smaller CMOS node, reducing the switching energy of logic gates.

With all this, we believe it is best that the analog ROIC is only tasked with converting the resistance value of the sensor into a frequency modulated digital signal which is then digitized by a microcontroller, as in Fig. 5.1 (c). Again, the scheme of Fig. 5.1 (a) is also compatible with ultra-low-power applications especially when using fully integrated oscillators.

5.1.4 Sampling frequency

Besides the counter structure, two modes of operation are possible. Firstly, one can measure the period of the ROIC by counting how many edges of

5 Period Digitization and System-Level Considerations

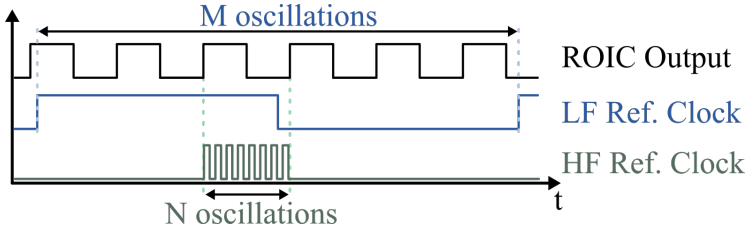


Fig. 5.6 Quantization of the ROIC oscillation.

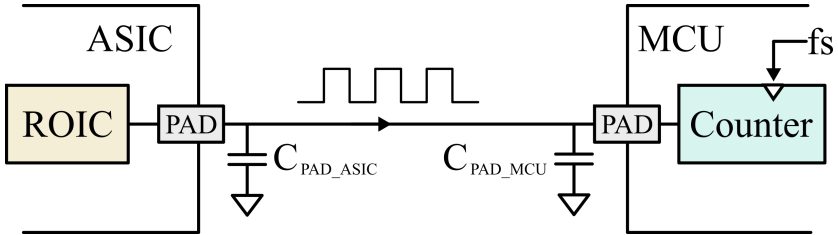


Fig. 5.7 Stray capacitances in the connection between the ASIC and the microcontroller.

a reference clock, with significantly higher frequency, fit in the period to be measured. This method, called input capture, is widely used in embedded systems. On the other hand, one can use a reference clock with much lower frequency and count how many ROIC oscillations fit in the reference period which we call: cycle counting, Fig 5.6.

With regards to sensor ROICs, the first approach has the advantage that it requires only one cycle of the ROIC to perform the measurements at the expense of running the counter at a very high frequency. The second approach, demands 2^n ROIC cycles for a n -bit, resolution, and in turn, a longer measurement time. Both approaches benefit from inherent first-order noise-shaping due to oversampling, as seen for VCO-based ROICs in chapter 3.

Fig. 5.7 models the stray capacitances found in the connection between both integrated circuits. The total EPC (accounting for the interconnection energy) is calculated in equations: 5.4 and 5.5 for both methods respectively.

$$E_{MeasurementHF} = E_{PC_{ROIC}} + E_{PAD} + 2^{\frac{SNR-1.76}{6.02}} E_{Count} \quad (5.4)$$

5.2 Conclusions

$$E_{MeasurementLF} = E_{Count} + (EPC_{ROIC} + E_{PAD})2^{\frac{SNR-1.76}{6.02}} \quad (5.5)$$

Where E_{Count} is the energy consumed per cycle by the counter and the timing source, and E_{PAD} the energy required to toggle from one to zero or vice versa the state of the communication path (pads, PCB traces, stray capacitances), Fig. 5.7, equation 5.6. The datasheet of the STM32U575ZIT6Q reports an input capacitance of 5 pF per pad, while the output capacitance for the X-FAB digital PAD is 0.618 pF (3.68 pF for analog PADs). Every centimetre of PCB trace (1 oz/ft² thick) separated 100 μ m from a ground plane or trace in the same PCB layer adds an additional 78 fF of parasitic capacitance to ground.

$$E_{PADSwitching} = \frac{C_{PAD}V_{DD}^2}{2} \quad (5.6)$$

The ratio of 5.4 and 5.5, designated β , is calculated in equation 5.7 and plotted in Fig. 5.8. From the plot it is observed that the most efficient approach is to quantify a single ROIC period by running the counter at a high frequency, determined by the target quantization noise. This is explained by the charge stored in the pads that is discharged to ground on every oscillation period.

$$\beta = \frac{E_{MeasurementHF}}{E_{MeasurementLF}} = \frac{EPC_{ROIC} + E_{PAD} + 2^{\frac{SNR-1.76}{6.02}} E_{Count}}{E_{Count} + (EPC_{ROIC} + E_{PAD})2^{\frac{SNR-1.76}{6.02}}} \quad (5.7)$$

In Fig. 5.9, the EPC of ULP1 and ULP2 is again plotted in the dashed line and the EPC including the estimated digitization energy in a solid line when using the timer of the STM32U575ZIT6Q microcontroller. We see that the for low resistances, the ROIC is already consuming significantly less energy than the timer, and the latter sets the limit. For higher resistances, the ROIC still dominates, and we conclude that the adaptive biasing techniques are effective and relevant in terms of diminishing the power consumption of the system as a whole.

5.2 Conclusions

From the perspective of the oscillator, we see a compromise between frequency stability and power consumption, with Pierce oscillator being the

5 Period Digitization and System-Level Considerations

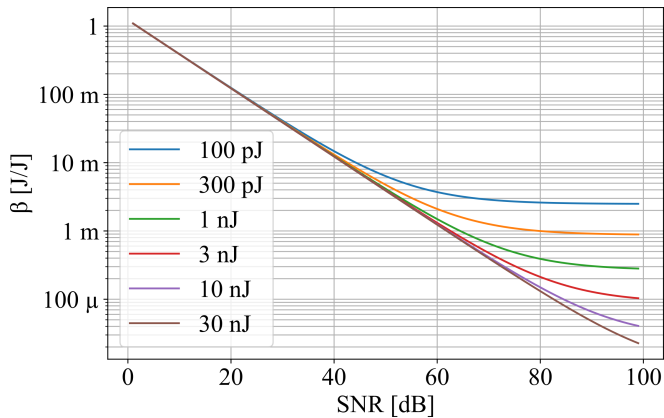


Fig. 5.8 Ratio of Energy for the two modes of period quantization for several ROIC EPC values.

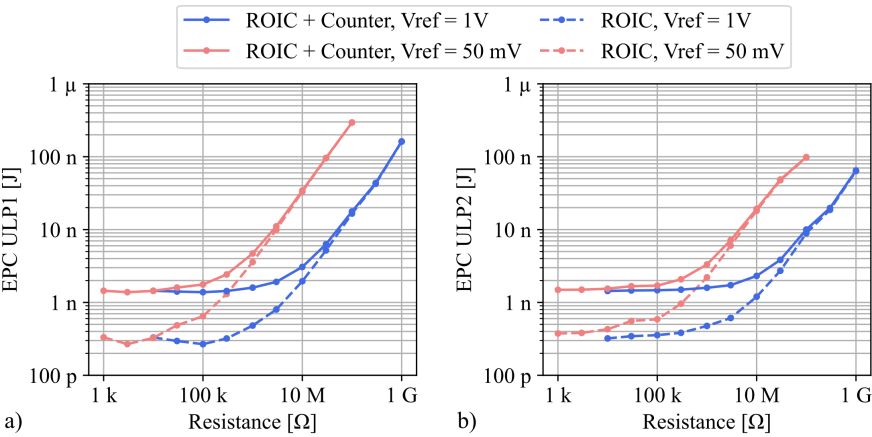


Fig. 5.9 a) EPC, with and without counter, for ULP1 b) EPC, with and without counter, for ULP2.

5.2 Conclusions

sturdiest choice, and integrated oscillators the most efficient. Given the CMOS process used for the ROIC and the performance of state-of-the-art microcontrollers, we conclude that it is best to use the clock source and timers already available in the microcontroller. Furthermore, we estimated the impact on the EPC of ULP1 and ULP2 and see a degradation for the highest frequencies where the ROICs EPC is lowest. The power consumption of the digital part sets a limit of the total EPC of the system at 1.12 nJ per conversion. Regarding the digitization method, input capture, a common feature found in most microcontroller embedded timers, takes less time per conversion and consumes less energy than cycle counting. This observation, once again, emphasizes the energy burden of high-speed chip-to-chip communications caused by PADs and PCB traces.

6

Conclusions and Perspectives

6.1 Conclusions

WHEN compared to current industrial gas sensing technologies, multi-sensor chemiresistive systems have the potential and the technological maturity level to offer a breakthrough solution in terms of power consumption, size and cost. As explained in this dissertation, the proper characterization and modelling of the sensors is needed to determine the architecture and design space of the ROIC.

After reviewing the state of the art, relaxation-oscillator based ROIC were found to be an efficient and flexible solution for the requirements of gas sensors given the large dynamic range achievable and sufficient SNR throughout an input range of over five decades. The power consumption profile of the ROIC was found to be proportional to the oscillation frequency and two adaptive biasing mechanism were proposed and tested to reduce the power consumption. The two techniques improved efficiency for large resistance sensors in a similar scale, reducing energy per conversion down to a tenth versus standard DC biasing. No other published worked presented simultaneously ultra-low power, an input range of over five decades and an SNR above 55 dB as seen in table 6.1. Biasing

6 Conclusions and Perspectives

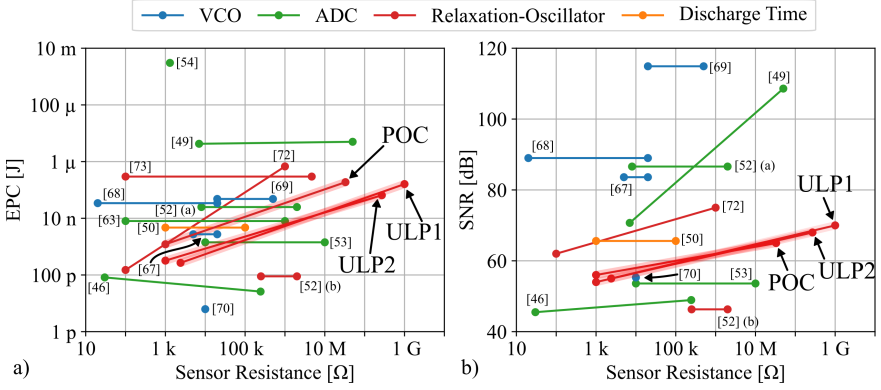


Fig. 6.1 a) EPC for several published ROICs separated by types. b) SNR for several published ROICs separated by types.

the comparators with a current proportional to that of the sensor (ULP2) yielded the largest improvement on EPC and was least affected by process variations. The zero-crossing bias current generator (ZXB) offers a significant improvement throughout the whole frequency range, at the expense of larger PVT variations. In all, these techniques are not mutually exclusive and a combination of both could be feasible. The effect of active and passive current mirrors was also researched, and we can conclude that active current mirrors can diminish nonlinearity for nA sensor currents with a small penalty in the EPC of 9 nJ (in a total of 161 nJ for 1 G Ω) and additional die area.

In table 6.1 and Fig. 6.1 we compare our three ROICs (POC, ULP1 and ULP2) against the most relevant cited works. VCO-based ROICs achieved the best SNR, at 114.9 dB for [69] and 89 dB for [67], at the expense of a high EPC (over 47.2 nJ in [67]) and an input range not greater than three decades in both cases. With regards to ADC-based ROICs, one remarkable work is that of [46], where by using a traditional Kelvin four-wire setup achieved a very low EPC, between 85.3 pJ - 23.8 pJ but lagging in SNR, reaching only 49 dB, reducing the limit-of-detection of many chemiresistive sensors which have a lower noise. As seen in Fig. 6.1 (a), all ROICs struggle to achieve low EPC or high SNR at high resistances, still time-based approaches benefit from the inherent time averaging, reducing noise at the expense of bandwidth.

In the beginning of this thesis, we laid out three questions that served

Table 6.1 Review of the state-of-the-art ROICs for resistive sensors.

Ref.	Type	Technology (nm)	Input Range (Ω)	Power (W)	EPC (J)	FOM (J/c.s.)	SNR (dB)	Nonlinearity
POC	RO	180	1 k - 33 M	$364 \mu^3$	1.21 n - 188 n ¹	2.09 p - 129 p ¹	56 - 65	-
ULP1	RO	180 SOI	2.4 k - 1 G	$1.8 \mu^3$	268 p - 161 n ¹	232 f - 78 p ¹	55 - 70	10 %
ULP2	RO	180 SOI	1 k - 270 M	$900 n^3$	321 p - 65 n ¹	219 f - 177 p ¹	54 - 68	10 %
[68]	VCO	180	20 - 20 k	171μ	34.2 n	1.48 p	89	0.005 %
[72]	RO	130	100 - 1 M	$680 \mu^3$	150 p - 675 n	32.6 f - 148 p	62-75	0.4 %
[53]	ADC	180	10 k - 10 M	1.7μ	1.42 n	-	53.6 ²	-
[50]	DT	65	1 k - 100 k	$44.5 n - 69.5 n^3$	4.7 n	2.54 p - 3.03 p	56 - 65	-
[52] a)	ADC	180	8 k - 2 M	$2.51 m^4$	$24.9 n^4$	1.43 p	86.6	-
[52] b)	RO	180	250 k - 2 M	$2.95 \mu^4$	$89.5 p^4$	0.53 p	46.3	-
[74]	DT	180	49.5 k - 98 k	15μ	2.73 n	14	47.5	-
[49]	CDS+ADC, $\Delta\Sigma$	180	7 k - 50 M	$425 \mu (\Delta\Sigma) - 500 \mu (CDS + ADC)$	$4.25 n (\Delta\Sigma) - 0.5 n (CDS + ADC)$	$91.9 (7 k\Omega) - 108.6 (120 k\Omega)$	-	-
[69]	VCO	110	20 k - 500 k	95μ	47.5 n	0.1 p	98.1	-
[56]	ADC	65	4.5 k - 1.53 M	85 n	-	-	-	-
[46]	ADC	65	30 k - 240 k	-	$85.3 p - 23.8 p^5$	552 f - 103 f	45.5 - 49	-

¹ Does not include the energy consumed by the counter.

² Not reported directly, calculated from the worst-case measurement error of 0.21 %.

³ Does not include the power dissipated in the sensor.

⁴ Sensor heater current of 118.8 μA not included.

⁵ Depends on the measurement rate, from 0.1 S/s up to 25 kS/s.

6 Conclusions and Perspectives

as a guide for our research. Now, after much work and many fruitful discussions, we return to these questions and offer some reflections on how our work contributed in these aspects.

1-What are the electrical characteristics of chemiresistive gas sensors of relevance to electronic interfaces?

The three main attributes that have to be considered when selecting a ROIC for chemiresistive gas sensors are the resistance input range, the biasing voltage and the SNR. The first two are important because of the large spread in the sensor characteristics and the maximum allowable voltage for each type of sensing layer, which will significantly affect the design space of the ROIC. The SNR is relevant because it is closely related to power consumption as both the ROIC noise and the quantization noise require additional power to be reduced.

From the point of the view of the sensor's modelling, as presented in chapter two, we were able to fully characterize several types of sensors and derive models that are useful for the ROIC design, that have not been published elsewhere. In a broader sense we learned that sensors are complex devices and there is no "one size fits all" solutions. Even though, our modelling of noise achieved similar results through two different methods, it is still unclear how the manufacturing process could alter the performance. Therefore, we feel that the likelihood of meeting the design performance of the system as a whole, is increased by diminishing the uncertainty of the many steps involved in the fabrication of such systems. Ideally, the ROIC should be designed once the manufacturing process of the sensors reaches a significant industrial maturity level, and the sensor parameters and its statistical profile had been measured.

2-What ROIC architecture is most suitable for a multi-sensor system?

We think we made the right decision by selecting relaxation-oscillator-based ROICs for a multi-sensor application. The designed ASICs validated this, and performed as predicted by simulations. The power consumption of relaxation-oscillator-based ROICs is in line with that achieved by best-in-class ADC-based designs. Given the sensors I-V curves, linearity was not a priority in the ROIC. Even though, we feel this is the right path, it brings to light the fact that gas sensors systems need to be calibrated at least at the factory. Allowing for nonlinearity reduced the energy consumption, because linearity is best achieved by overshooting the specification of a given parameter and adjusting its value by means of a negative feedback loop. We feel that with proper circuit design and layout, nonlinearity can be bounded and offer a predictable resistance-to-frequency curve. Once

again, the sensors sensitivity consistency plays a role in reaching an acceptable uncertainty level in instrumentation applications.

3-What are the implications in terms of power consumption and SNR for a relaxation-oscillator-based ROICs interfacing chemiresistive gas sensors?

As is frequent in engineering design, several parameters trade-off with each other. To fit more sensors in a die of a given size, they have to be made smaller. This implies that the sensor current will be smaller and hence the ROIC will operate at a lower frequency, with a higher EPC and a longer measurement time. Ultimately, a sweet spot has to be found where the benefit is maximized for the given sensing requirements.

Adaptive biasing techniques were successful in reducing power consumption for long measurement times, demanding little additional circuit area and being transparent to the user. Additionally, when the ROIC power consumption decreases other parts of the system become proportionally more power demanding imposing a lower limit, as we saw for the counter and digital blocks. Therefore, below 1 nJ, we were faced with a situation of diminishing returns when reducing the EPC of the ROIC.

Another aspect closely related to the power consumption is the bandwidth of the ROIC which is determined by the oscillation frequency. Decreasing the frequency, decreases the bandwidth and therefore the integrated noise, increasing the SNR. This has to be kept in mind when sizing the sensors, as the limit-of-detection, at least in part, will be in part set by the power consumption.

6.2 Future Work

As we progressed with the research, new question arose that were beyond the scope of this thesis but we feel they are important for future researchers to consider.

Beginning with the sensors, we feel that the work done has contributed to the characterization and modelling of these devices, especially in terms of noise which is often overlooked in publications. Besides this, given the vast amount of sensing layer formulations, and fabrication methods of sensors, further measurements could help validate the results found in this work. One dimensions that was not explored is the ageing of the sensors. Even though they are not consumable, like electrochemical gas sensors, the materials are still prone to oxidation, dust, UV light and the environment in general.

6 Conclusions and Perspectives

Even though the sensor's fabrication process was not one of the objectives of this thesis, it is interesting to note that the developed ROICs can be used during the electropolymerization to control the amount of polymer formed and therefore the resistance of the resulting sensor.

There are a few directions in which the research of the ROIC could proceed further. Firstly, relaxation oscillators are not exhausted and techniques of smart biasing to reduce the power in the sensor could be explored as has been done in publications for temperature sensors. Secondly, we focussed on 180 nm CMOS processes because its performance is sufficient for the design requirements. Using smaller, justified by the inclusion of digital IP like CPU or RF, may offer benefits like reducing the power of the comparators but may put a burden in the design if the supply voltage is restricted to e.g. 1.2 V. Thirdly, the tuning of parameters, like the size of the integration capacitor, has been left out of this work to make the ROIC as autonomous as possible but it has been proposed and tested in publications with good results to reduce the counter frequency and save energy.

Annex A - Aluminium Electrodes for Chemiresistive Sensors

IN both fabricated ASICs described in chapters 3 and 4 we included IDEs in the top metal layer. Normally, semiconductor foundries apply layers of silicon dioxide and/or silicon nitride, which are electrical insulators, as a last deposition in the fabrication process. This layer, a few micrometers thick, is mainly used as protection from the environment and post-dicing processing. After the deposition, a final etching step is done to expose the bonding pads to produce a good contact with the bonding wire. In prototyping runs, both UMC and X-FAB agreed to have pad openings besides that of the pads, to expose a IDEs placed in the middle of the ASIC and build chemiresistive sensors on top of them. In the first ASIC, a set of interdigitated electrodes with size $200\text{ }\mu\text{m}$ by $200\text{ }\mu\text{m}$ was included, Fig. 7.1. The finger width is $2\text{ }\mu\text{m}$ and the spacing between them $6\text{ }\mu\text{m}$. In the second ASIC, 16 sets of electrodes were included per ROIC (ULP1 and ULP2) with 4 different dimensions each:

- A- $50\text{ }\mu\text{m}$ by $60\text{ }\mu\text{m}$
- B- $50\text{ }\mu\text{m}$ by $350\text{ }\mu\text{m}$
- C- $200\text{ }\mu\text{m}$ by $250\text{ }\mu\text{m}$
- D- $250\text{ }\mu\text{m}$ by $350\text{ }\mu\text{m}$

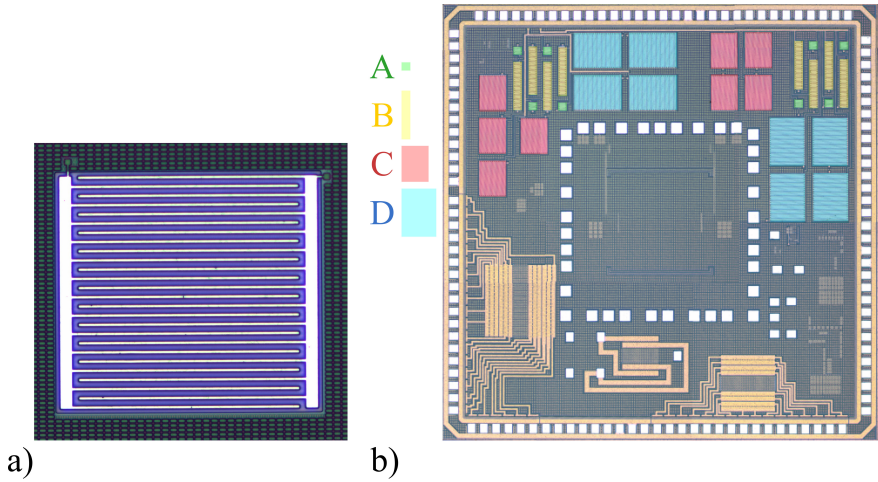


Fig. 7.1 a) Optical microscope photograph of the IDEs of the UMC ASIC. b) Optical microscope photograph of the IDEs of the X-FAB ASIC.

The fingers have an equal width and spacing of $3\ \mu\text{m}$. According to the process documentation, in both UMC and X-FAB, the top metal is made of aluminium. In X-FAB, it is alloyed with 0.5 % copper, for UMC no alloying element is indicated in the datasheets.

We did several attempts to try to fabricate a sensor with these electrodes by dropcasting but only measured non-significant leakage currents, mostly from the pads structures. The main culprit is most likely the native oxide layer that forms on the aluminium. As the etching on the electrodes was not standard in the fabrication process, we did an extensive characterization of the electrodes with: SEM + FIB (Zeiss Auriga), profilometry (Veeco Dektak 150), white-light interferometry (Polytec MSA-500) and optical microscopy (Zeiss Axio Imager Vario A2).

Figs: 7.2, 7.3, 7.4 were taken with the SEM and without any cleaning or etching of the IDEs. The electrodes are fully exposed on the three sides and there is significant over etching of the oxide layer below the top metal. A thin layer, 200 nm thick, with a different shade is seen at the top, Fig. 7.4. We did a wet etch cleaning with hydrofluoric acid and a plasma etching (EmiTech K1050X) but the conductance did not increase.

Profilometry (Fig. 7.5) and white-light interferometry (Fig. 7.6) did not provide enough resolution. The tip radius in the former was not able to

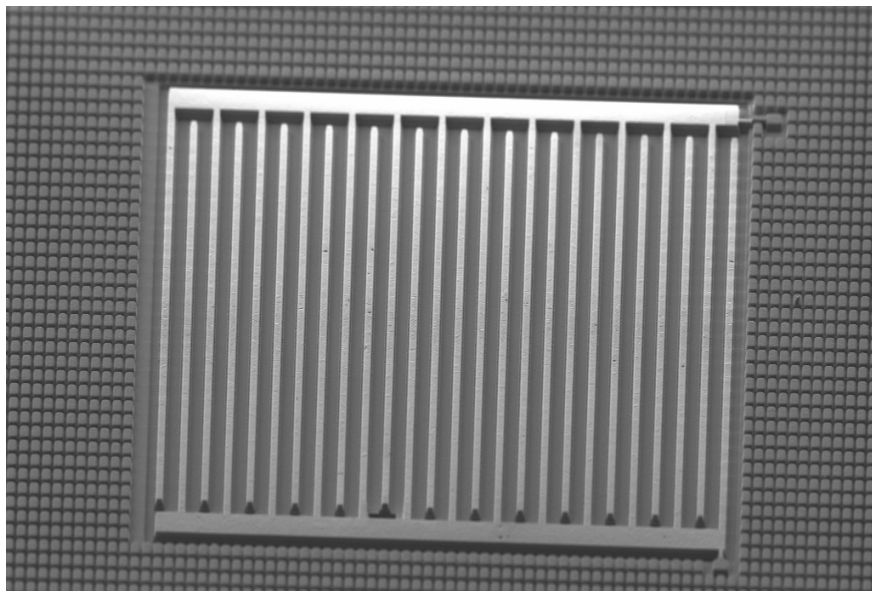


Fig. 7.2 SEM image of the interdigitated electrodes of the first ASIC

reach the bottom of the trench between the fingers due to the over-etching. With regards to interferometry, it does not produce a good representation of the edges, therefore, the fingers width is significantly overestimated to roughly the double of what was measured with optical and SEM microscopy.

We did one last attempt by increasing the electrical conductivity of PANI with gold nanoparticles, Fig. 7.7. In this way, the voltage drop occurs mainly at the oxide layer, consequently increasing the magnitude of the electric field. The dielectric breakdown of AlO_2 is below 5 MV/cm, [100] and the native oxide thickness of aluminium is between 2 nm and 4 nm [101]. Therefore, a voltage of 4 V could cause electrical breakdown if the conductivity of the sensing layer is high enough. The gold nanoparticles, 50 nm in diameter, are seen in Fig. 7.7 as small shiny dots. Unfortunately, the conductivity was once again very large resulting in a nonfunctional device.

As a closing remark, we repeated some of these experiments with the electrodes in the second ASIC and achieved comparable results. Therefore, we can conclude that aluminium electrodes are not viable for produc-

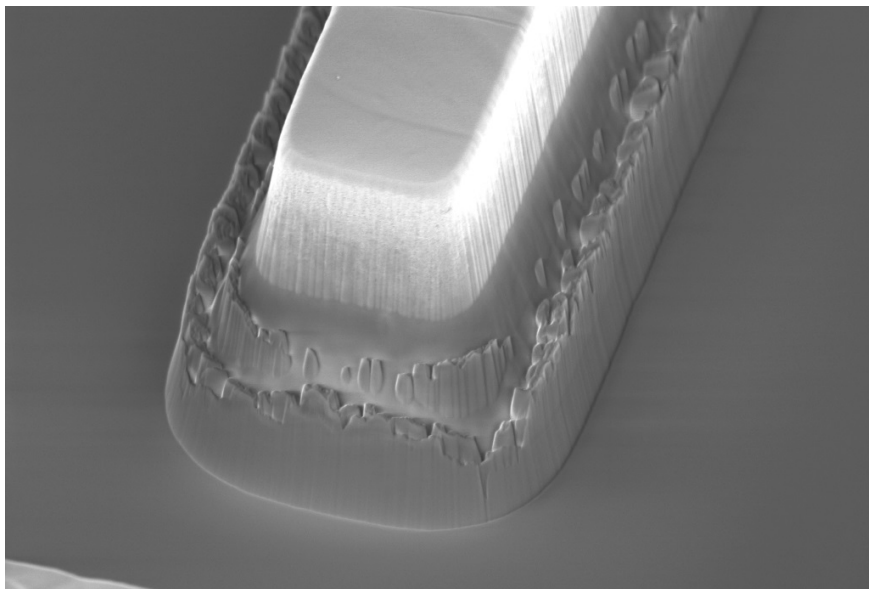


Fig. 7.3 SEM image of the end of one fingers of the first ASIC

ing resistive sensors electrodes, even when alloyed with small amounts of copper.

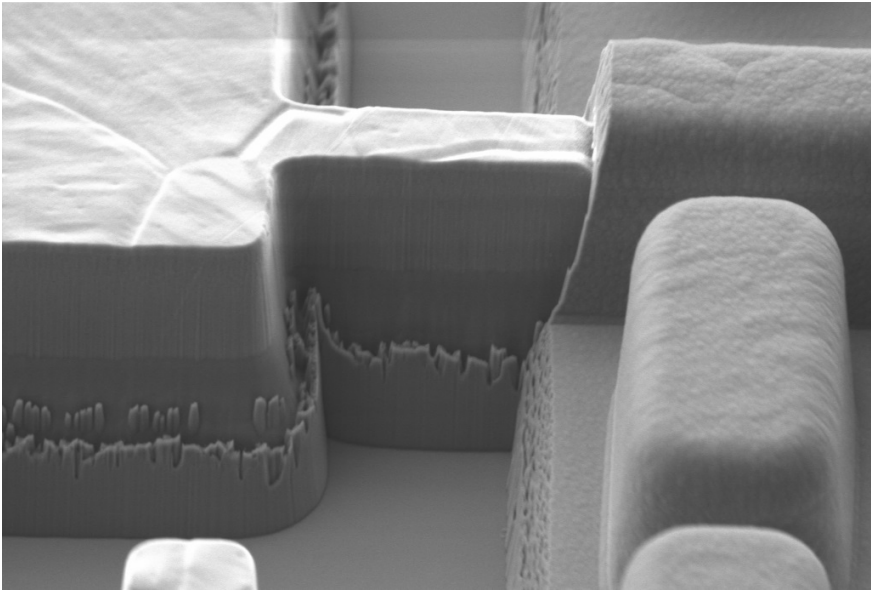


Fig. 7.4 SEM image of the end of the contact of one of the electrodes of the first ASIC

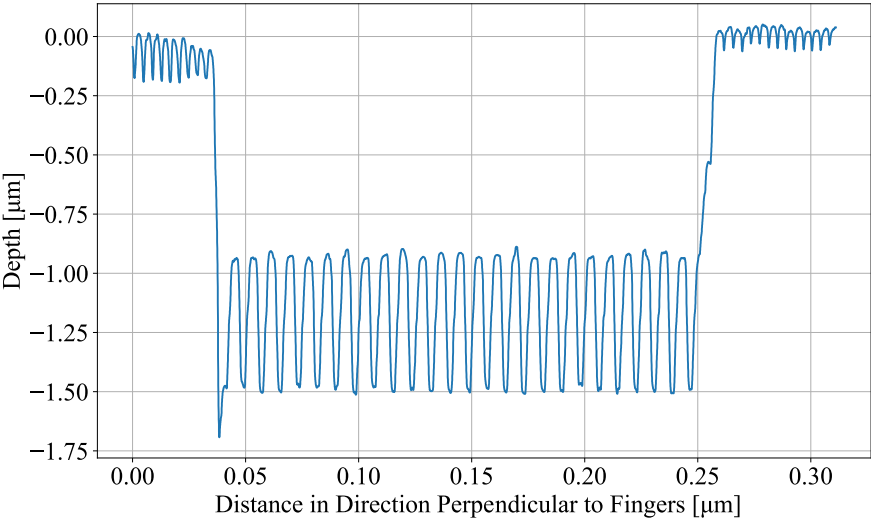


Fig. 7.5 Profilometry measurement of the electrodes of the first ASIC.

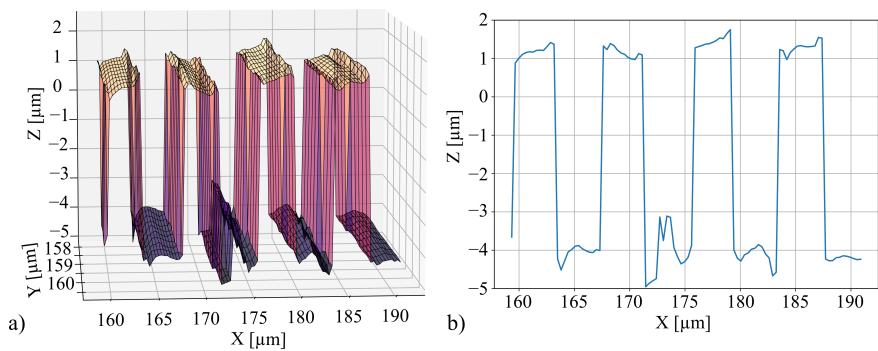


Fig. 7.6 White-light interferometry measurements of the electrodes of the first ASIC.

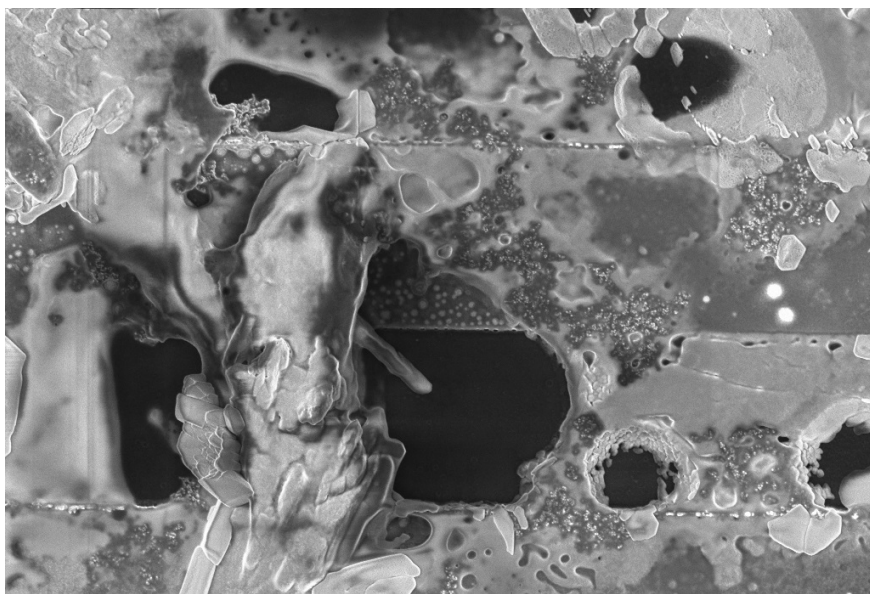


Fig. 7.7 SEM image of the interdigitated electrodes of the first ASIC after the application of the PANI sensing layer with gold nanoparticles.

Annex B - Linearity of Relaxation Oscillator-Based Interfaces

THE large input range required for chemiresistive sensors puts a strain on the linearity of the sensor interface. Even though the nonlinearity of the sensor and that of the ROIC can be compensated with a single linearization curve in the post processing of the measurement, it is useful to understand what factors affect the linearity of the ROIC and how it could be reduced.

In Figs. 8.1 and 8.2 the deviation of the sensor bias voltage from the reference voltage are plotted for several resistances and reference voltages as measured for ULP1 and ULP2. At the same time, in Figs. 8.3 and 8.4 we present the deviation of ULP1 and ULP2 from the ideal transfer function for all the combinations of reference voltage and resistances tested.

As previously noted for the POC ROIC in chapter 3, in ULP1 and ULP2 there are also two regions where nonlinearity is greatest: for the largest and the lowest sensor resistances. In the case of ULP1, we see that the biasing circuit is still operating with sufficient linearity when the the gain of the ROIC ψ ($T = \psi R$, equation 3.11) increases. This means that the bottleneck is in the comparators, whose delay becomes noticeable. As was explained in chapter 4, the efficiency of ZXD is diminished when the slew rate at the input of the comparator increases, reducing the real bias current of the comparator in this situation. This phenomenon does not happen in ULP2, where we see that the linearity degradation is closely correlated to the bias

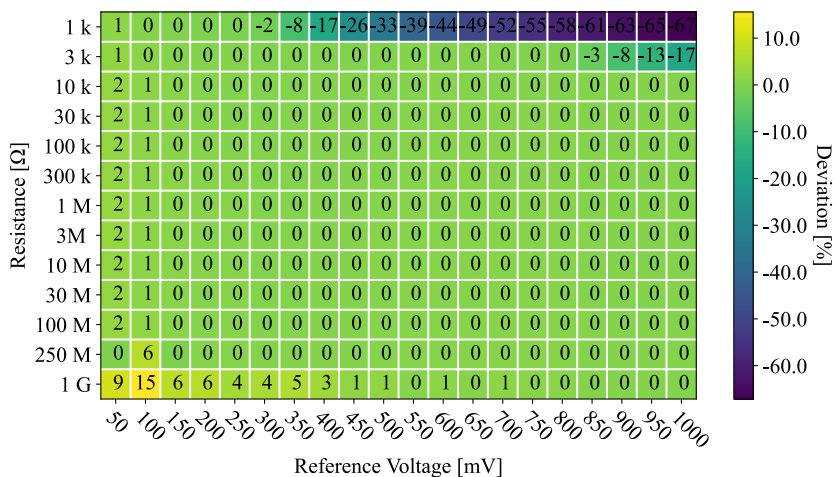


Fig. 8.1 Heatmap of the mismatch between the sensor and reference voltage for ULP1.

voltage error, therefore in this situation the comparator's delay is not of concern.

For the second situation, when the sensor current is lowest (less than 5 nA) we see that usage of regulated cascodes significantly improves the linearity between 1 nA and 5 nA. On the other hand, below 1 nA both interfaces are very nonlinear. This is not explained by a mismatch in the bias voltage of the sensor alone but is also caused by the nonlinearity in the current mirrors, leakage currents and charge injection in the MOS switches that cause the integration capacitor to charge faster than expected.

Lastly, we can compare the measurement to what was expected from simulations. In Figs. 8.5, 8.6, 8.7 and 8.8 the simulated transfer function for ULP1 and ULP2 and V_{ref} of 50 mV and 1 V are plotted. The simulation predicted accurately the operation of ULP2 in all situations. On the other hand, ULP1 did show a deviation from the simulation at high frequencies, not reaching the expected frequency in measurements. For example, with $V_{ref} = 1$ V, and $R = 30\Omega$ the simulation deviation from the ideal transfer function was 7 %, while 19 % was measured. One possible explanation, besides the inaccuracy of the EDA models, is the fact that the ZXD relies on fast pulses of current to bias the comparators. Even though internal decoupling capacitors were used, it is not possible to reliably include in

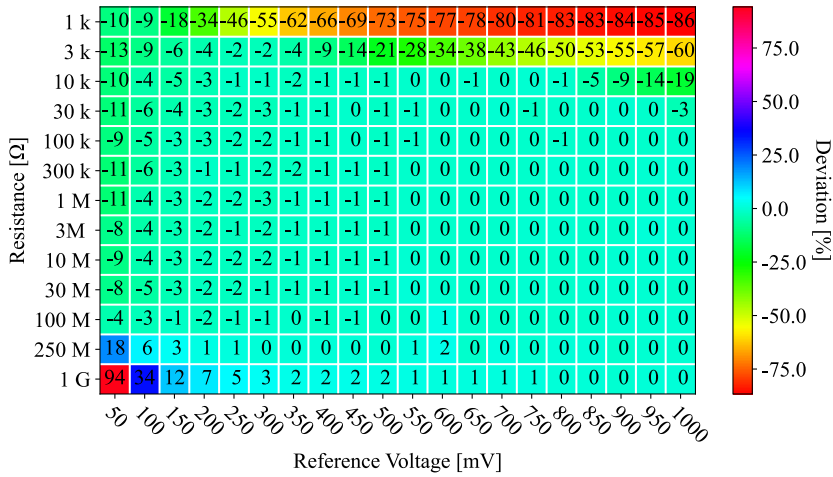


Fig. 8.2 Heatmap of the mismatch between the sensor and reference voltage for ULP2.

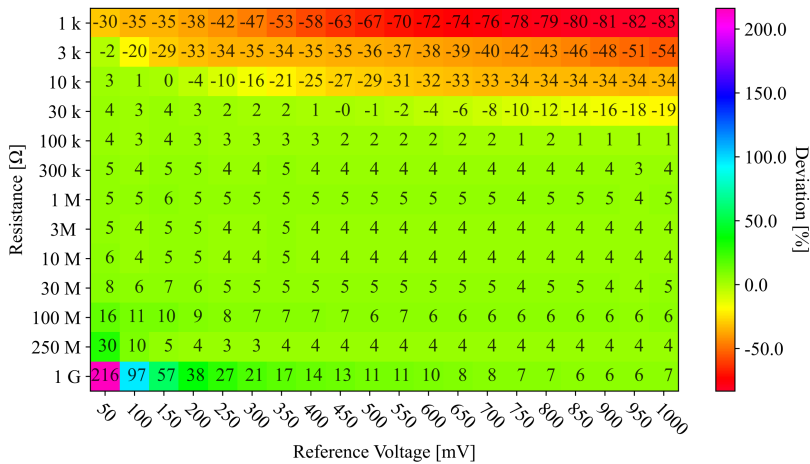


Fig. 8.3 Heatmap of the deviation from an ideal transfer function for ULP1.

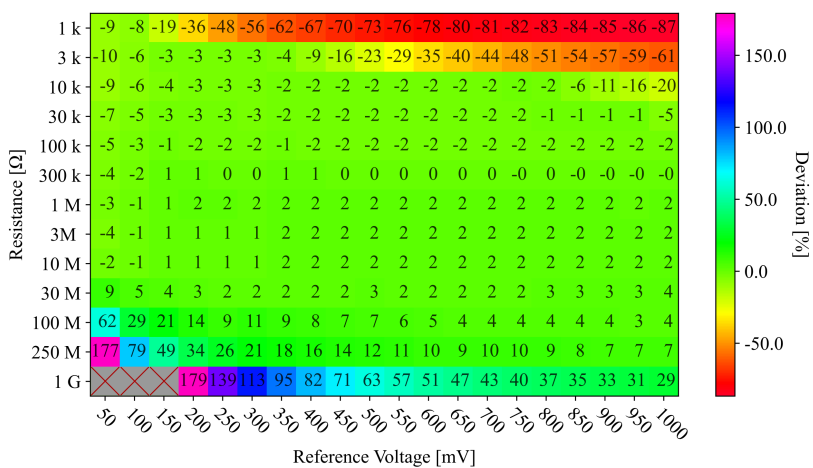


Fig. 8.4 Heatmap of the deviation from an ideal transfer function for ULP2.

the simulations the effects of external parasitics like wire bonding and PCB traces.

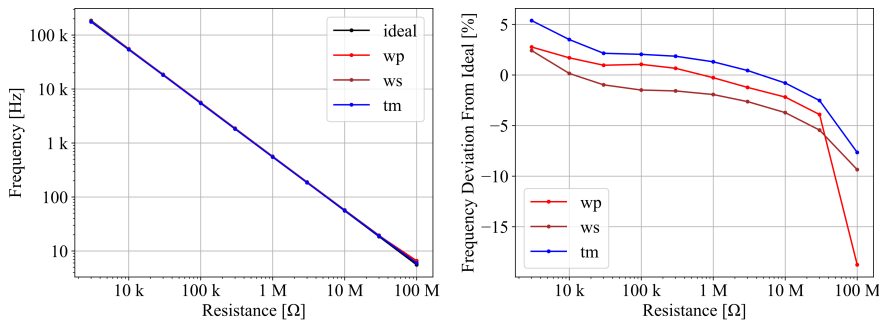


Fig. 8.5 Simulated transfer function of ULP1 for typical, worst power and worst slow corners for $V_{ref} = 50$ mV.

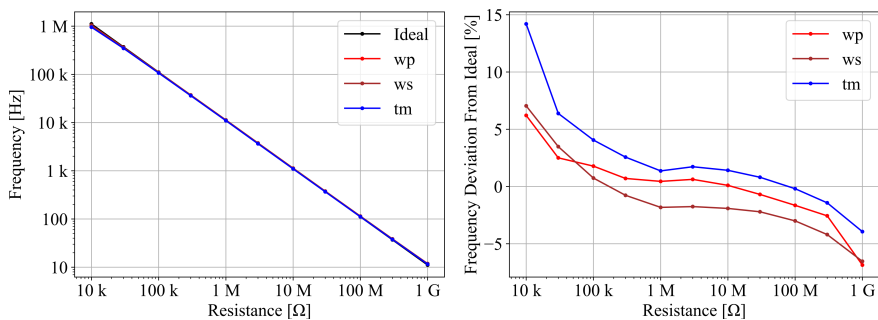


Fig. 8.6 Simulated transfer function of ULP1 for typical, worst power and worst slow corners for $V_{ref} = 1$ V.

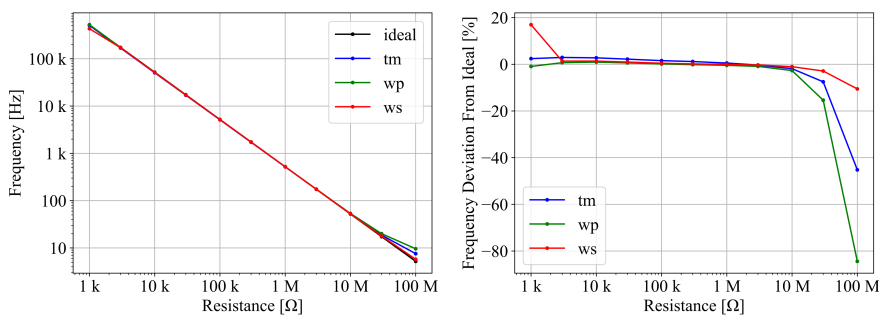


Fig. 8.7 Simulated transfer function of ULP2 for typical, worst power and worst slow corners for $V_{ref} = 50$ mV.

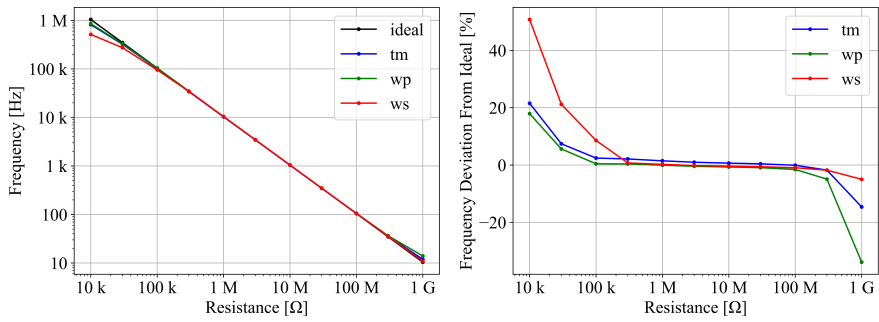


Fig. 8.8 Simulated transfer function of ULP2 for typical, worst power and worst slow corners for $V_{ref} = 1$ V.

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