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Designing Microsystems for Liquid Electrical Properties and Variable Magnetic Fields Measurements

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Abstract

The contents of the present thesis relate the design, characterization, and integration of sensing systems in order to obtain portable, simple, and power efficient solutions. A methodology of instrumentation system design is proposed in order to push the maturity of the systems technology at the end of the fundamental research. In order to evaluate the design progress of any instrumentation solution, we built our methodology in comparison with the Technology Readiness Level (TRL) scale. This indicates the TRL level we have to reach in order to prove the possibility of the industrial development of any technology. We assume in the present context that every instrumentation system includes a sensing element and a circuit interface; the design of such systems therefore requires making a judicious choice of the transduction principle of the sensed information.

The scientific challenges faced here focus on the interfacing principles of sensing elements by their insertion into triangular voltage oscillating circuits. The data delivered by a sensor deforms the analog output of the oscillator and is extracted by mathematical methods. We investigate the use of such oscillators in two well defined contexts: the interfacing of interdigitated electrode sensors for liquid sensing applications and the interfacing of inductive devices for the detection of abrupt variations of magnetic field.

The first sensing system we developed according to the design methodology we evoked before is an R-C impedance liquid sensing system. Consisting in an interdigitated electrodes (IDE) structure, the sensing device is studied and modeled using FEM tools. The proposed interfacing principle consists in stimulating the device with a square current wave; this results in a triangular output voltage of which shape contains the expected liquid parameters. These parameters are extracted by mathematical curve fitting. The interfacing method is implemented on a PCB platform and tested for two applications; the saline concentration and water-glycerol mixtures characterization. The architecture of the integrated circuit interface is next proposed in order to improve the sensing performances of the system in terms of parameters extraction accuracy and power consumption reduction. The results obtained for the fluid sensing system development brought improvements in terms of resistive resolution ($100\ \Omega$) and capacitive resolution ($120\ \text{fF}$).

According to our design methodology, we developed a detection system of magnetic field variations. The sensing element considered in this context is a square planar inductive element which is modeled by FEM tools similar to the IDE system. Being also based on a triangular voltage oscillating system, the

interfacing method we built consists in varying the triangular signal frequency for each detected magnetic perturbation. Mathematical output characteristics are consequently defined to quantify the amount of magnetic perturbations caught by the inductor. The system is tested for the Arc Fault Detection (AFD) application which consists in measuring the abrupt variations of current on a power line. The circuit interface is finally simulated on a CMOS integrated technology to improve the power consumption ($\approx 50 \mu\text{W}$) and portability of the solution ($\approx 2 \text{ cm}^2$) in comparison with existing systems that often consume several mW to detect magnetic field variations with acceptable sensitivity.

List of Publications

Journal papers

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N. André, S. Druart, P. Gérard, R. Pampin, L. Moreno Hagelsieb, T. Kezai, L. A. Francis, D. Flandre, J.-P. Raskin, *Miniaturized wireless sensing system for real-time breath activity recording*, IEEE Sensors Journal, Vol. 10, pp.178-184, 2010.

Conference papers

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N. André, L.A. Francis, B. Rue, S. Druart, P. Dupuis, D. Flandre, J.P. Raskin, *Ultra low power SOI transducer for flow and dew-based humidity sensing*, Proceedings of the 2011 CMOS Emerging Technologies Workshop, Whistler (Canada), 2011.

N. André, L.A. Francis, S. Druart, P. Dupuis, D. Flandre, J.P. Raskin, *Portable wireless microsensing system for human breath monitoring*, Proceedings of the 2010 CMOS Emerging Technologies Workshop, Whistler (Canada), 2010.

L. Moreno-Hagelsieb, N. André, T. Kezai, O. Bulteel, S. Druart, R. Pampin, B. Rue, B. Olbrechts, J.P. Raskin, D. Flandre, *UCL SOI Sensors Advantages for Consumer and Health Monitoring Applications*, Proceedings of the Commercialization of Micro and Nano Systems Conference, Mexico D.F. (Mexico), 2008.

N. André, P. Gérard, P. Drochmans, T. Kezai, S. Druart, L. Moreno-Hagelsieb, L.A. Francis, D. Flandre, J.P. Raskin, *Wireless microsensors system for monitoring breathing activity*, 4th European Conference of the International Federation for Medical and Biological Engineering IFMBE Proceedings, Antwerpen (Belgium), Vol. 22, pp.2396-2399, 2008.

T. Kezai, N. André, P. Gérard, P. Drochmans, S. Druart, L. Moreno-Hagelsieb, D. Flandre, J.P. Raskin, *Wireless Sensor Network For Breathing Activity Monitoring*, Proceedings of the SR05 Satellite Conference - Topics in Electrical Circuits and Systems, Fez (Marocco) 2008.

L. Moreno-Hagelsieb, R. Pampin, S. Druart, D. Flandre, *Al/Al₂O₃ Biosensors for DNA and Medical Applications: Materials and Design Concepts, Methodology and Perspectives*, Proceedings of PRO-BIOSYS workshop, Micro-Nanotechnology for Biomedical Applications (Bio-chips), Bucharest (Romania) 2008.

S. Ranvier, V. Rochus, S. Druart, H. Lamy, P. Rochus, L.A. Francis, A

MEMS-based xylophone bar magnetometer for pico satellites, Proceedings of the 61st IAC, International Astronautical Congress, Prague (Czech Republic), 2010.

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List of Acronyms

ADC	Analog to Digital Converter
AFD	Arc Fault Detection
ALD	Atomic Layer Deposition
ASIC	Application-Specific Integrated Circuit
CCO	Current-Controlled Oscillator
CMOS	Complementary Metal-Oxide Semiconductor
CPE	Constant Phase Element
EDL	Electrical Double Layer
FEM	Finite Element Model
IDE	InterDigitated Electrodes
LTI	Linear Time Invariant
MEMS	MicroElectroMechanical System
MOS	Metal-Oxide Semiconductor
MOSFET	Metal-Oxide Semiconductor Field Effect Transistor
NMOS	N-type MOSFET
OTA	Operational Transconductance Amplifier
PCB	Printed Circuit Board
PMOS	P-type MOSFET
RF	Radio-Frequency
RFID	Radio-Frequency Identification
TRL	Technology Readiness Level
TSMC	Taiwan Semiconductor Manufacturing Company
TWV	Triangular Waveform Voltage
ULP	Ultra Low Power

CHAPTER 1

Introduction: context and motivations

1.1 Instrumentation and technology

The improvement of the scientific knowledge is motivated by the wish for understanding the world in which we are living. Indeed, the environment is governed by a multitude of physical laws our civilization constantly tries to master and use for improving our quality of life. The study of a physical phenomenon is an activity that requires us to detect or measure some specific features related to this phenomenon; sensing methods are therefore in constant development and improvement.

For the past few years, our human civilization entered into the era of advanced electronics; the major trend of this era is to develop efficient, low-power, portable, and non-invasive systems. This trend, undoubtedly motivated by the growing market of instrumentation solutions, is especially due to the user's requirements: performing tests directly in a relevant and operational environment, having the opportunity of placing sensors in confined areas, and avoiding the use of a non-portable power source [1].

Therefore, the design of an instrumentation solution is a rigorous activity that has to respect a well thought out methodology of development which defines without any ambiguity what we want to measure, how and where to perform this measurement, and what kind of output signal the system will deliver. Important questions consequently have to be solved: what is the most adapted sensing element? What is the most efficient interfacing circuitry? How to reduce its power consumption and size? How to retrieve the sensed data using the most accurate and immediate method?

The main objective of this thesis is to present a design methodology of miniaturized sensing solutions; this methodology has a specific challenge: answering all those questions while pushing the technology maturity of a downscaled sys-

tem up to the beginning of its industrial development. The technology maturity of a given system is evaluated by the well-known Technology Readiness Level (TRL) scale (Tab. 1.1); the definition of the TRL scale is mainly used by American governmental agencies and is provided by several websites (ex.: NASA or U.S. Army websites). An example of the TRL scale application is noted for the maturity evaluation of CMOS circuits to be radiation tolerant [2]. The scale is also often used for project monitoring and management [3]. We describe here the nine steps of the TRL scale in the context of a sensing application.

TRL level	Description
TRL 1	Identification of the phenomenon to sense and governing laws.
TRL 2	Review of the technology and definition of the sensing elements.
TRL 3	Simulation of the sensor response and mathematical modeling.
TRL 4	System implementation in lumped blocks and technology characterization in laboratory.
TRL 5	Laboratory tests of the sensing solution using certified equipment.
TRL 6	Test of the sensing system in a relevant environment.
TRL 7	Test of the sensing system in an operational environment (ex: airplane).
TRL 8	The sensing system satisfies to industrial constraints and performances.
TRL 9	Final verifications (bug fixing) for industrial production.

Table 1.1: TRL scale described for the design of instrumentation systems.

1.2 Sensing system design methodology

In the present thesis, the goal of the design methodology is to prove that reducing the complexity, size, and power consumption of the existing instrumentation systems is achievable; opening thus the possibilities for the industrial development. From the TRL scale point of view, the principal tasks we conduct in the present thesis are focused on the maturation of the technology from TRL1 to TRL5. From the researcher's point of view, the maturity of the technology is clearly proved at this level. Once the research activity is pushed beyond TRL6, the resulting demonstrators are developed into industrial prototypes.

1.2.1 Sensing chain description

The role of an interfacing system is to provide a specific environmental set of parameters $(\varepsilon, \alpha, \sigma, \beta, \dots)$ to a user. Actually, the output of the instrumentation system is an estimation $(\hat{\varepsilon}, \hat{\alpha}, \hat{\sigma}, \hat{\beta}, \dots)$ of the expected real values of the sought parameters. The estimation is performed by a sensing chain especially designed to extract one or several well-defined environmental features.

The design of a sensing system requires first to identify the different blocks composing the complete system chain and secondly how those blocks interact with each other. The head block is the one which directly senses the environmental data. The task of the second block of a sensing system is to collect the data sensed by the head element and to transform it into an electrical signal; this block is the direct electrical interface. The last block of the sensing chain reads the electrical signal provided by the direct interface and extracts the features related to the sensed data by mathematical processing functions. The global view of the sensing system chain is depicted on Fig. 1.1.

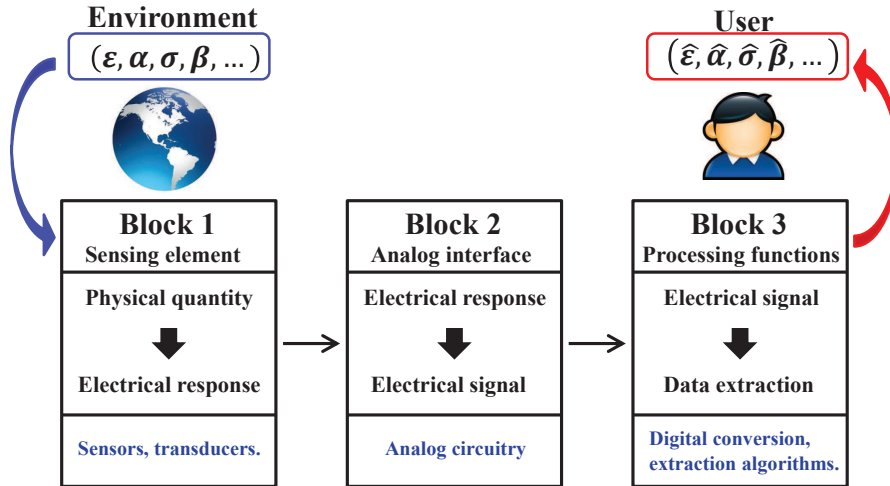


Figure 1.1: Sensing system global block diagram.

1.2.2 Methodology steps

The system design methodology is divided into steps or stages which describe each advancement that has to be performed to develop the whole solution [4]. The design approach proposed throughout this work is divided into seven well

defined steps. In order to compare the research progression to the maturity of the technology, a parallelism is made between those steps of research and the five first TRL levels.

1.2.2.1 Step 1: Context and review of existing solutions (before TRL 1)

Before entering in the TRL scale, the principle of the detection system is first chosen considering a specific sensing application. The design of a novel instrumentation system then requires an analysis of existing solutions. This allows for positioning the present work and for choosing an innovative research direction.

1.2.2.2 Step 2: Detection principle (TRL 1 $\rightarrow$ TRL 2)

The detection or measuring principle is proposed and explained. The head element of the sensing chain is analyzed in terms of structure and materials and is next modeled as an equivalent electrical circuit. The electrical response of the sensing element is therefore defined.

1.2.2.3 Step 3: Sensing element physical modeling (TRL 2 $\rightarrow$ TRL 3)

This step consists in the calculation of each element that composes the electrical model resulting from the previous step. This calculation is based on the physical laws that govern the sensing principle. The head element structure is represented in a mathematical domain in which the governing equations are solved. Since sensing element geometries are complex, the use of Finite Element Modeling (FEM) tools is preferred. The electrical model operation is therefore predicted numerically and compared to existing analytical formulas.

1.2.2.4 Step 4: Sensing element sizing (TRL 3)

The calculation tools being established, an optimization approach of the sensing element geometry has to be proposed. Considering the position of the head block in Fig. 1.1, this sizing approach has to make assumptions on the input data (i.e. environmental data) and output data (i.e. the signal provided to the interface block). In other terms, the signal provided by the sensing element has to give representative information of the whole range of sensed data. This output signal also has to be measurable by the electrical interface. The sizing method result is an algorithm that gives the sensor dimensions as a function of the expected sensed data range and circuit interface detection abilities.

1.2.2.5 Step 5: Interfacing principle (TRL 3 $\rightarrow$ TRL 4)

This fifth step of the design methodology is the definition of the interfacing principle of the head element. The role of the circuit interface is to deliver an electrical signal from which the sensed data can be extracted. Considering Fig. 1.1, this step focuses on block 2 and block 3; the analog signal provided by the interface is described by formal equations. Processing functions or extraction formulas are next defined to retrieve the sensed data.

1.2.2.6 Step 6: Implementation and measurements (TRL 4 $\rightarrow$ TRL 5)

The five previous steps were the theoretical part of the sensing system design method. All these concepts have to be implementable on a physical platform. The goal of this step is to present the circuit architecture that implements the sensing system. The chosen direction is to test the system using off-the-shelf discrete components. The system is next characterized by the reproduction of a representative environment in a laboratory context. This approach is low-cost and validates all the theoretical results developed in the previous steps. In addition, the discrete implementation allows for highlighting all the limitations that have to be considered in the final integration step.

1.2.2.7 Step 7: Integration: sensing performances (TRL 5)

The sensing system being designed and validated using a discrete implementation, the objective of this last step is to integrate the circuit interface on miniaturized circuit technologies (i.e. CMOS). The major contribution of this step is the huge improvements of the interfacing performances and therefore the sensing performances. Indeed, the miniaturization attenuates a lot of parasitic effects reported in the discrete implementation, reduces the size of the system, and drastically decreases its power consumption.

1.2.2.8 Development research work (after TRL 5)

The TRL 5 indicates the end of the fundamental research activity and therefore the validation of the proposed sensing context. The next part of the work is considered as development research and principally consists in optimization that pushes further the solutions performance in terms of power consumption and compactness. As discussed further in this dissertation, we focus on the functional part of the sensing system. All the other system components which regroup power supply blocks and processing digital circuits are part of the development research work that starts from the TRL 6. The end of the TRL scale, i.e. TRL 8 and TRL 9, is industrial development. For example, these last TRL steps include the adaptation of the systems to specific safety norms or user's requirements.

1.3 Sensing applications and system integration

The research work presented in this dissertation focuses on two sensing contexts: the liquid impedance measurement and the detection of magnetic field variations. These two research directions are motivated by the possibilities of developing sensing systems useful for the industry and the research activities. Those contexts have also been chosen to be fundamentally different from each other in order to prove the efficiency of the design methodology.

1.3.1 System 1: Liquid impedance sensor

The first sensing system consists in measuring the permittivity and the conductivity of any liquid using interdigitated electrode structures. The principal reason that motivates the choice of this context lies in the possibilities of extracting a lot of useful parameters from the conductivity and the permittivity of any liquid. As examples, the conductivity is used to evaluate the salinity of the solution and the permittivity measurement can be used to identify some specific compounds of a liquid mixture. Many other examples justify the usefulness of measuring the impedance of a liquid.

1.3.2 System 2: Magnetic field variation sensor

The goal of the second sensing system is to detect the sharp variations of a magnetic field using an inductive sensor. Such a system could be used to detect the electric arc faults on a power current line. This application context is present in the domain of the health monitoring of high powered installations and equipment. The consequences of failures are indeed more important in high voltage installations; justifying therefore the high expansion of the health monitoring solutions market.

1.3.3 System integration

Since we entered into the era of microsystems design, there are real interests of downscaling any functional circuit in order to obtain low-power, compact and smart systems. The possibility of sensor interface integration is therefore investigated in the present work. The development of an integrated system is a complete activity that includes several skills in the circuit design field. Indeed, the design of an ASIC (Application-Specific Integrated Circuit) requires several operating blocks. In order to give a more concrete explanation of the integration problematic, the illustration of Fig. 1.2 shows the functional diagram of a

fully-integrated instrumentation system.

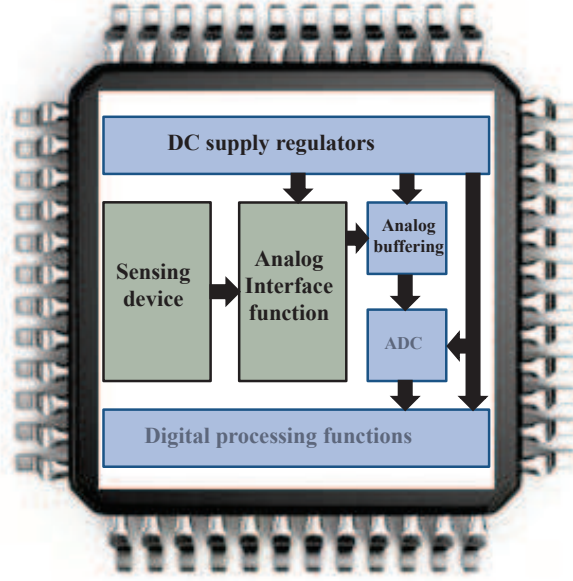


Figure 1.2: Functional diagram of a fully-integrated instrumentation circuit.

Power regulation blocks are required to provide stable and accurate supply to the different circuit components and digital units are often used to directly convert the analog information into numerical data which is easier to process by external high-level computer tools. In the present context, we focus on the core of the analog interface (central block of Fig. 1.2); the idea is to analyze this component in order to prove its expected functionality after integration; the other circuit block are considered as ideal components. For specific applications, the possibility of sensing device co-integration with the analog interface is investigated, thus resulting in a small-sized system in package.

1.4 Scientific challenges and innovations

Since the present sensing systems design activity starts from the first level of the technology readiness scale (TRL1), all the possibilities of scientific challenges and innovations are open. In the context of this thesis, we consider that the sensing elements we have chosen to develop for the liquid and magnetic sensing

applications are based on existing devices of which we propose to extend the functionalities. The major innovation of our research work lies in the manner by which we interface the devices and how we extract the useful information from the system output signal. Therefore, the novelties proposed in this thesis are more focused on the manner used to analyze the sensing data than the fabrication or improvement of interdigitated electrodes or inductive devices.

1.4.1 Why sensor interfaces ?

In electronic systems, the choice, design, and implementation of a sensor interfacing principle are of major importance in order to transfer the sensed information in a representative and easily processable output signal. Since the trend of the high technology era is to reduce the complexity of instrumentation systems while improving their functionalities, an efficient and well thought out sensor interface is a powerful asset in terms of portability and power consumption [5]. In addition, the conquest of the electronic system market requires designing smart, attractive, and practical products. As an example, the illustration depicted in Fig. 1.3 shows an instrumentation system inserted in an RF application. This example of instrumentation system consists in an active RFID tag; the whole sensor interface includes all the baseband circuitry. Those systems are designed to be compact ($< 10 \text{ cm}^3$) and power efficient ($<< 1 \text{ W}$). Considering such systems, the main research activity is focused on the analog interface design and particularly on the mathematical analysis of its analog output signal.

1.4.2 Oscillating circuit interfaces

From a global point of view, a multitude of circuit architectures can be used to capture the information provided by a sensing element. As examples, we note the use of Wheatstone bridges [6, 7], potentiostats for studying chemical phenomena [8], oscillators, or any other signal conveyor. The research direction we have investigated to interface the sensors is the use of self-oscillating circuits. From the electronic point of view, an oscillator is a circuit that provides a periodic signal when supplied by a DC power source. The use of those circuits to interface sensors is of major interest for two reasons. First, the time-varying signals provided by the oscillator can be used to stimulate capacitive or inductive components. Secondly, an oscillator can be used as a transducer that converts a given physical data into a frequency or a period; these quantities can be estimated efficiently by digital back-end units.

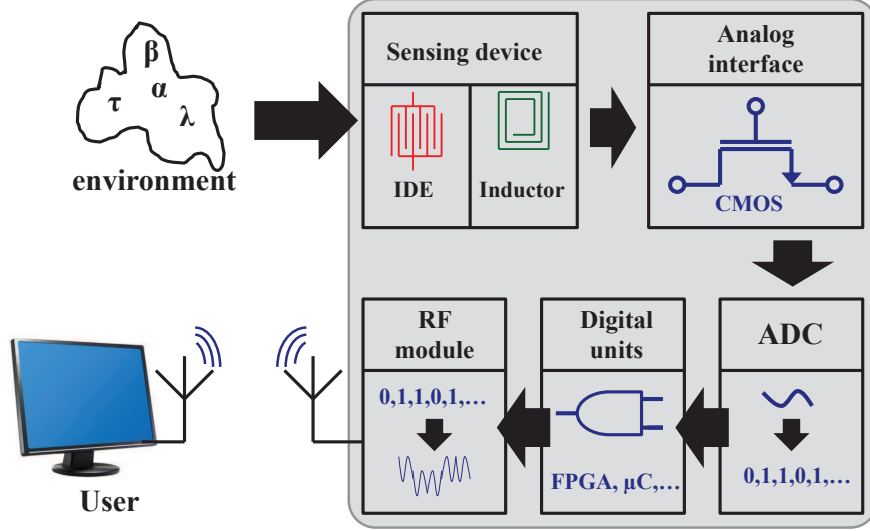


Figure 1.3: Sensing system smart prototype integrated on a RFID tag circuit (gray frame).

Among all the existing oscillators, two kinds are listed: the resonant or harmonic oscillators (i.e. Colpitts [9], Hartley [10], etc) and the relaxation oscillators (i.e. ring oscillators [11], delay line oscillators [12]). Being represented in Fig. 1.4, the proposed interfacing system is based on current-controlled oscillators (CCO) [13]. These consist in loading a current into a capacitor; resulting in a triangular voltage across the capacitor. The oscillation is assured by a Schmitt trigger that resets the circuit when the output voltage reaches a given comparison threshold. The oscillating voltage equation form is given by (1.1); I_0 and C being respectively the controlling current and the load capacitance; the voltages v_- and v_+ are the two levels that mark out the signal swing range.

$$v(t) = \frac{I_0}{C}t \quad v(t) \in [v_-, v_+] \quad (1.1)$$

The illustration depicted in Fig. 1.5 gives an overview of the oscillating triangular wave. Thanks to the possibility of adjusting the rising and falling current amounts in the oscillator, the rising and falling periods shown in this example have not the same value.

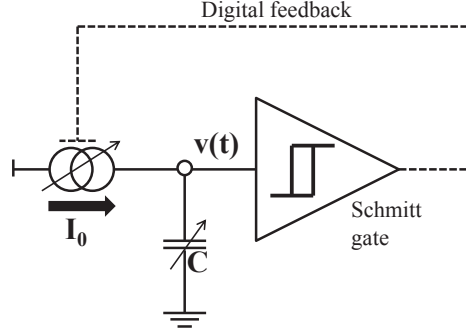


Figure 1.4: System overview of a current-controlled triangular voltage oscillator.

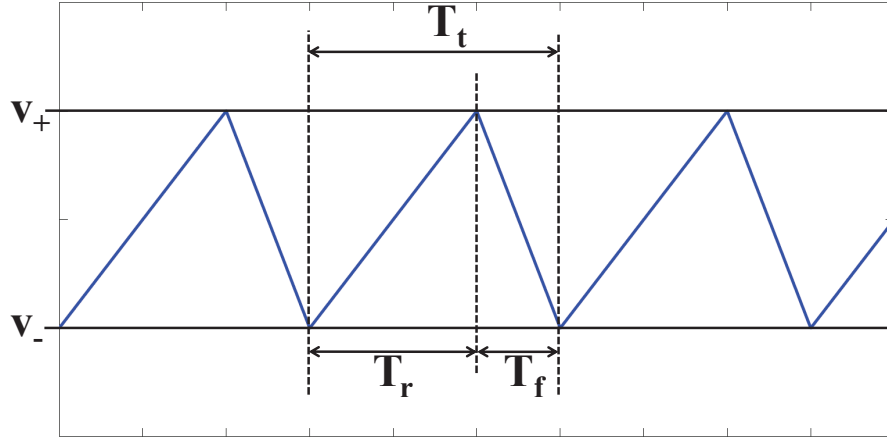


Figure 1.5: Triangular oscillating wave overview.

$$v(t) = \begin{cases} \frac{I_{0,r}}{C} t + v_- & \text{if } t \in [0, T_r] \\ -\frac{I_{0,f}}{C} (t - T_r) + v_+ & \text{if } t \in [T_r, T_t] \\ \frac{I_{0,r}}{C} [t - T_t] + v_- & \text{if } t \in [T_t, 2T_r + T_f] \\ -\frac{I_{0,f}}{C} [t - (2T_r + T_f)] + v_+ & \text{if } t \in [2T_r + T_f, 2T_t] \\ \frac{I_{0,r}}{C} [t - nT_t] + v_- & \text{if } t \in [nT_t, (n+1)T_r + nT_f] \\ -\frac{I_{0,f}}{C} (t - (n+1)T_r - nT_f) + v_+ & \text{if } t \in [(n+1)T_r + nT_f, (n+1)T_t] \end{cases} \quad (1.2)$$

The main idea presented in this thesis is to prove that the modulation of a triangular signal is an efficient method to transduce the information provided

by a sensing element. The efficiency of the method means that the resulting system is implemented by the most simple and low-power architecture.

1.4.3 Liquid sensing system innovations

Focusing on the liquid sensing system, the interfacing methodology is the stimulation of the interdigitated structure by a time-varying square auto-regulated current. Considering equation (1.1), we replace the load capacitance C by the interdigitated sensor equivalent impedance. The challenges faced in this context consist in studying and modeling the sensor triangular response in order to extract in a single output signal both the conductivity and the permittivity of liquids. The auto-regulation of the system is a major advantage for the portability of electronic solutions since a single battery is required to supply the circuit. The proposed solution considerably simplifies implementation and requires only a few electrical components to be fabricated.

1.4.4 Magnetic sensing system innovations

The interfacing principle of the inductive sensor is also based on the generation of a triangular signal. The difference lies in the fact that we use the factor I_0 of equation (1.1) to transduce the sensed information. Indeed, we explain in the magnetic system chapter that the inductive signal is transformed into a current that loads the capacitance of the triangular voltage oscillator. This interfacing principle allows for defining output characteristics directly representative of the sensed magnetic variations.

1.4.5 Circuit integration innovations

The integration of a circuit interface requires designing all the blocks composing the system on an analog CMOS technology. Since the triangular oscillator requires a Schmitt trigger that controls the voltage oscillation, this block is improved in the CMOS integration step. The challenge we present is to design a Trigger of programmable hysteresis. The architecture of the system is chosen in order to release the sizing constraints of the transistors, thus improving the robustness of the circuit.

1.5 Thesis contents

This thesis manuscript is composed of two chapters describing respectively the liquid sensing system (chapter 2) and the magnetic variations detection system (chapter 3). Those chapters are first introduced in order to describe

the sensing context and are next divided into sections that follow the steps of the proposed instrumentation system design methodology. The following description gives a brief overview of the thesis content. This is given in more details in the introduction sections of the chapters.

In chapter 2, the liquid sensing context is motivated by the large field of applications for those the proposed system is of interest. A review of the interdigitated sensor usage for the fluid detection is next presented. The detection principle of interdigitated structure takes into consideration the possibility of passivating the electrode fingers; a generic equivalent impedance model is obtained from the analysis of the interdigitated sensor structure. All the electrical elements composing the model are computed by a finite element approach. The interfacing methodology, based on the generation of an oscillating triangular voltage, consists in charging and discharging the sensing device with the controlling current of the oscillator. The response of the system is next analyzed in details by a mathematical demonstration. The implementation of the sensor and its interface is then performed using *off-the-shelf* components and the sensing system theoretical model is validated by measurements in electrolyte solutions and liquid mixtures. The proposed integration of the sensor interface is finally presented and analyzed in terms of improved performances.

Chapter 3's structure is presented similarly to the previous one. The introduction of the chapter points at a well-defined application: the arc fault detection (AFD) which consists in the detection of the sharp current variations occurring on high-powered conductors. The research review exposes next the principal magnetic detection principles and the most relevant and innovative solutions; this allows for justifying the use of inductors as current variation sensors. The detection principle provides the inductor induced voltage as a function of the magnetic flux caught by the device; the electrical equivalent model of the sensor is consequently established and each element of the model is computed by finite element tools. The sensor is next inserted in a triangular saw-tooth current-controlled oscillator; the controlling current of the oscillator is modulated by the magnetic flux perturbations caught by the sensing device. Output characteristics are therefore defined according to the measured period of the saw-tooth wave. The system is then implemented by lumped elements and tested on a testbench of electrical arcs generation. Finally, the circuit interface is simulated and integrated.

CHAPTER 2

Fluid impedance sensing system

2.1 Introduction

The quantification and qualification of the conductivity and permittivity of fluids is of major interest for the fluid identification in several domains like medicine, biology or chemistry. These parameters can give access to many properties like ionic concentration [14] or pH [15], metallic and non-metallic particles contamination [16, 17], DNA detection [18], or bacterial presence in biological fluids [19]. Measuring the conductivity and permittivity of any liquid with accuracy is therefore a complex challenge. Instrumentation solutions are thus constantly developed.

We present here the modeling and the design of an R-C miniaturized sensing system (i.e. the sensing elements and its circuit interface) and its application for liquids conductivity and permittivity measurements. The sensing element is a solid-state interdigitated electrodes (IDE) structure deposited on an insulating substrate. The proposed sensor interface implements the principle of the triangular wave voltage (TWV) generation to stimulate the IDE device. All the information we seek is included in the shape of the voltage response of the system. As presented in the introduction chapter of the thesis, one of our objectives is to apply a scientific methodology of instrumentation system development which consists in pushing a fundamental instrumentation concept up to the end of the research activity; meaning that the system is ready to be tested in a relevant environment.

The chapter is divided into sections which follow the steps of the scientific development approach. In this context, the detection principle gives the electrical representation of the sensing device in order to highlight the parameters that have to be measured: the liquid resistance and capacitance. The physical modeling step explains the fundamental laws that govern the sensing device operation; a finite element modeling (FEM) tool is used to extract the electrical

impedance of an IDE structure assuming the liquid as a parallel R-C dipole of finite conductivity σ_L and permittivity ε_L . The sizing approach of the sensor is then described according to a key parameter of IDE structures: the geometrical cell constant.

The interfacing principle is developed by a mathematical study of the modeled impedance stimulated by a square current wave. This leads to a formal expression of the system output signal as a function of the sensed $R-C$ parameters. This technique has been used successfully to sense constant phase element impedances. Nevertheless, the extraction of the sensed data required to count all the elapsed periods of the triangular signal after its generation [20]. The present work uses this technique for measuring the conductivity and permittivity of liquids modeled as $R-C$ dipole. The main amelioration we brought for our system consists in isolating each voltage segment into a compact equation independent from the previous states of the oscillating wave. Therefore, the signal analysis is highly simplified and only requires sampling the data of a signal period to get all the expected parameters of a tested liquid.

The next step explains how the triangular wave sensing technique is implemented using a current-controlled oscillator (CCO) circuit interface. This interface and the IDE sensor are next manufactured on PCB using off-the-shelf components. Measurements are performed and presented for two specific applications: the characterization of saline concentrations of NaCl and KCl and the quantification of water-glycerol mixtures. The CMOS integration of the circuit interface is investigated in the last part of this chapter; the aim of the CMOS design is to drastically improve the interfacing performances of the circuit.

2.2 Context and state of the art

The measurement of the conductivity and permittivity requires the injection of an electrical stimulus in the sensed liquid which impedance constitutes the electrical response. The analysis starts by giving the general concept of the liquid measurement cell and defines the general mathematical links between the sensed impedance parameters and the global conductivity and permittivity of any liquid. Focusing on the interdigitated configuration of a sensing cell, we next look into the research solutions which used this configuration to perform liquid characterization. Since a fluid is a material in which all the diluted species can move freely, the existing parasitic effects of charge motion are explained. Solutions are next given in order to neglect these effects in the models developed in the next sections of this chapter.

2.2.1 Concept of the cell constant

In every liquid sensing system based on electrical stimulation, the detection of the conductivity and the permittivity ($\sigma_L - \varepsilon_L$) needs the injection of an external input signal in the tested liquid applied on two electrodes. The response of the stimulation illustrated in Fig. 2.1 consists in a complex impedance Z_L in which a resistive and a capacitive component ($R_L - C_L$) can be identified.

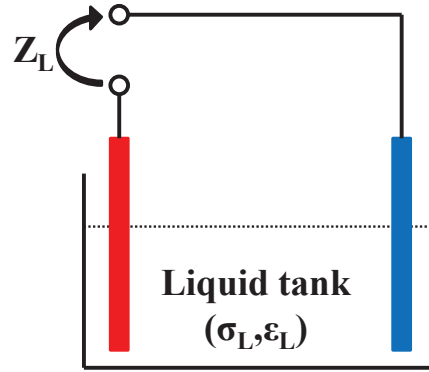


Figure 2.1: Liquid impedance sensing cell representation.

The respective links between the pairs $\sigma_L - R_L$ and $\varepsilon_L - C_L$ are given by a transformation factor known as the sensor cell constant; this key parameter is only fixed by the geometrical configuration of the sensing electrodes. The cell

constant, denoted as c_k and expressed in $[\text{m}^{-1}]$, is defined by equation (2.1); the parameters σ_L and ε_L respectively represent the conductivity and the permittivity of any tested liquid in which the sensing element is immersed, R_L and C_L are the sensed resistance and capacitance given by the liquid external stimulation. In other terms, the cell constant represents the ratio between the distance that separates the two sensing electrodes and their effective area. The concept of the sensor cell constant is commonly used in impedimetric applications [21]; the design of gas concentration sensors also use this concept [22]. The cell constant also gives the link between the sensed liquid impedance Z_L and the $\sigma_L - \varepsilon_L$ parameters (2.2).

$$c_k = \sigma_L R_L = \frac{\varepsilon_L}{C_L} \quad (2.1)$$

$$Z_L = R_L // C_L = \frac{R_L}{j\omega R_L C_L + 1} = \frac{c_k}{j\omega \varepsilon_L + \sigma_L} \quad (2.2)$$

2.2.2 Interdigitated electrode structures

The choice of the sensing device geometry sets the cell constant value and therefore a detection range for the conductivity and the permittivity. The conductivity of any liquid can vary on a wide range of values; the deionized water presents a conductivity of $5.5 \mu\text{S.m}^{-1}$ while the mercury has a value of more than 1 MS.m^{-1} ! The situation is totally different for the permittivity. The lowest possible value is 1 for the vacuum while one of the greatest is 80 for water; the rare materials presenting higher value of permittivity are compounds of titanium [23].

The low variation range of permittivities is the major constraint for the design of the sensing device. According to the definition of the cell constant, we consider as an example two sensing electrodes of 1 cm^2 of effective area separated by a distance of 1 cm; the electrodes are then immersed in pure water; the resulting capacitance between the electrodes is less than 10 fF. This value of capacitance is very small and needs the design of very costly techniques to be detected. The only way to improve the value of the capacitance is to decrease considerably the distance between the electrodes and to increase their effective area: the interdigitated electrodes (IDE) structure (Fig. 2.2) is the most suitable geometry to improve the capacitance values [28].

IDE sensors can be manufactured using different kinds of metals and insulating materials. We denote as examples the fabrication of aluminum fingers deposited on oxidized silicon bulk wafer or glass wafer [24], platinum [26], gold [28], or polysilicon fingers IDE sensors [27] and many others.

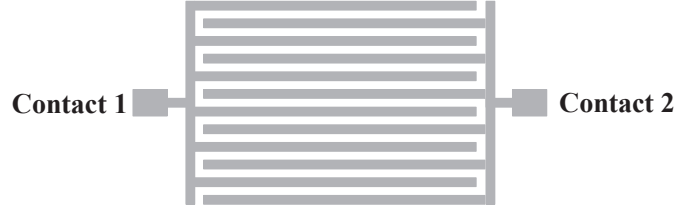


Figure 2.2: Interdigitated configuration of the sensing electrodes (top view).

2.2.3 Liquid-solid interface effects

Because of the electrical nature of the conductivity and permittivity, the literature shows many research activities on the liquid impedance extraction modeled as a R-C parallel dipole [29]. The R-C representation of a given liquid has unfortunately its limits of validity that have to be discussed before going further in the physical modeling of the sensor.

2.2.3.1 Faradaic current

A liquid is a material containing moving diluted ion species. Supposing the sensing electrodes biased by a given electrical potential, those ionic species are attracted and move towards the electrode of opposite electrostatic charge. According to the oxido-reduction chemical reaction, a charge transfer occurs at the solid-liquid interface; this charge exchange is well known as the faradaic current. Depending on the metallic nature of the electrodes, the phenomenon of corrosion is more or less important. The noble-metals electrodes like platinum, gold or silver are highly cathodic; it means that their tendency of being ionized is very low in comparison with other materials. This tendency to corrosion is quantified by the standard electrode potential E° . We give some examples of potentials at 25°C in Tab. 2.1.

The design of a liquid sensing cell using noble metals considerably limits the apparition of the faradaic current and consequently the metal corrosion. Research works which use sensing cells fabricated with classic metals must take into account the faradaic current in their electrical modeling [27]. This phenomenon is often represented by an abstract circuit element called the Constant Phase Element (CPE). This element models the faradaic current and also takes into account the quality of the electrode surface (i.e. roughness or porosity). Another example described in [30] uses an IDE sensor especially for

Metal type	Oxidation	Potential E° [V]
Gold	$Au \rightarrow Au^{3+} + 3e^-$	+1.498
Platinum	$Pt \rightarrow Pt^{2+} + 2e^-$	+1.200
Silver	$Ag \rightarrow Ag^+ + e^-$	+0.799
Copper	$Cu \rightarrow Cu^{2+} + 2e^-$	+0.337
Hydrogen	$H_2 \rightarrow 2H^+ + 2e^-$	0.000 (ref.)
Zinc	$Zn \rightarrow Zn^{2+} + 2e^-$	-0.763
Aluminum	$Al \rightarrow Al^{3+} + 3e^-$	-1.662

Table 2.1: Metallic electrodes standard potential E° at 25°C.

monitoring the redox reactions occurring on the electrode-liquid interface. This parasitic effect can be avoided by the use of noble metal electrodes or by the passivation of the electrodes explained further in the section.

2.2.3.2 Electrical double layer

The other parasitical phenomenon occurring on the solid-liquid interface is the formation of the Electrical Double Layer (EDL). The first part of the EDL is formed by ions adsorbed on the solid surface while the second part of the EDL layer, called the diffuse layer, is formed of diluted species attracted to the surface charge according to the Coulomb force. Unlike the ions of the first part of the EDL, the ions composing the second part of the EDL are not firmly anchored and present a space charge gradient decreasing exponentially towards the liquid bulk (Fig. 2.3); this part of the EDL is therefore called the diffuse layer. The impact of the EDL is also dependent on the tendency of the electrode materials to attract the diluted ions. As observed for the faradaic current phenomenon, the noble metals are less impacted by the double layer formation.

In most of electrochemical cells stimulated by a time-varying signal, the EDL electrical model is given by a R-C dipole [29] placed in series with the liquid bulk R-C model. The effect of this layer can be observed according to the input stimulus frequency. The typical EDL thicknesses observed rarely exceeds a few dozens of nanometers [31]; the EDL capacitance is therefore very high and its influence is quickly lowered when the sensing cells dimensions exceeds a few microns.

2.2.4 Electrode passivation

Depending on a given sensing application, the IDE sensor can be passivated. The passivation cancels the faradaic current effect and therefore prevents the

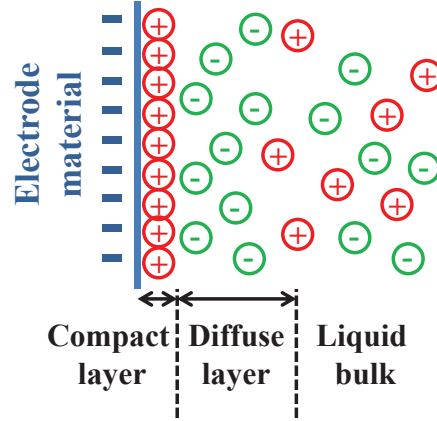


Figure 2.3: Ionic repartition of the electrical double layer (EDL); the electrode is charged negatively for this example.

electrode corrosion. This process involves the deposition of a thin insulating layer that separates the liquid from the electrode metal (Fig. 2.4). Several chemical processes can be used to perform the passivation deposition. The anodization process [32] is the insulating layer formation by the oxidation of the electrodes into a chemical solution by current injection on the metal surface. The atomic layer deposition process is another way to passivate the interdigitated electrodes with a very accurate passivation thickness control [33]. The consequence of the metal passivation is to cut the DC path between the electrodes formed by the liquid finite conductivity.

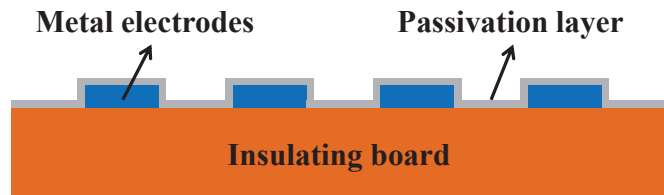


Figure 2.4: Representation of the passivation layer covering the metal electrodes.

2.2.5 Measurement techniques

The research works cited in the previous section were principally limited to the design and the fabrication of IDE sensors; the measurement techniques men-

tioned in these works involve the use of laboratory equipment like LCR meters, lock-in amplifiers, or other bulky benchtops. As explained in the first chapter, one of the scientific goals of this thesis is the design of a sensor interfacing principle in order to fabricate a portable and power-efficient instrumentation solution.

The interfacing techniques of IDE sensors have been investigated in the context of a breathing detection application [24]. The concept is to detect the capacitance variations due to the presence of expurgated water on the sensing device area.

As explained in [24], an IDE interface consisting in a Schmitt Trigger oscillator has been implemented. The schematic diagram showed in Fig. 2.5 illustrates the conversion principle of the sensed capacitance into a frequency modulated signal. The action of the Schmitt Trigger is to maintain the oscillation between its two threshold voltages, the output signal frequency is thus proportional to the time-constant provided by the R-C pair of the system.

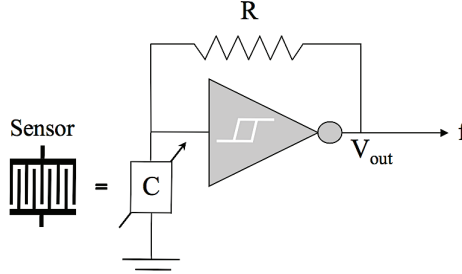


Figure 2.5: Schmitt Trigger based oscillating interface [24].

Another principle of IDE interface has also been tested directly on an integrated circuit technology [34]. As depicted by Fig. 2.6, an odd number of inverters are connected in cascade in order to enforce the system to oscillate, each inverter output is connected to an IDE sensor. The output signal results in a square wave which frequency is proportional to the sensed capacitance. These two interfacing principles showed excellent results for the capacitive component detection of the IDE sensor; it also showed the efficiency of their oscillating principles. The main weakness of these approaches is that the resistive part of the IDE sensors cannot be discriminated from the IDE sensor capacitive variations.

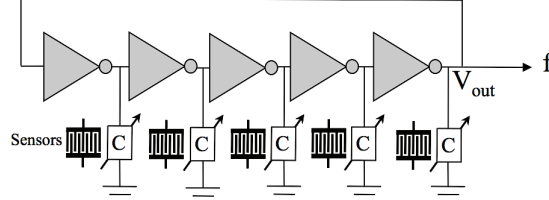


Figure 2.6: Ring oscillator IDE interface [34].

2.2.6 Work position

As shown throughout this review, the possibilities of the use of IDE structures for sensing applications are very large. Our goal is to present a system which uses this kind of structure to quantify the liquid conductivity and permittivity using an original and effective interfacing approach. We revisit the model of the cell constant calculation of such a structure by finite element modeling (FEM) approach. This allows for obtaining better impedance estimations than the previous analytical models [36].

The major innovation we propose in this chapter is the stimulation approach of the sensor used to extract the liquid data in a single signal. The technique is based on the Triangular Wave Voltage (TWV) generation and has been tested in [35]. The sensor is stimulated by a triangular voltage and the information is provided by the measured current flowing through the sensing cell. The fundamental difference with the technique we present in this chapter is that we reverse the principle; the sensor is stimulated by a square current in order to generate the triangular output voltage. The advantage of our approach is its considerable ease of circuit implementation.

In order to summarize the most interesting research works on IDE sensors, a classification is reported in Tab. 2.2; this classification allows us for giving our work position. Among all these works, the majority have developed interfacing principles mainly based on a unique capacitive detection. One of the most original interfacing principle is given by [39]; the sensor interface extracts both the resistance and capacitance components of an IDE sensor. The shortcomings are the complexity of the sensor interface (digital units are required) and the absence of any output signal modeling. The global position of our work is that we propose a more simple and auto-regulated circuit interface for the evaluation of IDE sensors R-C components. We also provide a method to optimize the sensor dimensions for a given fluid application and finally derive a detailed

expression form of the analog output voltage of our circuit interface.

Ref.	Applications	Interface	Models	Sensed elem.
[45]	A CMOS humidity sensor with on-chip calibration.	Switched capacitor CMOS interface.	no	C
[24]	Capacitive detection of the air humidity.	Delay-time oscillator.	no	C
[34]	Capacitive detection of the human breathing.	Integrated ring oscillator.	no	C
[43]	Capacitive DNA measurements.	Analog to digital capacitive CMOS interface.	no	C
[42]	Detection of the dielectric constants of chemical solutions.	CMOS interface for charge-based capacitive measurements.	Model of the output response for non-conductive fluids.	C
[39]	R-C detection of DNA hybridization.	Integrated current-voltage converter, level detector, and ADC.	no.	R-C
This work	R-C measurements of fluid properties.	Integrated current-controlled oscillator.	IDE sensor electrical and FEM models. Interfacing transient response model. Sizing methodology.	R-C

Table 2.2: IDE research works table summary and present work position.

2.3 Interdigitated sensor detection principle

As revealed in the previous section, the interdigitated electrodes (IDE) structures are easy to fabricate using insulating substrates on which the metal patterns are deposited and can be passivated by a thin layer of material. The detection principle explained in the present section assumes each insulating material as perfectly dielectric, homogeneous, dense and therefore purely capacitive. The tested liquid is considered as an imperfect dielectric of finite permittivity and conductivity; resulting in other terms in an R-C parallel dipole. The IDE device is thus represented by an electrical model of which the equivalent impedance is represented in the frequency domain. Throughout this section, we consider the use of techniques and materials that allow us to neglect the solid-liquid interface effects (i.e. the faradaic current and the electrical double layer formation).

2.3.1 IDE structure description

The structure global view is illustrated in Fig. 2.7; it depicts on the left the device top and cross views and the periodic cell structure cross-section on the right. The periodic cell is geometrically composed of two halves of a finger width and the space that separates two fingers.

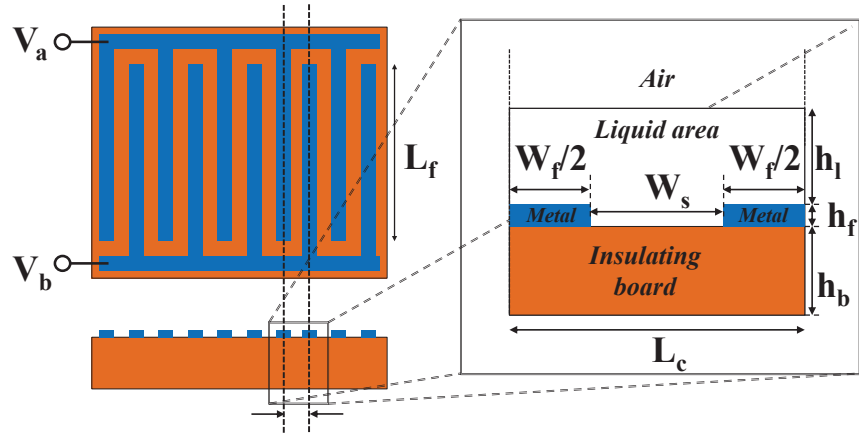


Figure 2.7: Global view and cross-section of the interdigitated electrodes (IDE) structure. The periodic cell is described and annotated on the right.

The representation of the IDE passivation is shown in Fig. 2.8 by the gray

layer that covers the IDE fingers. The electrical model developed here considers the passivation material as homogeneous and compact. The description and the symbols used for the geometrical parameters are reported in Tab. 2.3.

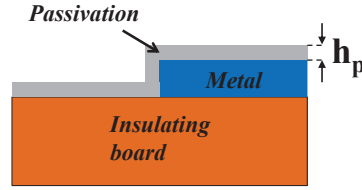


Figure 2.8: Metal passivation thickness representation.

Name	Symbol	Units
Board thickness	h_b	[m]
Electrodes height	h_f	[m]
Electrodes width	W_f	[m]
Electrodes spacing	W_s	[m]
Cell width	L_c	[m]
Electrodes length	L_f	[m]
Electrodes number	N_f	[–]
Passivation thickness	h_p	[m]

Table 2.3: Parameters description.

2.3.2 Equivalent electrical model

The general form of the IDE electrical model is extracted from the cross view of the periodic cell (Fig. 2.9); three accesses are considered, the two electrode contacts and the backside or substrate contact. This representation assumes that the fingers are passivated; a capacitance C_P is thus added in series with the tested liquid dipole $R_L - C_L$. The substrate capacitances are given by the lumped elements C_B et C_S to complete the electrical path between the device fingers.

In theory, the backside contact is not necessary to assure the good working of the device; the bias of the substrate is nevertheless recommended for oxide-on-silicon substrate structures [24, 34]. The variations of the signal applied to the electrodes can indeed generate charge displacements in the semiconductor

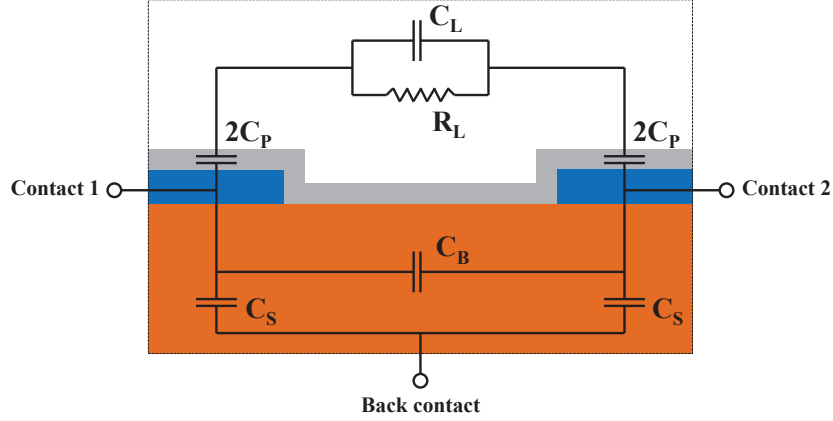


Figure 2.9: Electrical equivalent model of the IDE structure.

close to the oxide-semiconductor interface. Enforcing a well-defined potential on the substrate allows for creating a virtual ground on this material interface and therefore to annihilate this charge modulation. In practice, the bias of the backside allows for creating a ground plane that reduces the noise undesirable effects on the sensor response. Therefore, we always consider that the device backside is grounded for the description of the sensing principle. The modeling and the interfacing principle of the device exposed further respectively consider the two interdigitated electrodes as connected on a stimulation potential and the ground. This means that two of the three accesses illustrated on Fig. 2.9 are merged into one terminal: the ground node (Fig. 2.10). In order to simplify the mathematical expressions of the sensor impedance, all the substrate capacitances, i.e. C_B and C_S , are included into a single capacitance denoted C_B .

The expression of the equivalent impedance seen between the two electrodes is therefore given by equation (2.3) in the Laplace domain and is composed of a zero-pole pair. The expressions (2.4), (2.5) and (2.6) represent respectively the zero frequency ω_z , the pole frequency ω_p and the initial amplitude coefficient C_0 .

$$Z_{IDE} = \left[\frac{1}{sC_0} \right] \left[\frac{s + \omega_z}{s + \omega_p} \right] \quad (2.3)$$

$$\omega_z = \frac{1}{R_L (C_P + C_L)} \quad (2.4)$$

$$\omega_p = \frac{C_B + C_P}{R_L (C_B C_P + C_L C_P + C_L C_B)} \quad (2.5)$$

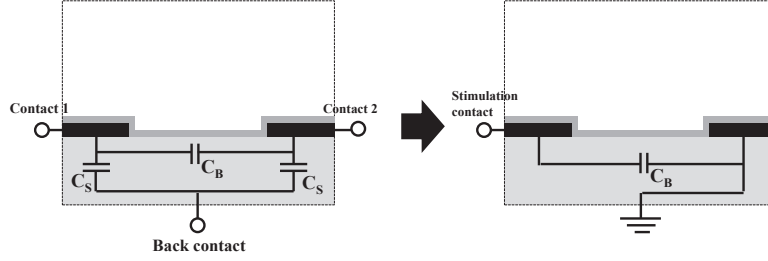


Figure 2.10: IDE structure connection configuration.

$$C_0 = \frac{C_B C_P + C_L C_P + C_L C_B}{C_P + C_L} \quad (2.6)$$

It is of interest to analyze in detail the positions of the zero and the pole. Indeed, we observe that the pole frequency position is always higher than the zero one; the differential frequency $\Delta\omega_{pz}$ given by the equation (2.7) is positive for this electrical model. This theoretical result means that the impedance resistive component is predominant in comparison with the capacitive one between the zero and the pole according to the presence of a liquid of finite conductivity. The bode asymptotic diagram of the IDE impedance illustrates this last observation by a flat area as depicted by Fig. 2.11; where $|Z_{\omega=1}|$ is expressed by (2.8). The value of $|Z_R|$ is the amplitude of the resistive flat area given by equation (2.9).

$$\Delta\omega_{pz} = \omega_p - \omega_z = \frac{C_P^2}{R_L (C_P + C_L) (C_B C_P + C_L C_P)} \quad (2.7)$$

$$|Z_{\omega=1}| = \frac{\omega_z}{C_0 \omega_p} \quad (2.8)$$

$$|Z_R| = \frac{1}{C_0 \omega_p} \quad (2.9)$$

2.3.3 Unpassivated IDE impedance

The general expression (2.3) of the IDE impedance is the starting point at which the interfacing methodology is based. This method must also take into account the absence of the passivation layer that covers the metal fingers. From a mathematical point of view, the absence of the passivation layer can consider the capacitance C_P as infinite (i.e. short-circuit) and therefore causes the DC

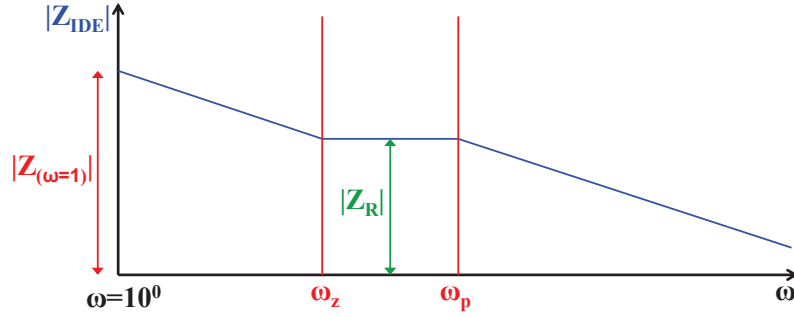


Figure 2.11: Asymptotic Bode diagram of the IDE sensor impedance magnitude.

path between the electrodes to be open according to the liquid resistance. However, the interfacing method described further requires a finite capacitance that cuts the DC path between the electrodes. This serial capacitance, noted as C_A , is therefore added externally for unpassivated devices as shown by Fig. 2.12. The expressions of the impedance parameters ω_z , ω_p et C_0 are then rewritten in equations (2.10), (2.11) and (2.12).

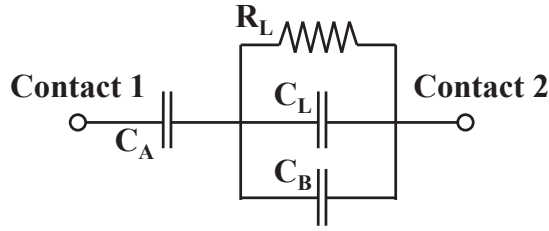


Figure 2.12: Electrical representation of the unpassivated IDE electrical model with the serial capacitance C_A added externally.

$$\omega_z = \frac{1}{R_L (C_A + C_L + C_B)} \quad (2.10)$$

$$\omega_p = \frac{1}{R_L (C_L + C_B)} \quad (2.11)$$

$$C_0 = \frac{C_A(C_L + C_B)}{C_A + C_L + C_B} \quad (2.12)$$

2.4 Interdigitated structure physical model

The detection principle proposed in the previous section allows for establishing the mathematical form of the IDE sensor frequency response in terms of lumped component parameters. The goal of the physical modeling section is to quantify these parameters and therefore to validate this equivalent model using finite element modeling (FEM) tools. The use of numerical tools is principally motivated by the high accuracy and ability of computing complex geometrical structures. The software **COMSOL Multiphysics** includes an AC-DC simulation module allowing for solving the electric potential profile of a biased device in the frequency domain and also offers the possibility of extracting the electrical impedance seen from the device ports.

2.4.1 IDE solving domain and boundary conditions

The modeling equation used to solve the harmonic potential profile is based on the theory of electric currents explained in [25]. Because of the periodical aspect of IDE structures, it is not necessary to solve the Laplace potential equation (2.13) on the whole domain of a given device, only a cell periodic element can be used to define the solving domain of the FEM simulator (Fig. 2.13). All the materials composing the IDE structure are configurable in terms of electrical conductivity and permittivity. The FEM system is next biased on the three ports (i.e. the two electrodes and the backside). One of the two electrodes and the backgate are tied to ground and the other electrode is biased by a sine voltage; the other boundaries are electrically insulated. Since the electric field influence does not spread infinitely above the sensing plane, the solving domain height is limited to twice the cell width in the fluid sensing area; 99% of the electric energy is included below this limit [26]. The metal areas have been excluded from the solving domain; they are indeed considered as perfect conductors; the electrical potential is constant in these areas.

$$-\nabla \cdot (\sigma + j\omega\epsilon_0\epsilon_r)\nabla V(\omega) = 0 \quad (2.13)$$

As depicted on Fig. 2.13, the boundary condition $c1$ is the stimulation node; its equation is given by (2.14); where V_{max} and ω are respectively the amplitude and frequency of the stimulating signal.

$$V(\omega) = V_{max} \cos(\omega t) \quad (2.14)$$

The contacts $c2$ and $c4$ are ground contacts (2.15).

$$V(\omega) = 0 \quad (2.15)$$

The electric insulation condition $c3$ applied on all the left boundaries is mathematically described by equation (2.16). Thanks to the system periodicity, this condition can be enforced on the vertical edges of the structure cell. Indeed, the electric field profile is parallel on those edges.

$$\nabla V(\omega) \cdot \mathbf{n} = 0 \quad (2.16)$$

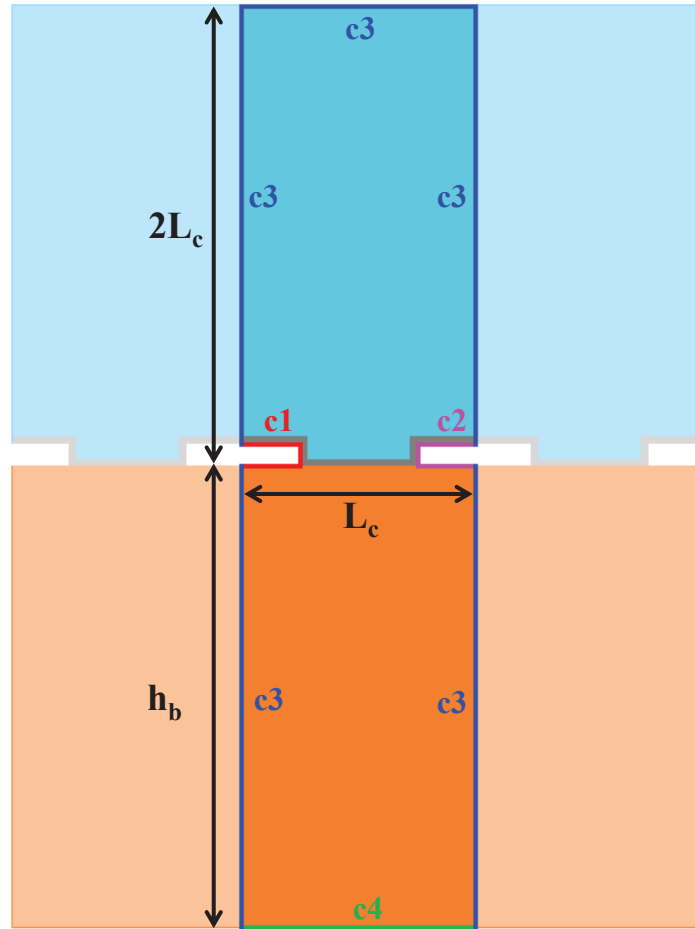


Figure 2.13: Sensing cell periodic domain and applied conditions on the system; $c1$ is the stimulus contact, $c2$ and $c4$ are ground contacts and $c3$ represents the electric insulation. The electrodes are considered as perfect electric conductors and are excluded out of the solving domain.

2.4.2 Global impedance simulations

The present simulations have been performed in order to observe the influence of the sensor parameters on the output impedance. These impedance results are extracted from the global evaluation of the complex current flowing through the IDE structure cell after the computation of equation (2.13). Therefore, the impedance results obtained by this global evaluation of the current are the most realistic in terms of electrical phenomena modeling. The computation of the electrical model components of Fig. 2.9 are based on hypotheses that involve errors that will be discussed further.

The simulated parameters are reported in Tab. 2.4. The choice of the parameter values corresponds to the integrated interdigitated sensor used in the context of a breathing application developed in [24] and [34]; the structure is made of aluminum electrodes deposited on an oxidized silicon wafer and the passivation thickness is obtained from an anodization process. This choice of materials is arbitrary and can be adapted for any kinds of materials.

Name	Symbol	Value	Units
Electrodes height	h_f	0.5	$[\mu\text{m}]$
Electrodes width	W_f	2	$[\mu\text{m}]$
Electrodes spacing	W_s	2	$[\mu\text{m}]$
Electrode fingers length	L_f	200	$[\mu\text{m}]$
Electrode fingers number	N_f	100	$[-]$
Sensor surface	A_s	$N_f L_f (W_f + W_s)$	$[\text{mm}^2]$
Passivation thickness	h_p	20	$[\text{nm}]$
Substrate thickness	h_b	8	$[\mu\text{m}]$
Liquid conductivity	σ_L	10^{-4}	$[\text{S.m}^{-1}]$
Liquid relative permittivity	ε_L	80 (water)	$[-]$
Passivation relative permittivity	ε_P	8.9 (aluminum oxide)	$[-]$
Substrate relative permittivity	ε_B	3.9 (silicon oxide)	$[-]$

Table 2.4: Device parameters default values.

The influence of the liquid conductivity is first observed at Fig. 2.14. As predicted by the equations (2.4) and (2.5), the zero and the pole frequency positions are both inversely proportional to the liquid resistance R_L resulting in the shift of the resistive flat area along the impedance modulus curve.

The second simulation shows the impact of the liquid permittivity on the

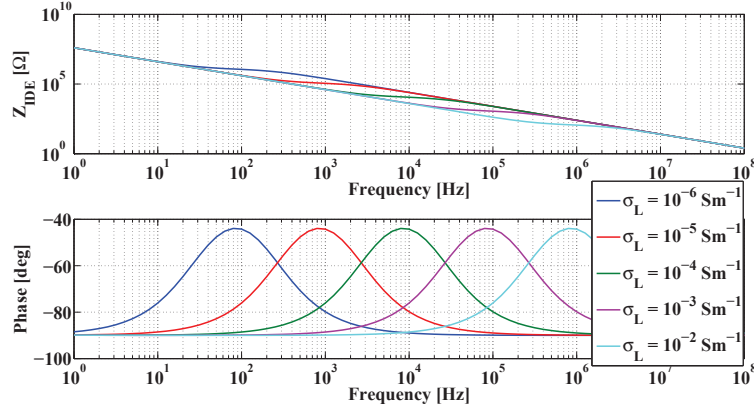


Figure 2.14: IDE sensor impedance for different values of the liquid conductivity σ_L ; all the other parameters are set to the default values of Tab. 2.4.

IDE impedance. Since the range of the permittivity of a liquid is much narrower than the conductivity, a realistic choice of simulated value has been set from 40 to 80 as depicted by Fig. 2.15. We note a displacement of the pole position and the invariance of the zero. The modeling equation (2.4) presents however a dependence of the zero position with the liquid capacitance. In practice, a weak dependence of the zero is observed for the following reason: the passivation capacitance is directly linked to the passivation layer thickness which has to be the smallest in order to increase the impact of the liquid presence on the sensitive area. The serial capacitance C_P is therefore high in comparison with the other capacitances C_L and C_B of the system; this makes the zero position of the IDE impedance present a negligible shift to the right with the increase of the liquid permittivity.

The passivation thickness has to be the smallest to improve the impact of the liquid on the device sensitive area. The simulation illustrated on Fig. 2.16 shows the evolution of the impedance spectrum for several test values of the passivation thickness. The thicker the passivation layer, the narrower the difference between the pole and zero frequency positions. The principal indicator of the liquid-structure interaction is the maximum value of the phase impedance; this parameter is an input of the sizing algorithm developed in the next section.

2.4.3 Cell constant calculation

The previous simulations described the evolution of the global impedance Z_{IDE} of an IDE sensor. As the goal of the sensing system is to deliver the conductivity σ_L and permittivity ε_L of any liquid, a mathematical link between

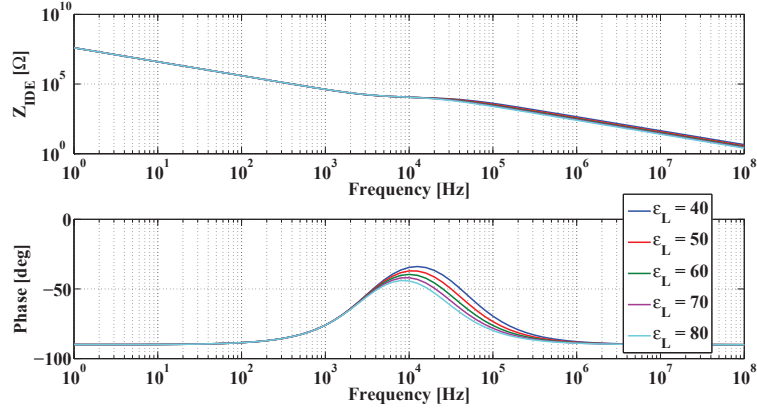


Figure 2.15: IDE sensor impedance for different values of the liquid permittivity ε_L ; all the other parameters are set to the default values of Tab. 2.4.

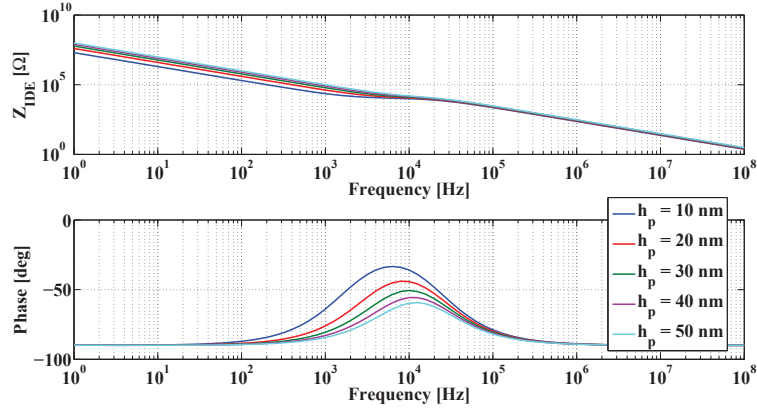


Figure 2.16: IDE sensor impedance for different values of the passivation thickness h_p ; all the other parameters are set to the default values of Tab. 2.4.

$R_L - \sigma_L$ and $C_L - \varepsilon_L$ must be established; this link is known as the sensor cell constant c_{kL} as defined in the review section by the general equation (2.1). The calculation of the cell constant therefore allows for evaluating the lumped elements of the electrical model depicted in Fig. 2.9.

The cell constant calculation is given on a single periodic element; the adaptation of its value for the whole IDE sensor is simply performed by its multiplication by the total electrode number N_f and the finger length L_f . This constant is also invariant with the scaling factor of the system; it means that

the multiplication of all the geometric dimensions of the IDE structure by a given factor does not change the cell constant value.

In the case of IDE sensors, the electrical field flowing between the sense contact to the ground one is divided in two parts: the fluid area and the substrate area. Two cell constants must therefore be defined; the liquid area cell constant c_{kL} and the substrate area cell constant c_{kB} . The equations (2.17) and (2.18) express the cell constants of the sensor, making thus the link between the material properties σ_L , ε_L and ε_B with the circuit lumped model components R_L , C_L and C_B . As shown in Fig. 2.17, the separation of the cell constant calculations in the liquid and substrate areas states that the electric field flowing between the electrodes is tangential to the substrate surface. This last assertion is not actually exact, this calculation technique involves small errors on the device impedance estimation; those errors are consequently discussed further.

$$c_{kL} = \sigma_L R_L = \frac{\varepsilon_L}{C_L} \quad (2.17)$$

$$c_{kB} = \frac{\varepsilon_B}{C_B} \quad (2.18)$$

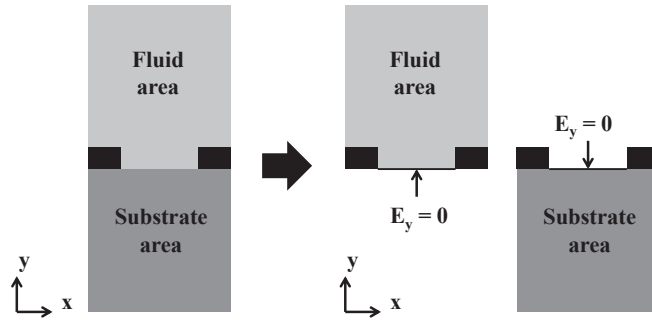


Figure 2.17: Separation of the fluid and substrate area; the electric field is supposed to be tangential to the fluid-substrate interface.

The calculation of IDE cell constants has been investigated in [26] and given the analytical formulas (2.19), (2.20) and (2.21). This last expression considers the electrical field profile as elliptic between the electrodes. The shortcomings of this approach is the disregard of the metal thickness and the limitation of the calculation domain to one half of the cell cross view; the elliptic profile of the electric field is therefore supposed as symmetric in both the liquid and substrate areas. As performed for the global IDE impedance, the cell constant

FEM calculation is computed on the whole cell cross view of the IDE system, giving thus more realistic results.

$$c_k = \frac{1}{(N_f - 1)L} \frac{2K(k)}{K[\sqrt{(1 - k^2)}]} \quad (2.19)$$

$$k = \cos \left(\frac{\pi}{2} \frac{W_f}{W_f + W_s} \right) \quad (2.20)$$

$$K(k) = \int_0^1 \frac{dt}{\sqrt{(1 - t^2)(1 - k^2 t^2)}} \quad (2.21)$$

The FEM approach consists in solving the stationary Laplace potential equation (2.22); the application of the Gauss law (2.23) allows for evaluating the charge Q_L and Q_B present at the electrode surface respectively in contact with the fluid and substrate areas (the indexes L and B denote respectively the liquid and substrate parts). Therefore, the Gauss surface used for the normal electric field integration has been splitted into two surface parts S_L and S_B on the fluid-substrate interface. The capacitances C_L and C_B are then evaluated by (2.24) and the cell constant is extracted by the formulas (2.17) and (2.18) according to the user-defined permittivities in the FEM system and the polarization ΔV between the electrodes.

$$\nabla^2(\varepsilon_0 \varepsilon_r V) = 0 \quad (2.22)$$

$$Q_{L,B} = \varepsilon_0 \varepsilon_{L,B} \oint_{S_L, S_B} \mathbf{E} \cdot d\mathbf{S} \quad (2.23)$$

$$C_{L,B} = \frac{Q_{L,B}}{\Delta V} \quad (2.24)$$

The simulation of the cell constants of the IDE sensor are performed by varying two ratios of the geometrical parameters: the finger spacing and width ratio and the finger width and height ratio. The results in Fig. 2.19 are compared to the use of formula (2.19) in order to highlight the limitation of the elliptic analytical calculation.

We can assert from these simulation results that the FEM calculation approach brings a more accurate quantification of the IDE cell constant; the elliptic formula (2.21) ignoring the finger thickness h_f influence. This loss of accuracy is less influent in the substrate area of the device; the metal thickness influence being nonexistent. For extreme values of the W_f/W_s ratio, the quality of the analytical formula fitting with the FEM results is poor; the configuration of the electrodes indeed becomes incompatible with the elliptic form predicted by the formula.

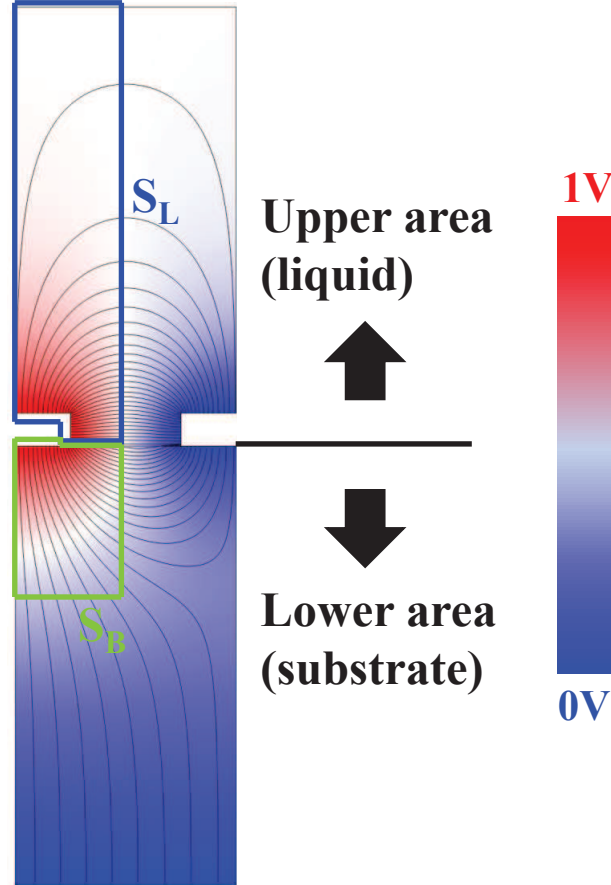


Figure 2.18: Cross view of the simulated IDE cell for the cell constants calculation; the potential profile is given by the color shading with the electric field lines. The Gauss integration surfaces S_L and S_B are schematized in the liquid and substrate areas.

2.4.4 Electrical model error discussion

The previous simulation results validated the evolution of the Z_{IDE} impedance in the frequency domain and gave a calculation technique to evaluate the cell constants of the device sensing area and substrate. It is important to recall the following assertion: the lumped model of the IDE electrical diagram exposed in the previous section (Fig. 2.9) is exact only if the fluid and substrate areas are distinct. In other terms, the electric field close to the sensor fluid-substrate interface is considered as purely parallel to it. In a more realistic context, this

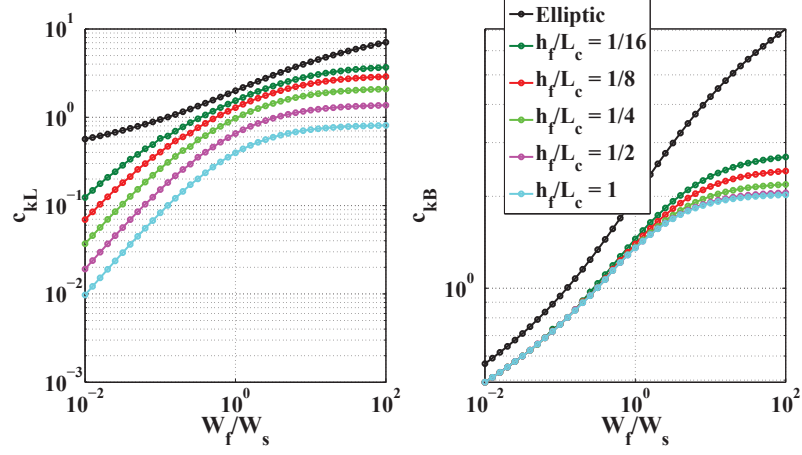


Figure 2.19: Evolution of the IDE cell constants for several W_f/W_s and H_f/L_c ratios, the results are compared with the elliptic formula (2.21). The simulated fluid properties are those of saline water ($\sigma_L = 0.1 \text{ mS.m}^{-1}$ and $\varepsilon_L = 80$)

hypothesis is not quite right. As shown on Fig. 2.20, we observe that the electric field lines are not totally horizontal on the fluid-substrate interface. This electric field distortion is caused by the non-negligible electrode thickness and also by the dissymmetry of the electric potential profile.

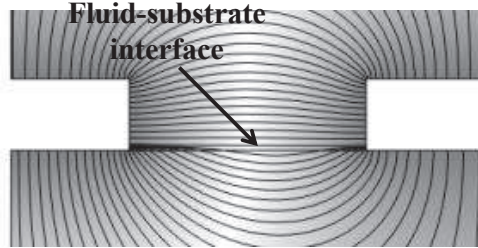


Figure 2.20: Zoom on the fluid-substrate interface; the electric field lines are not perfectly horizontal.

The error encountered by the hypothesis of tangential electric field on the fluid-substrate interface can fortunately be evaluated and therefore allows for fixing the limits of the electrical model validity. This error is extracted from the comparison between the global impedance calculations as performed in subsection 2.4.2 and the impedance (2.3) obtained by the evaluation of the device

cell constants that provides each circuit element of the electrical lumped models. In order to provide a more compact form of this error, equation (2.25) gives the error mathematical expression; $Z_{IDE,glo}$ and $Z_{IDE,imp}$ being respectively the global evaluated impedance and the lumped model impedance.

$$err(Z_{IDE}) = \frac{|Z_{IDE,glo} - Z_{IDE,imp}|}{|Z_{IDE,glo}|} \quad (2.25)$$

The simulation plotted in Fig. 2.21 illustrates the effect of the fluid permittivity and conductivity variations on the IDE impedance error. We observe that the error increases with the decrease of the fluid permittivity ϵ_L . The curvature of the electric field in the fluid area is indeed stronger with low fluid permittivities close to the substrate interface causing thus the error to increase. The maximum of the error is located around the cut off-frequency fixed by the $R_L - C_L$ dipole. The fluid conductivity σ_L does not affect the error in terms of amplitude, only a frequency shift is observed. This shift directly follows the impedance cut off-frequency fixed by the $R_L - C_L$ dipole.

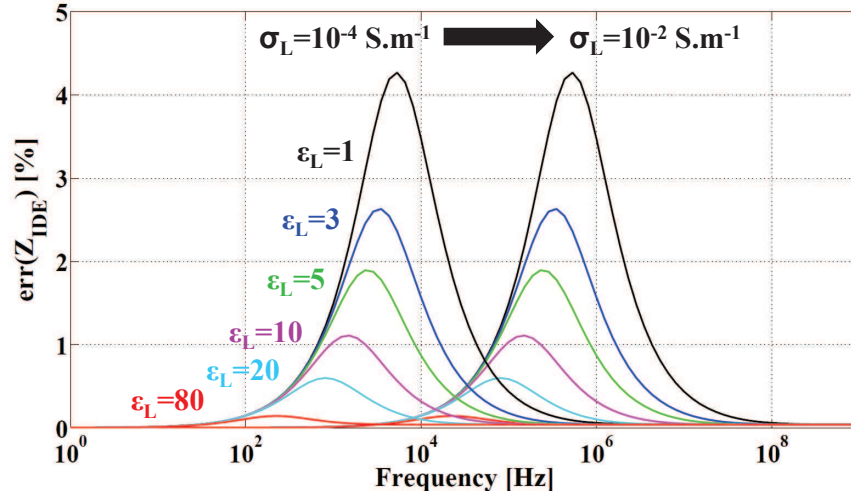


Figure 2.21: Impedance error observed for several fluid permittivities and conductivities versus the voltage stimulus frequency; all the other parameters are set to the default values of Tab. 2.4.

A second effect has also to be notified: the error caused by the passivation material. This error is due to the non-constant value of the electric potential at the passivation-fluid interface, causing thus a loss of accuracy during the

lumped element extraction. The major influent parameter on this error is the passivation layer thickness h_p ; the greater this parameter value, the stronger the perturbation of the electric field close to the passivation layer interface and therefore the greater the error encountered on the IDE impedance. In this context, the liquid cell constant c_{kL} , being extracted from the Gauss integral in the fluid area, gives access to the global capacitance C_L^* seen between the electrodes. Because of the passivation layer influence on the cell constant calculation, the capacitance C_L^* does not correspond exactly to the expected liquid capacitance C_L . Consequently, the C_P capacitance of the lumped electrical model is extracted using the classical planar capacitance formula (2.26). The liquid capacitance C_L is next calculated by considering the serial configuration of C_P and C_L^* (2.27). This allows for recalculating the fluid cell constant more accurately and therefore the fluid resistance R_L . We simulated the phenomenon in Fig. 2.22, the error impact grows quickly with the passivation thickness.

$$C_P = \frac{\varepsilon_0 \varepsilon_P L_f N_f \left(\frac{W_f}{2} + h_f \right)}{2h_p} \quad (2.26)$$

$$C_L = \left(\frac{1}{C_L^*} - \frac{1}{C_P} \right)^{-1} \quad (2.27)$$

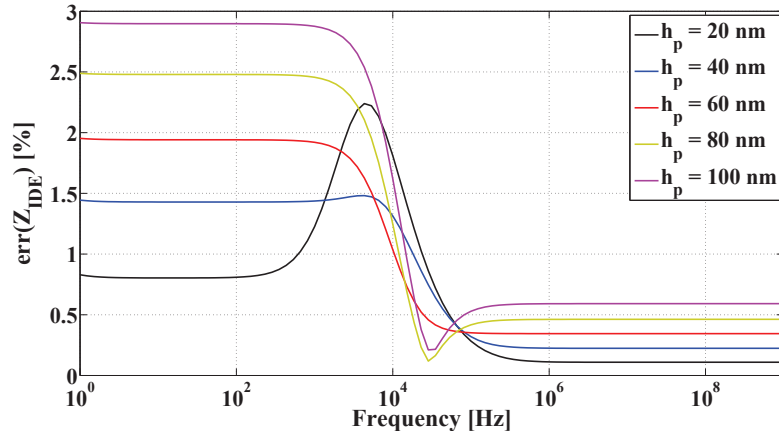


Figure 2.22: Impedance error observed for several passivation thicknesses versus the voltage stimulus frequency; all the other parameters are set to the default values of Tab. 2.4.

In order to close the lumped impedance error discussion, the electric representation of the sensing element given in Fig. 2.9 can be considered as valid since the measured fluid permittivities are sufficiently high, as example, this

parameter has to respect $\varepsilon_L > 3$ to keep the impedance error under 3% as shown in Fig. 2.21. On the other hand, the choice of passivating the IDE sensor has to respect a maximal passivation thickness.

2.5 IDE sensor sizing methodology

The present section explains how the IDE sensor dimensions are chosen depending on a specific sensing application. Starting from a sensor technology and the constraints imposed by this application, a sizing algorithm is established and gives the IDE sensor optimal dimensions for specific detection ranges of fluid permittivity and conductivity.

2.5.1 Methodology description

Prior to describing the algorithm, an important point has to be clarified about the liquid sensing applications. Varying the conductivity of a liquid by keeping its permittivity to a constant value is possible. For example, a given salt concentration is dissolved in water; the water permittivity will be kept close to 80. The contrary is not possible. The only way to make the liquid permittivity varying is to perform mixtures of several fluids, and this will inevitably change the conductivity of the mixture assuming that each liquid has its own conductivity. Consequently, the sizing algorithm starting point considers two possible application modes: mixed-mode $\sigma_L - \varepsilon_L$ or conductivity measurement. These modes give the minimal permittivity or conductivity the sensing system must detect. Depending on these sensing modes and also on the performances of the sensing interface exposed further, a minimal resolution of the sensed liquid resistance or capacitance is next chosen. This allows to fix the desired IDE cell constant c_{kL} for the application.

- Mixed-mode: the required IDE cell constant is fixed by the liquid permittivity. As said before, its variation range is indeed much narrower than the liquid resistivity and therefore more binding for the IDE sizing. The interfacing principle developed further also gives the possibility of adjusting the liquid resistance range by varying the stimulating current injected in the sensor.
- Conductivity mode: the application requires a maximum liquid conductivity σ_L to be quantified, the minimum of measurable resistance R_L is constrained by the sensing system, the required cell constant is thus yielded by $c_{kL} = \sigma_L R_L$. According to the permittivity of the tested liquid, this cell constant gives access to the nominal value of the liquid capacitance C_L which has to be high enough to be detected by the sensing interface. If this capacitance is too small to be quantified, the addition of a capacitor C_M is required in parallel with the IDE sensor.

Once the required cell constant c_{kL} is determined from one of the two application modes cited before, the sizing of the interdigitated structure depends

on the fabrication technology being used. A technology node fixes its own minimal metal pattern width W_f , spacing W_s and height h_f ; setting thus the geometrical dimensions of a metal finger. The substrate of the material is also an input of the chosen technology; having its own dielectric constant. The FEM tools developed in the physical modeling section consequently allows for computing the normalized cell constants c_{kLn} and c_{kBn} for a periodic cell element of the structure per units of finger length for the chosen technology. The effective length of a finger length is thus given by the ratio of the required cell constant c_{kL} by the normalized cell constant of the technology c_{kLn} , as expressed by (2.28).

$$L_{f,eff} = \frac{c_{kLn}}{c_{kL}} \quad (2.28)$$

The goal of the device interdigitation is to avoid long structures and therefore to optimize the area required by the fabrication of the device. The formula (2.29) consists in the calculation of the number of interdigitated electrodes N_f in the case of a square structure; the length of an interdigitated finger L_f is thus calculated by (2.30). The interdigitation of the electrodes in a square shape is the most common but interdigitating a device into a rectangle or spiral shape [37] is also a possibility.

$$N_f = \text{round} \left(\sqrt{\frac{L_{f,eff}}{W_f + W_s}} \right) \quad (2.29)$$

$$L_f = \frac{L_{f,eff}}{N_f} \quad (2.30)$$

The interdigitated sensor is now sized, the last step is to calculate the minimal serial capacitance to add to the system in order to respect the global form of the IDE impedance (2.3). As explained previously in the detection principle section, the serial capacitance can be made of an insulating material that covers the metal fingers (the C_P component of Fig. 2.9) or can be added externally by a physical component (the C_A component of Fig. 2.11). The choice of passivating the IDE electrodes generally depends on the kind of metal in which the electrodes are fabricated. Indeed, a more sensitive metal to corrosion phenomena (for example: aluminum) should be passivated. In the case of passivated interdigitated structures, the serial capacitance directly fixes the material passivation thickness h_p according to its dielectric constant.

The calculation of the serial capacitance C_P or C_A is based on the global impedance formula maximum of phase. In order to have a sufficiently wide flat area on the bode diagram of the IDE impedance modulus, the maximum of phase has to be fixed as high as possible. Being derived from equation (2.3)

in appendix C.1, the mathematical expression of the phase maximum is given by (2.31); the frequency point of this maximum being given by (2.32).

$$M_{\Phi} = \max(\Phi_{Z_{IDE}}) = \arctan\left(\frac{2\sqrt{\omega_z\omega_p}}{\omega_z - \omega_p}\right) \quad (2.31)$$

$$\omega_{max} = \sqrt{\frac{\omega_z^2\omega_p - \omega_z\omega_p^2}{\omega_z - \omega_p}} \quad (2.32)$$

Replacing the zero and pole frequencies by their modeling expressions of passivated and unpassivated IDE structures, we can extract the C_P and C_A capacitances by solving the equation (2.33); C_P or C_A being the unknown. Since the mathematical development led to a second order polynomial for unpassivated devices, the solution for C_A is expressed by (2.33). The expression of the equation for the passivated structures led to a fourth order polynomial, the best manner is thus to solve the equation numerically.

$$C_A = 2(C_L + C_B) \frac{1 + \sqrt{1 + \tan^2 M_{\Phi}}}{\tan^2 M_{\Phi}} \quad (2.33)$$

2.5.2 Sizing algorithm

In order to summarize the sizing algorithm, a global sizing diagram is depicted on Fig. 2.23. The blue rectangular boxes represent the inputs of the algorithm: the application requirements, the circuitry ability of detection and the used IDE technology; all these inputs fix several parameters. The brown rectangular boxes are the intermediate steps: the FEM box consists of the normalized cell constant calculation for the given technology, the IDE box is the application of equation (2.29) and (2.30), the impedance phase box makes the calculation of the serial capacitance C_A or C_P according to formula (2.31), and the passivation thickness is provided depending on the choice of insulating the metal fingers. All the simulation tools used by this algorithm allows therefore for providing all the value of the electrical models parameters.

2.5.3 Sizing examples

The sizing algorithm has been tested for the two application modes. We give examples of two fabrication technologies, the first is a Printed Circuit Board (PCB) technology and the second is a miniaturized silicon technology. In order to be in the most realistic context, we also suppose that the silicon technology is interfaced by a more efficient circuitry which involves the possibility of sensing smaller capacitance values. The resistive resolution of the circuitry is the same for the two technologies. The used materials for the substrate of the sensors

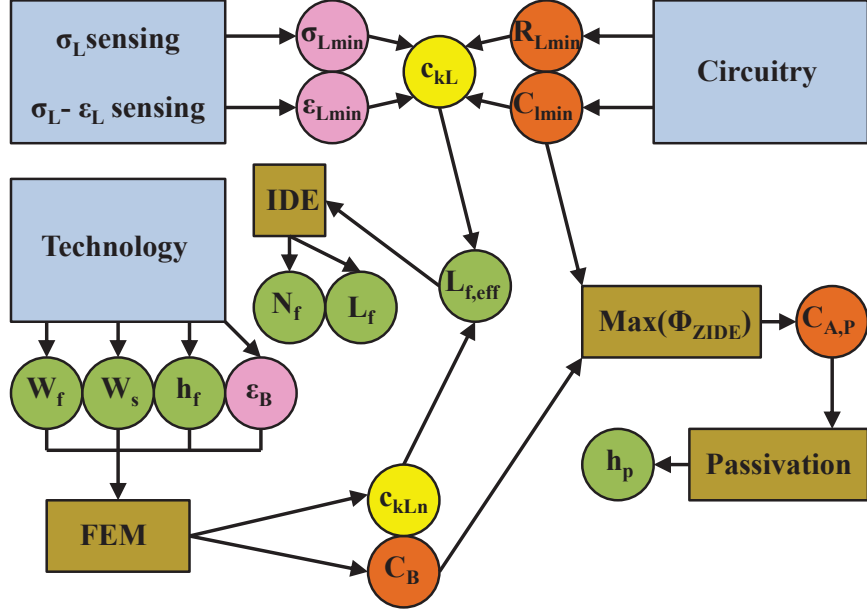


Figure 2.23: Sizing algorithm diagram.

are the *FR4* material for the PCB technology ($\varepsilon_{r,box} = 4.1$) and the silicon oxide for the Si technology ($\varepsilon_{r,box} = 3.9$). The eventual passivation layer is supposed to be of aluminum oxide $\alpha - Al_2O_3$ ($\varepsilon_{r,pass} = 8.9$).

The first resulting values reported in Tab. 2.5 have been calculated for the mixed-mode detection. We supposed an application that requires a minimal permittivity of 20. It is of interest to analyze the impact of the capacitance resolution of the circuitry; the better the resolution of this capacitance (cf. 3rd line of Tab. 2.5), the higher the maximum detectable conductivity of the liquid (cf. last line of Tab. 2.5). Indeed, the choice of passivating the electrodes in the case of the silicon technology requires very thin layers of insulating material of a few dozens of nanometers. The best way to obtain sufficient deposition accuracy for such thicknesses is to use the Atomic Layer Deposition (ALD) process [33].

The second example is reported in Tab. 2.6; it illustrates the conductivity detection mode. As explained at the beginning of this section, the starting point of this mode is based on the maximal detectable liquid conductivity which has been set to $\sigma_L = 0.01 \text{ S.m}^{-1}$ for this example. We supposed the liquid per-

	PCB technology	Si technology
Minimal permittivity	$\varepsilon_{r,L} = 20$	$\varepsilon_{r,L} = 20$
Resistance resolution	$R_{L,min} = 1000 \Omega$	$R_{L,min} = 1000 \Omega$
Capacitance resolution	$C_{L,min} = 100 \text{ pF}$	$C_{L,min} = 10 \text{ pF}$
Metal resolution	$W_{f,s} = 150 \mu\text{m}$	$W_{f,s} = 2 \mu\text{m}$
Metal thickness	$h_f = 30 \mu\text{m}$	$h_f = 1 \mu\text{m}$
Substrate permittivity	$\varepsilon_{r,box} = 4.1$	$\varepsilon_{r,box} = 3.9$
Required cell constant	$c_{kL} = 1.77 \text{ m}^{-1}$	$c_{kL} = 17.7 \text{ m}^{-1}$
Resulting finger number	$N_f = 52$	$N_f = 119$
Resulting finger length	$L_f = 0.016 \text{ m}$	$L_f = 0.48 \text{ mm}$
Resulting sensor surface	$A_s = 250 \text{ mm}^2$	$A_s = 0.2285 \text{ mm}^2$
Serial capacitance	$C_{A,P} = 2.88 \text{ nF}$	$C_{A,P} = 285 \text{ pF}$
Passivation permittivity	$\varepsilon_{r,pass} = 8.9$	$\varepsilon_{r,pass} = 8.9$
Passivation thickness	$h_p = 1.2 \mu\text{m}$	$h_p = 15.8 \text{ nm}$
Maximal conductivity	$\sigma_L = 1.77 \text{ mS.m}^{-1}$	$\sigma_L = 17.7 \text{ mS.m}^{-1}$

Table 2.5: Sizing algorithm application for the mixed-mode detection.

mittivity as 80 (i.e. water). The cell constant is therefore fixed by the maximal detectable conductivity. As noted for the mixed-mode example, the capacitive resolution of the circuitry has a major impact on the sensing system. If the liquid capacitance obtained from the calculated cell constant is smaller than the capacitive resolution of the circuit interface, an additional capacitance C_M is added in parallel with the sensor.

	PCB technology	Si technology
Maximal conductivity	$\sigma_L = 0.01 \text{ S.m}^{-1}$	$\sigma_L = 0.01 \text{ S.m}^{-1}$
Liquid permittivity	$\varepsilon_{r,L} = 80$	$\varepsilon_{r,L} = 80$
Resistance resolution	$R_{L,min} = 1000 \Omega$	$R_{L,min} = 1000 \Omega$
Capacitance resolution	$C_{L,min} = 100 \text{ pF}$	$C_{L,min} = 10 \text{ pF}$
Metal resolution	$W_{f,s} = 150 \mu\text{m}$	$W_{f,s} = 2 \mu\text{m}$
Metal thickness	$h_f = 30 \mu\text{m}$	$h_f = 1 \mu\text{m}$
Substrate permittivity	$\varepsilon_{r,box} = 4.1$	$\varepsilon_{r,box} = 3.9$
Required cell constant	$c_{kL} = 10 \text{ m}^{-1}$	$c_{kL} = 10 \text{ m}^{-1}$
Resulting finger number	$N_f = 22$	$N_f = 159$
Resulting finger length	$L_f = 6.6 \text{ mm}$	$L_f = 0.635 \text{ mm}$
Resulting sensor surface	$A_s = 43.6 \text{ mm}^2$	$A_s = 0.4 \text{ mm}^2$
Liquid capacitance	$C_L = 70.8 \text{ pF}$	$C_L = 70.8 \text{ pF}$
Additional capacitance	$C_M = 29.2 \text{ pF}$	not required
Serial capacitance	$C_{A,P} = 0.72 \text{ nF}$	$C_{A,P} = 0.504 \text{ nF}$
Passivation permittivity	$\varepsilon_{r,pass} = 8.9$	$\varepsilon_{r,pass} = 8.9$
Passivation thickness	$h_p = 0.85 \mu\text{m}$	$h_p = 15.8 \text{ nm}$

Table 2.6: Sizing algorithm application for the conductivity detection.

2.6 Interfacing principle: IDE transient response

The detection principle and the physical modeling of the sensor exposed in the previous sections described the sensitive element impedance in the frequency domain (2.3) and explained the FEM calculation method used to quantify the electrical model lumped component values. Starting from this expression, an original interfacing methodology of the sensor is proposed and demonstrated in the present section. As exposed in [38], the principle is based on the Triangular Waveform Voltage (TWV) generation. The technique we propose consists in the injection of a time-square current signal into one of the electrodes of the device, the other electrode and the backside being tied to the ground (Fig. 2.24). The liquid parameters are consequently included in the shape of the probed voltage wave. The transient response of an IDE structure is then reported.

2.6.1 Impedance Laplace transform

The first step of the mathematical demonstration is to establish the expression of a square-shaped current wave (2.34) of amplitude I_0 and period $2T$, $u(t)$ being the unit hard-step function. As shown in [20], the two first steps of the current wave have respectively $I_0/2$ and $3I_0/2$ of amplitude; this will keep the mean value of the voltage response at zero. The Laplace transform of the

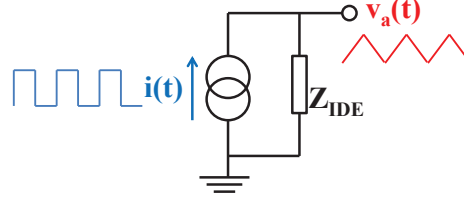


Figure 2.24: Square current injection in the IDE sensor impedance.

current wave is next expressed by (2.35).

$$\begin{aligned}
 i(t) &= \frac{I_0}{2}u(t) - \frac{3I_0}{2}u(t-T) \\
 &+ 2I_0 \sum_{n=2}^{\infty} (-1)^n u(t-nT)
 \end{aligned} \tag{2.34}$$

$$\begin{aligned}
 I(s) &= \frac{I_0}{2s} - \frac{3I_0 \exp(-Ts)}{2s} \\
 &+ 2I_0 \sum_{n=2}^{\infty} (-1)^n \frac{\exp(-nTs)}{s}
 \end{aligned} \tag{2.35}$$

In every LTI (Linear Time Invariant) system, the response $V_a(s)$ (2.36) is obtained by the multiplication in the Laplace domain of the system input $I(s)$ by the impulse response which is the sensor impedance Z_{IDE} given by equation (2.3). The triangular transient or time domain voltage wave $v_a(t)$ (2.37) is thus the result of the application of the inverse Laplace transform on the $V_a(s)$ system response. Three parameters are next defined to describe the shape of the triangular wave $v_a(t)$: the coefficients β , ζ et τ . These last quantities are given by the expressions (2.38), (2.39) and (2.40) making the mathematical link between the impedance zero-pole frequencies $\omega_z - \omega_p$ and initial amplitude C_0 . An example of the transient progression of the transient voltage $v_a(t)$ and current wave $i(t)$ is illustrated on Fig. 2.25.

$$V_a(s) = I(s)Z_{IDE}(s) \tag{2.36}$$

$$\begin{aligned}
v_a(t) &= 0.5u(t) \cdot \left[\beta t + \zeta - \zeta \exp\left(-\frac{t}{\tau}\right) \right] \\
&- 1.5u(t-T) \cdot \left[\beta(t-T) + \zeta - \zeta \exp\left(-\frac{t-T}{\tau}\right) \right] \\
&+ 2\beta \sum_{n=2}^{\infty} [(-1)^n (t-nT) \cdot u(t-nT)] \\
&+ 2\zeta \sum_{n=2}^{\infty} [(-1)^n \cdot u(t-nT)] \\
&- 2\zeta \sum_{n=2}^{\infty} \left[(-1)^n \exp\left(-\frac{t-nT}{\tau}\right) \cdot u(t-nT) \right] \quad (2.37)
\end{aligned}$$

$$\beta = I_0 \frac{C_0 \omega_z}{\omega_p} \quad (2.38)$$

$$\zeta = \frac{I_0}{C_0 \omega_p} \left(1 - \frac{\omega_z}{\omega_p} \right) \quad (2.39)$$

$$\tau = \frac{1}{\omega_p} \quad (2.40)$$

2.6.2 Links between the output wave and the sensor models

The shaping parameters β , ζ and τ are functions of the lumped elements included in the equivalent models of the IDE structures describes previously. Their mathematical expressions are reported in Tab. 2.7; the mathematical links between the passivated and the unpassivated model structures is summarized. The additional capacitance required for unpassivated structures is included in those expressions.

Symbol	Passivated IDE	Unpassivated IDE	Units
β	$\frac{I_0}{C_P + C_B}$	$\frac{I_0}{C_A}$	[V.s ⁻¹]
ζ	$\frac{I_0 R_L C_P^2}{(C_P + C_B)^2}$	$I_0 R_L$	[V]
τ	$\frac{R_L (C_P C_B + C_L C_P + C_L C_B)}{C_P + C_B}$	$R_L (C_L + C_B)$	[s]

Table 2.7: Expressions of the shaping parameters as functions of the lumped components values of the IDE electrical equivalents models.

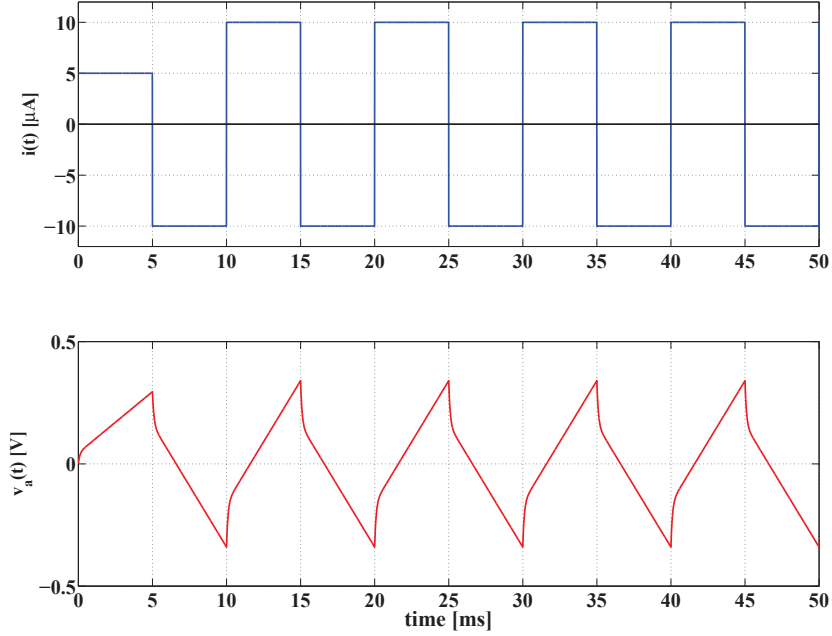


Figure 2.25: Square current signal $i(t)$ and output voltage $v_a(t)$; the simulation parameters, as depicted in Fig. 2.11, are: $\omega_z = 10^4 \text{ rad.s}^{-1}$, $\omega_p = 10^5 \text{ rad.s}^{-1}$ and $Z_{\omega=1} = 10^{10} \Omega$.

2.6.3 Output voltage wave segmentation

The transient development gave the output voltage wave as a serial expression. Each rising or falling wave segment contains all the information we seek about the tested impedance. A compact equation of wave segments thus needs to be calculated to extract the curve parameters β , ζ and τ . The demonstration reported in appendix C.2 derives the mathematical expression (2.41) of wave segments. The curve plotted in Fig. 2.26 summarizes the parametric representation of a wave segment with the definition of a variable ζ' (2.42).

$$v_u(t) = \beta t + \frac{2\zeta}{1 + e^{\frac{-t}{\tau}}} (1 - e^{\frac{-t}{\tau}}) \quad (2.41)$$

$$\zeta' = \frac{2\zeta}{1 + e^{-T/\tau}} \quad (2.42)$$

The segment wave comprises two distinct parts, an exponential one and a linear one. Their separation point is equal to the time where 99% of the exponential part has been attenuated; this corresponds approximately to 5τ .

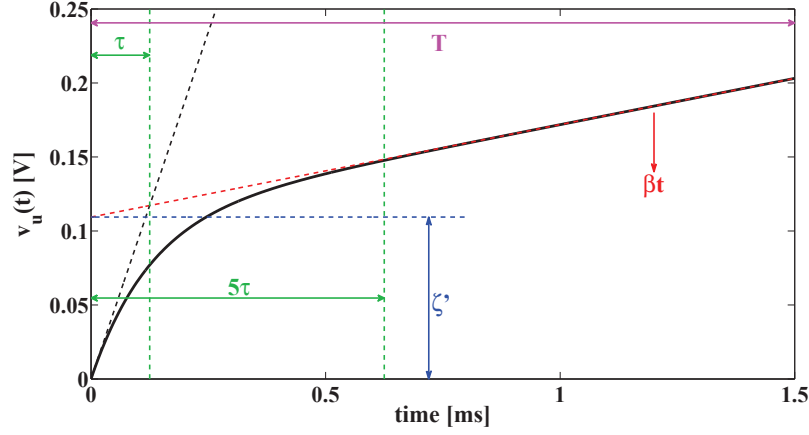


Figure 2.26: Output voltage wave segment illustration predicted by (2.41) and (2.42).

2.6.4 Segmented wave interpretation

The expression (2.41) is the compact mathematical form of each wave segment. Three cases can be observed from this expression.

2.6.4.1 The exponential part of $v_u(t)$ is predominant

This first case appears when the time constant τ becomes many times greater than the current wave oscillation half-period ($\tau \gg T$); the exponential part is therefore approximated by a straight line. The top curves of Fig. 2.27 illustrate the exponential part growing and how this first situation is reached. The derivative of (2.41) at the time $t = 0$ is computed to get the equation of the approximating straight line, resulting to the initial slope expression of the wave segment function (2.43).

$$\left. \frac{dv_u}{dt} \right|_{t=0} = \beta + \frac{-2\zeta}{\tau \left(1 + e^{-\frac{T}{\tau}} \right)} \quad (2.43)$$

According to the comparison between τ and T , the quantity $e^{-\frac{T}{\tau}}$ can be approximated to one. The approximating straight voltage equation is thus given by (2.44). The parameters β , ζ and τ are replaced by the frequency model expressions (2.38), (2.39) and (2.40).

$$v_u(t) = \left(\beta + \frac{\zeta}{\tau} \right) t = I_0 C_0 t \quad (2.44)$$

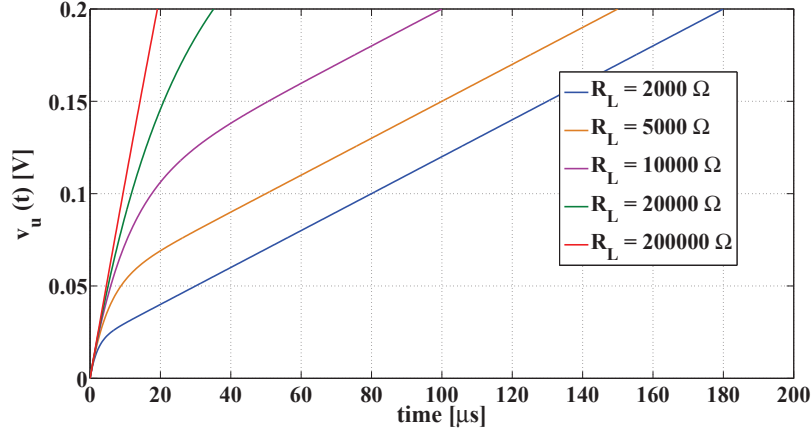


Figure 2.27: Increase effect of the exponential part of the analog voltage illustrating the case $\tau \gg T$.

From the liquid sensing point of view, the situation occurs when the tested liquid resistivity is too high. Mathematically, it is explained by the ζ parameter expression as a function of the IDE model parameters given in Tab. 2.7. This parameter is directly proportional to the liquid resistance R_L and the injection current amplitude I_0 . A too high value of the R_L resistance gives a too high value to the ζ shaping parameter; considering Fig. 2.26, the linear part of the segmented voltage disappears. The simple manner to fix this *out-of-range* situation is to decrease the injection current amplitude I_0 .

Though the liquid resistance R_L cannot be quantified in the present situation; it can be of interest to enforce this case in order to measure directly the liquid capacitance C_L . Indeed, if we replace C_0 by its expression as a function of the circuit elements in equation (2.44), we get new expressions (2.45) for passivated devices and (2.46) for unpassivated ones. These last expressions are straight line equations which slope contains the liquid capacitance C_L .

$$v_u(t, pass) = \left(\beta + \frac{\zeta}{\tau} \right) t = I_0 t \frac{C_P + C_L}{C_B C_P + C_L C_P + C_L C_B} \quad (2.45)$$

$$v_u(t, unpass) = \left(\beta + \frac{\zeta}{\tau} \right) t = I_0 t \frac{C_A + C_L + C_B}{C_A (C_L + C_B)} \quad (2.46)$$

2.6.4.2 The linear part of $v_u(t)$ is predominant

This case is the opposite of the previous one, it occurs when ($\tau \ll T$). As illustrated in Fig. 2.28, the exponential part amplitude becomes negligible;

the signal is thus perfectly triangular and has a slope equal to the β shaping parameter. The $v_u(t)$ equation can therefore be simplified in equation (2.47).

$$v_u(t) = \beta t = I_0 C_0 \frac{\omega_z}{\omega_p} t \quad (2.47)$$

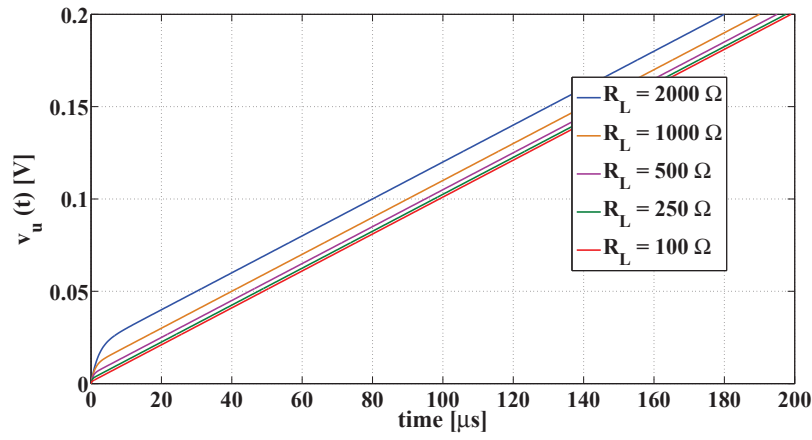


Figure 2.28: Decrease effect of the exponential part of the analog voltage illustrating the case $\tau \ll T$.

Unlike the previous situation, the exponential part becomes negligible when the liquid resistance R_L becomes too low. The ζ parameter being the one which expresses directly the exponential amplitude, an increase of the injection current I_0 is needed to make the two parts of the segmented wave $v_u(t)$ detectable. The injection of the expressions given in Tab. 2.7 in equation (2.47) gives the output segmented voltage equations (2.48) and (2.49) as a function of the circuit elements for both IDE passivation cases. Those mathematical expressions demonstrate the absence of the liquid component's influence on the segmented voltage as shown in Fig. 2.28 when the liquid resistance R_L is too small.

$$v_u(t, pass) = \beta t = \frac{I_0 t}{C_P + C_B} \quad (2.48)$$

$$v_u(t, unpass) = \beta t = \frac{I_0 t}{C_A} \quad (2.49)$$

2.6.4.3 Both parts of $v_u(t)$ appear distinctly

This case corresponds to the general form of the segmented voltage $v_u(t)$ and is required to extract the fluid resistance R_L . Focusing on the ζ' parameter,

the exponential term $e^{-\frac{t}{\tau}}$ can be approximated to zero with an error of less than 1% when the condition $T > 5\tau$ is satisfied. The parameter ζ' is therefore substituted by 2ζ (2.50), the ζ parameter being directly linked to the fluid resistance R_L .

$$v_u(t) = \beta t + 2\zeta(1 - e^{-\frac{t}{\tau}}) \quad (2.50)$$

2.7 Interface characterization

The experiments presented in this section are conducted to validate the detection methodology applied by the oscillating system proposed in the previous section. The system circuit architecture is first described and next implemented on PCB (Printed Circuit Board) with discrete components. The CMOS integration is simulated in the next section; this will improve the system performances in terms of detection resolution, power consumption and size reduction. This section first describes the characterization setup and the PCB platform. Preliminary tests on lumped capacitance and resistance elements are next performed to illustrate the evolution of the analog output voltage and to test the accuracy of the components value extraction. Two kinds of liquid measurements are finally presented to extract the conductivity and permittivity of liquids. The first experiment characterizes the conductance variation of NaCl and KCl solutions while the second test highlights the ability of the system to detect both the permittivity and the conductivity of glycerol-water mixtures. Finally, a discussion highlights circuit limitations.

2.7.1 System block diagram

Thanks to the easiness from which a square signal current can be physically generated, it is of major interest to represent the system input and output signal in the time domain. The sensed information is retrieved from the charge and discharge of the sensitive device by this square current. The system we propose implements the TWV technique with a current controlled oscillator (CCO) so that no variable signal source is needed to stimulate the circuit; the oscillating wave being generated by the system itself. Two opposite current sources are alternatively enabled and disabled by a digital command $v_d(t)$ which is the system feedback control signal. The swing of the analog wave $v_a(t)$ is limited by the hysteresis of a Schmitt trigger. The system polarity changes each time $v_a(t)$ reaches the upper or lower limit of the hysteresis voltage. The system interface block diagram is illustrated in Fig. 2.29.

2.7.2 Circuit implementation

The CCO architecture of the closed-loop system generates an oscillating signal according to a controlled square current signal. This kind of system is easily transferable on CMOS advanced technologies [47]. As our goal is to prove the liquid instrumentation concept, we design a similar architecture using discrete components with bipolar transistors to mirror the load current; the discrete platform allows for highlighting all limitations of the system that the CMOS integration must improve. The diagram depicted in Fig. 2.30 represents the discrete architecture of the sensing system. The MOSFET devices act as

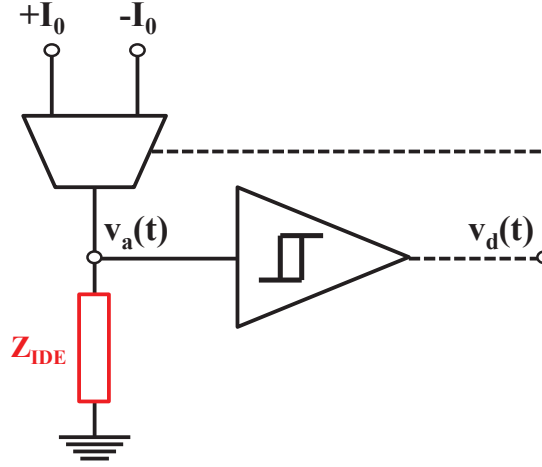


Figure 2.29: System block diagram representation.

switches that invert the circuit polarity. The analog voltage is monitored by a Schmitt NAND gate which yields the feedback digital control to change the controlling current polarity. A high input impedance analog buffer has been added to avoid the deformation of the triangular wave due to the relatively low input impedance of the probing equipment.

2.7.3 Characterization setup

The circuit architecture depicted in Fig. 2.30 has been implemented on a standard PCB platform (*Multi Circuit Boards Ltd.*) using discrete components listed in Tab. 2.8. The board includes two output connectors used to probe the analog voltage and the digital control signal. The analog buffer and the Schmitt trigger are supplied by the voltage values listed in Tab. 2.9. The trigger hysteresis amount V_h is fixed by the supply voltage amplitude. The serial capacitance C_S value of the system is set to 90 nF to fix an output frequency range starting from 100 Hz to 10 kHz respecting the load current amplitude range of 10 μ A to 1 mA. The interdigitated structures are fabricated by gold deposition directly printed on the circuit board; the passivation layer that covers the metal fingers is removed. A picture of the discrete circuit is depicted in Fig. 2.31.

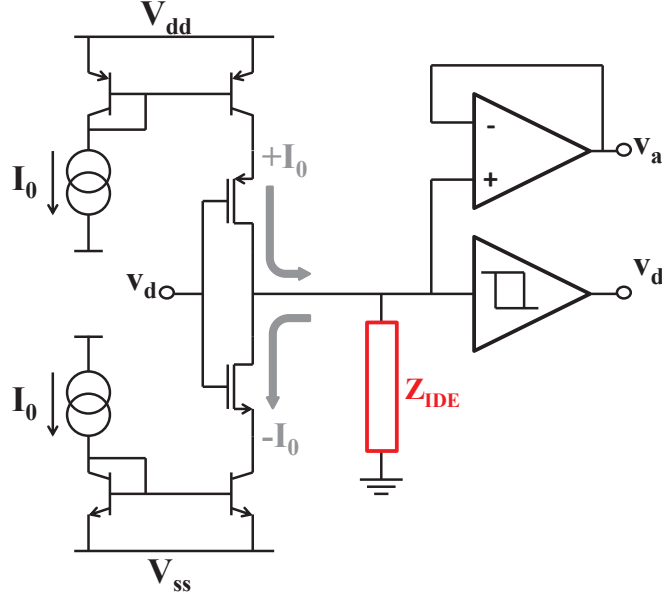


Figure 2.30: System block diagram.

Component type	Technical name
PNP bipolar mirror	BCV62 [55]
NPN bipolar mirror	BCV61 [56]
Analog CMOS switch	ADG636 [57]
Schmitt NAND gate	CD4093BE [58]
Analog amplifier	AD8661ARZ [59]

Table 2.8: PCB prototype components annotations and models.

Description	Name	Value
Positive supply voltage	V_{dd}	5 V
Negative supply voltage	V_{ss}	0 V
Hysteresis voltage amplitude	V_{hst}	0.9 V
Default serial capacitance	C_A	90 nF

Table 2.9: System setup parameters.

2.7.4 Lumped elements values extraction

The analog voltage measurements have been performed using lumped components that represent the IDE sensor electrical model. The tested components

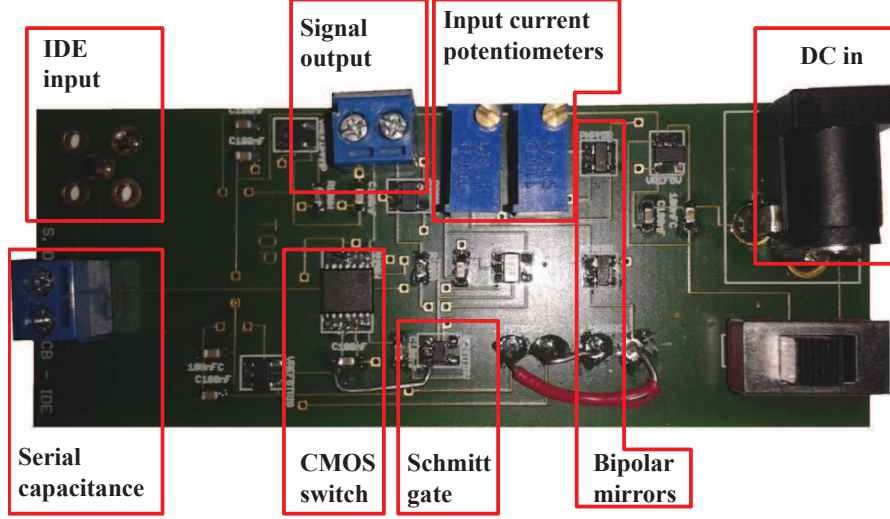


Figure 2.31: PCB discrete circuit illustration.

are listed in Tab. 2.10, the analog voltage is probed with the *Agilent Infiniium MSO8104A* oscilloscope and plotted in Fig. 2.32. These results validate the analog output wave shape variations as predicted by the analytical model of section 2.6. As the system works in closed-loop, the analog wave oscillates between the two hysteresis levels enforced by the Schmitt gate. The values of the lumped elements have been extracted according to the compact expression of the analog voltage (2.50). The measured values given in Tab. 2.10 are compared to the values of the lumped components used for this experiment. Most errors are lower than 5%; this proves thus the interfacing method efficiency. As the C_L capacitance is the smallest to quantify, it presents the highest extraction error. As discussed further in more detail, the justification of the observed errors is divided into two categories. The first one regroups all the static errors due to the circuit accuracy; the parasitic capacitances and the quality of the current copy in the sensing device. The second category of errors is the repeatability of the measurements, i.e. the jitter and noise errors caused by the probing equipment.

2.7.5 Liquid conductivity measurements

The objective of the electrolyte conductivity tests is to prove the efficiency, repeatability and linearity of the present sensing system; several NaCl and KCl solutions are then prepared. The setup of the experiment is schematized in

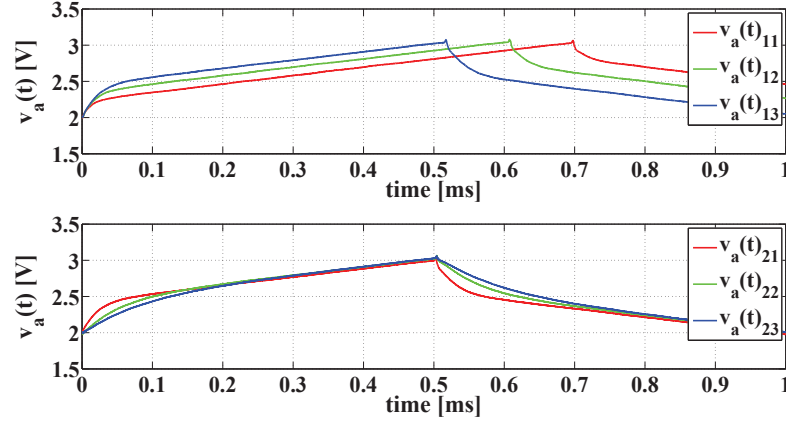


Figure 2.32: Analog voltage versus time for different values of R_L (top view) and C_L (bottom view), the legend details are reported in Tab. 2.10.

wave index	R_L [Ω]	C_L [nF]	C_A [nF]	
$v_a(t)_{11}$	1100	10	90	calibrated
	1140	10.5	86.5	measured
	3.6%	5%	3.9%	error
$v_a(t)_{12}$	1600	10	90	calibrated
	1670	10.7	86.7	measured
	4.2%	7%	3.7%	error
$v_a(t)_{13,21}$	2100	10	90	calibrated
	2049	10.7	87.2	measured
	2.4%	7%	3.1%	error
$v_a(t)_{22}$	2100	25	90	calibrated
	2152	24.7	87.9	measured
	2.4%	1.2%	2.3%	error
$v_a(t)_{23}$	2100	40	90	calibrated
	2187	38.7	88.7	measured
	4%	3.2%	1.4%	error

Table 2.10: Tested lumped components expected values and measured fitting equation (2.50).

Fig. 2.33; a commercial conductivity meter *PC700 EUTECH* probes the solution conductivity while the IDE sensing system measures the conductance of the solution. The full conductivity measurement range of the equipment is included in the $[0.0 \mu\text{S} - 200 \text{ mS}]$ interval and split into five scales; the equipment accuracy is 1% of each conductivity scale.

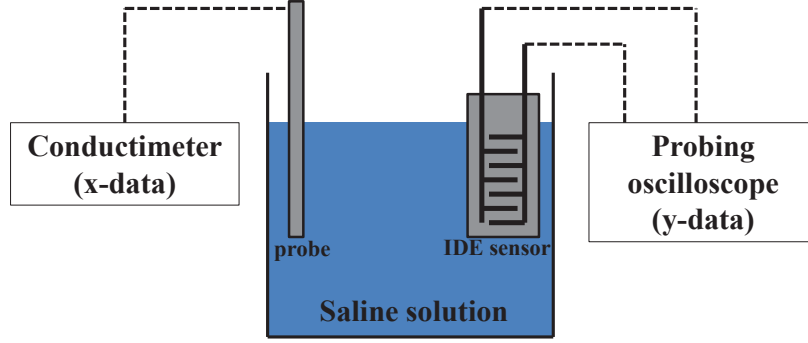


Figure 2.33: Setup overview of the saline solutions characterization.

In practice, the experiment starts by dissolving a small quantity of salt in order to obtain a first reference conductivity point from the *EUTECH* meter. After a few seconds, the measurement is stabilized. The corresponding salt concentration is evaluated according to formula (2.51).

$$c_{NaCl|KCl} = \sigma_L (\lambda_{Na+|K+} + \lambda_{Cl-}) \quad (2.51)$$

During the acquisition of the reference conductivity point, the response $v_a(t)$ of the IDE system is monitored and displayed by the *Agilent Infiniium MSO8104A* oscilloscope. After the *EUTECH* conductivity meter display stabilization, the probed wave data is saved a dozen of times. Another small quantity of salt is next added to the solution and the same measurement procedure is applied several times in order to get a sufficient number of measurements. This results in a linear characteristic; the x-coordinates of the measured points are consequently defined by the concentration of the electrolyte samples calculated according to formula (2.51) and the values of the ionic molar conductivities reported in Tab. 2.11; σ_L being measured by the commercial conductimeter. The y-coordinates of the measured points correspond to the liquid conductance, i.e. R_L^{-1} , sensed by the IDE system; the standard deviation of the multiple saved data is given for each measured point.

In the case of the salt concentration measurements, the IDE sensor dimensions (Tab. 2.12) are obtained from the sizing algorithm developed in section 2.5 while considering the range of the electrolyte samples conductivities; the resulting cell constants c_{kL} and c_{kB} have been computed using the FEM tools. According to the water and the substrate permittivities, the respective nominal capacitances C_L and C_B of the liquid and the substrate are evaluated using the

Ion type	λ [mS.m ² .mol ⁻¹]
Na^+	5.01
K^+	7.35
Cl^-	7.63

Table 2.11: Ionic molar conductivities.

cell constant formulas (2.17) and (2.18). As explained in the sizing methodology description, if the IDE capacitance is too small to be detected by a system designed with discrete components, an additional capacitance C_M must be added in parallel with the IDE sensor. The serial capacitance C_A of the electrical model of Fig. 2.12 is finally chosen to assure a sufficient phase maximum of the sensed impedance. The picture presented in Fig. 2.34 shows the interdigitated sensor printed in gold. It is also important to specify that the dimensions values of the electrodes undoubtedly allow for neglecting the electrical double layer effects.

Parameter	Value	Unit
W_f	500	[μm]
W_s	500	[μm]
N_f	20	[—]
L_f	11.5	[mm]
c_{kL}	7.5	[m ⁻¹]
c_{kB}	7.5	[m ⁻¹]
C_L	3.07	[pF]
C_B	157	[fF]
C_M	5	[nF]
C_A	30	[nF]

Table 2.12: IDE parameters for saline concentration tests.

The results plotted on Fig. 2.35 and 2.36 show the progression of the liquid conductance R_L^{-1} extracted from the system triangular output voltage probed by the *Agilent* oscilloscope. As mentioned before, the x-axis scale is the electrolyte concentration which corresponds to the calibrated conductivity of the solutions. We observe, at first sight, an excellent linearity and repeatability of the liquid measurements. The standard deviation of the measurements increases with the electrolyte concentration; a high electrolyte concentration means that the liquid resistance is small, difficult to extract, and more sensitive to the probing equipment noise and jitter during the falling and rising trigger digital signal events. The results are also compared to simulated values (red curves)

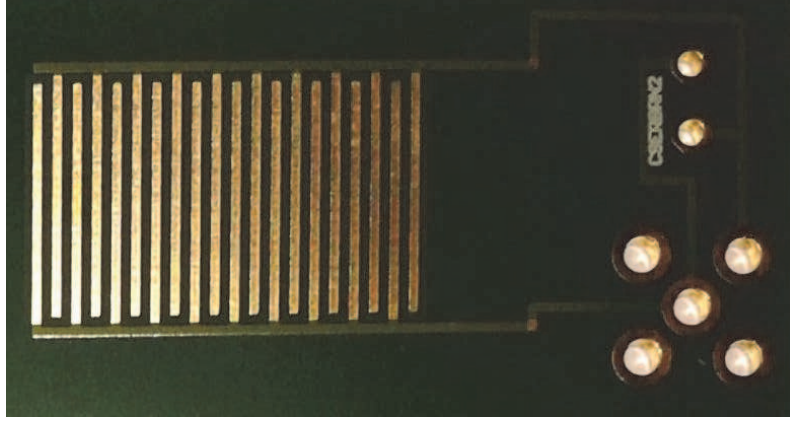


Figure 2.34: Printed interdigitated sensor used for the saline measurements application.

obtained by the FEM calculation of the cell constant and the application of formula (2.17); linking the cell constant c_{kL} with the liquid conductance. This comparison allows for evaluating the static error of the system; it corresponds to the small difference between the measured and simulated lines. The static error is explained by all the imperfections of the circuit prototype including the parasitic capacitances, the quality of the current copy, and the device fabrication which alters the cell constant value. On the other hand, the FEM calculation errors discussed in subsection 2.4.4 also alter the IDE cell constant value. Nevertheless, the observed error is small and does not exceed 3% of error.

The sensitivity s_{σ_L} of the IDE sensing system is given by the slope of the straight characteristics of Fig. 2.35 and 2.36. If we have a closer view on a measurement point (Fig. 2.37), we observe, as measured previously, the conductivity deviation error Σ and the concentration reading error δ which corresponds to the ratio of the conductivity deviation error Σ by the system sensitivity s_{σ_L} (2.52). In conclusion, the greater the system sensitivity, the lower the error encountered for a measurement of any fluid saline concentration.

$$\delta = \frac{\Sigma}{s_{\sigma_L}} \quad (2.52)$$

2.7.6 Liquid permittivity measurements

The characterization of liquid permittivity is performed by mixing two liquids of different permittivities. The two liquids have to be miscible in each other to be homogeneous and thus to create the most linear permittivity range

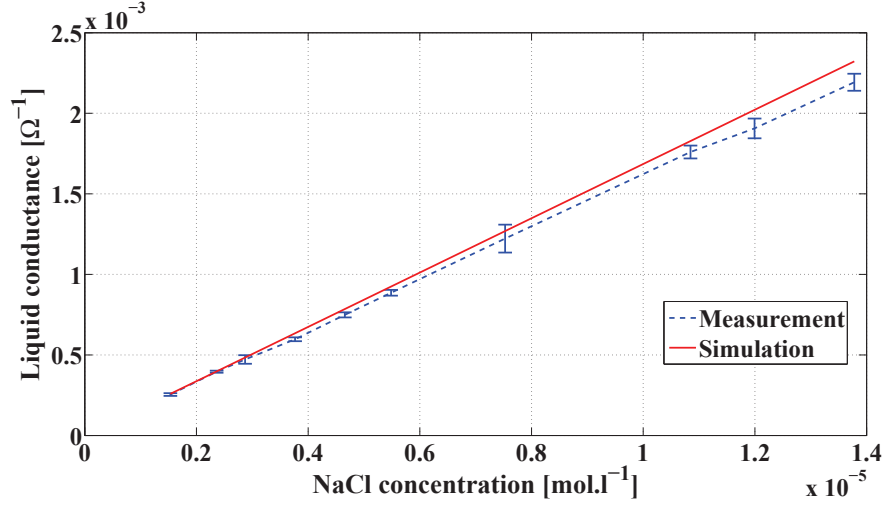


Figure 2.35: Evolution of the NaCl electrolyte conductivity for several salt dilutions.

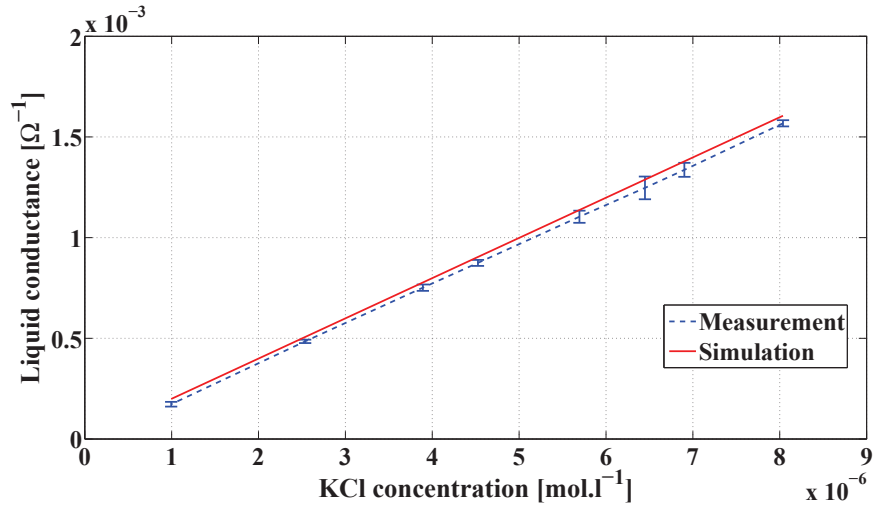


Figure 2.36: Evolution of the KCl electrolyte conductivity for several salt dilutions.

of mixtures. We start from pure glycerol in which we add water to prepare several mixtures of different water-glycerol ratio. Knowing that the relative permittivities are about 41 for glycerol and 80 for water, the IDE sensor cell constant has to be chosen to have a significant capacitance to detect. The IDE parameters to achieve permittivity measurements are listed in Tab. 2.13; a picture of the sensor is shown in Fig. 2.38. The measurements depicted on Fig. 2.39 show the evolution of the dielectric constant versus the water-glycerol

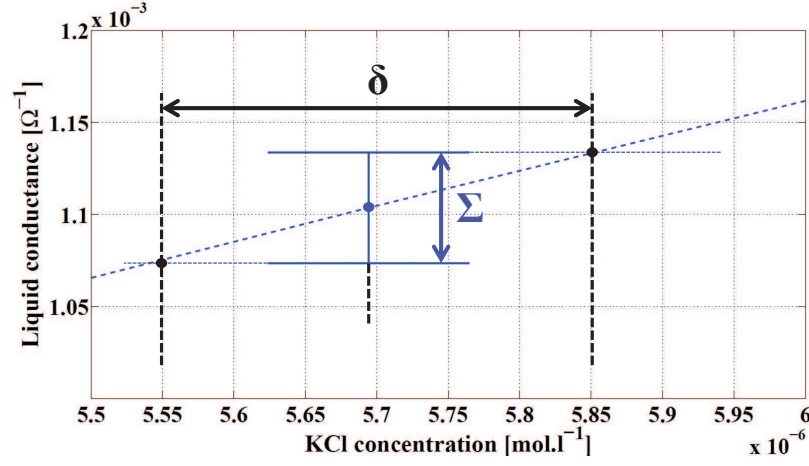


Figure 2.37: Zoom on a measured conductance point of Fig. 2.36.

ratio of the mixture. As these two liquids have a different natural conductivity, the evolution of conductivity is also shown for the mixture. Similar to the salt measurement application, the double layer effects can also be neglected according to the IDE electrodes dimensions.

Parameter	Value	Unit
W_f	150	$[\mu\text{m}]$
W_s	150	$[\mu\text{m}]$
N_f	300	$[-]$
L_f	3	$[\text{cm}]$
$c_{k,L}$	0.14	$[\text{m}^{-1}]$
$c_{k,B}$	0.14	$[\text{m}^{-1}]$
C_B	260	$[\text{pF}]$
C_A	5	$[\text{nF}]$

Table 2.13: IDE parameters for water-glycerol tests.

The permittivity curve versus the water-glycerol ratio was supposed to be linear due to the homogeneous prepared mixtures. The progression of the permittivity and resistivity presents a slight curvature which can be explained by difference between the physical features of the two liquids : glycerol is much more viscous than water and the electrode metal hydrophobic properties are different for the two fluids.

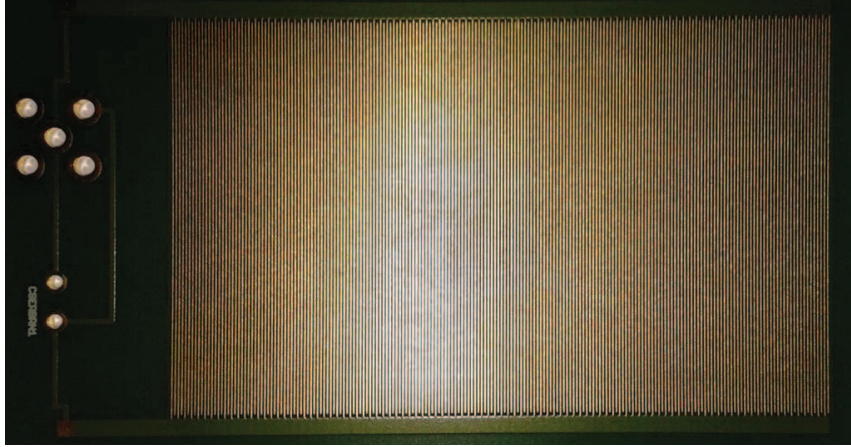


Figure 2.38: Printed interdigitated sensor used for the water-glycerol mixture measurements application.

2.7.7 Circuit interface limitations

In every system implementation, the circuit elements are non-ideal and limit system performances. Our major concern is to insure the good quality of the analog output signal to detect capacitances and resistances on the largest measurement range. All parasitic impacts on the analog output voltage have therefore to be discussed. As the system is implemented on a PCB platform by discrete components, it will be easier to highlight the system limitations. Having a good identification of these problems is important for system implementation on any technology.

2.7.7.1 Clock feedthrough and charge injection

The first error that impacts the measurement results is the voltage drop ΔV_{dp} observed on the analog voltage (Fig. 2.40). In order to quantify this error, we replaced the sensor by a pure capacitor; this generates a pure triangular analog output. The rising and falling slopes of the signal are next stretched as shown by the black lines in Fig. 2.40; the voltage drop is next extracted by taking the difference between the intersection points of the slopes prolongations on the vertical edge of the digital control signal.

The source of this error is located in the analog switch formed by the MOS pair depicted in Fig. 2.30 and is due to two effects: the clock feedthrough [48] and the charge injection [49].

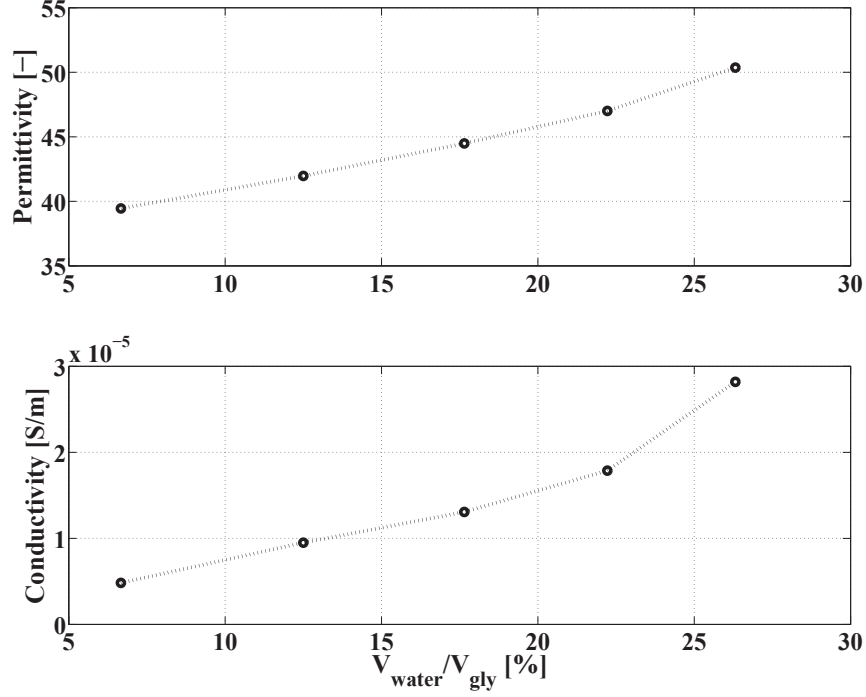


Figure 2.39: Characterization of the permittivity (top curve) and conductivity (bottom curve) of water-glycerol mixtures; the mixture ratio is given by $V_{\text{water}}/V_{\text{gly}}$.

The clock feedthrough effect is mainly caused by the overlap gate-to-drain capacitance C_{gd} . Each time the system polarity is changed by the Schmitt gate, a capacitive divisor is formed by the test capacitance C_L and the overlap capacitance C_{gd} . On the other hand, the schematic representation illustrated in Fig. 2.41 also takes into account the contribution of the C_{sd} capacitance; the v_s node corresponds to the source nodes of the MOS switches depicted in Fig. 2.30. Similar to the clock feedback voltage $v_d(t)$ applied on the gates of the switches, the voltage of these source nodes presents amplitude steps that involve a feedthrough on the analog output voltage $v_a(t)$. The total amount of voltage drop ΔV_{cf} caused by the clock feedthrough is evaluated by equation (2.53); C_L being the tested capacitance.

$$\Delta V_{cf} = |v_d| \frac{C_{gd}}{C_{gd} + C_L} + |v_s| \frac{C_{sd}}{C_{sd} + C_L} \quad (2.53)$$

The second source error, identified as the charge injection effect, is due to the elimination of the charges present in the transistors channel. Considering

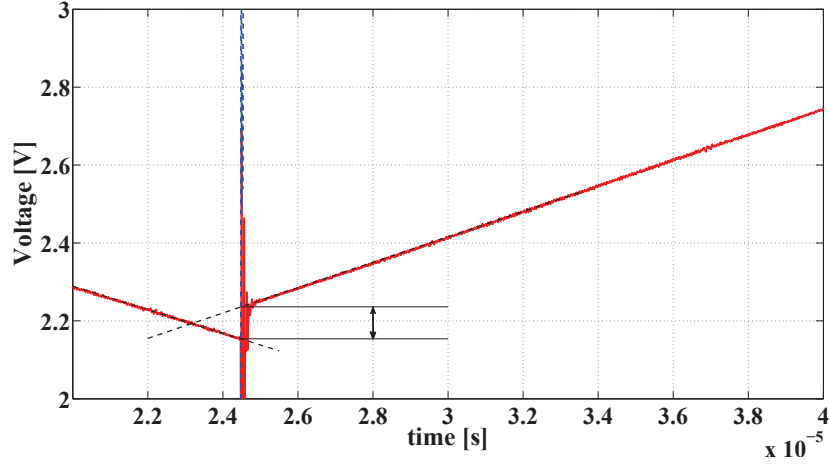


Figure 2.40: Observed voltage drop ΔV_{dp} denoted by the black arrow; $v_a(t)$ (continuous red line), $v_d(t)$ (discontinuous blue line).

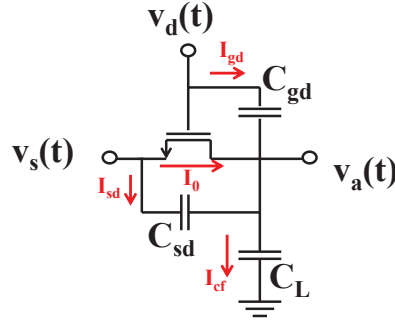


Figure 2.41: Schematic illustration of the clock feedthrough effect; a parasitic current I_{cf} is added in the test capacitance during each clock event.

Fig. 2.42, the channel charges are evacuated in the source and drain of the device when a switch-off voltage is applied on its gate; this effect thus involves a voltage difference ΔV_{ci} on the analog output wave according to equation (2.54); Q_i and C_L being respectively the amount of charge injection and the sensed capacitance.

$$\Delta V_{ci} = \frac{Q_i}{C_L} \quad (2.54)$$

The total amount of drop voltage ΔV_{dp} observed on the measurements is therefore given by the sum of the clock feedthrough and charge injection contributions (2.55).

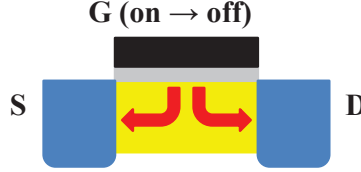


Figure 2.42: Representation of the charge injection when a switch MOS transistor is toggled off; the charges (red arrows) are evacuated in the source and drain terminals.

$$\Delta V_{dp} = \Delta V_{ef} + \Delta V_{ci} \quad (2.55)$$

Considering the example of the clock feedthrough and charge injection voltage drop V_{dp} measurement depicted in Fig. 2.40, we observe a total drop of 90 mV for a 200 pF test capacitance. This resulting drop voltage amplitude is the contribution of both clock feedthrough and charge injection effects. Considering equation (2.54) the charge injection amount of 0.3 pC given in the datasheet [57], the resulting voltage drop is only of 1.5 mV using a test capacitance of 200 pF. We can therefore assert that the clock feedthrough effect is clearly predominant. The Switched Capacitor (SC) circuits are often impacted by this kind of parasitic effect; solutions using bootstrapping techniques and addition of dummy switches have been designed to compensate the clock feedthrough effect [50].

Anyway, the consideration of those parasitic effects is of major importance. The consequence of this voltage drop error can be disastrous for the extraction of any fluid conductance and capacitance. For example, a voltage drop of more than 10% of the total available amplitude of the triangular signal $v_a(t)$ may completely hide the exponential part of the signal. This causes the fluid parameters ε_L and σ_L hard to estimate; especially for low fluid resistances and low capacitances that respectively involve low values of ζ and τ . The only way to attenuate those effects on the signal is to lower as most as possible the feedthrough capacitance of the analog CMOS switch. This can be achieved by downscaling the transistors of the analog switch.

2.7.7.2 Parasitic capacitance effects

In every circuit implementation, limitations are caused by the parasitic capacitances of the components. These capacitances affect the analog measure-

ment and make the sensor capacitance more difficult to be quantified, lowering the sensitivity. The parasitic effects are represented in Fig. 2.43 with all the capacitances connected to the analog voltage node. The Schmitt trigger and the analog buffer have their own input capacitances noted C_i , the capacitances C_{ds1} , C_{ds2} and C_{dg} are the MOSFET switches internal capacitances. Depending on where these capacitances are connected, their impact is different on the analog voltage.

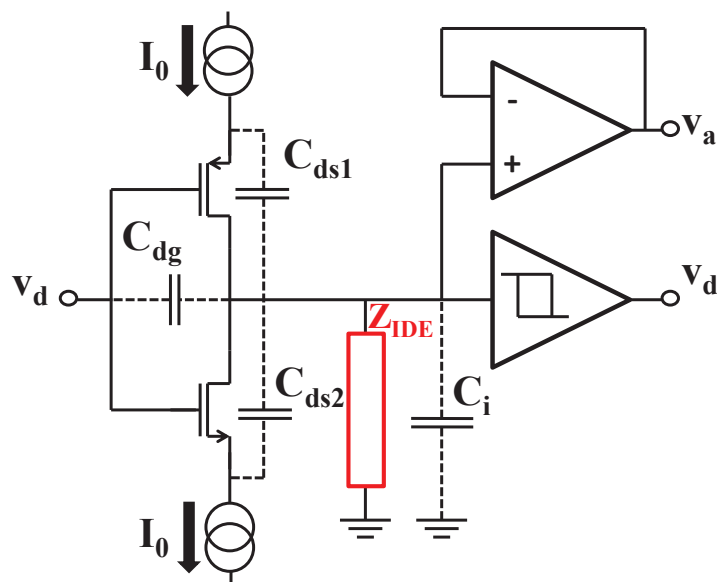


Figure 2.43: Analog voltage node parasitic capacitances.

The capacitance C_i is connected to a DC node and only impacts the system sensitivity. As discussed before, the effect of the C_{dg} , C_{ds1} , C_{ds2} capacitances was proved to be the cause of the clock feedthrough effect. In addition, these elements are the source of second-order effects. Being observed in Fig. 2.40, these effects are identified as perturbations at the beginning of the rising or falling times of the feedback digital signal, the analog output response presents small peaking effects compensated by small oscillations which disappear after a few microseconds. Regarding the system frequency bandwidth, the limiting factors are the Schmitt trigger switching times and the second order oscillations caused by the parasitic capacitances of the switches. The PCB implementation of the interface includes a *4093 NAND* gate component as a Schmitt trigger [58]. Its transition time depends on the supply voltage. For the test conditions presented

in the measurement section, both the rising and falling times of the trigger are of 600 ns. In comparison with the duration of the second order oscillations which exceeds 1 μ s, the bandwidth limitation is therefore due to the parasitic capacitances of the switches. For the proposed PCB implementation, we set the working frequency limit under 50 kHz; the quality of the analog output voltage is considered as to be too deteriorated beyond this limit.

2.7.7.3 Current mirrors

The last impact on the measurements is due to the current copy performed by the bipolar transistors. Regarding the datasheet of the bipolar transistors [55, 56], the minimum value of the forward characteristic β_f is 100; this means that the base current error is lower than 1%. The quality of the current copy is also altered by the Early effect and the mismatch between the transistors. We compensated this error by performing a calibration which consists in measuring the mirrored amount of current injected in a reference test capacitor.

2.7.8 Error discussion conclusions

In order to close the measurement errors discussion, the most impacting parasitic effects on the fluid properties characterization are the clock feedthrough and charge injection phenomena. The resulting voltage drop engendered by those effects may completely hide the exponential part of the analog voltage segment and thus cause the fluid parameters estimation to be impossible. For all cases where this error does not exceed the exponential part amplitude of the analog signal, the system can be calibrated to fix the error we observed for the saline concentration measurements. The calibration can be performed by testing fluids of well-known properties; an experimental cell-constant of the sensor is estimated and thus allows for correcting the data extracted by the sensing system.

The most suitable and smartest solution to lower all the parasitic effects is to integrate the interfacing circuitry and therefore to downscale all the CMOS transistors. The next section clearly proves that the system integration on advanced CMOS technologies drastically attenuates these undesirable effects.

2.8 IDE interface CMOS integration

Since the sensing system has been modeled and tested using discrete elements on a PCB technology, the last step of the scientific methodology is to prove that the CMOS integration of the circuit interface brings much better sensing performances. The CMOS technology chosen for the system integration is the TSMC 250 nm technology. This technology node on which the circuits are designed is especially adapted for mixed-signal circuits [51]. The technology is fabricated on bulk silicon wafers and counts five metal levels. The major advantage of this technology is the large number of transistor doping levels allowing for choosing several threshold voltages.

IDE structures are only made of metal patterns, the possibility of device and circuitry co-integration is therefore possible. The present section first gives an overview of the CMOS technology layers and specifies how the IDE sensor can be merged with the CMOS circuitry. The proposed integrated circuit, described further, is composed of two blocks: the switched current mirrors required to inject a specific level of current into the device and the Schmitt trigger. For this last block, the present architecture allows for programming the hysteresis level externally. Simulations are finally performed to foreground the contributions of the circuit integration in terms of accuracy, power consumption, and effective system surface reduction.

As specified in the introduction of this manuscript, the study is focused on the functional part of the analog interface. The system DC source supplies are supposed to be ideal and stable; the study of the DC sources impacts is a perspective of the present work. All the simulations presented in the present section have been performed after circuit layout generation; the parasitic elements extracted after the post-layout netlist generation are therefore considered.

2.8.1 Sensor and circuit co-integration

Every CMOS integrated technology is made of several material layers. These layers are fabricated using clean-room processes to implement the expected circuit architecture. Focusing on the TSMC 250 nm technology, the structure is made of five stacked metal levels. These metal layers are deposited above the CMOS devices which are buried under all the oxide layers that separate each metal level. Thanks to the large number of metal layers, it becomes possible to draw the IDE structure on the top metal level while insulating the circuit influence from the IDE sensor by a ground plane. The technology cross view schematized in Fig. 2.44 illustrates the stacking of the different layers of material that compose the structure.

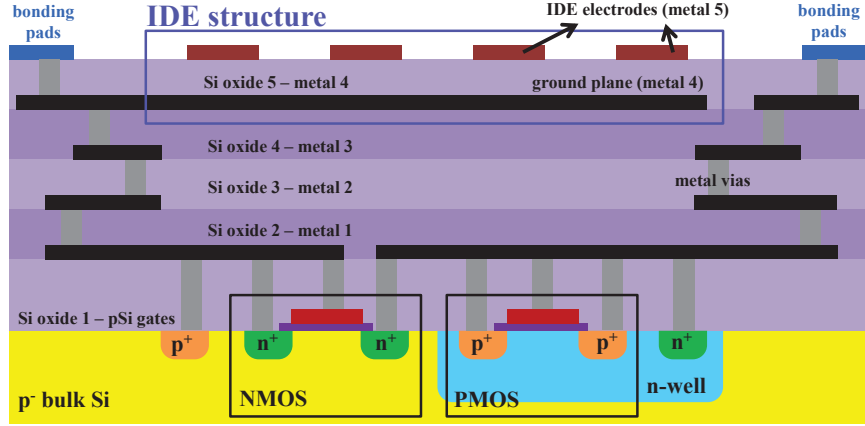


Figure 2.44: Co-integration of the IDE sensor and its interfacing circuitry in the TSMC 250 nm technology: cross view of the structure. The IDE sensor and its ground plane are respectively fabricated in metal 5 and metal 4.

2.8.2 Circuit block 1: current mirrors

Performing the copy of a current from a circuit branch to another is typically done by MOSFET transistors connected in diode operation. As discussed after the measurements of the previous section, the principal concerns are the clock feedthrough and the charge injection that occurs on the analog switches. The architecture of Fig. 2.45 is composed of cascoded MOSFET pairs; this configuration assures a much better quality of the current copy required for such a sensing system. The analog switch is the CMOS current-controlled inverter with which the gates are connected to the digital output of the Schmitt trigger.

The sizing of all the mirroring transistors has the advantage to be free of constraints. The dimensions are therefore chosen in order to consume a minimal design area. Their threshold voltage is chosen to obtain a sufficient output swing range for the triangular voltage while guaranteeing a low leakage current. The medium threshold devices are the most appropriate for this architecture. The switching transistors M_{SP} and M_{SN} must be chosen to minimize the charge injection effect observed in the measurements section. The normal threshold voltage presents the lowest charge injection effect and the sizing dimensions are set to symmetrize the rising and falling times of the inverter formed by the M_{SP} and M_{SN} transistor pair. All the transistor sizes and layout of the switched mirror stage are reported in appendix E.

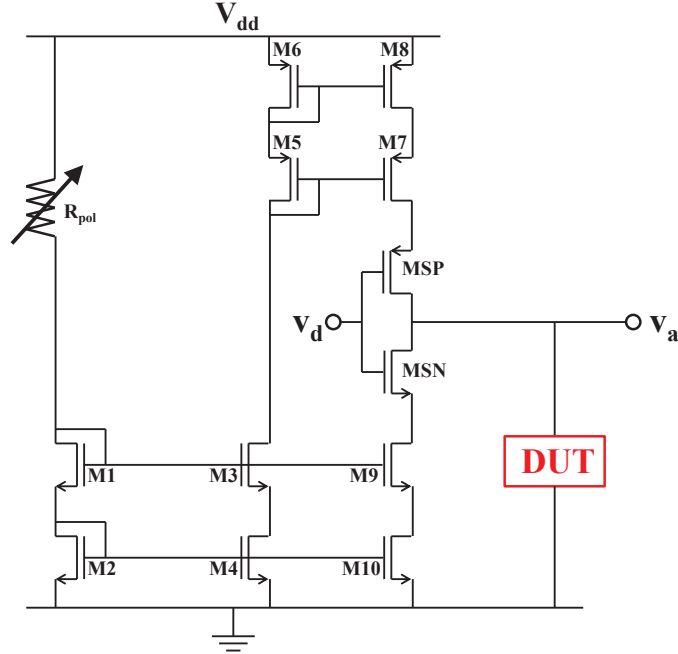


Figure 2.45: CMOS Switched current source architecture.

2.8.3 Circuit block 2: Schmitt trigger

As explained in the system description section, the feedback control which causes the polarity of the system to be inverted is assured by a Schmitt Trigger. In this section, we propose an extension of the component possibilities to a programmable hysteresis trigger. The programming of a Schmitt trigger can be performed by digital control transistors [53] or using current conveyors [54]. We chose to modify an implementation based on the architecture proposed in [52]. The trigger hysteresis can be adjusted externally according to a simple resistor.

2.8.3.1 Global system view

The system view depicted on Fig. 2.46 illustrates the two stages composing the circuit block: a differential comparator (sub-circuit A) and the hysteresis adder (sub-circuit B). In actuality, this kind of system acts exactly like a differential comparator of which the reference voltage is shifted by the hysteresis adder stage; the reference voltage being tied to the supply voltage half.

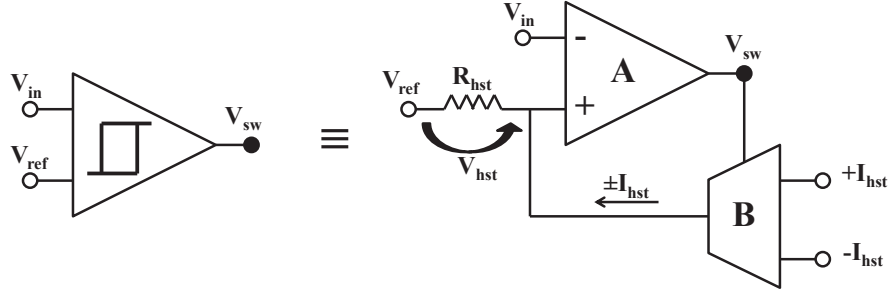


Figure 2.46: Schmitt trigger block diagram.

2.8.3.2 Differential comparator

The comparator stage architecture is a classic differential pair on which a decision stage has been connected as shown on Fig. 2.47. The design of this component is of major importance since it sets the trigger speed and the output swing for the hysteresis voltage range. In our application context, the following constraints have been fixed:

- Maximum hysteresis amount: 1 V
- Maximum working frequency: 50 kHz

The choice of these specifications is based on the possibility of digitizing the triangular output voltage of the sensing system. If the system works at its maximal frequency of 50 kHz, an analog to digital converter (ADC) will have to sample the voltage at a frequency of $5 \cdot 10^6$ megasamples per second in order to provide a minimum of 100 data points. The large targeted amount of hysteresis of 1 V is also an advantage for the ADC accuracy; a small hysteresis amount requires indeed a high conversion accuracy (i.e. the Effective Number of Bits (ENOB) of the ADC).

The sizes of the inverting decision stage transistors can be chosen to optimize the switching delay of the comparator. Indeed, if the transistors $M7$ and $M8$ are too long, the inverter drives too small current and causes the component to be slower. High transistor width allows consequently for driving more current but also increases the input capacitance of the inverter causing thus the slew rate of the node v_A (cf. Fig. 2.45) to be smaller. The sizes of the differential

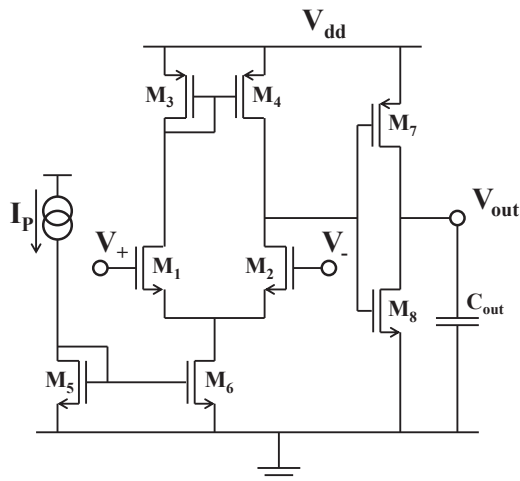


Figure 2.47: Input comparator stage.

2.8.3.3 Hysteresis adder

2.8.3.3 Hysteresis adder

The last circuit stage of the Schmitt trigger is the hysteresis adder. Its objective is to provide a well-defined level of current which directly fixes the amount of hysteresis for the trigger. As presented in [52], the implementation of the hysteresis stage consists of a differential pair which output is connected to the resistance R_{hst} which is added externally. The transistor dimensions are reported in appendix E.

2.8.4 Schmitt trigger simulations

The principal merits presented by our Schmitt trigger architecture are the accuracy of the hysteresis voltage, the power consumption and the working frequency bandwidth. The goal of the simulations we performed is to prove that the objective fixed by the trigger performances has been reached.



This first simulation considers the trigger in open-loop; a sine test voltage is applied at its negative input. The goal of this simulation is to perform a preliminary study of the V_{hst} evolution as a function of the sine test voltage frequency applied to the negative input of the trigger. The hysteresis amount is next shown for several values of the external programming resistance R_{hst} . The resulting plots illustrated on Fig. 2.49 prove that V_{hst} is of good stability until 100 kHz; this largely exceeds our specification of 10 kHz. This frequency limit is fixed by the input capacitance of the hysteresis block formed by the $M11$ and $M12$ pair shown in Fig. 2.48. Furthermore, the linearity of V_{hst} with the programming resistance is perfect. The simulations have been performed considering the process deviation models provided by the TSMC foundry.

The previous simulation result does not take into account the impact of the trigger insertion in the whole sensor interface. As explained before, the triangular output voltage of the system is generated by the injection of a fixed amount

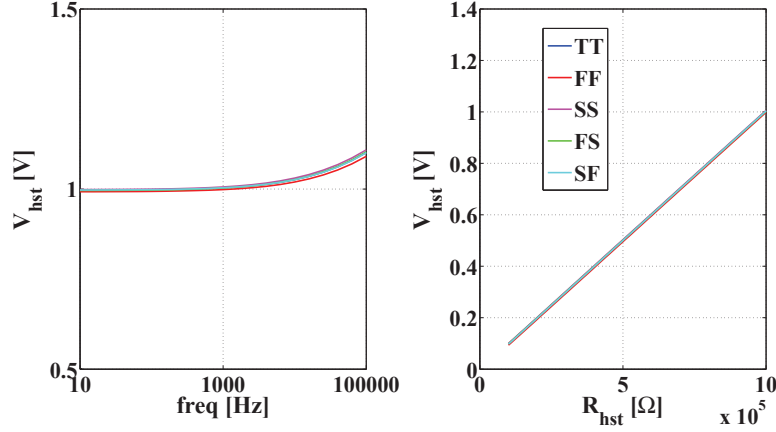


Figure 2.49: Trigger hysteresis voltage versus the input signal frequency on the left ($R_{hst} = 1 \text{ M}\Omega$) and versus the external resistance on the right ($\text{freq} = 1 \text{ kHz}$); the legend shows all the possible process deviations (T, F and S respectively mean typical, fast and slow, the two letters are used to indicate both the NMOS and PMOS devices).

of current in the IDE device. If the device is replaced by a pure capacitor C_{IDE} , the shape of the signal is perfectly triangular and its resulting oscillation frequency depends on the capacitor value; the smaller the capacitance, the higher the triangular signal frequency. As demonstrated by the previous simulation (left plot of Fig. 2.49), the accuracy of the hysteresis voltage V_{hst} starts to decrease when the working frequency of the system exceeds 100 kHz. Therefore, an evaluation of the error on V_{hst} has been performed using several values of C_{IDE} . This error is the result of the comparison between the expected value of V_{hst} (right plot of Fig. 2.49) and the value of V_{hst} extracted after the insertion of the Schmitt trigger in the whole sensor interface. Indeed, we observe on Fig. 2.50 that the error is dependent on the triangular signal frequency f_{TWV} . This means, in other terms, that the Schmitt trigger switching delay becomes too high in comparison with the triangular signal period for low C_{IDE} values. Fortunately, high errors on V_{hst} do not destroy the system functionality; this proves that the system remains robust against those errors.

2.8.4.3 Schmitt trigger consumption

Since the goal of the CMOS integration is indeed to reduce the power consumption of the liquid instrumentation system, a simulation versus the input signal frequency is therefore shown on Fig. 2.51. For low input signal frequencies up to 500 Hz, the trigger switching rate is relatively low. The dynamic power

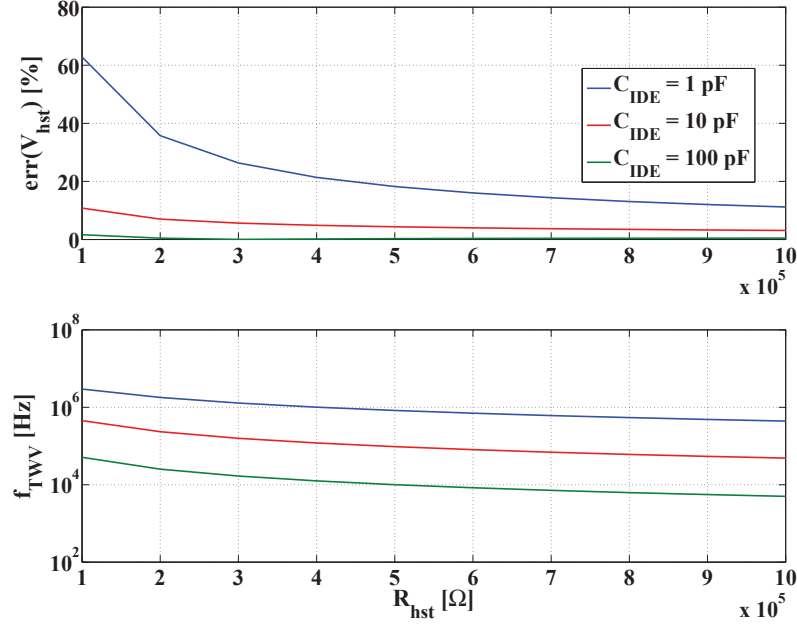


Figure 2.50: Simulation of the error on the hysteresis voltage V_{hst} versus the external programming resistance R_{hst} for several values of the capacitance C_{IDE} .

consumption is thus negligible in comparison to the static one. Beyond 500 Hz, the dynamic is indeed increasing but fortunately never reaches high power levels under 100 kHz.

2.8.5 System detection resolution

Having designed the entire analog CMOS interface of the IDE sensors, the sensing performances are now tested combining the results provided by the FEM simulations and the circuit simulator tools. The minimal capacitance and resistance the interface is able to detect are evaluated.

2.8.5.1 Capacitive detection resolution

As discussed in the measurement section, the minimal capacitance that can be detected is fixed by both the clock feedthrough and the charge injection phenomena. The amplitude of those effects depend on the sizes and models of the switching transistors of Fig. 2.45. In the case of the proposed architecture, we can give an estimation of the voltage drop using the extraction tools of the

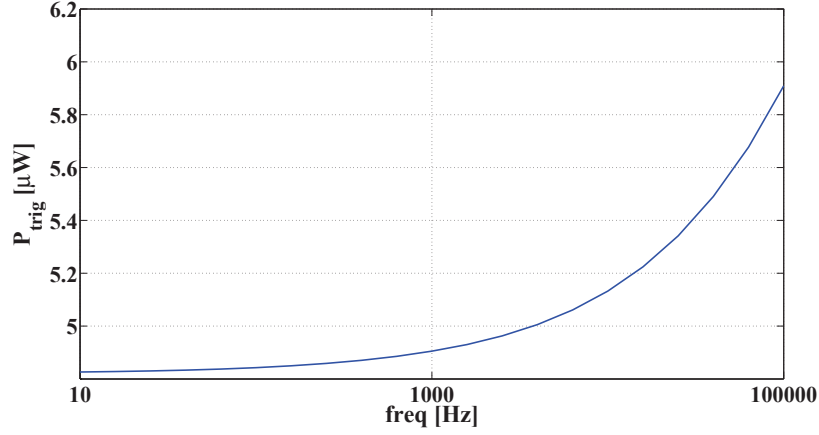


Figure 2.51: Trigger power total consumption versus the input signal frequency on the left ($R_{hst} = 1 \text{ M}\Omega$).

MOSFET parasitic capacitances. We simulated the system by replacing the IDE impedance by a 100 fF test capacitance.

Considering first the clock feedthrough effect, values of 0.5 fF and 0.005 fF have been extracted for the C_{gd} and C_{sd} capacitances of the NMOS and PMOS switches. Due to the very small value of C_{sd} , the contribution of this capacitance is neglected. The voltage drop calculated by equation (2.53) is approximately 12 mV under a DC bias of $V_{DD} = 2.5 \text{ V}$. The charge injection amount is next evaluated using the technology parameter C_{ox} ; evaluated according to the gate oxide thickness of 5 nm. A total gate charge of 5 fC is considered, leading to a voltage drop of 50 mV according to equation (2.54).

Summing the contribution of those two effects, a total voltage drop of 60 mV is evaluated using a 100 fF of test capacitance. If we accept a voltage drop of less than 5% of the total hysteresis amount using all the transistors sizes set previously, the minimal capacitance that can be detected is 120 fF. In comparison with the discrete interface capacitive resolution of 200 pF, we strongly improve the detection quality thanks to the CMOS integration.

2.8.5.2 Resistive detection resolution

In order to estimate the resistance resolution, we have to determine the minimal value of the ζ voltage (2.39) that can be detected. Considering the model equations reported in Tab. 2.7, the ζ potential is linearly proportional to the product $R_L I_0$ in both IDE models. For example, the amplitude of ζ is

equal to 100 mV for a tested liquid resistance of 100 k Ω loaded by a current of 1 μ A. It is also possible to reach very low resistance resolutions by pushing large amounts of currents into the sensing device; a resolution of 100 Ω can be reached using a load current of 10 μ A while measuring a ζ potential of 1 mV.

In the most complete instrumentation system, the analog quantities are digitalized by analog to digital converters (ADC). These components have their own maximal sampling frequency and accuracy.

2.8.6 Liquid parameter extraction accuracy

The largest advantage of the CMOS integration is the accuracy of the liquid parameters σ_L and ε_L measurements. This accuracy can be predicted by combining all of the equation models and simulations tools developed throughout this chapter. The calculation of the estimation error on the liquid parameters is performed by extracting the fitting parameters of the segmented curve $v_u(t)$ simulated by the circuit tools using our CMOS interface. An estimation of the liquid conductivity $\hat{\sigma}_L$ and permittivity $\hat{\varepsilon}_L$ is consequently provided respecting the liquid and substrate cell constants c_{kL} and c_{kB} calculated by the FEM tools. The FEM model also computes the interdigitated impedance Z_{IDE} required by the circuit simulator. The estimation errors are thus given by the equations (2.56) and (2.57) according to the expected conductivity and permittivity injected as inputs of the error calculation algorithm summarized in Fig. 2.52.

$$err(\sigma_L) = \frac{|\hat{\sigma}_L - \sigma_L|}{\sigma_L} \quad (2.56)$$

$$err(\varepsilon_L) = \frac{|\hat{\varepsilon}_L - \varepsilon_L|}{\varepsilon_L} \quad (2.57)$$

2.8.6.1 Passivated structures accuracy

We first tested our error calculation algorithm on the passivated electrode structures of which the electrical model is described in section 2.3. The results illustrated on Fig. 2.53 have been sketched as a function of the input permittivity ε_L for several values of the passivation thickness h_p . The error on the conductivity (i.e. the left part of Fig. 2.53) remains constant. Indeed, the variation of the liquid permittivity does not change the ζ parameter (cf. Tab. 2.7) which is independent of the liquid capacitance C_L , this constant error is only due to the passivated electrical model which considers the interface passivation-liquid as an equipotential node.

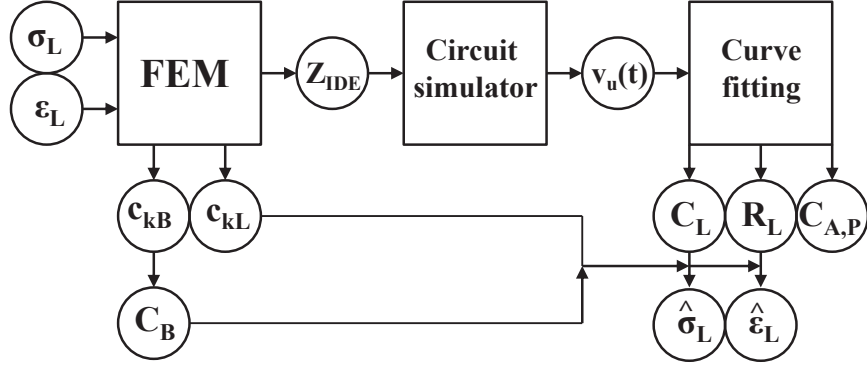


Figure 2.52: Algorithm of liquid parameters error estimation.

The variations of the estimation error on the extracted permittivity $\hat{\epsilon}_L$ (i.e. the middle part of Fig. 2.53) are explained regarding the τ/T ratio (i.e. the right part of Fig. 2.53). Since the fitting of $v_u(t)$ makes a discretization of the wave segment, a small τ/T ratio means a small number of data points on the exponential part of $v_u(t)$ causing the estimation error increase. Working with thin passivation layers makes the C_P capacitance to increase and therefore the β fitting parameter decreases. The voltage wave consequently reaches the hysteresis limit fixed by the trigger in a longer time and causes the ratio τ/T to be lowered so that the estimation error is most important for thin passivation layers.

The simulated case of Fig. 2.54 shows the evolution of the estimation errors by varying the expected liquid conductivity σ_L . The error evolution explanation is the same as the simulation case of the permittivity variation. For both cases the parameter estimation can be improved by adjusting the injection current I_0 in the device in order to maintain a sufficiently large τ/T ratio. This ratio must however be kept under 1 : 5 to assure the visibility of the linear and exponential parts of the segmented wave as explained in section 2.6; the estimation errors indeed increase when this condition is unsatisfied (cf. the left and middle parts of Fig 2.54).

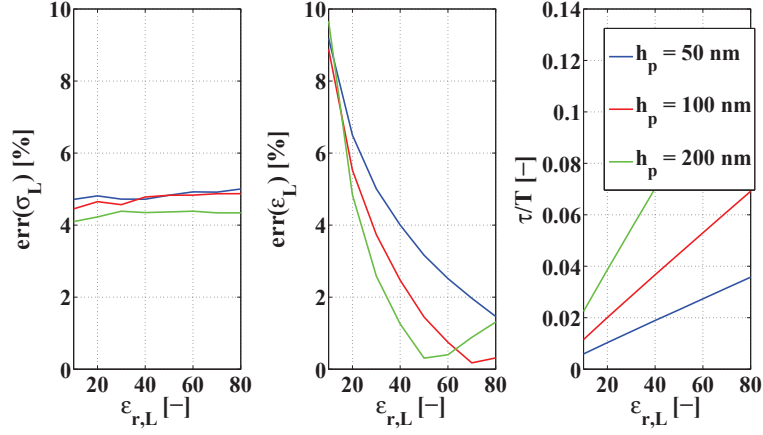


Figure 2.53: Liquid parameter estimation error as a function of the expected permittivity ε_L for several passivation thickness values (tested parameters: $L_c = 20 \mu\text{m}$, $W_f = 10 \mu\text{m}$, $h_f = 3 \mu\text{m}$, $N_f = 200$, $L_f = 150 \mu\text{m}$), $\varepsilon_{r,box} = 3.9$, $\varepsilon_{r,pass} = 8.9$, $\sigma_L = 10^{-5} \text{ S.m}^{-1}$ and $I_0 = 500 \text{ nA}$.

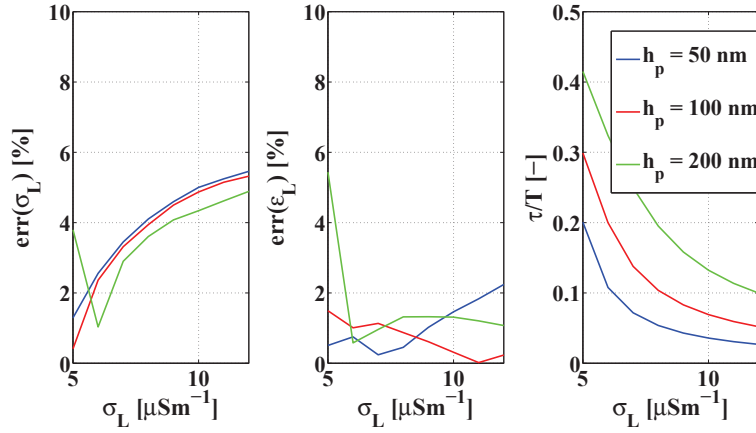


Figure 2.54: Liquid parameter estimation error as a function of the expected conductivity σ_L for several passivation thickness values (tested parameters: $L_c = 20 \mu\text{m}$, $W_f = 10 \mu\text{m}$, $h_f = 3 \mu\text{m}$, $N_f = 200$, $L_f = 150 \mu\text{m}$), $\varepsilon_{r,box} = 3.9$, $\varepsilon_{r,pass} = 8.9$, $\varepsilon_L = 80$ and $I_0 = 500 \text{ nA}$.

2.8.6.2 Unpassivated structures accuracy

The error estimation algorithm is tested for unpassivated sensors for which a serial capacitance C_A has been added to create the triangular shape of the signal. Unlike passivated sensors, the electrical model of unpassivated sensors

is exact. The errors observed in Fig. 2.55 and 2.56 are caused by the same reasons as the passivated sensors. Nevertheless, the exactitude of the electrical model causes those errors to be very small (i.e. smaller than 1%).

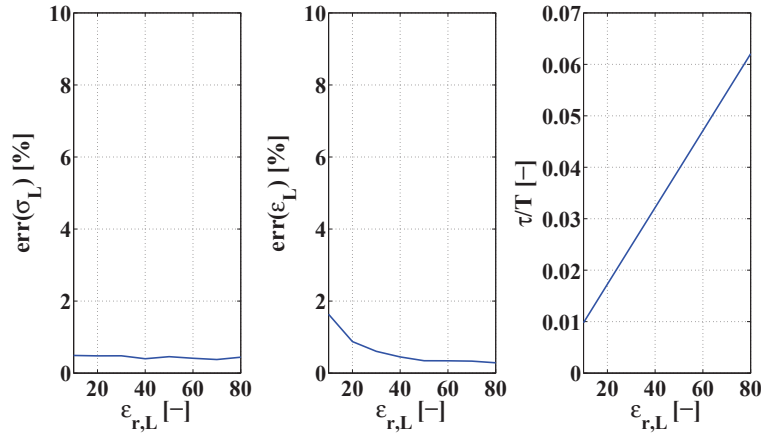


Figure 2.55: Liquid parameter estimation error as a function of the expected permittivity ε_L (tested parameters: $L_c = 20 \mu\text{m}$, $W_f = 10 \mu\text{m}$, $h_f = 3 \mu\text{m}$, $N_f = 200$, $L_f = 150 \mu\text{m}$), $\varepsilon_{r,box} = 3.9$, $C_A = 1 \text{ nF}$, $\sigma_L = 10^{-5} \text{ S.m}^{-1}$ and $I_0 = 500 \text{ nA}$).

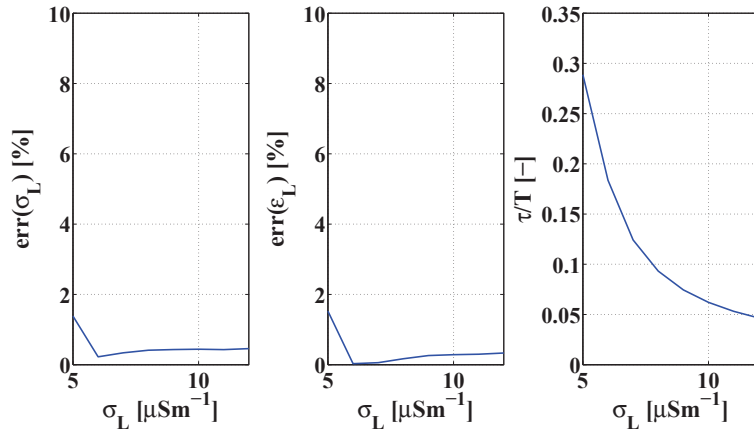


Figure 2.56: Liquid parameter estimation error as a function of the expected conductivity σ_L (tested parameters: $L_c = 20 \mu\text{m}$, $W_f = 10 \mu\text{m}$, $h_f = 3 \mu\text{m}$, $N_f = 200$, $L_f = 150 \mu\text{m}$), $\varepsilon_{r,box} = 3.9$, $C_A = 1 \text{ nF}$, $\varepsilon_L = 80$ and $I_0 = 500 \text{ nA}$).

2.8.7 Global contributions of the CMOS integration

The contribution of the CMOS circuit integration brings several major advantages for the workings of the liquid sensing system. We reported all of the system performances in Tab. 2.14 for the discrete and the CMOS implementations. We conclude that CMOS integration offered strong performances compared to the discrete implementation. This is actually not a problem since the detectable resistance range can be easily adjusted by the injection current I_0 and the sensor dimensions. The major contribution of CMOS integration is the considerable reduction of the switching clock feedthrough and charge injection phenomena; this allows for quantifying small capacitances with high accuracy and therefore highly releases the design constraints of the sensors. The power consumption of the CMOS circuit has obviously been decreased. For the present system, the total power consumption should never exceed $30 \mu W$ while considering the co-integration of power supply references and digital units designed using ULP (Ultra Low Power) techniques [60].

	Discrete circuit	CMOS circuit
Power consumption	$> 10 \text{ mW}$	$< 30 \mu W$
Circuit area	$> 30 \text{ cm}^2$	$< 10 \text{ mm}^2$
Capacitance resolution	$\approx 300 \text{ pF}$	$\approx 120 \text{ fF}$
Resistance resolution	$\approx 300 \Omega$	$\approx 100 \Omega$

Table 2.14: Comparison between the CMOS integration and discrete implementation.

Considering the other IDE on-chip interface research works, we can conclude, from the results reported in Tab. 2.15, that our system clearly brought performances in terms of sensing functionalities and resolutions.

	[45]	[34]	[39]	this work
Detection ability	Capacitive	Capacitive	R-C	R-C
Capacitance resolution	$> 10 \text{ pF}$	$\approx 1 \text{ pF}$	$\approx 1 \text{ nF}$	120 fF
Resistance resolution	—	—	$> 10 \text{ k}\Omega$	$\approx 100 \Omega$

Table 2.15: Comparison between the proposed system and other research works.

2.9 Conclusions

The present chapter was the first application of the sensing system design methodology described in the introduction of the thesis. Throughout the chapter, several important and innovative results have been obtained proving that the sensing system has reached the expected readiness level of technology.

The first point was about the electrical stimulation of interdigitated structures. Generic model impedance has been established for passivated and unpassivated structures, allowing for defining an original interfacing principle which can be applied for all kinds of IDE sensors. Being based on the triangular wave voltage (TWV) generation, the stimulation of the sensor with a square-current signal results in an output triangular voltage of variable shape. The mathematical analysis of the voltage shape proved the possibility of extracting both the fluid under test resistance and capacitance.

The prototyping of the sensing system on a PCB platform validated the interfacing principle. The proposed architecture of the circuit interface is simple and only required a few components to be implemented. Despite the important parasitic impacts of the discrete elements (i.e. circuit node capacitances and charge injection effects), we obtained excellent results in terms of saline liquid measurements and permittivity of mixtures quantification.

As suggested by the design methodology of instrumentation systems, the improvement of the sensing performances has been demonstrated by the circuit CMOS integration. The TSMC 250 nm technology was proved to be efficient in terms of system power consumption ($< 30 \mu W$) and accuracy on the extracted results. We demonstrated that the measurement errors are very small; we obtained less than 5 % and 0.5 % respectively for the passivated and the unpassivated IDE devices. One of the most important contributions of the CMOS integration is the high attenuation of the clock feedthrough and charge injection effects on the analog switch of the system. This allows for the possibility of measuring capacitances down to 120 fF.

In order to further improve sensing performances, the most interesting perspective of this work is to design and optimize all the other blocks of the CMOS interface (power supply references and analog buffers). This will result in a full-integrated system-on-chip of which the area does not exceed 1 cm².

CHAPTER 3

Sensing system of varying magnetic field

3.1 Introduction

Electrical power installations are present in all fields of the human activity; industry, mobility and domotic domains all require the monitoring of electric energy supply networks. In every power installation, several kilometers of wire are needed to supply pieces of equipment and the probability of encountering troubles is proportional to the size and voltage level of the installation. Bad wire insulation or defective welding may generate electrical arcs that considerably damage equipment or cause fire in the worst cases. Fortunately, these kinds of problems can be prevented by monitoring the current carried by the power conductors thanks to the Ampere's and Biot-Savard's laws. These state that a magnetic field is generated by a current conductor according to equation (3.1) and Fig. 3.1; where I_w is the current carried by the conductor, μ_0 is the magnetic permeability of the vacuum, and R the distance from the conductor where the magnetic field is probed.

$$B = \frac{\mu_0 I_w}{2\pi R} \quad (3.1)$$

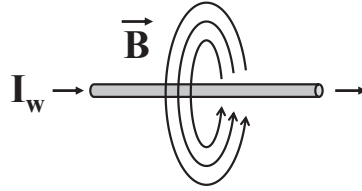


Figure 3.1: Magnetic field generated by a conducting wire.

This chapter proposes a novel approach to characterize and analyze the mag-

netic field created by a power line in order to detect critical variations which may be caused by possible problems on the power line. The proposed system is based on inductive detection and comprises a circuit interface of which the goal is to deliver the sensed data into an easy interpreted output signal. An innovative interfacing methodology is therefore proposed to detect and quantify the variation of the magnetic field caught by an inductive sensing element.

Starting from the well-known physical laws of Lorentz and Faraday to detect a magnetic field, a literature review first presents the major and most recent results of magnetic sensor design. A classification of research works discriminates sensing solutions following their physical structure and governing law.

The detection principle is next described; the sensing element is represented by an equivalent electrical model in which all the physical parameters are calculated using a finite element modeling (FEM) approach. The influence of the geometrical parameters is studied in terms of output voltage amplitude and frequency bandwidth. A sizing methodology is next established according to a formula validated by the FEM results.

The circuit interface of the sensor consists in a current-controlled saw-tooth oscillator; the output voltage shape is deformed by the sensed data which modulates the control current of the circuit. The transfer functions of the interface are next mathematically derived, building the links between the sensed magnetic flux and the circuit output voltage.

An implementation of the system on a printed circuit board (PCB) is therefore proposed using *off-the-shelf* components. The system is measured to validate the theoretical assumptions of the other sections and is tested in the framework of a specific power monitoring application: arc fault detection (AFD).

The last part of the chapter covers sensor interface integration on the mixed-signal TSMC 250 nm CMOS technology. The simulation results show that the CMOS integration of the interface considerably improves the sensing system in terms of compactness and power consumption.

3.2 Magnetic sensors review

In every magnetic sensing system, the governing physical principle is based on one of these two fundamental laws. The first is the Lorentz's law (3.2) where $\mathbf{F}$ is the force generated by the presence of a moving charge $q\mathbf{v}$ immersed in a magnetic field $\mathbf{B}$. The second is the Faraday's law (3.3) where $\mathbf{E}$ is the electric field generated by the presence of a time-varying magnetic field $d\mathbf{B}/dt$.

$$\mathbf{F} = q\mathbf{v} \times \mathbf{B} \quad (3.2)$$

$$\nabla \times \mathbf{E} = -\frac{d\mathbf{B}}{dt} \quad (3.3)$$

The design of magnetic sensing devices is a research area on which a large number of scientific publications have been written. All those solutions have their own governing principle and fabrication method. The physical output data delivered by a sensor can be of various physical natures (voltage, displacement, capacitance, etc.). The present review classifies the magnetic sensing solutions by their fabrication method (i.e. integrated or not) and governing principle (i.e. Lorentz or Faraday). Because of the large number of research publications, the following analysis highlights the most interesting solutions in terms of transduction principle.

3.2.1 Integrated MEMS sensors

As noted in the review of chapter 2, the integration of a device means that its miniaturization uses clean room processes. Unlike the interdigitated impedance sensors which are relatively easy to integrate, the fabrication techniques of magnetic microsensors are extremely varied starting from classical CMOS to complex MEMS etching processes. The acronym MEMS is based on the mechanical motion aspect of the structures and also on the micro-scale device fabrication technology. The most common MEMS devices used for the magnetic detection is a 3D structure of which the response is a displacement generated by the Lorentz force. Nevertheless, the MEMS fabrication technologies can be used to fabricate microcoils for a Faraday-based detection; the main reason of the MEMS technology usage lies in the improvement of the quality factor of the devices. As the micromechanical systems are fabricated using complex and custom processes, the solutions exposed in the literature are based on research activities meaning that no industrial development is considered in these works.

3.2.1.1 Lorentz-based MEMS sensors

The sensing devices based on the Lorentz force consist of three dimensional beams [61, 62], cantilevers [63, 68], or membranes [64] (cfr. Fig. 3.2 and 3.3). These kinds of structures need to be actuated by an electrical current to create the mechanical force as expressed by the Lorentz principle. The moving element displacement can be amplified by enforcing the bias current to the resonant frequency of the structure [65]. The sensor response can be transformed into an electric signal by piezoresistances deformation [66] or capacitive detection [67]. The major limitation notified for the MEMS structures is the need of a high actuation current; amounts of more than 10 mA are required to obtain an acceptable sensitivity of detection. The illustrations of Fig. 3.2 and Fig. 3.3 represent respectively the fabrication results of a cantilever beam and xylophone bar magnetometer.

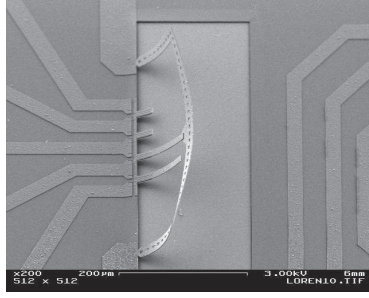


Figure 3.2: Lorentz magnetic field sensing cantilever fabricated on the UCL 2 μm technology [63].

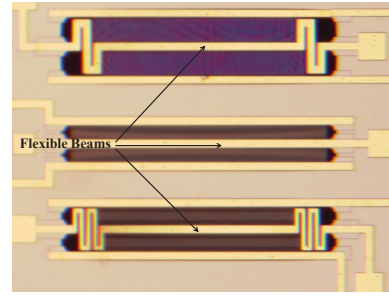


Figure 3.3: Interdigitated capacitive transducer used as magnetic sensor based on the Lorentz force [65].

3.2.1.2 Faraday-based MEMS sensors

The sensing devices using the Faraday's principle are mainly the electrical inductances (cfr. Fig. 3.4 and 3.5). Those devices can be of several geometrical configurations starting from the macro-sized solenoid to the on-chip spiral micro-coil. A large number of applications require the presence of inductors: the RF data transmission [69], the design of CMOS filters [70], or the magnetic sensing systems [71]. The fabrication methodology of inductors is directly related to the size of the device. The micro-sized structures require clean-room processes of which the complexity depends on the used materials, geometry, quality factor, presence of a magnetic core, and targeted applications. Research publications present several complex processes of microcoil fabrication consisting of flat spiral [72] or solenoidal coils [73] with or without an integrated magnetic core [74].

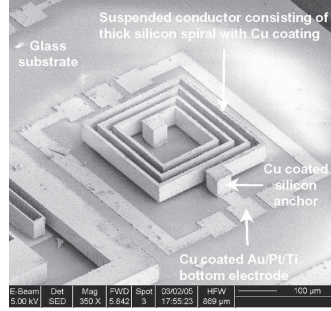


Figure 3.4: Suspended spiral inductor used for RF applications [69].

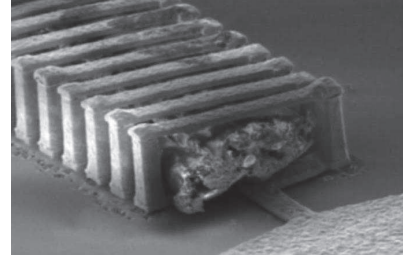


Figure 3.5: Solenoidal micro-coil used in a LC tank resonator [74].

Another kind of magnetic sensing device working under the Faraday's law is the fluxgate sensor which is a combination of magnetic materials with a pair of inductors. The miniaturization of such devices is relatively complex and was investigated in a few research works. The integration of a micro-fluxgate has been achieved in [75] and [76]; the results showed excellent sensitivities of detection ($> 400 \text{ V/T}$) because of the presence of co-integrated CMOS amplification. The weakness of these solutions is the power consumption of such systems, reaching values of several dozens of milliwatts. In addition, one of the fluxgate inductors has to be actuated with a high current ($> 100 \text{ mA}$) to set an efficient system sensitivity.

3.2.2 Integrated solid-state sensors

The category named *solid-state* regroups all of the sensors which are built using solid materials to allow the charge carriers motion; the detection phenomena are therefore purely electrical. Since their design flow does not require any custom MEMS processes, they can be much more easily fabricated and industrialized.

3.2.2.1 Lorentz-based solid-state sensors

The action of the Lorentz principle on solid-state structures is better known as the Hall Effect. The magnetic force deflects the trajectory of the electrons and the holes in a semiconductor device, generating a transverse electric field (cfr. Fig. 3.6). The Hall Effect devices present the advantage of fabrication simplicity, robustness, and high frequency bandwidth in comparison with MEMS motion devices. The basic integration on a classical CMOS technology is therefore trivial. Nevertheless, sensitivity optimization of Hall integrated devices

sometimes requires more complex processes; the Lorentz-based sensors can be of different natures: simple silicon tracks [77], split-drain MOSFET devices [78] (cfr. Fig. 3.7), magnetoresistances [79], or III-V semiconductor devices [80]. Despite their easy fabrication and high bandwidth, Hall sensing device sensitivities never exceed a few Volts per Tesla without any amplification stages.

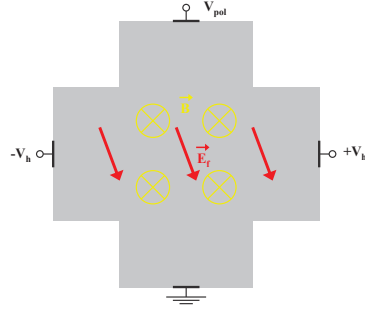


Figure 3.6: Hall effect device consisting of a polarized plate; the sensed magnetic field is transduced into a transversal voltage ΔV_h .

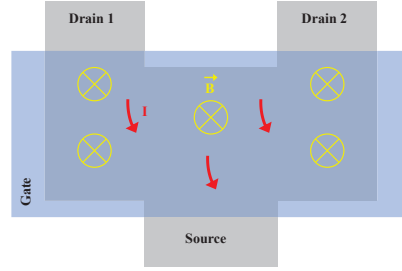


Figure 3.7: Split-drain MOSFET transistor; the current is deflected proportionally to the sensed magnetic field.

3.2.2.2 Faraday-based solid-state sensors

Some specific RF applications use inductors fabricated using solid-state processes meaning that the inductive device consists of classical metal tracks on a silicon substrate. Depending on the technology, the sensor can be stacked according to the different metal levels that the technology provides. This improves the net inductance of the device using a minimum of fabrication area. The quality factor of those devices remains however poor in comparison with the MEMS inductors [81].

3.2.3 Macro-scale sensors

As reported before, the integration of magnetic devices on microscale technologies is indeed advantageous for Hall effect sensors or MEMS motion sensors. For Faraday-based devices (i.e. inductors and fluxgates), the integration presents three weaknesses: the difficulty of magnetic material integration, the technology costs, and the low quality factor of inductive devices. In a large field of applications, the design of macro-scaled sensors is consequently a more judicious choice according to the low price of the technology (like a classic Printed Circuit Board technology) and the possibility of fabricating high quality factor devices. The literature presents a large number of macro-scaled sensing

solutions; as examples, we denote the Rogowski coils (Fig. 3.8) or the fluxgate systems (Fig. 3.9).

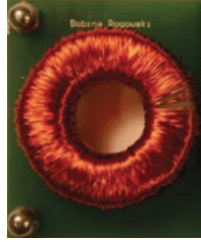


Figure 3.8: Macro-sized Rogowski coil [82].

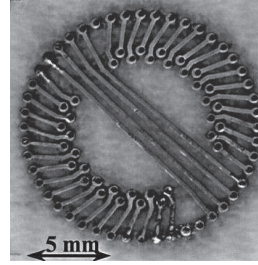


Figure 3.9: PCB printed fluxgate sensor [83].

3.2.4 Work position

Similar to the IDE system developed in the previous chapter, the research summary presented in Tab. 3.1 allows for giving our work position in terms of sensing structure, circuit interface, sensing models, system area, and power consumption. Several structures of various sizes and detection principles are reported. The first conclusion we can draw from this summary is that the measurement of a magnetic phenomenon requires large amount of power. We prove in this chapter that a well thought out interfacing principle can drastically reduce the power consumption of the whole system while extending the use of inductors as magnetic sensors.

Inductors are devices that can adopt several geometrical configurations; we denote the solenoidal coil, the toroidal coil, and the planar spiral-shaped structures. Depending on a specific application, the inductance can be used as a basic circuit element (for example in analog filters), a magnetic sensor, or a RF antenna. Since our goal is to use inductance as a magnetic sensing device, the choice of its geometry is essential. The inductance has to catch a maximum of magnetic flux in order to generate an acceptable voltage amount. It also has to be the most compact in terms of volume. Starting from the spiral-shaped geometry, the objective of this chapter is to prove the feasibility of such a structure to be used for the detection of magnetic perturbations generated by the current carried by an electric high power conductor.

The planar geometrical configuration is advantageous because of its two-dimensional aspect and its easy fabrication. The device dimensions are opti-

mized depending on the frequency bandwidth required by a specific application. In our case, the proposed model of the inductance is general; it can be applied for an integrated or a non-integrated fabrication technology. Since our goal is to develop a system to monitor the power lines activity, the working frequency bandwidth will never exceed 1 MHz. The sensor dimensions will therefore be large enough to avoid integration processes.

Ref.	Structure	Interface	Models	Size	Power
[83]	Fluxgate	no	Magnetic sensitivity expression.	$\approx 2 \text{ cm}^2$	$> 300 \text{ mW}$
[71]	LC tank	LC resonant oscillator	Magnetic transducer gain expression.	$\approx 1.5 \text{ mm}^2$	$\approx 73 \text{ mW}$
[75]	Fluxgate	Excitation and readout circuit interfaces.	no	$\approx 1 \text{ mm}^2$	$> 50 \text{ mW}$
[61, 63, 68]	Cantilever	Wheatstone bridge	Mechanical response.	$\approx 0.01 \text{ mm}^2$	$> 10 \text{ mW}$
[64]	Mag. Res.	no	Magneto-resistance model based on Hall effect.	$\approx 10 \text{ } \mu\text{m}^2$	$> 10 \text{ mW}$
[82]	Rogowski coil	Discrete adaptive filter	Sensor output response.	$\approx 2 \text{ cm}^2$	$> 10 \text{ mW}$
This work	Faraday	Integrated current-controlled oscillator	Sensor electrical response and FEM model. Interface output characteristic definition. Sizing methodology.	$\approx 2 \text{ cm}^2$	$< 50 \text{ } \mu\text{W}$

Table 3.1: Magnetic detection research works summary table and positioning of our work.

3.3 Inductive detection principle

The present section first gives an overview of the inductive sensor in terms of geometrical configuration and highlights all the structure design parameters. An electrical representation of the sensing device is consequently established and allows for describing the detection principle by formal expressions.

3.3.1 Device geometry and fabrication

Among the passive linear electrical components, inductors are the most complex to fabricate because of the three-dimensional aspect of the governing Maxwell equations. Because of the simplicity of their geometry and compactness, the preferred geometry is the planar device configuration. Several parameters describe the device geometry and have their own influence on the sensor response. The structural representation of Fig. 3.10 depicts all the views of the inductance and highlights all the geometrical parameters; the inductor metal is deposited on an insulating box. The description of each geometrical parameter is reported in Tab. 3.2.

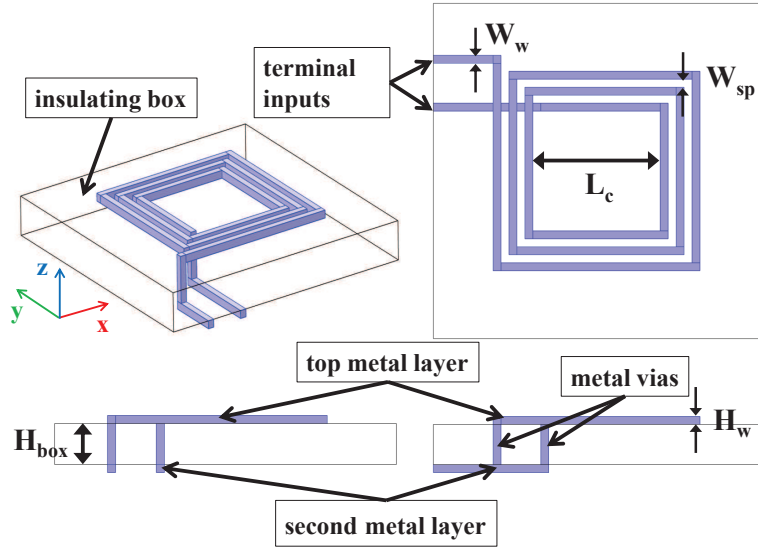


Figure 3.10: Inductive device views.

Description	Notation	Units
Track width	W_w	[m]
Track spacing	W_{sp}	[m]
Track thickness	H_w	[m]
Turn number	N_s	[m]
Inner width	L_c	[m]
Box thickness	H_{box}	[m]
Metal conductivity	σ_m	[Sm ⁻¹]
Box permittivity	$\varepsilon_0 \varepsilon_{box}$	[Fm ⁻¹]

Table 3.2: Spiral inductor geometrical-material parameters.

Planar inductor fabrication requires at least two metal levels in any technology. The insulating box of Fig. 3.10 is the material that separates the two metal levels. The metal pattern which forms the spiral shape is deposited on the top metal level. In order to give access to the internal terminal of the device, metal vias are fabricated through the insulating box making out-of-plane connections. The inductor is therefore connected by metal tracks drawn on the second metal level. The three-dimensional view of Fig. 3.10 shows the global structure of the device fabricated in a two sided metal level technology like a classic printed circuit board (PCB) technology. The difference between printed and integrated inductors is the substrate; the metal levels are buried into oxide layers and fabricated on silicon wafers at smaller sizes.

3.3.2 Inductive sensor electrical model

Despite the complexity of the geometrical configuration of the system, the electrical model is quite simple. It consists of a dipole including the self-inductance L_S of the sensor, a resistive part R_S representing the wire resistance, a component C_P which is the parasitic distributed capacitance between the metal turns and the voltage source component ε_{ind} provided by the external magnetic field variations. As reported in [84], the equivalent schematic model is shown in Fig. 3.11.

The induced voltage ε_{ind} is generated on the device by the variation of the net magnetic flux Φ_e caught by the inductor according to the Faraday's law (3.4). The induction phenomenon is modeled as a voltage source in series with the inductive component of the sensor.

$$\varepsilon_{ind} = -\frac{d\Phi_e}{dt} \quad (3.4)$$

Because of the induced voltage generation phenomenon, the sensor is as-

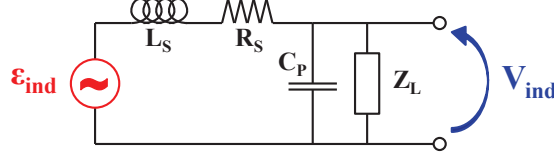


Figure 3.11: Inductive sensor electrical model, the input ε_{ind} and the output V_{ind} are respectively annotated in red and blue.

simulated to active component and can therefore provide electrical power to a specific load Z_L . The sensor response is the voltage drop across the load and is given in frequency domain by equation (3.5). The magnetic signal angular frequency is represented by ω . The induced voltage ε_{ind} is given by (3.6) according to the Faraday's law in frequency domain.

$$V_{ind} = -\frac{j\omega |\Phi_e| Z_L}{R_S + j\omega L_S + j\omega C_P Z_L R_S - \omega^2 C_P Z_L L_S + Z_L} \quad (3.5)$$

$$\varepsilon_{ind} = -j\omega |\Phi_e| \quad (3.6)$$

In order to simplify the response interpretation of the output voltage equation, a pure resistive load is considered and the capacitive parasitics of the device are neglected. The voltage response is then rewritten as (3.7) where the pole of the denominator presents the high-pass filtering action of the device.

$$V_{ind} = \frac{j\omega |\Phi_e| R_L}{R_S + j\omega L_S + R_L} \quad (3.7)$$

The sensor output voltage real and imaginary parts are then isolated into (3.8) and (3.9).

$$\Re(V_{ind}) = \frac{\omega^2 |\Phi_e| R_L L_S}{(R_S + R_L)^2 + \omega^2 L_S^2} \quad (3.8)$$

$$\Im(V_{ind}) = \frac{\omega |\Phi_e| R_L (R_S + R_L)}{(R_S + R_L)^2 + \omega^2 L_S^2} \quad (3.9)$$

3.4 Inductive sensor physical FEM model

The physical model reported here has been developed to evaluate each component of the equivalent electric model of the sensor using the *Comsol* FEM tools. The numerical results provided by the FEM tools are compared with analytical formulas in order to validate their accuracy. Simulations are also performed to analyze the sensor output voltage in the frequency domain.

3.4.1 Finite element system description

Evaluating each circuit component of the sensor electrical model requires the coupled local equations (3.10) and (3.11) to be solved. These two equations, based on the Maxwell-Ampère's law are developed in [85] and reported in appendix D.1. The two variables are respectively the magnetic potential $\mathbf{A}$ and the electric potential V . All the equation parameters are summarized in Tab 3.3.

$$-\nabla \cdot \left[(j\omega\sigma - \omega^2\varepsilon_0\varepsilon_r) \mathbf{A} + (\sigma + j\omega\varepsilon_0\varepsilon_r) \nabla V \right] = 0 \quad (3.10)$$

$$(j\omega\sigma - \omega^2\varepsilon_0\varepsilon_r) \mathbf{A} + \nabla \times \left(\frac{1}{\mu_0\mu_r} \nabla \times \mathbf{A} \right) + (\sigma + j\omega\varepsilon_0\varepsilon_r) \nabla V = 0 \quad (3.11)$$

Description	Notation	Units
Input signal frequency	ω	$[\text{rad.s}^{-1}]$
Material conductivity	σ	$[\text{S.m}^{-1}]$
Material permittivity	$\varepsilon_0\varepsilon_r$	$[\text{F.m}^{-1}]$
Material permeability	$\mu_0\mu_r$	$[\text{H.m}^{-1}]$
Electric potential	V	$[\text{V}]$
Magnetic potential	$\mathbf{A}$	$[\text{T.m}]$

Table 3.3: Equations (3.10) and (3.11) parameters.

The generation of the FEM structure of spiral inductance using **Comsol** has been automated [86]. The finite element algorithm is summarized in Fig. 3.12. It explains the calculation process of the **Comsol** software. The computation of the sensor inductance L_S and resistance R_S is performed by applying an inward current I_{inw} into one of the two terminals; the other one being tied to ground. The calculation of the net induced voltage is performed by the generation of uniform magnetic field perpendicular to the sensor plane. The goal of this technique is to evaluate the inductance global sensitivity.

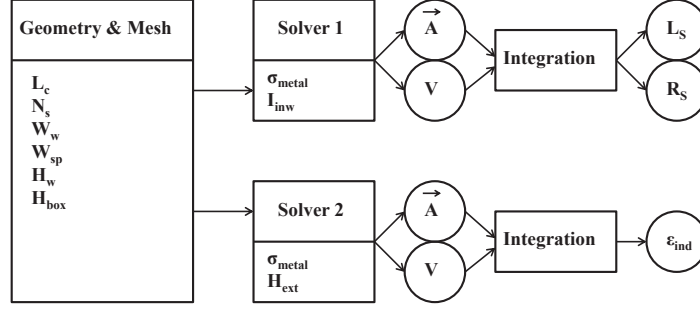


Figure 3.12: Comsol solution computing algorithm.

3.4.2 Simulation results

Simulations are performed to analyze the impact of the device geometric parameters and to show the device frequency properties. Since three-dimensional finite element structures are very memory and time consuming, the tested parameter values are adapted in order to make costly realistic FEM problems for the computing unit memory.

3.4.2.1 Central square length and turns number

The resulting plots in Fig. 3.13 show the evolution of the sensor serial self-inductance and resistance. As the empirical evaluation of planar inductor has been developed in many research works, we have made a comparison between our FEM inductance values and a literature formula called the Modified Wheeler formula [87]; where the mathematical expressions are written in accordance with Tab. 3.2 notations in equations (3.12), (3.13), and (3.14). The good fit of the simulated results with the modified Wheeler formula proves the accuracy of the Finite Elements tools.

$$L_s = 2.34\mu_0 \frac{N_s d_{avg}}{1 + 2.75\rho} \quad (3.12)$$

$$d_{avg} = \frac{2L_c + W_w N_s + W_{sp}(N_s - 1)}{2} \quad (3.13)$$

$$\rho = \frac{W_w N_s + W_{sp}(N_s - 1)}{2L_c + W_w N_s + W_{sp}(N_s - 1)} \quad (3.14)$$

The amplitude of the induced voltage is the key parameter that has to be optimized to insure acceptable performance for a given inductive sensing application. The efficiency of an inductive sensing system depends on the ability

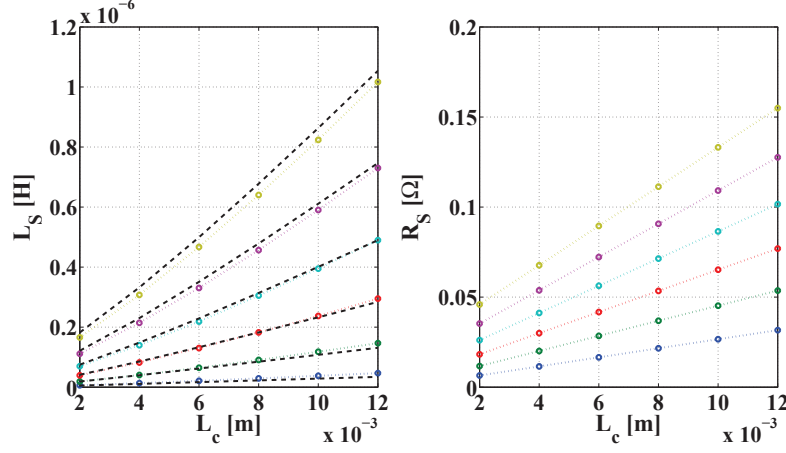


Figure 3.13: Inductive and resistive results for different values of the central square length L_C ; each curve corresponds to a fixed number of turns N_s varying from 1 (blue curve) to 6 (yellow curve); the inductance L_S is compared with the modified Wheeler formula [87] (discontinuous black lines); the track width and spacing are set to $W_w = W_{sp} = 200 \mu\text{m}$; the wire permeability is set to $\mu_r = 1$ (copper); and the box permittivity is set to $\varepsilon_r = 4.1$ (FR4 substrate).

of a given inductor to catch a sufficient magnetic flux and therefore to generate acceptable amplitude of induced voltage. The ability of the device to produce an induced voltage is therefore evaluated by the generation of an uniform external magnetic field according to equation (3.15) perpendicular to the sensor plane; this equation is a boundary condition applied for each outer surface. The H_0 vector z-component is enforced to μ_0^{-1} for the lateral boundary surfaces; the cross product $\mathbf{n} \times \mathbf{H}$ vector components are equal to zero for the upper and lower surfaces. This configuration allows for creating a perfect uniform field of 1 T in the z direction.

$$\mathbf{n} \times \mathbf{H} = \mathbf{n} \times \mathbf{H}_0 \quad (3.15)$$

The induced voltage amplitude is linearly dependent on the amount and frequency of the magnetic flux. We express therefore the induced voltage in Volts per Tesla and per Hertz ($\text{V} \cdot \text{T}^{-1} \cdot \text{Hz}^{-1}$) in the simulations. As a first approximation, the induced voltage can be computed by equation (3.16) which considers the device as a perfect square; the number of turns N_s multiplies the magnetic flux caught by the sensor. The spiral shape of the device is taken into account in formula (3.17). The simulation results presented in Fig. 3.14 show excellent fit with formula (3.17).

$$|\varepsilon_{ind}| = \omega |\Phi_e| = \omega N_s L_c^2 |\mathbf{B}_z| \quad (3.16)$$

$$|\varepsilon_{ind}| = \omega |\Phi_e| = \omega N_s [L_c + (W_w + W_{sp})N_s]^2 |\mathbf{B}_z| \quad (3.17)$$

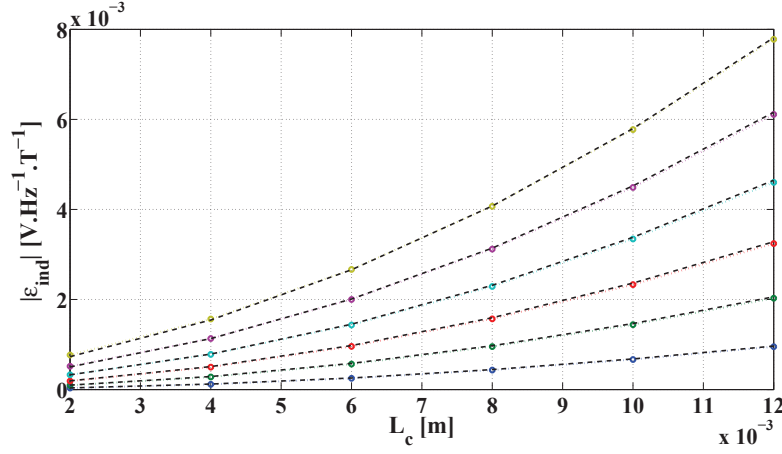


Figure 3.14: Induced voltage magnitude for different values of the central square length L_c ; each curve corresponds to a fixed number of turns N_s varying from 1 (blue curve) to 6 (yellow curve); the simulated results are compared with formula (3.17) (discontinuous lines).

3.4.2.2 Scaling factor

As explained previously, spiral inductors can be fabricated on any technology that includes at least two metal levels. Each technology has its own resolution in terms of metal pattern width and spacing. The adaptation of a specific device from a given technology to another requires evaluating the impact of the technology transfer due to the change of the metal pattern resolution. The technology transfer impact can be evaluated by multiplying all of the dimensions by a scaling factor; the ratio between the metal pattern resolutions of two different technologies. The simulation results depicted in Fig. 3.15 is an example of how the induced voltage magnitude varies for different scaling factors. The lowest technology resolution consists of a 0.02 mm metal pattern width and spacing and the scaling factor is varied from 1 to 10. A perfect quadratic dependence is observed between the induced voltage magnitude and the scaling factor.

3.4.2.3 High-frequency device simulation

Since the device is sensitive to a varying magnetic field, simulations in frequency domain are necessary to analyze its response evolution. The first

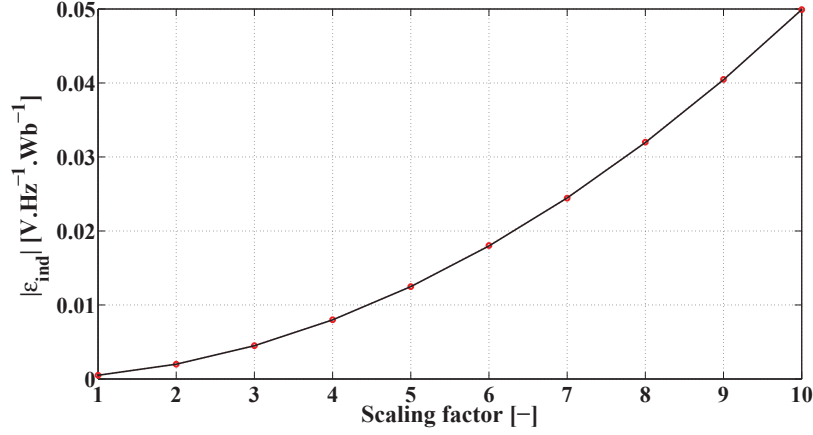


Figure 3.15: Induced voltage magnitude for different values of the scaling factor, the lowest technology resolution is a 0.02 mm metal pattern width and spacing; the other parameters are set to $N_s = 3$ and $L_c = 4$ mm; the red plots are the FEM results and the black line is a quadratic fitting curve.

observation of the frequency impact is the skin effect increase on the device. As the simulated case in Fig. 3.16 is performed for a 200 μm copper wire width and thickness, the skin effect appears between 100 kHz and 1 MHz as predicted by skin depth calculation formula (3.18) where the skin depth δ presents a value of 66 μm at 1 MHz in copper. Beyond approximately 5 MHz, we observe an alteration in the progression of the resistive and inductive parameters of the device. The source of this error lies in the mesh quality; the higher the frequency, the thinner the required mesh to be in accordance with the skin depth effect.

$$\delta = \sqrt{\frac{2}{\omega \sigma \mu_r \mu_0}} \quad (3.18)$$

As the sensor response given by equation (3.7) depends on the load impedance, a pole located in high frequency is observed. For ideal sensing devices, the response is equal to the net induced voltage ε_{ind} with a purely imaginary complex value. The presence of the device self-inductance and its parasitic resistance cause the real part of the response to have its own influence so that the smaller the load resistance, the more impact the real part of the response has. Results of simulations shown in Fig. 3.17 illustrate the impact of several load resistance values on the sensor response V_{ind} in the frequency domain.

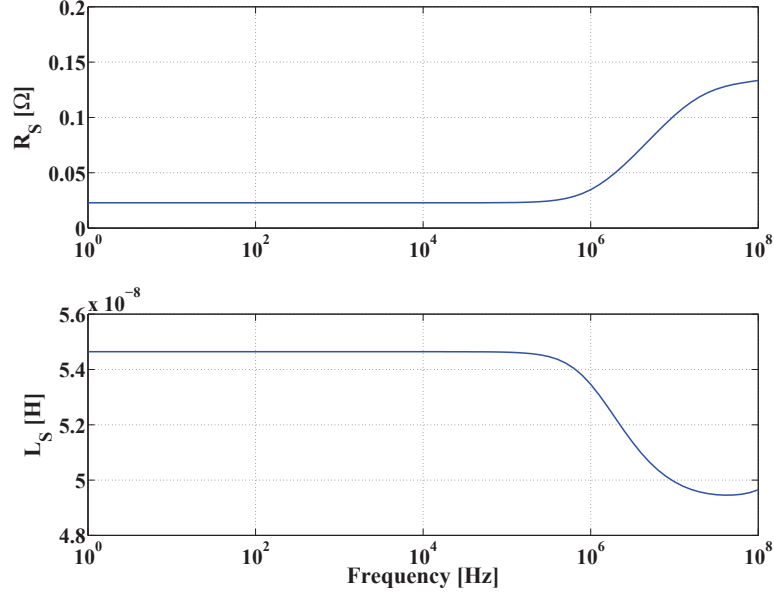


Figure 3.16: Evolution of the resistive and inductive parts of the sensor in the frequency domain.

3.4.2.4 Wire thickness

The impact of the wire thickness on the sensor inductance value can be neglected [87]. The amount of magnetic flux caught by the device remains indeed unchanged. Since the inductor wire section is proportional to the wire thickness, the serial resistance presents a linear dependence with this parameter according to the formula (3.19), l_{ap} being the approximate length of the inductor track.

$$R_S = \frac{l_{av}}{\sigma_m W_w H_w} \quad (3.19)$$

The quality factor of the device (3.20) is therefore affected by the resistance variation; the real part of the sensor impedance is the serial resistive part of the device. This is why planar inductors are manufactured with thick metal layers [69].

$$Q = \frac{\omega L_S}{R_S} \quad (3.20)$$

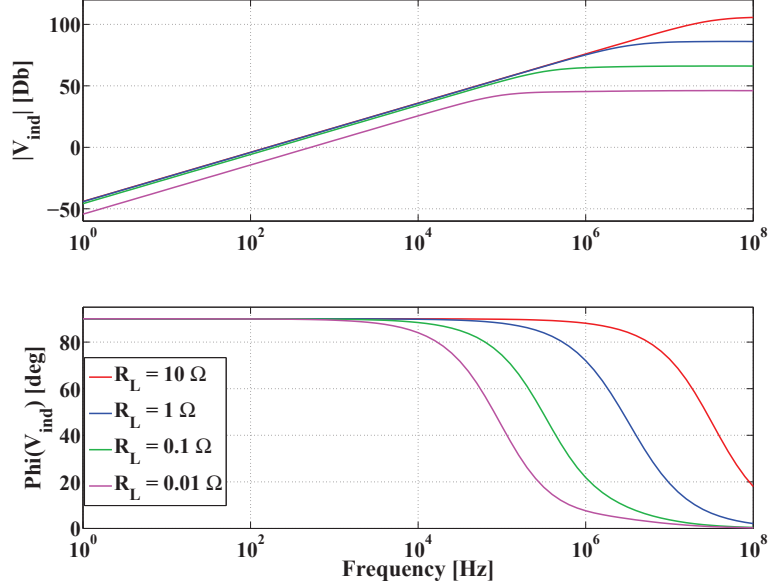


Figure 3.17: Sensor voltage frequency response for several values of the load resistance R_L .

3.4.2.5 Wire turns spacing

Wire separation distance acts on the capacitive component of the device. The electrical model of the sensor has been simulated neglecting the parasitic capacitance of the device. The response voltage equation (3.7) is consequently valid for frequency bandwidths which do not reach the RF domain ($f < 100$ MHz). Wave transmission applications can therefore only be envisaged considering the capacitive component of the inductive device in order to predict a realistic frequency working limit.

The parallel capacitance of the electrical model depicted in Fig. 3.11 can be calculated by solving the finite element model ignoring magnetic effects. The global impedance of the device is then reconstructed combining the FEM calculation of the $R_S - L_S$ pair, the skin effect described by (3.18) and the C_P calculation which is performed by the cancellation of the magnetic effects in the FEM structure of the device. The result of Fig. 3.18 shows the progress of the quality factor considering its general expression (3.21), Z_L being the global impedance of the inductive device without any induced voltage generation.

$$Q = \frac{\Im(Z_L)}{\Re(Z_L)} \quad (3.21)$$

A maximum value of the quality factor appears at the resonant frequency of the device. It means that the capacitive component cannot be neglected beyond this frequency; this has to be considered especially in the case of RF applications. Since the sensing context of the present chapter does not reach the RF domain, the turn spacing W_{sp} influence is negligible. This parameter will be always considered equal to the turn width W_w .

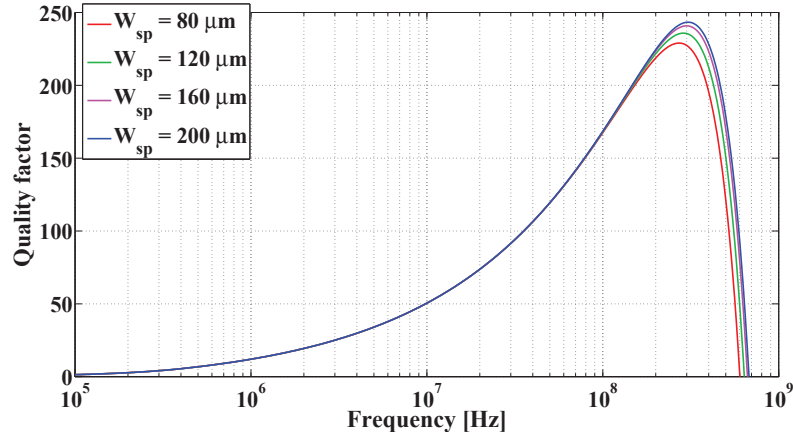


Figure 3.18: Inductance quality factor versus frequency for several values of the turn spacing W_{sp} ($W_w = H_w = 200\mu m$; $L_c = 2mm$; $N_s = 4$).

3.5 Inductor sizing approach

The choice of the sensor dimensions requires analyzing the induced voltage amplitude on a wide range of input sizing parameters. Since the formula (3.17) has been validated by the FEM model, the sensor induced voltage can be calculated without the use of FEM tools which involve a large cost of computing time and memory. The values of the sensor dimensions result in a sizing window that gives all the geometrical parameters which satisfy the constraints and minimum performances of a specific application.

3.5.1 Sensor sizing constraints

Thanks to the fitting formula (3.17) established in the previous section, we are now in position to calculate the sensor sizing laws. Four input constraints must be satisfied to obtain a suitable set of geometrical parameters.

- The critical dimensions of the wire tracks are given by the fabrication technology: $(W_w, W_{sp}) > d_{min}$;
- The effective surface of the sensor cannot exceed a given value depending on a specific application: $A_{eff} < A_{max}$. The objective of the effective surface constraint is to obtain acceptable sensing performances while keeping the device dimensions as small as possible to justify the portativity of our system;
- A minimal induced voltage amplitude is required depending on a specific application: $|\varepsilon_{ind}| > |\varepsilon_{min}|$. This value is chosen according to the amplitude and frequency of the external magnetic field caught by the device;
- A maximal number of turns has to be fixed for a given central square length to avoid oversized devices. This maximum is given by the ratio of the central square length by the track width: $N_s < \frac{L_c}{W_w}$. This constraint is a choice based on the turn number optimization which has been discussed in [88] and [89] in order to improve the device's quality factor. The quality factor value saturates with an increase in the number of turns.

3.5.2 Sensor sizing window

The application of these criteria in the case of a given application fixes a sizing window of allowed geometrical dimensions. We illustrate in Fig. 3.19 the sizing window for two examples of critical technology dimensions. The effective area constraint A_{max} has been respectively set to 160 mm² for the 0.2 mm technology node and 40 mm² for the 0.05 mm one. For both technologies, we

set minimal induced voltage amplitude at $1 \text{ mV} \cdot \text{Hz}^{-1} \cdot \text{T}^{-1}$.

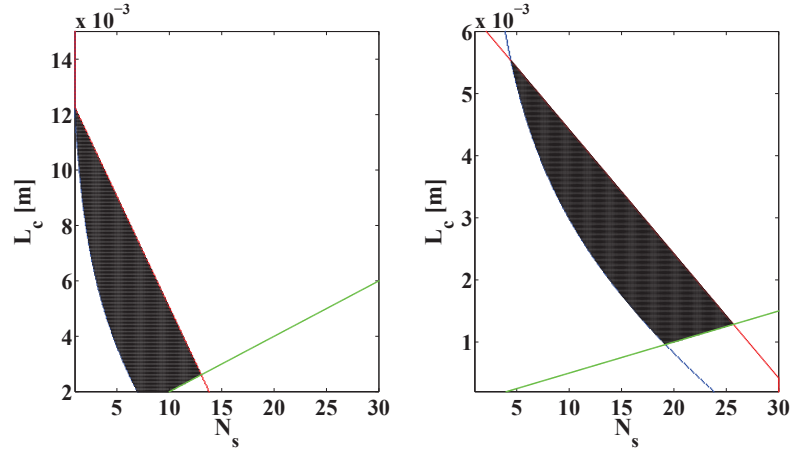


Figure 3.19: Sizing window of allowed dimensions (0.2 mm minimal metal track dimension on the left, 0.05 mm on the right); maximal effective area (red), minimum induced voltage performance (blue), oversizing number of turns limit (green).

We can conclude from the results of Fig. 3.19 that the proposed sizing methodology is the best way to see the impact of a specific technology choice on the sizing window surface. Smaller technologies allows for toughening the effective area constraint while keeping the sensing performance of the device. This observation means that the smaller the technology, the better the device performance. The lower limit of the technology choice in terms of critical dimension is purely economic; the smaller the technology, the higher the price of this technology.

3.6 Circuit interface principle and implementation

The goal of a circuit interface is to transform the information provided by a sensor into a signal from which the sensed data can be extracted and interpreted. The circuit architecture is then chosen to implement the function that transforms the sensed information.

The sensor model developed in the previous section delivers a voltage proportional to the net-magnetic flux variations. The role of the proposed interface is to produce an output voltage proportional to the magnetic flux caught by the inductive sensor. Since magnetic flux is differentiated according to the Faraday's law (3.4), an integration stage is required to compensate the time differential. The approach proposed in this section differs from the architectures of integrators composed of operational amplifiers and integration capacitances (OTA-C) [90]; no operational amplifier is implemented in the present circuit. The interfacing principle is based on a current-controlled oscillator (CCO); the sensor response modulates the circuit oscillating wave by modifying the current controlling the CCO.

3.6.1 Principle of transduction

The physical quantity to evaluate is the net magnetic flux caught by the inductor. The first stage of the circuit interface transforms the sensor voltage into a current via the circuit input impedance. This current is next copied into a capacitance in order to be integrated into an output voltage which contains an image of the sensed magnetic flux. The idea of the system implementation starts from a saw-tooth wave generator system as shown in Fig. 3.20. This principle of transduction is different and innovative in comparison with the classical solutions which consist of amplifying the induced voltage with an operational amplifier [93].

The current source I_L loads the capacitance C_L to create the linear increasing voltage V_{osc} . A Schmitt trigger gate is used to restrain the saw-tooth oscillation into a voltage range defined by the trigger hysteresis. The advantage of such a system is the pure linearity of the triangular signal meaning that the current integration perfectly follows equation (3.22).

$$V_{osc}(t) = \frac{1}{C_L} \int_0^t I_L(t) dt \quad (3.22)$$

The system output signal is depicted in Fig. 3.21; each time the output voltage reaches the upper threshold of the trigger v_+ , it resets the node to

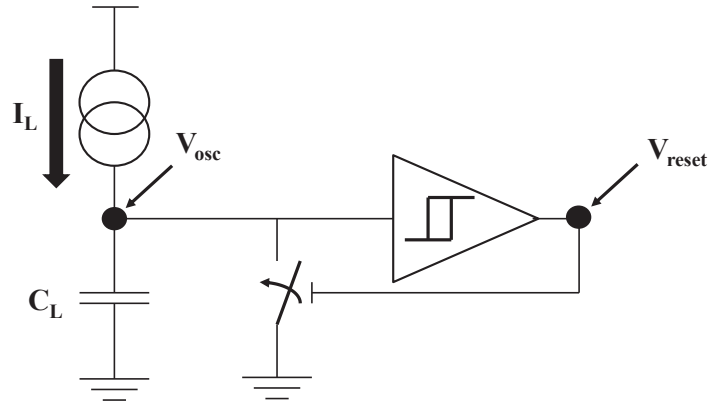


Figure 3.20: Saw-tooth oscillator diagram.

the lower threshold voltage v_- resulting in a signal of period T_0 . The sensor information is conveyed by the current source, meaning that each variation of the current I_L provided by the source is integrated into the capacitance and deforms the saw-tooth shape of the output voltage.

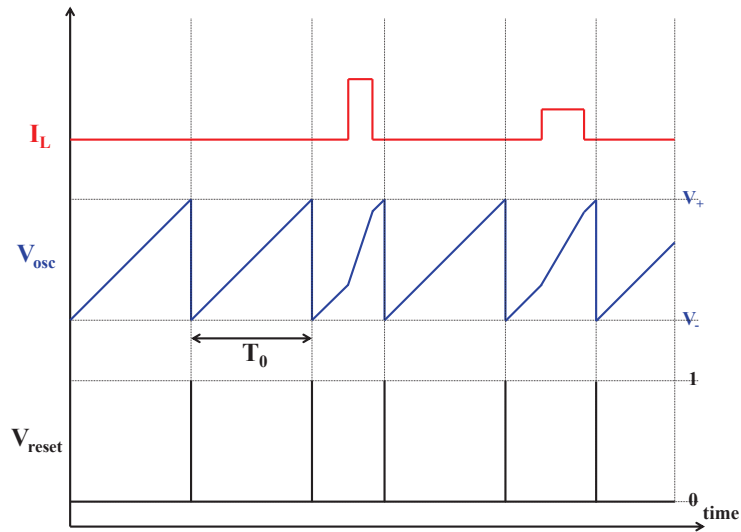


Figure 3.21: Circuit interface signals.

3.6.2 Inductive sensor position

As previously noted, the sensed information is included in the control current I_L of the saw-tooth oscillator. This current is generated by a cascoded current mirror on which the sensing inductance is connected as shown in Fig. 3.22; the two input transistors of the current source operate like two serial diodes. A reference voltage V_{ref} must therefore be chosen to create a DC bias point for the circuit and therefore an initial saw-tooth oscillation period.

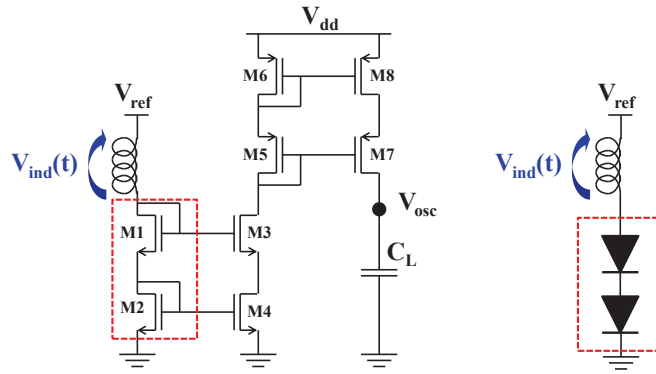


Figure 3.22: Current mirrors architecture and inductive device position.

3.6.3 Interface transfer functions

The presence of an external and time-varying magnetic flux in the sensitive device generates an induced voltage V_{ind} as described in the detection principle section. The circuit interface transfer function is studied following two directions.

The first one, the large-signal mode, is based on the DC current characteristic of the input cascoded MOS diode. The large-signal approach explains how the saw-tooth period of the circuit is altered when a sinusoidal induced voltage of given amplitude and frequency is generated on the inductive device. An output characteristic is then mathematically defined; its value is a function of the external magnetic flux intensity and frequency.

The second one is the small-signal mode and gives the amplification factor of the magnetic perturbations caught by the inductance for a given bias point

of the circuit.

3.6.3.1 Large-signal transfer function

The first step of the large-signal approach is to provide the expression of the saw-tooth voltage initial period noted T_0 . The input diode is biased by a well-defined reference voltage V_{ref} , generating a DC current I_{d0} in the diode. According to the current mirrors, this current is copied into the integrating capacitor C_L . The saw-tooth oscillation period is then fixed by (3.23). The wave amplitude is given by the Schmitt trigger hysteresis voltage V_{hst} .

$$T_0 = \frac{C_L(v_+ - v_-)}{I_{d0}} = \frac{C_L V_{hst}}{I_{d0}} \quad (3.23)$$

If a time-varying induced voltage $V_{ind}(t)$ is generated on the inductive device, a time-dependent current $I_d(t)$ is created in the input diode resulting in a time-dependent output saw-tooth period (3.24).

$$T_{out}(t) = \frac{C_L(v_+ - v_-)}{I_d(t)} = \frac{C_L V_{hst}}{I_d(t)} \quad (3.24)$$

Depending on the reference voltage V_{ref} amplitude, the MOS diode operates in strong or weak inversion. The current expression is therefore given with respect to the operation mode of the input diode. The strong inversion operation current is given by (3.25), due to its quadratic dependence with the reference voltage V_{ref} . Being explained in [91], the weak inversion current presents an exponential dependence with the reference voltage (3.26). These two equations take into account the cascode configuration of the MOS diode, their calculation detail is reported in appendix D.2; n is the substrate factor of the transistors and I_s is the current coefficient given by (3.27).

$$I_{d0,SI} = \frac{\beta}{8n} [V_{ref} - 2V_{th}]^2 \quad (3.25)$$

$$I_{d0,WI} = I_s e^{\frac{V_{ref}}{2nU_t}} \left[1 - e^{\frac{-V_{ref}}{2U_t}} \right] \quad (3.26)$$

$$I_s = \mu_n C_{ox} \frac{W}{L} (n-1) U_t^2 e^{\frac{-V_{th}}{nU_t}} \quad (3.27)$$

The introduction of the time-varying induced voltage $V_{ind}(t)$ into expressions (3.25) and (3.26) leads to the time-varying equations of the diode current in the two operation modes (3.28) and (3.29). The contribution of the induced voltage is simply added to the reference voltage source in relation with the circuit in Fig. 3.22.

$$I_{d,SI}(t) = \frac{\beta}{8n} [V_{ref} + V_{ind}(t) - 2V_{th}]^2 \quad (3.28)$$

$$I_{d,WI}(t) = I_s e^{\frac{V_{ref} + V_{ind}(t)}{2nU_t}} \left[1 - e^{\frac{-V_{ref} - V_{ind}(t)}{2U_t}} \right] \quad (3.29)$$

Supposing now a pure sinusoidal variation of the induced voltage as expressed by (3.30), the diode currents in strong and weak inversion are averaged on one oscillation period $T_e = f_e^{-1}$ of the input sine induced voltage $V_{ind}(t)$ by equation (3.31). The induced voltage amplitude has been replaced according to Faraday's law in order to highlight the sensed magnetic flux amplitude Φ_e .

$$V_{ind}(t) = |V_{ind}| \cos(2\pi f_e t) = 2\pi f_e \Phi_e \cos(2\pi f_e t) \quad (3.30)$$

$$\bar{I}_{d,SI,WI} = \frac{1}{T_e} \int_0^{T_e} I_{d,SI,WI}(t) dt \quad (3.31)$$

Therefore, the oscillation period of the circuit saw-tooth wave in the presence of an averaged magnetic perturbation is rewritten as (3.32).

$$\bar{T}_{out,SI,WI} = \frac{C_L V_{hst}}{\bar{I}_{d,SI,WI}} \quad (3.32)$$

The generic definition of the circuit interface output characteristic is given by (3.33). This equation provides a dimensionless quantity which is equal to zero in the absence of any induced voltage on the circuit input; it expresses the global sensitivity of the system to detect a magnetic perturbation on the inductive device. The objective of the square root present in the Ω_{out} expression is to linearize the quadratic dependence of the diode current with the reference voltage V_{ref} in strong inversion. As demonstrated further by graphical plots, the curvature of the output characteristic Ω_{out} with the induced voltage amplitude $|V_{ind}|$ allows for distinguishing the weak and strong operation modes of the input diode.

$$\Omega_{out} = \sqrt{\left| \frac{T_0}{\bar{T}_{out,SI,WI}} - 1 \right|} \quad (3.33)$$

The time-average of the output characteristics in the strong and weak inversion operations are mathematically demonstrated in appendix D.3 and are respectively given by equations (3.34) and (3.35); $I_\nu(x)$ being the modified Bessel function of the first kind [92].

$$\Omega_{out,SI} = \frac{|V_{ind}|}{\sqrt{2} (V_{ref} - 2V_{th})} \quad (3.34)$$

$$\Omega_{out,WI} = \sqrt{\left| \frac{I_{\nu=0} \left(\frac{|V_{ind}|}{2nU_t} \right) - 1 + e^{\frac{-V_{ref}}{2U_t}} \left[1 - I_{\nu=0} \left(\frac{|V_{ind}(1-n)|}{2nU_t} \right) \right]}{1 - e^{\frac{-V_{ref}}{2U_t}}} \right|} \quad (3.35)$$

The curves plotted in Fig. 3.23 illustrate the graphical progression versus the induced voltage amplitude $|V_{ind}|$ of the output characteristics given by equations (3.34) and (3.35) for several bias voltages V_{ref} of the diode circuit branch. The strong inversion curves increase linearly while the weak inversion curves present an exponential progression. The split voltage point between the two diode operations is given by twice the threshold voltage V_{th} of a NMOS transistor. The separation voltage between the two inversion modes is detailed further in the CMOS integration section.

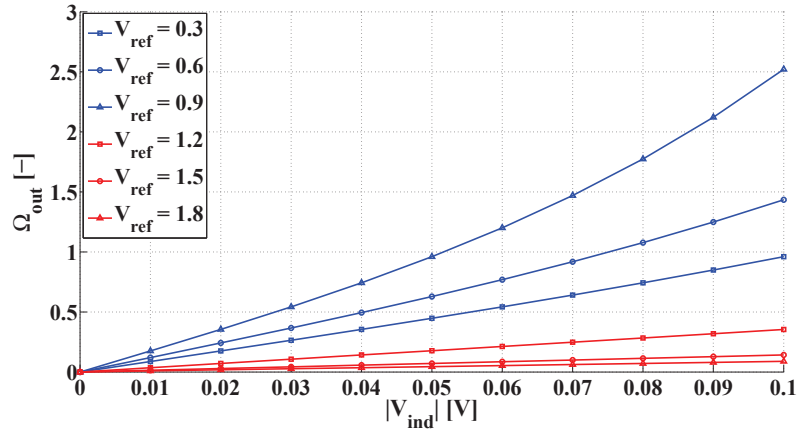


Figure 3.23: Output characteristic Ω_{out} as a function of the induced voltage amplitude for several reference voltages; the strong (equation (3.34)) and weak (equation (3.35)) inversion operations are respectively sketched in red and blue.

The choice of working in weak inversion is advantageous to get more output sensitivity; the linearity of the transfer characteristic is unfortunately lowered. For example, the weak inversion bias point is a more suitable choice for applications based only on the detection of a phenomenon. Moreover, working in strong inversion guarantees a better output linearity and therefore provides the possibility of quantifying the magnetic phenomenon with better accuracy.

3.6.3.2 Small-signal transfer function

The small-signal approach starts with the representation of the analog output voltage V_{osc} by a steady-state term (SS) and a varying term $v_{osc,AC}$ as given by expression (3.36). The SS component, represented with a capital letter, is the natural oscillation of the circuit without any magnetic signal; this state is fixed by the DC bias point of the circuit. The varying component is the small-signal that modulates the saw-tooth shape of the analog output voltage.

$$V_{osc}(t) = V_{osc,SS} + v_{osc,AC} \quad (3.36)$$

The small-signal transfer function is defined by the ratio between the varying component of the output voltage and the magnetic flux signal caught by the device. Supposing a perfect copy of the current in the mirror circuits and neglecting all the parasitic elements of the circuit, the transfer function A_Φ is expressed by (3.37) and is given in Volts per Weber (V.Wb^{-1}), g_{in} being the equivalent small-signal conductance of the two input serial transistors M1 and M2 (Fig. 3.24). The representation shown in Fig. 3.25 is the system small-signal simplified architecture. The transfer function is calculated from this representation.

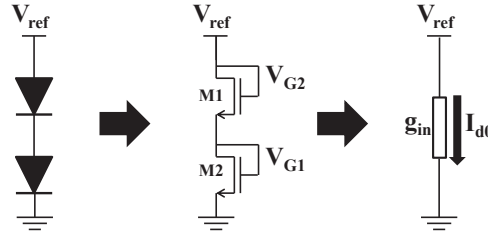


Figure 3.24: CMOS cascoded diode input conductance representation

$$A_\Phi = \frac{v_{osc,AC}}{\Phi_e} = \frac{g_{in}}{C_L} \quad (3.37)$$

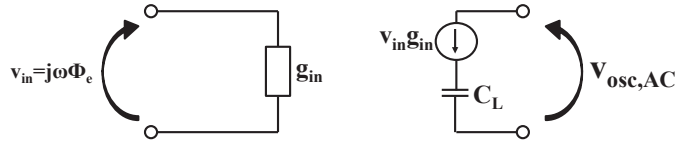


Figure 3.25: Small-signal representation of the circuit interface.

The small-signal input conductance g_{in} is fixed by the DC bias of the circuit. The strong and weak inversion mode of this parameter are respectively given by (3.38) and (3.39), I_{d0} being the DC current flowing through the two

cascoded CMOS diodes when no induced voltage is generated on the sensor; the calculation of these expressions are detailed in the appendix section D.4.

$$g_{in,SI} = \frac{2I_{d0,SI}}{V_{ref} - 2V_{th}} \quad (3.38)$$

$$g_{in,WI} = \frac{I_{d0,WI}}{2nU_t} \left(1 + \frac{ne^{\frac{-V_{ref}}{2U_t}}}{1 - e^{\frac{-V_{ref}}{2U_t}}} \right) \cong \frac{I_{d0,WI}}{2nU_t} \quad (3.39)$$

The CMOS integration section shows simulation results of the small-signal transfer function. The internal nodes of the interfacing circuit are corrupted by parasitic capacitances which limits the frequency bandwidth of the transfer function.

3.7 Measurements and applications

Having described the sensing principle and the sensor interface, the goal of the measurements presented in this section is to validate the theoretical developments exposed in the previous sections. The experimental approach proposed here consists of the reproduction of the sensing system using existing components. The inductive sensor is directly fabricated on the *Multi CB* printed circuit board technology [94].

The first step of the characterization setup is to create an external and time-varying magnetic field. According to the formula (3.1), a straight conductor generates a magnetic field $\mathbf{B}$ directly proportional to the current I_w carried by this conductor, the parameter R being the distance from the conductor to the point where the field is evaluated. The spiral inductor is next placed close to the wire in order to capture the maximum amount of magnetic flux. Measurements are next performed to validate the system global response Ω_{out} described by the demonstration reported in the previous section. The AFD tests of the inductive sensing system are finally presented; their objective in terms of application is to prevent installation from damage that can be caused by electrical arcs.

3.7.1 Measurement setup description

The description of the measurement setup explains first how the sensing device is placed relative to the current line in order to give an approximation of the net magnetic flux caught by the device. This allows for evaluating the coupling between the current line and the device. As the target application of the sensing system is to detect arc fault on power lines, specifications are given in order to find the most suitable dimensions for the sensing device. In this context, those specifications are based on aeronautic data [95]. The discrete implementation of the circuit interface is next presented and each component is described.

3.7.1.1 Sensor and current line coupling

The electrical current carried by a straight conductor wire generates a magnetic field according to equation (3.1); the magnetic field magnitude being proportional to the wire current amplitude I and decreasing with the distance R from the wire. Knowing the circular profile of the magnetic field, the sensing device plane must be parallel to the wire so that the field normal component is maximal on the device plane as depicted in Fig. 3.26 showing the top (a) and the cross views (b) of the system geometrical configuration.

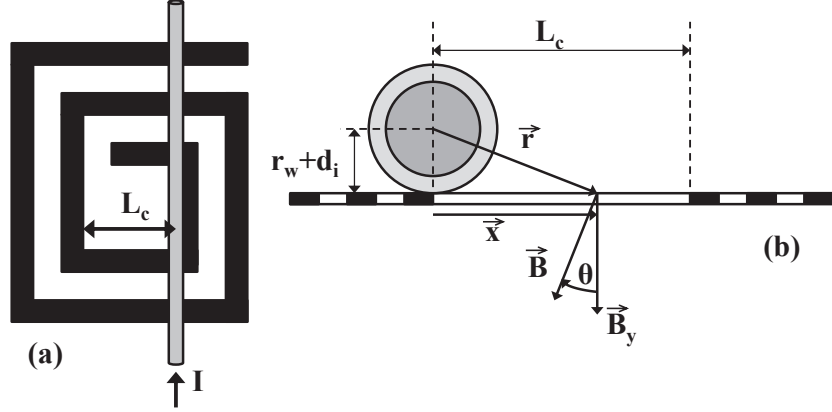


Figure 3.26: Sensing device and current line positions.

An approximation of the net magnetic flux caught by the device surface can be calculated according to the geometric configuration illustrated in Fig. 3.26b resulting in the expression (3.40). This formula, being detailed in Appendix D.5, does not consider the spiral shape of the device. The magnetic coupling between the sensor and the wire is given by the ratio between the magnetic flux caught by the inductor and the current carried by the straight conductor (3.41); this ratio is known as the mutual inductance. The system view depicted in Fig. 3.27 illustrates the electrical representation of the sensor using the concept of the mutual inductance.

$$\Phi_e = \frac{I_w \mu_0 L_C N_s}{4\pi} \ln \left(\frac{L_C^2 + (r_w + d_i)^2}{(r_w + d_i)^2} \right) \quad (3.40)$$

$$M_S = \frac{\Phi_e}{I_w} = \frac{\mu_0 L_C N_s}{4\pi} \ln \left(\frac{L_C^2 + (r_w + d_i)^2}{(r_w + d_i)^2} \right) \quad (3.41)$$

$$|\varepsilon_{ind}| = \omega M_S I_w \quad (3.42)$$

3.7.1.2 Sensor sizing

The sizing methodology developed previously is used to define a set of geometrical dimensions in order to respect the specifications of the current variations sensing application:

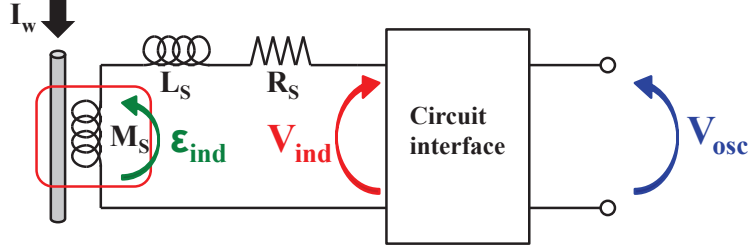


Figure 3.27: System view of the power line current sensing system, the mutual inductance is highlighted by a red frame.

- Minimal induced voltage required: $\varepsilon_{ind} = 1 \text{ mV}$;
- Minimal current detected by the system: $I_w = 500 \text{ mA}$;
- Minimal frequency detected by the system: $f_w = 100 \text{ kHz}$;
- Maximal sensor effective area: $A_{max} = 400 \text{ mm}^2$.

In this context, we ensure a minimal induced voltage value of 10 mV for those specifications on a $200 \mu\text{m}$ printed board technology. The geometric parameters obtained by the application of the sizing methodology are reported in Tab. 3.4. The track thickness H_w is fixed by the PCB technology. From the mutual coupling point of view given by formula (3.42), these specifications lead to a minimum required mutual inductance of $M_s = 31.8 \text{ nH}$.

Sizing parameters	Notations	Values
Track width	W_w	$200 \mu\text{m}$
Track spacing	W_{sp}	$200 \mu\text{m}$
Track thickness	H_w	$30 \mu\text{m}$
Turn number	N_s	10
Inner width	L_c	8 mm

Table 3.4: Inductor dimensions.

The application of the formula (3.41) allows for approximating the mutual inductance according to the parameters of Tab. 3.4 and taking into account the power wire dimensions of $r_w = 0.25 \text{ mm}$ and $d_i = 0.5 \text{ mm}$. This approximation has been measured probing the voltage on the inductive device. The numerical data reported in Tab. 3.5 summarizes the calculated mutual inductance values for the present system.

Minimal required mutual inductance (3.42)	$M_S = 31.8 \text{ nH}$
Calculated mutual inductance (3.41)	$M_S = 37.9 \text{ nH}$
Measured mutual inductance	$M_S = 39.6 \text{ nH}$

Table 3.5: Mutual inductance values.

3.7.1.3 Discrete interface components and supply conditions

The first discrete component is the inductive sensitive device; its design is based on the *Multi CB* industrial technology. The circuit interface prototype is the reproduction of the current mirrors block with discrete MOS transistors. The Schmitt trigger is a classic 4093 inverter gate that provides a well-defined hysteresis voltage. The power supply conditions are established with consideration to the discrete device electrical properties. All the discrete component models and the power supply conditions are summarized in Tab. 3.6. External accesses are provided to probe the saw-tooth voltage generated by the system. A view of the discrete system is depicted in Fig. 3.28.

Polarisation parameters	Notations	Values
Supply voltage	V_{dd}	5 V
Load capacitance	C_L	10 nF
Discrete component type	References	Parameters
N channel MOSfet	<i>BSS138</i> [96]	$V_{th} = 1.2 \text{ V}$
P channel MOSfet	<i>BSS84</i> [97]	$V_{th} = -1.7 \text{ V}$
Schmitt gate	<i>MC14093B</i> [98]	$V_{hst} = 0.7 \text{ V}$

Table 3.6: Discrete component and polarisation setup parameters.

3.7.2 System response characterization results

The objective of these experiments is to observe the system global response Ω_{out} versus the current line frequency and amplitude as described by equations (3.34) and (3.35).

The first test illustrates the evolution of the system response Ω_{out} for several diode reference voltages V_{ref} with the current line frequency by fixing the power current line amplitude to 1 A. The results are shown on the left part in Fig. 3.29 in logarithmic scales. Because of their curvature, the weak inversion operation is observed on the $V_{ref} = 2 \text{ V}$ and $V_{ref} = 2.3 \text{ V}$ characteristics. This second test highlights the influence of the current line amplitude on the system response;

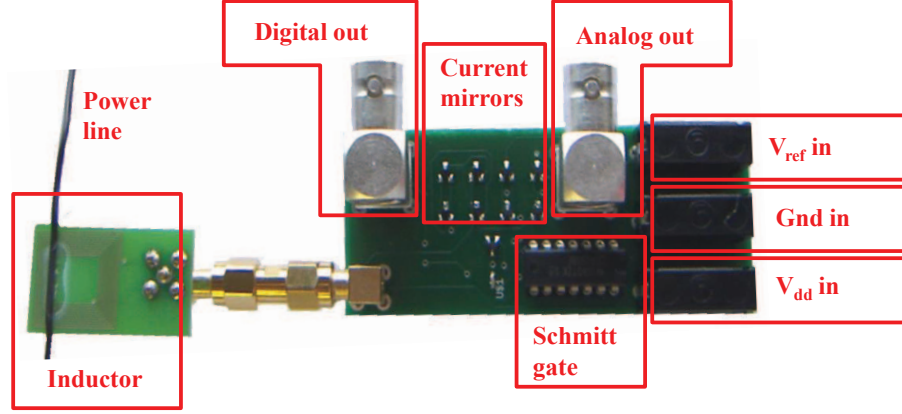


Figure 3.28: Magnetic sensing system PCB prototype picture.

the current line frequency is maintained constant to 100 kHz. As the magnetic field generated by the line is directly proportional to the current line, the results plotted on the right of Fig. 3.29 show the evolution of the system response versus the current line amplitude. As modeled by equation (3.34), response linearity is observed in the strong inversion operation of the input diode while the weak inversion operation results in an exponential curvature of the characteristics.

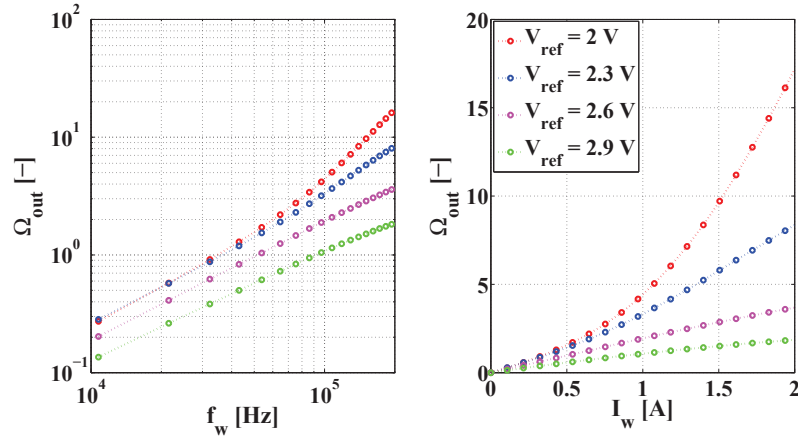


Figure 3.29: Output characteristic Ω_{out} as a function of the current line frequency on the left ($I_w = 2$ A) and current line amplitude on the right ($f_w = 200$ kHz).

3.7.3 Current variations detection for Arc Fault Detection (AFD)

The final measurements presented in this section are the detection of sharp variations occurring on a current line. These variations can be caused by bad electrical contacts such as a defective weld and can therefore create electrical arcs that considerably damage power installations. In most research results, the detection of electrical arcs is performed by analyzing the frequency spectrum of the measured signal [82]. Our detection methodology focuses on the temporal domain of the output signal by the extraction of Ω_{out} .

3.7.3.1 Electrical arc bench test

The generation of electrical arcs is performed on a bench test designed by the *Centre Spatial de Liège*. The principle consists of the abrupt creation of a short circuit with a metallic blade on an operating power line supplying various types of electrical loads; a similar example of arc generator is described in [99]. The schematic representation depicted in Fig. 3.30 gives a brief description of the arc creation bench test where the arc creation can be represented by a switch that abruptly enforces a short-circuit on the current source.

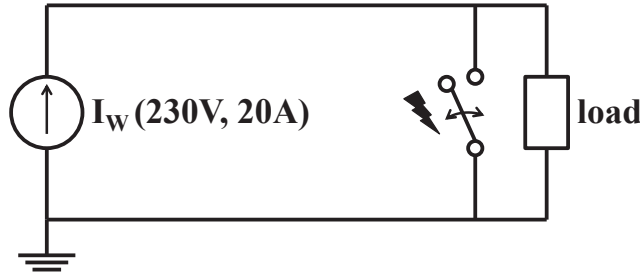


Figure 3.30: Schematic representation of the AFD bench test.

3.7.3.2 Arc impact overview

The experimental results presented in Fig. 3.31 are the first example showing the impact of an electrical arc on the power line current. A reference *Tektronix* Hall Effect probe measures the line current in order to make a preliminary identification of the variations caused by the arcs. Three phenomena are observed on the current plots measured by the *Tektronix* probe: the load supply operating regime, the short circuit regime and the current peak caused by the

electrical arc itself. In this case, the line is loaded by an open circuit, meaning that the supply regime current amplitude is close to zero. The short-circuit regime is established after the arcing peak and ends when the contact formed by the blade of the test bench is interrupted. The view on the right in Fig. 3.31 is a closer zoom on the arc area: the saw-tooth voltage of the inductive system presents a perturbation exactly at the same time as the beginning of the arc.

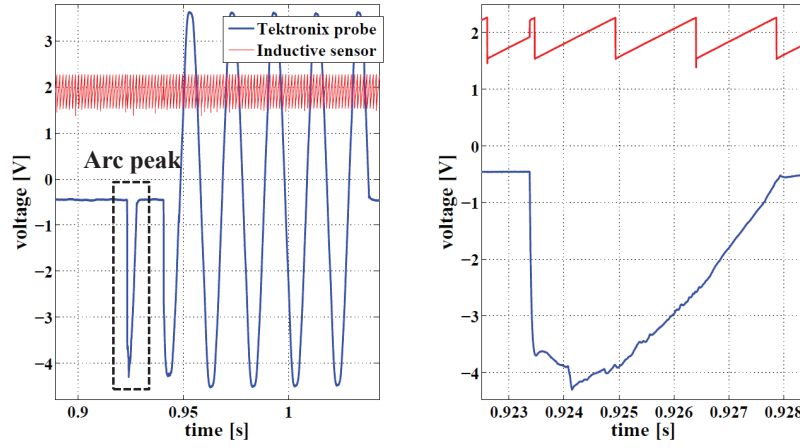


Figure 3.31: Characterization of the power line current loaded by an open circuit.

3.7.3.3 Arc detection based on the large-signal approach

The application of the large-signal approach on AFD measurements offers a global possibility of identifying the impact of the electric arcs on the power line current wave. The principle is to observe each period of the saw-tooth voltage. If a shorter period is detected, the value of the large-signal transfer characteristic Ω_{out} is increased. We also set the reference voltage of the diode to $V_{ref}=1.55$ V in order to work in the weak inversion range; the output sensitivity of the system is therefore increased.

As tested previously, the first result shown in Fig. 3.32 has been performed without any load connected to the power line. The high values of Ω_{out} are clearly observed on the critical variations of the current line. Because of the

high current values during the short circuit operation, we note light variations of Ω_{out} . The end of the short circuit operation is also detected by the system, marking the stopping limit of the impact caused by the electric arc on the power line.

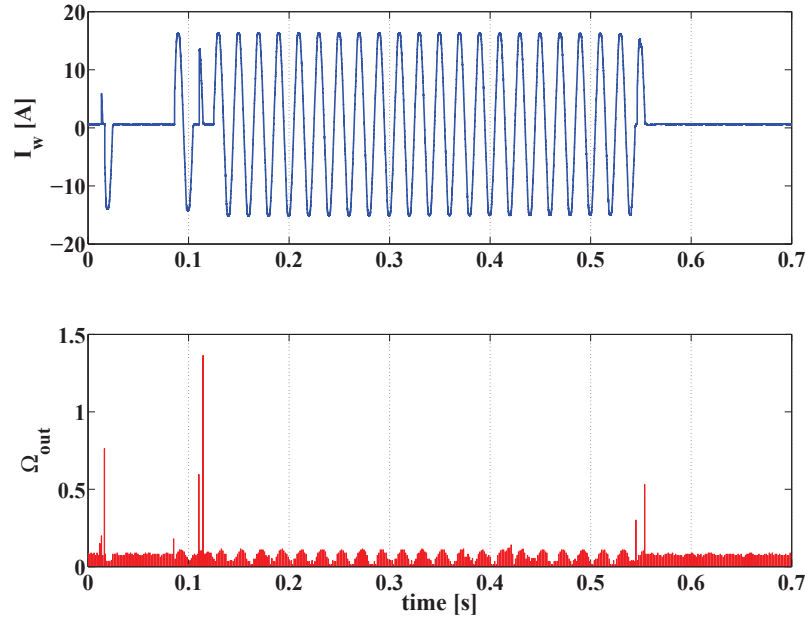


Figure 3.32: Test 1: Ω_{out} characteristic for an open circuit load.

The results in Fig. 3.33 were obtained using a resistive load on the power line. The short-circuit operation caused by the electric arc is shorter than the previous test. The observed characteristic Ω_{out} is increased on the critical current variations and therefore proves the arc detection.

A motor combined with a chopper amplifier has been used as an electrical load for the test exposed in Fig. 3.34. In this case, the nominal operation due to the presence of the motor shows a more complex current wave shape. The output characteristic of the system presents a good discrimination between the electric arc impacts and the chopper amplifier activity on the power line. A small oscillation of the output characteristic Ω_{out} is also noted during the short circuit operation on the power line. Though the inductive sensor acts as a high-pass filter, the amplitude of the 50 Hz abnormal short-circuit current becomes sufficiently high to generate this small oscillation on the output characteristic

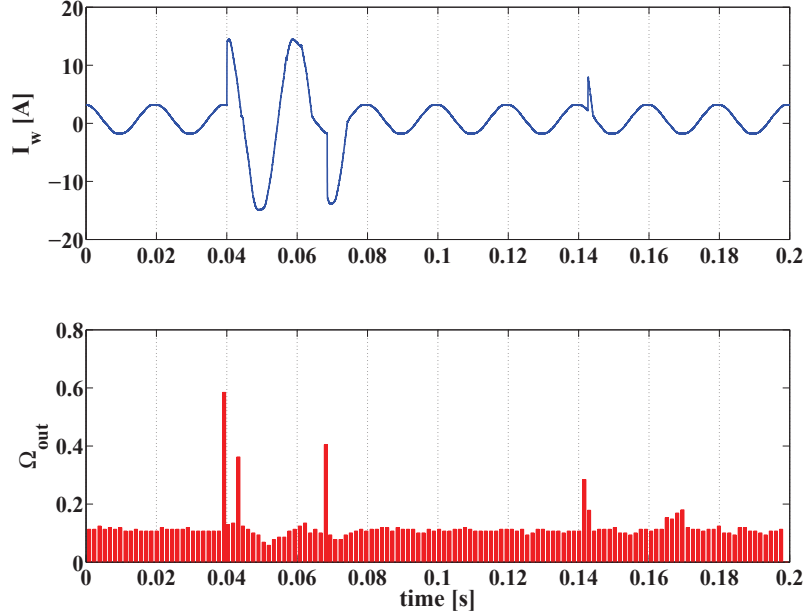


Figure 3.33: Test 2: Ω_{out} characteristic for a resistive load.

of the system. However, this last observation could be useful to identify the short-circuit arc phenomenon with a higher probability.

The last test of Fig. 3.36 is the result of the combination of two loads on the power line: the resistive load of the second test and the motor. As observed in previous tests, sharp current variations are detected with good sensitivity. The impact of the chopper amplifier is ignored by the detection system.

3.7.3.4 Small-signal detection approach

Since the large-signal approach gave the time locations of the critical current variations, the small-signal analysis allows for locally highlighting these variations and to quantify them by voltage amplitude. As described by equation (3.36), these variations are isolated by removing the natural oscillation of the sensing system. The result illustrated in Fig. 3.37 is a zoom on two current peaks caused by an electrical arc. Two steps are observed on the output voltage v_{osc} corresponding to the most abrupt variation of the current peaks.

The amplitude of these voltage steps allows for the quantification of the

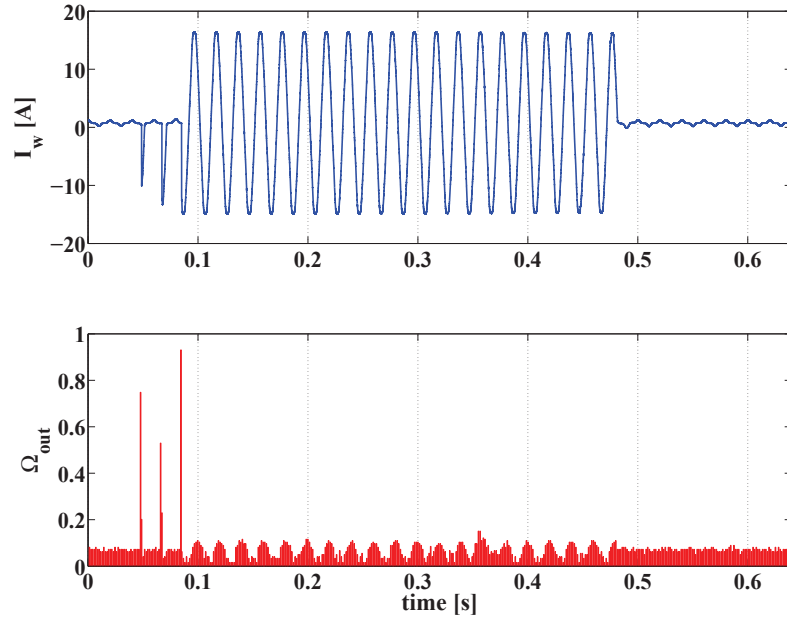


Figure 3.34: Test 3: Ω_{out} characteristic for a motor load.

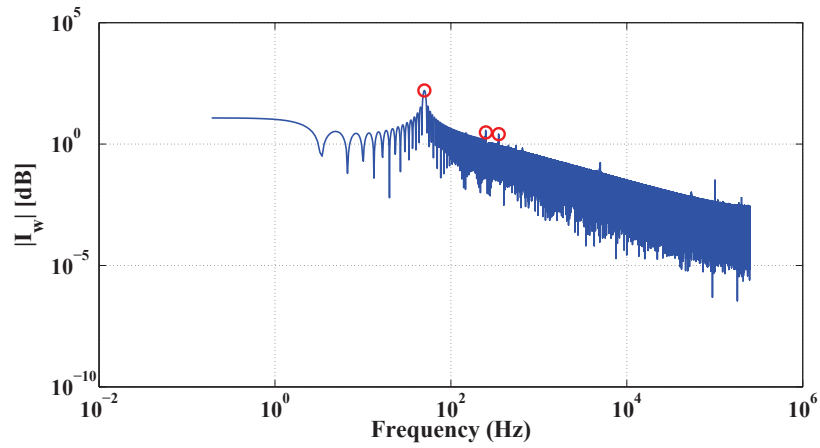


Figure 3.35: Fast Fourier transform of the short circuit current probed by the Tektronix sensor, the red circles highlight the amplified harmonics of the fundamental 50 Hz wave.

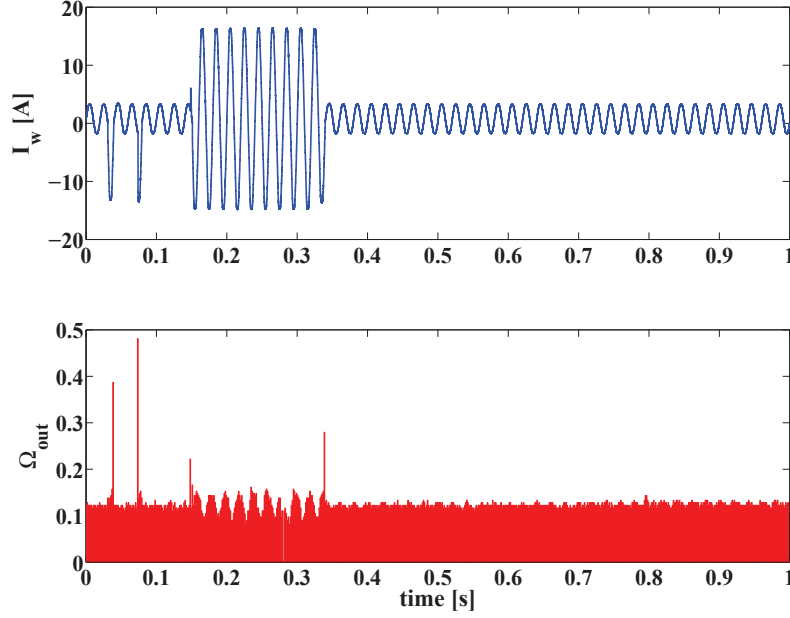


Figure 3.36: Test 4: Ω_{out} characteristic for the combination of a resistive load and a motor load.

arcing current. As the circuit reference voltage is biased at 1.55 V corresponding to a weak inversion current of $5.16 \mu\text{A}$, the input conductance g_{in} of the circuit given by equation (3.39) is approximately equal to 9.10^{-5} S . The equation (3.37) yields thus $A_\Phi = 9000 \text{ V.Wb}^{-1}$. The measured pulse amplitudes in Fig. 3.37 present a value of $v_{osc} = 30 \text{ mV}$. Putting all these quantities in the small-signal modeling equations and taking into account the measured mutual inductance of the power line on the sensor, the arcing current flash is evaluated by the following formula:

$$I_{arc} = \frac{|v_{osc}|}{M_S A_\Phi} = \frac{1}{9000} \frac{30 \text{ mV}}{39.6 \text{ nH}} \cong 85 \text{ A} \quad (3.43)$$

Since this strong amount of current is beyond the detection range of the *Tektronix* probe, we can conclude from this result that the inductive sensing approach allows for observing the real amplitude of an electric arc flash in the present conditions (230 V-20 A). The measurement of the arc fault detection current created by a short-circuit can reach values of several dozens of Amperes; exceeding the fuse limit during a few milliseconds. Though this result remains

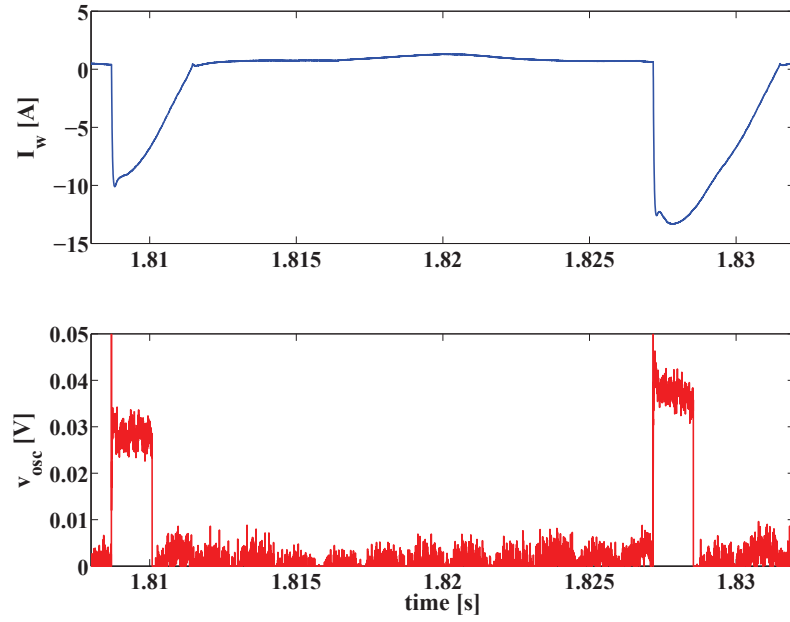


Figure 3.37: Zoom on current peaks caused by electrical arcing.

a coarse estimation of the arcing flash, the study reported in [100] presented an arc fault analysis in which the arcing currents reached comparable amplitude levels in similar test conditions.

The TSMC 250 nm technology offers four values of threshold voltage: a

native, a low, a medium, and a high voltage. The threshold voltage choice must take into account the CMOS diode abruptness and the possibility of biasing them in weak or strong inversion. The best compromise is therefore the use of the medium voltage which allows for using the system in both of the inversion modes and consequently ensures an acceptable input swing range for the induced voltage V_{ref} .

The sizing of the transistor is relatively free of constraints; the dimensions are chosen to minimize the circuit die area. The sizes of the PMOS transistors pairs are fixed with respect to the hole and electron mobility values to keep the symmetry between the NMOS and PMOS pairs. The feedback transistor is chosen to be wide enough to set the reset time of the saw-tooth wave under 100 ns. Numerical data shown in appendix E contains all transistor sizes and threshold voltages for the input current source.

3.8.2 NMOS diodes inversion factor

Depending on the reference voltage V_{ref} , the large-signal transfer characteristic is split into two different operation modes corresponding to the weak and strong inversion operations of the input diode. The simulator tools allow for calculating the reference voltage for each inversion mode with accuracy using the definition of the inversion factor (3.44); U_t being the thermal voltage equal to 0.026 V at 25°C.

$$IC = \frac{I_{d0}}{2n\mu_n C_{ox} \frac{W}{L} U_t^2} \quad (3.44)$$

The inversion limits are then given by the following conditions:

- Weak inversion: $IC < 1$;
- Strong inversion: $IC > 10$;
- Moderate inversion: $IC \in [1, 10]$.

The inversion factor is also a function of the substrate factor n which presents important variations with low reference voltage amplitudes. The substrate factor can be evaluated by (3.45) according to the different conductances of the MOS transistor model. The resulting plots sketched in Fig. 3.39 show the evolution of the inversion and substrate factors for the two cascoded transistors of the input diode. The inversion limits are marked with vertical lines. The moderate inversion area has not been considered in the model equations of the large-signal transfer characteristic Ω_{out} . In practice, the reference voltage is chosen to avoid the moderate inversion mode in which the model equations defined previously are invalidated.

$$n = \frac{g_m + g_{mb} + g_d}{g_m} \quad (3.45)$$

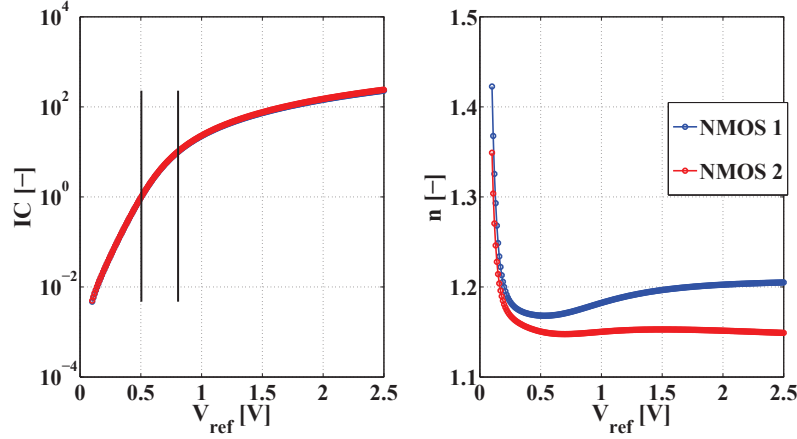


Figure 3.39: Evolution of the inversion IC and substrate n factors versus the reference voltage V_{ref} .

3.8.3 System response simulations

Considering the theoretical developments for the large and small-signal transfer functions, the most influential parameter is the bias voltage of the input cascoded NMOS diodes. The transfer functions calculated in large and small-signal modes are simulated using the *AMS ELDO* software for different values of the reference voltage V_{ref} .

3.8.3.1 Large-signal simulations

The large-signal simulation plotted in Fig. 3.40 and Fig. 3.41 shows the large-signal response Ω_{out} evolution versus the magnetic field frequency and current amplitude of the power line. As predicted by equation (3.34), the linearity of the response is observed for the strong inversion curves; the strong inversion voltage point being located at $V_{ref} = 0.8$ V. We observe that the linearity of the curve located in the moderate inversion ($V_{ref} = 0.6$ V) begins to disappear, thus resulting in the transition between the weak and strong inversion modes.

3.8.3.2 Small-signal simulations

Small-signal simulations illustrate frequency response approximated by the equation (3.37). As the voltage reference value sets the diodes small-signal

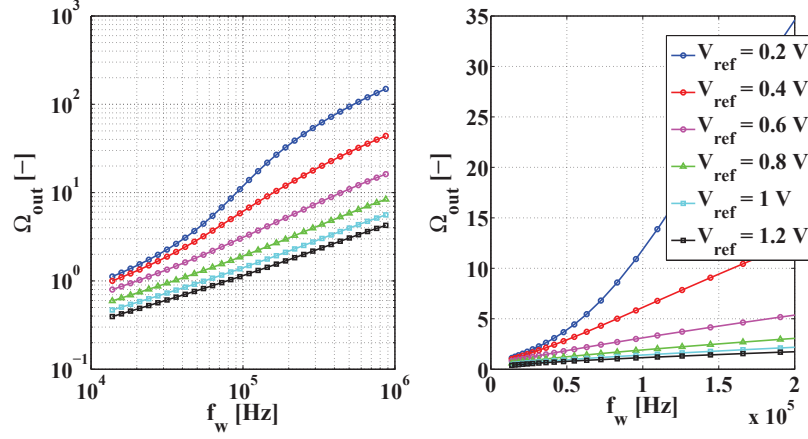


Figure 3.40: Simulation of the large-signal response Ω_{out} for several values of V_{ref} ($N_{sp} = 4$; $L_c = 2$ mm; $W_w = W_{sp} = 200$ μ m, $I_w = 10$ A).

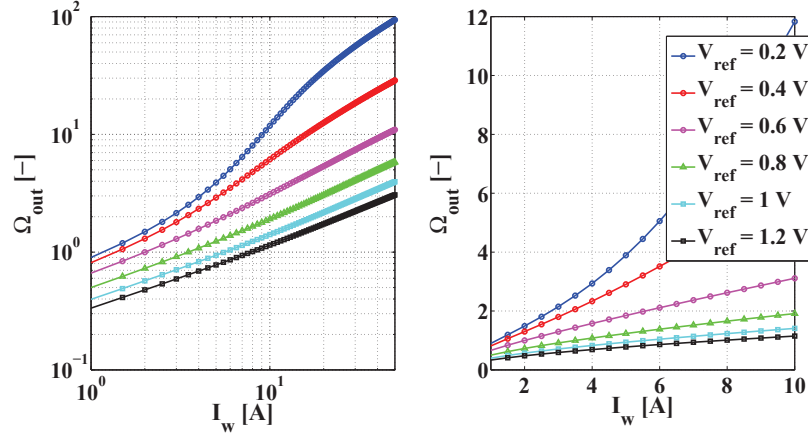


Figure 3.41: Simulation of the large-signal response Ω_{out} for several values of V_{ref} ($N_{sp} = 4$; $L_c = 2$ mm; $W_w = W_{sp} = 200$ μ m, $f_w = 100$ kHz)

conductance, the amplification gain must logically increase with V_{ref} . The plots sketched in Fig. 3.42 show the bode diagram of the transfer function.

The frequency range on which the phase is close to zero for each simulation represents the range on which the approximation made by equation (3.37) is valid. The two dominant poles delimiting the validity of the transfer function approximation are due to the circuit interface internal nodes. The lowest frequency pole is due to the output branch of the circuit. The load capacitance

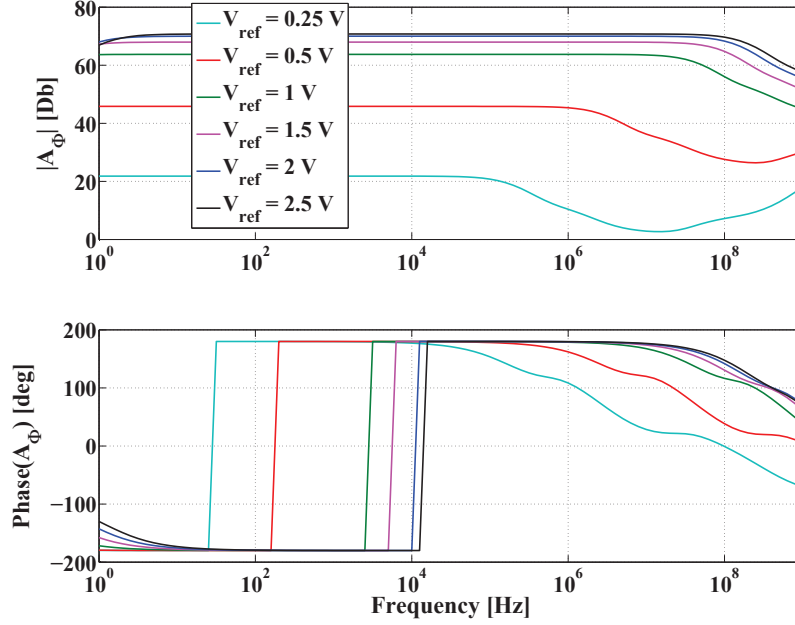


Figure 3.42: Simulation of the small-signal response A_Φ for several values of V_{ref} ($N_{sp} = 4$; $L_c = 2$ mm; $W_w = W_{sp} = 200$ μm).

C_L forms an R-C pair with the output transistor conductances. The upper frequency limit is set by the parasitical capacitances of the interfacing circuit; the internal nodes of the circuit have their own bandwidth limited by the R-C parasitical load seen from the nodes. In this case, the most limited frequency node is located between the transistor pairs M3-M4 and M5-M6 in Fig. 3.22. The third pole formed by the R-L serial pair of the sensing element is non-dominant. The resistive component of this pair is indeed a function of the input conductance g_d which is low enough to push this pole at high frequencies.

3.8.3.3 CMOS System power consumption

The last advantage to integrating the inductive detection system is the considerable decrease of the power consumption. In such a system, the only parameter which causes the power consumption to increase is the reset frequency f_{osc} of the Schmitt trigger due to the strong current drops during the switching time of the system. The other part of the power consumption is only due to the DC bias point of the circuit. The results illustrated in Fig. 3.43 are the total power consumption of the system including the Schmitt trigger versus the

saw-tooth oscillation frequency f_{osc} .

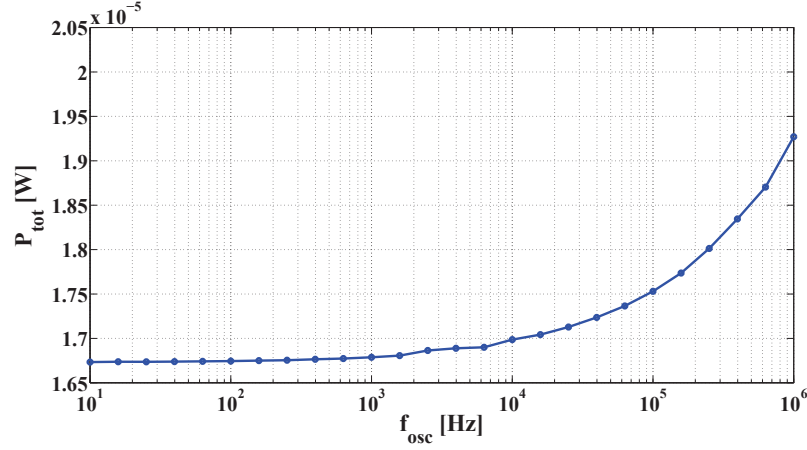


Figure 3.43: Total system power consumption as a function of the saw-tooth frequency f_{osc} .

3.8.4 Global contributions of the CMOS integration

In order to highlight the advantages of the CMOS integration of the system, a comparative table have been established in Tab. 3.7. This table exposes the figures of merit that has been improved from the discrete implementation of the system to its CMOS integration. The first improved performance is the power consumption which has been reduced by three orders of magnitude. All of the discrete component power contributions indeed quickly reach several dozens of milliwatts. The other improvements have been explained in the liquid sensing system chapter.

	Discrete circuit	CMOS circuit
Power consumption	> 50 mW	< 50 μ W
Circuit area	> 30 cm ²	< 2 cm ²
Parasitic effects	high	low

Table 3.7: Comparison between the CMOS integration and discrete implementation.

3.9 Conclusions

The results obtained in the present chapter are of real interest from which several conclusions and perspectives of future work can be drawn.

The first ascertainments we can make from what has been exposed earlier are the novelty and originality of the detection system. The main innovation is the methodology of interfacing the classical planar inductor devices. Based on a saw-tooth oscillator, the circuit architecture provides a signal from which the sensed data can be easily extracted using the modeling equations and transfer functions developed before.

The approach we used to simulate the devices using the finite element tools has been proven to be accurate and validates the analytical formulas. It also allowed performing simulations that highlighted all of the parameters that influence the magnetic sensor response. The sizing method we developed to fix the dimensions of the sensor has been proven to be efficient and fast for finding a set of valid geometrical parameters of the sensing element.

The mathematical modeling of the circuit interface, given by both large and small-signal points of view, summarized in two compact formulas the entire transducing principle of the circuit.

The implementation of the entire magnetic sensing system in discrete elements proved the efficiency of the sensing methodology and presented the expected linearity in the transfer characteristic response. The idea of arc fault detection (AFD) application finally brought a real value to the use of such a system to sense the sharp variation of the magnetic field generated by a supply current line. The system also discriminated between the nominal operations of the line depending on the supplied load and the sharp variations of the current due to the electrical arcs. The calculation of the magnetic flux amplification factor allowed for evaluating the arcing current peak amplitude.

The simulations performed with accurate simulator tools on the TSMC technology validated the large-signal and small-signal transduction formulas and gave their limitations in terms of linearity and frequency bandwidth as a function of the input diodes reference voltage. The CMOS integration of the circuit interface clearly proved the possibility of improving circuit performances in terms of power consumption and accuracy.

All of these conclusions prove the simplicity and the ease with which the proposed sensing solution is implemented. As the challenges of today are to

miniaturize and optimize power consumption within circuits, the design and integration of CMOS advanced technology on the proposed system are the best improvements that can be performed.

CHAPTER 4

Conclusions

This PhD thesis exposed a scientific design methodology of instrumentation systems. All of the results obtained in terms of sensor and circuit interface design, simulations, and measurements proved the methodology efficiency. The design steps of the instrumentation systems we described in the introduction chapter brought scientific innovations we highlighted in this conclusion.

Considering first the design methodology, each of the seven steps brought its own contribution for the development of the liquid and magnetic sensing systems:

- The review of the sensing solutions allowed for positioning the sensing contexts and for providing the direction in research.
- The detection principle descriptions resulted in equivalent models which specified explicitly the input and the output data captured and delivered by the sensing element. The compactness of those models was an advantage when used by circuit simulator tools; each circuit element being clearly identified.
- The sensor physical modeling using an FEM approach presented accurate results and described the operation of the sensing elements with realism.
- The sizing algorithms we developed gave the most suitable dimensions for each sensor in order to obtain the best sensing performance while considering circuit interface abilities (i.e. sensing resolution).
- The step of the interfacing principle technique is the most interesting in terms of scientific innovations. The interfacing approaches were summarized into compact mathematical formulas from which the sensed data can be easily read. As said further, those approaches investigated the technique of triangular voltage deformation.
- The implementation of the systems using discrete components validated the theoretical models studied for each application context. The poor resistance of the discrete components to parasitic effects was advantageous to highlight all of the limitations of the circuit architectures.

- The final CMOS circuit integration step considered all the limitations of the discrete prototyping and resulted in excellent performance; the parasitic effects have been highly attenuated, the power consumption of the systems has been decreased down to $50 \mu W$, and the interface circuit chip sizes never exceed 1 cm^2 .

Focusing on both of the detection systems we described in the chapters 2 and 3, the major novel contribution we foregrounded is the manner of which the sensing element is electrically interrogated. The interfacing circuits of both sensing systems were based on the modulation of a triangular signal. As noted in the thesis introduction, the generation of a triangular voltage is performed using a current-controlled oscillator (CCO); a capacitance is charged by a constant current. The modulation of the output triangular voltage is therefore feasible according to two methods: varying the load capacitance or varying the load current. The triangular modulation technique that governed the liquid sensing system was based on the replacement of the loading capacitance by the IDE sensor complex impedance. The magnetic sensing system investigated the second possibility of a triangular voltage modulation; the sensed information was carried by the oscillator loading current. Since we also showed the considerable ease of implementing triangular voltage oscillating circuits, the use of CCO in sensor interfaces is highly efficient and advantageous in terms of architectural simplicity.

Analyzing the achievements of chapter 2, the results obtained from the design of the liquid sensing system proved the reliability of the proposed technique of the sensing electrodes electrical stimulation. The linearity of the saline NaCl and KCl concentrations characterization is excellent and close to the theoretical model. Furthermore, the permittivity measurement of the water-glycerol mixture gave good results in comparison with the value found in the literature. In conclusion, such a system can be qualified by its genericity (i.e. capacity of interfacing both passivated and unpassivated interdigitated structures), and its ability to "kill two birds with one stone" (i.e. detection of both the liquid conductivity and permittivity in a single signal).

The system presented in chapter 3 extended the field of applications of planar inductors. The literature showed that the most common use of those devices lies in RF applications. We proposed a novel usage of planar inductors to detect strong current variations on power lines for the Arc Fault Detection (AFD) application. The experiments conducted on the electrical arc generator testbench proved the sensing system could discriminate the nominal operation from the arcing activity. We also gave an estimation of the arcing current amount according to the transfer function formula defined in the chapter. The major innovative contributions of the magnetic sensitive system are assuredly

the sensor interfacing method and the technique we have designed to interpret the sensed data.

The systems developed throughout this thesis can be qualified by three keywords: simplicity, efficiency, portability. The simplicity means that the sensor interfacing techniques are of considerable ease to implement since very few resources are required to fabricate the system architecture. The systems detection principles are efficient meaning the useful information can be extracted with accuracy according to the mathematical methods demonstrated for the two applications. Finally, we can assert that the instrumentation solutions are portable in that the circuits are autonomous and require a single power source to operate. The size of the systems remains small ($< 1 \text{ dm}^3$) even when implemented using discrete components. The CMOS integration improves this figure of merit down to less than 1 cm^3 .

Having successfully faced the challenges presented at the beginning of this work, we note that all of the solutions have reached technology readiness levels which are close to the end of the research activity; TRL4 and TRL5 have been respectively reached for the fluid and magnetic variations detection systems. The next phase of work, principally consists in industrial development in order to push the technology further. The transition between fundamental research work and industrialization has been completed according to our design methodology of portable instrumentation systems.

The principal perspectives of this research work lie in the CMOS circuit integration of the proposed systems. Indeed, we focused the analysis on the direct analog interface of the sensor. Actually, a fully integrated system comprises supply blocks, buffering and digital units. The design and optimization of all those components constitute the principal development tasks to reach the end of the TRL scale.

References

- [1] M. Lee, J. Bae, J. Lee, C.S. Lee, S. Hong, Z.L. Wang, "Self-powered environmental sensor system driven by nanogenerators", *Energy and Environmental Science*, Vol. 4, pp.3359-3363, 2011.
- [2] M.A. Xapsos, "Technology Readiness Overview: CMOS Ultra-Low Power Radiation Tolerant Integrated Circuits", NASA Goddard Space Flight Center.
- [3] J.E. Ramirez-Marquez, B.J. Sauser, "System Development Planning via System Maturity Optimization", *IEEE Transactions on Engineering Management*, Vol. 56, pp.533-548, 2009.
- [4] A. Finegan, "A Methodology to Design an Expert System for Remote Sensing Technology Management", RMIT Centre of Remote Sensing, Dept. of Land Information, 1993.
- [5] N. André, L.A. Francis, S. Druart, P. Dupuis, D. Flandre, J.P. Raskin, "Portable wireless microsensing system for human breath monitoring", *Proceedings of the 2010 CMOS Emerging Technologies Workshop*, 2010.
- [6] A. Thanachayanont, S. Sangtong, "Low-Voltage Current-Sensing CMOS Interface Circuit for Piezo-Resistive Pressure Sensor", *ETRI Journal*, vol. 29, pp. 70-78, 2007.
- [7] A. Morgenshtein, L. Sudakov-Boreysha, U. Dinnar, C.G. Jakobson, Y. Nemirovsky, "Wheatstone-Bridge readout interface for ISFET/REFET applications", *Sensors and Actuators B*, vol. 98, pp. 18-27, 2004.
- [8] B. Lakard, D. Magnin, O. Deschaume, G. Vanlancker, K. Glinel, S. Demoustier-Champagne, B. Nysten, A.M. Jonas, P. Bertrand, S. Yunus, "Urea potentiometric enzymatic biosensor based on charged biopolymers and electrodeposited polyaniline", *Biosensors and Bioelectronics*, vol. 26, pp. 4139-4145, 2011.
- [9] A. Arnau, "A Review of Interface Electronic Systems for AT-cut Quartz Crystal Microbalance Applications in Liquids", *Sensors*, Vol. 8, pp.370-411, 2008.
- [10] N.B. Yusof, N. Soin, S.Z.M. Dawal, "Capacitive interfacing for MEMS humidity and accelerometer sensors", *Technical Postgraduates (TECHPOS) International Conference*, pp.1-5, 2009.

-
- [11] B. Olbrechts, B. Rue, G. Rinaldi, I. Stiharu, D. Flandre, J.P. Raskin, "SOI Devices and Ring Oscillators on Thin Dielectric Membranes For Pressure Sensing Applications", *SOI Conference, IEEE International*, pp.109-110, 2007.
- [12] S.W. Lee, J.W. Rhim, S.W. Park, S.S. Yang, "A Novel Micro Sensor using a Surface-Acoustic-Wave (SAW) Delay-line Oscillator", *Sensors, IEEE Conference*, pp.1156-1159, 2007.
- [13] S. Cui, T.L. Viswanathan, T.R. Viswanathan, B. Banerjee, "A highly linear CMOS current-controlled oscillator using a novel frequency detector", *Circuits and Systems ISCAS, IEEE International Symposium*, pp.2841-2844, 2008.
- [14] M. Futagawa, MD.I. Hussain, K. Kamado, F. Dasai, M. Ishida, K. Sawada, "Multimodal Water Content and Nutrient Concentration Sensor for On-Site Soil Monitoring", *SENSORDEVICES 2011*, pp. 50-53, 2011.
- [15] N.F. Sheppard, M.J. Lesho, P. McNally, A.S. Francomacaro, "Microfabricated conductimetric pH sensor", *Sensors and Actuators B*, vol. 28, pp. 95-102, 1995.
- [16] A. V. Mamishev, Y. Du, "Development and Applications of Fringing Electric Field Dielectrometry Sensors and Parameter Estimation Algorithms", *Journal of Electrostatics*, vol. 46, pp.109-123, 1999.
- [17] A.T. Pérez, M. Hadfield, "Low-Cost Oil Quality Sensor Based on Changes in Complex Permittivity", *Sensors 2011*, vol. 11, pp. 10675-10690, 2011.
- [18] E. Finot, E. Bourillot, R. Meunier-Prest, Y. Lacroute, G. Legay, M. Cherkaoui-Malki, N. Latruffe, O. Siri, P. Braunstein, A. Dereux, "Performance of interdigitated nanoelectrodes for electrochemical DNA biosensor", *Ultramicroscopy* vol. 97, pp.441-449, 2003.
- [19] M. Varshney, Y. Li, "Interdigitated array microelectrodes based impedance biosensors for detection of bacterial cells", *Biosensors and Bioelectronics*, vol. 24, pp.2951-2960, 2009.
- [20] J. Lario-García, R. Pallas-Areny, "Constant-phase element identification in conductivity sensors using a single square wave", *Sensors and Actuators A*, vol. 132, pp. 122-128, 2006.
- [21] J. Hong, D.S. Yoon, S.K. Kim, T.S. Kim, S. Kim, E.Y. Pak, K. No, "AC frequency characteristics of coplanar impedance sensors as design parameters", *The Royal Society of chemistry - Lab Chip*, vol. 5, pp.270-279, 2005.

- [22] K.G. Ong, C.A. Grimes, "A Carbon Nanotube-based Sensor for CO_2 Monitoring", *Sensors*, vol. 1, pp.193-205, 2001.
- [23] Clipper Control, Dielectric constant Values, available on <http://www.clippercontrols.com/pages/Dielectric-Constant-Values.html>.
- [24] N. André, S. Druart, P. Gérard, R. Pampin, L. Moreno-Hagelsieb, T. Kezai, L.A. Francis, D. Flandre, J-P. Raskin, "Miniaturized Wireless Sensing System for Real-Time Breath Activity Recording", *IEEE Sensors Journal*, vol. 10, pp.178-184, 2010.
- [25] **Comsol** User's Guide Version 4.3 "AC/DC Module", pp. 172-178, 2012.
- [26] W. Olthuis, A.J. Sprenkels, J.G. Bomer, P. Bergveld, "Planar interdigitated electrolyte-conductivity sensors on an insulating substrate covered with Ta_2O_5 ", *Sensors and Actuators B*, vol. 43, pp. 211-216, 1997.
- [27] Roberto de la Rica, César Fernandez-Sánchez, Antonio Baldi, "Polysilicon interdigitated electrodes as impedimetric sensors", *Electrochemistry Communications*, vol. 8, pp. 1239-1244, 2006.
- [28] K.B. Hendrick, "Planar Interdigitated Dielectric Sensor", *United States Patent*, Patent number: 5045798, 1991.
- [29] Vadim F. Lvovich, C.C. Liu, Matthew F. Smiechowski, "Optimization and fabrication of planar interdigitated impedance sensors for highly resistive non-aqueous industrial fluids", *Sensors and Actuators B*, vol. 119, pp. 490-496, 2006.
- [30] E.D. Goluch, B. Wolfrum, P.S. Singh, M.A.G. Zevenbergen, S.G. Lemay, "Redox cycling in nanofluidic channels using interdigitated electrodes" *Anal Bioanal Chem*, vol. 394, pp.447-456, 2009.
- [31] K. Bohinc, V. Kralj-Iglic, A. Iglic, "Thickness of electrical double layer. Effect of ion size", *Electrochimica Acta*, vol. 46, pp.3033-3040, 2001.
- [32] L. Moreno-Hagelsieb, B. Foutier, G. Laurent, C. Poleunis, P. Bertrand, J-P. Raskin, J. Remacle, D. Flandre, "Aluminium anodizing process characterization for DNA attachment and electrical detection", *MICROTAS 2004*, pp.395-397, 2004.
- [33] X. Tang, L.A. Francis, P. Simonis, M. Haslinger, R. Delamare, O. Deschaume, D. Flandre, P. Defrance, A. Jonas, J. Vigneron, J-P. Raskin, "Room temperature atomic layer deposition of Al_2O_3 and replication of butterfly wings for photovoltaic application" *Journal of Vacuum Science and Technology A. Vacuum* vol. 30, pp.01A146-1-5, 2012.

-
- [34] N. André, S. Druart, P. Dupuis, B. Rue, P. Gérard, D. Flandre, Jean-Pierre Raskin, Laurent Francis, "Dew-Based Wireless Mini Module for Respiratory Rate Monitoring", *IEEE Sensors Journal*, vol. 12, pp. 699-706, 2012
 - [35] J. Wu, J. P. W. Stark, "A low-cost approach for measuring electrical conductivity and relative permittivity of liquids by triangular waveform voltage at low frequencies", *Measurement Science and Technology*, vol. 16, pp. 1234-1240, 2005.
 - [36] Peter Van Gerwen, Wim Laureyn, Wim Laureys, Guido Huyberechts, Maaïke Op De Beeck, Kris Baert, Jan Suls, Willy Sansen, P. Jacobs, Lou Hermans, Robert Mertens, "Nanoscaled interdigitated electrode arrays for biochemical sensors", *Sensors and Actuators B*, vol. 49, pp. 73-80, 1998.
 - [37] A.M. Sampson, E.T.K. Peterson, I. Papautsky, "Interdigitated array microelectrode capacitive sensor for detection of paraffinophilic Mycobacteria", *Proc. of SPIE*, vol. 6886, pp.68860X-1-9, 2008.
 - [38] J. Wu, J. P. W. Stark, "A low-cost approach for measuring electrical conductivity and relative permittivity of liquids by triangular waveform voltage at low frequencies", *Measurement Science and Technology*, vol. 16, pp. 1234-1240, 2005.
 - [39] K-H. Lee, J-O. Lee, M-J. Sohn, B. Lee, S-H. Choi, S. K. Kim, J-B. Yoon, G-H. Cho, "One-chip electronic detection of DNA hybridation using precision impedance-based CMOS array sensor", *Biosensors and Bioelectronics*, vol. 26, pp. 1373-1379, 2010
 - [40] N. Couniot, D. Flanre, L.A. Francis, A. Afzalian, "Signal-to-noise ratio optimization for detecting bacteria with interdigitated microelectrodes", *Sensors and Actuators B*, Article in press, 2013.
 - [41] M. Kitsara, D. Goustouridis, S. Chatzandroulis, M. Chatzichristidi, I. Raptis, T. Ganetsos, R. Igreja, C.J. Dias, "Single chip interdigitated electrode capacitive chemical sensor arrays", *Sensors and Actuators B*, vol. 127, pp. 186-192, 2007.
 - [42] E. Ghafar-Zadeh, M. Sawan, D. Therriault, "A 0.18 μm CMOS capacitive sensor Lab-on-Chip", *Sensors and Actuators A*, vol. 141, pp. 454-462, 2008.
 - [43] C. Stagni, C. Guiducci, L. Benini, B. Riccò, S. Carrara, B. Samorì, C. Paulus, M. Schienle, M. Augustyniak, R. Thewes, "CMOS DNA Sensor Array With Integrated A/D Conversion Based on Label-Free Capacitance Measurement", *IEEE Journal of Solid-State Circuits*, vol. 41, pp. 2956-2964, 2006.

- [44] L. Gu, Q.A. Huang, M. Qin, "A novel capacitive-type humidity sensor using CMOS fabrication technology", *Sensors and Actuators B*, vol. 99, pp. 491-498, 2004.
- [45] Y. Qiu, C. Azeredo-Leme, L.R. Alcácer, J.E. Franca, "A CMOS humidity sensor with on-chip calibration", *Sensors and Actuators A*, vol. 92, pp. 80-87, 2001.
- [46] X. Tang, D. Flandre, J.P. Raskin, Y. Nizet, L. Moreno-Hagelsieb, R. Pampin, L.A. Francis, "A new interdigitated array microelectrode-oxide-silicon sensor with label-free, high sensitivity and specificity for fast bacteria detection", *Sensors and Actuators B*, vol. 156, pp. 578-587, 2011.
- [47] C. Azcona, B. Calvo, N. Medrano, S. Celma, M. R. Valero, "A CMOS Micropower Voltage-to-Frequency Converter for Portable Applications", *Ph.D. Research in Microelectronics and Electronics (PRIME), 2011 7th Conference on*, vol. 26, pp. 141-144, 2011.
- [48] W. Xu, E.G. Friedman, "Clock Feedthrough in CMOS Analog Transmission Gate Switches", *Analog Integrated Circuits and Signal Processing*, vol. 44, pp. 271-281, 2005.
- [49] L. Dai, R. Harjani, "CMOS Switched-Op-Amp-Based Sample-and-Hold Circuit", *IEEE Journal of Solid-State Circuits*, vol. 35, pp. 109-113, 2000.
- [50] W. Xiaofeng; L. Hongxia, S. Li, H. Yue, L. Di, H. Shigang, "A bootstrapped switch employing a new clock feed-through compensation technique", *Journal of Semiconductors*, vol. 30, pp. 125007.1-10, 2009.
- [51] RG16100 digital to analog converter (DAC): available on *Ridgtop Group* website.
- [52] V.B. Kulkarni, P.M. Furth, "Low-Voltage CMOS Comparators With Programmable Hysteresis", Master thesis, New Mexico State University, 2005.
- [53] W. Prodanov, M.C. Schneider, "A Digitally Programmable Current Schmitt-Trigger", Laboratório de Circuitos Integrados, Universidade Federal de Santa Catarina (UFSC).
- [54] J. Misurec, J. Koton, "Schmitt Trigger with Controllable Hysteresis Using Current Conveyors", *International Journal of Advances in Telecommunications, Electrotechnics, Signals and Systems*, Vol. 1, No. 1, 2012.
- [55] BCV62 bipolar mirror PNP: available on *NXP Semiconductor* website.
- [56] BCV61 bipolar mirror NPN: available on *NXP Semiconductor* website.

- [57] ADG636 Analog Switch: available on *Analog Devices* website.
- [58] CD4093BE Schmitt NAND gate: available on *Texas Instruments* website.
- [59] AD8661ARZ CMOS amplifier: available on *Analog Devices* website.
- [60] D. Bol, "Pushing Ultra-Low-Power Digital Circuits into the Nanometer Era", *PhD thesis*, Université catholique de Louvain, 2008.
- [61] A. L. Herrera-May, L. A. Aguilera-Cortés, P. J. García-Ramírez and E. Manjarrez, "Resonant Magnetic Field Sensors Based On MEMS Technology", *Sensors*, vol. 9, pp. 7785-7813, 2009.
- [62] B. Bahreyni, F. Najafi, C. Shafai, "Piezoresistive sensing with twin-beam structures in standard MEMS foundry processes", *Sensors and Actuators A*, vol. 127, pp. 325-331, 2006.
- [63] M. El Ghorba, N. André, J.-P. Raskin, "Out-of-plane CMOS compatible and in plane ferromagnetic magnetometers", *Transducers Conference*, pp. 2373-2376, 2007.
- [64] C. Sirisathitkul, H. Longkullabutra, P. Kirdtongmee, S. Aslam and J. Gregg, "Magnetoresistance due to the Lorentz force in silicon membrane", *Songklanakarin J. Sci. Technol.*, vol. 24, no. 2, 2002.
- [65] P. Courtois, "Conception, réalisation, et caractérisation d'un magnétomètre MEMS en technologie SOIMUMPs", *Master thesis*, Université catholique de Louvain, 2011.
- [66] E. V. Thomsen, J. Richter, "Piezo Resistive MEMS Devices: Theory and Applications", *Thesis*, Department for Micro and Nano Technology, Technical University of Denmark (DTU), 2005.
- [67] S. Ranvier, S. Paquay, S. Requier, H. Lamy, V. Rochus, L. A. Francis, P. Rochus, "Influence of multiphysics couplings on the performance of a MEMS magnetometer", *EuroSimE Conference*, 2011.
- [68] M. El Ghorba, N. Andre, S. Sobieski, J.-P. Raskin, "CMOS Compatible Out-of-Plane and In-Plane Magnetometers", *Transducers conference*, 2007.
- [69] W. Wengang, H. Fengyi, L. Yi, Z. Shaoyong, H. Xiang, L. Zhihong, H. Yilong, W. Yangyuan, "RF Inductors With Suspended and Copper Coated Thick Crystalline Silicon Spirals for Monolithic MEMS LC Circuits", *IEEE Microwave and Wireless Components Letters*, vol. 15, pp. 853-855, 2005.
- [70] J.C. Kao, M. Wei-Hsiang, V.S. Sathe, M. Papaefthymiou, "Energy-Efficient Low-Latency 600 MHz FIR With High-Overdrive Charge-Recovery Logic", *IEEE Transactions on Very Large Scale Integration (VLSI) Systems*, vol. 20, pp. 977-988, 2012.

- [71] H. Wang, C. Sideris, A. Hajimiri, "A frequency-shift based CMOS magnetic biosensor with spatially uniform sensor transducer gain", *Custom Integrated Circuits Conference*, 2010.
- [72] C. Leroy, M. B. Pisani, C. Hibert, D. Bouvet, M. Puech, A. M. Ionescu, "High quality factor copper inductors integrated in deep dry-etched quartz substrates", *Microsyst Technol.*, vol. 13, pp. 1483-1487, 2007.
- [73] D. Flynn, M. P.Y. Desmulliez, "Design Methodology and Manufacture of a Microinductor", *Symposium on Design Test Integration and Packaging of MEMS/MOEMS (DTIP)*, 2008.
- [74] K. Yong-Seok, Y. Seong-Cho, L. Jeong-Bong, L. Heebok, "A new class of LC-resonator for micro-magnetic sensor application", *Journal of Magnetism and Magnetic Materials*, vol. 304, pp. 117-121, 2006.
- [75] A. Baschiroto, E. Dallago, P. Malcovati, M. Marchesi, E. Melissano, M. Morelli, P. Siciliano, G. Venchi, "An Integrated Micro-Fluxgate Magnetic Sensor With Front-End Circuitry", *IEEE Transactions on Instrumentation and Measurement*, vol. 58, pp. 3269-3275, 2009.
- [76] W. Pei-Ming, "Micromachined On-Chip Fluxgate Magnetometers with Low Power Consumption", *PhD thesis*, University of Cincinnati, 2009.
- [77] M. Blagojevic, M. Kayal, M. Gervais, D. De Venuto, "SOI Hall-Sensor Front End for Energy Measurement", *IEEE Sensors Journal*, vol. 6, pp. 1016-1021, 2006.
- [78] F.C. Castaldo, V.R. Mognon, C.A. dos Reis Filho, "Magnetically Coupled Current Sensors Using CMOS Split-Drain Transistors", *IEEE Transactions on Power Electronics*, vol. 24, pp. 1733-1736, 2009.
- [79] A. Barthélémy, A. Fert, J-P. Contour, M. Bowen, V. Cros, J.M. De Teresa, A. Hamzic, J.C. Faini, J.M. George, J. Grollier, F. Montaigne, F. Pailloux, F. Petroff, C. Vouille, "Magnetoresistance and spin electronics", *Journal of Magnetism and Magnetic Materials*, pp. 242-245, 2002.
- [80] G. Mihajlovic, P. Xiong, S. von Molnar, M. Field, G.J. Sullivan, K. Ohtani, H. Ohno, "Submicrometer Hall Sensors for Superparamagnetic Nanoparticle Detection", *IEEE Transactions on Magnetics*, vol. 43, pp. 2400-2402, 2007.
- [81] A. Nieuwoudt, M. S. McCorquodale, R. T. Borno, Y. Massoud, "Efficient Analytical Modeling Techniques for Rapid Integrated Spiral Inductor Prototyping", *Custom Integrated Circuit Conference*, 2005.

- [82] O. Poncelas, J. A. Rosero, J. Cusidó, J. A. Ortega, and L. Romeral, "Motor Fault Detection Using a Rogowski Sensor Without an Integrator", *IEEE Transactions on Industrial Electronics*, vol. 56, pp. 4062-4070, 2009.
- [83] O. Dezuari, E. Belloy, S. E. Gilbert, and M. A. M. Gijs, "New Hybrid Technology for Planar Fluxgate Sensor Fabrication", *IEEE Transactions on Magnetics*, vol. 35, pp. 2111-2117, 1999.
- [84] S. Tumanski "Induction coil sensors - a review", *Meas. Sci. Technol.*, vol. 18, pp. 31-46, 2007.
- [85] **Comsol** User's Guide Version 4.3 "AC/DC Module", pp. 256, 2012.
- [86] S. Druart, D. Flandre, L.A. Francis "A Methodology for the Simulation of MEMS Spiral Inductances used as Magnetic Sensors", *Comsol Conference*, 2010.
- [87] S. Mohan, M. del Mar Hershenson, S. Boyd, T. Lee, "Simple Accurate Expressions for Planar Spiral Inductances", *IEEE Journal of Solid-State Circuits*, vol. 34, pp. 1419-1424, 1999.
- [88] L. Ballicchia, S. Orcioni, "Methodology for Design and Modeling of Optimum Quality Spiral Inductors", *IEEE Topical Meeting on Silicon Monolithic Integrated Circuits in RF Systems (SiRF)*, Conference Publications, pp. 102-105, 2008.
- [89] G. Stojanovic, L. Zivanov, M. Damnjanovic, "Optimal Design of Circular Inductors", *Facta Universitatis, Elec. Energ.*, vol. 18, pp. 57-70, 2005.
- [90] R. Rieger, A. Demosthenous, J. Taylor, "A 230-nW 10-s time constant CMOS integrator for an adaptive nerve signal amplifier", *IEEE Journal of Solid-State Circuits*, vol. 39, pp. 1968-1975, 2004.
- [91] K. Roy, S. Mukhopadhyay, H. Mahmoodi-Meimand, "Leakage current mechanisms and leakage reduction techniques in deep-submicrometer CMOS circuits", *Proc. IEEE*, vol. 91, pp. 305-327, 2003.
- [92] Mathematics lessons, "Bessel Functions of the First and Second Kind", Univeristy of Waterloo (Canada).
- [93] Csanky et al., "Arc Fault Detection System", *United State Patent*, Pub. No. US 2004/0156153, 2004.
- [94] Multi Circuit Board, PCB Manufacturer, <http://www.multi-circuit-boards.eu>
- [95] T.E. Potter, M. Lavado, "Arc Fault Circuit Interruption Requirements for Aircraft Applications", *Texas Instruments*, www.sensata.com.

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- [96] BSS138 Datasheet: www.fairchildsemi.com/ds/BS/BSS138.pdf
 - [97] BSS84 Datasheet: www.fairchildsemi.com/ds/BS/BSS84.pdf
 - [98] MC14093B Datasheet: www.farnell.com/datasheets/76413.pdf
 - [99] S. Wang, C. Wu, Y. Wang "An Effective Detection Method of Serial Arc Fault on Low Voltage Power Circuits", *Latest Advances in Systems Science and Computational Intelligence*, pp. 177-181, 2012.
 - [100] C.E. Restrepo, "Arc Fault Detection and Discrimination Methods", *the 53rd ieee holm conference on Electrical contacts*, Inc., 2007.

APPENDIX A

List of Notations: chapter 2

Fluid parameters

ε_L	Fluid relative permittivity	$[-]$
σ_L	Fluid conductivity	$[\text{S.m}^{-1}]$
c_k	Cell constant general notation	$[\text{m}^{-1}]$
R_L	Fluid resistance	$[\Omega]$
C_L	Fluid capacitance	$[\text{F}]$
Z_L	Fluid impedance	$[\Omega]$

IDE sensor dimensions

h_b	Board thickness	$[\text{m}]$
h_f	Electrodes height	$[\text{m}]$
W_f	Electrodes width	$[\text{m}]$
W_s	Electrodes spacing	$[\text{m}]$
L_c	Cell width	$[\text{m}]$
L_f	Electrodes length	$[\text{m}]$
N_f	Electrodes number	$[-]$
h_p	Passivation thickness	$[\text{m}]$
A_s	Sensor area	$[\text{m}^2]$
c_{kL}	Fluid area cell constant	$[\text{m}^{-1}]$
c_{kB}	Substrate area cell constant	$[\text{m}^{-1}]$

IDE Electrical components

C_B	Substrate IDE inter-electrode capacitance	$[\text{F}]$
C_P	Passivation capacitance	$[\text{F}]$
C_A	Serial external capacitance	$[\text{F}]$
C_M	Sizing additional fluid parallel capacitance	$[\text{F}]$
C_S	IDE electrodes to backside capacitance	$[\text{F}]$

IDE impedance parameters

Z_{IDE}	IDE equivalent impedance	$[\Omega]$
ω_z	Zero frequency position	$[\text{rad.s}^{-1}]$
ω_p	Pole frequency position	$[\text{rad.s}^{-1}]$

C_0	Initial capacitive coefficient	[F]
$\Delta\omega_{pz}$	pole-zero frequency difference	[rad.s ⁻¹]

IDE FEM parameters

V	Electric potential	[V]
ω	FEM frequency solver	[rad.s ⁻¹]
σ	FEM solver conductivity	[S.m ⁻¹]
ε_r	FEM solver relative permittivity	[-]
ε_0	Absolute vacuum permittivity	[F.m ⁻¹]
ε_B	Relative substrate permittivity	[-]
ε_P	Relative passivation material permittivity	[-]
E_y	Vertical electric field component	[V.m ⁻¹]
k	Elliptic integral modulus	[-]
Q_L	Fluid area sensed electrode charge	[C]
Q_B	Substrate area sensed electrode charge	[C]
S_L	Fluid area Gauss integration surface	[m ²]
S_B	Substrate area Gauss integration surface	[m ²]

IDE sizing methodology parameters

$L_{f,eff}$	Electrodes total length	[m]
$c_{kL,n}$	Fluid area cell constant of a periodic IDE cell	[m ⁻¹]
M_Φ	IDE phasis maximum	[rad]
ω_{max}	Frequency position of M_Φ	[rad.s ⁻¹]

IDE interface methodology parameters

I_0	Square current signal amplitude	[A]
$v_a(t)$	Triangular oscillating voltage	[V]
$v_u(t)$	Triangular oscillating wave segment	[V]
$i(t)$	Square current signal	[A]
T	Half period of $v_a(t)$	[s]
ζ	Initial voltage of $v_u(t)$	[V]
β	Slope of the linear part of $v_u(t)$	[V.s ⁻¹]
τ	Initial slope of $v_u(t)$	[V.s ⁻¹]

IDE discrete circuit interface

$v_d(t)$	Digital Schmitt trigger output	[V]
V_{dd}	Upper circuit supply voltage	[V]
V_{ss}	Lower circuit supply voltage	[V]
V_{hst}	Hysteresis voltage	[V]

Conductivity experiments

λ	Ionic molar conductivity	[S.m ² .mol ⁻¹]
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λ_{Cl^-}	Chlorine ionic molar conductivity	[S.m ² .mol ⁻¹]
λ_{Na^+}	Sodium ionic molar conductivity	[S.m ² .mol ⁻¹]
λ_{K^+}	Potassium ionic molar conductivity	[S.m ² .mol ⁻¹]
c_{NaCl}	Sodium chloride molar concentration	[mol.l ⁻¹]
c_{KCl}	Potassium chloride molar concentration	[mol.l ⁻¹]
Σ	Conductance standard deviation	[S.m ⁻¹]
δ	Concentration standard deviation	[mol.l ⁻¹]
s_{sigma_L}	System sensitivity	[S.l.mol ⁻¹ .m ⁻¹]

Circuit error discussion

ΔV_{cf}	Clock feedthrough voltage drop	[V]
ΔV_{ci}	Charge injection voltage drop	[V]
ΔV_{dp}	Total voltage drop	[V]
v_s	Input source voltage	[V]
C_{ds}	Drain-to-source overlap capacitance	[F]
C_{sd}	Source-to-drain overlap capacitance	[F]
C_{dg}	Drain-to-gate overlap capacitance	[F]
C_{gd}	Gate-to-drain overlap capacitance	[F]
C_i	Schmitt trigger and analog buffer input capacitance	[F]
Q_i	MOS channel stored charge	[C]
β_f	Forward bipolar transistor gain	[—]

IDE CMOS interface

R_{pol}	Resistor for I_0 programming	[Ω]
I_P	Differential comparator bias current	[A]
I_{hst}	Hysteresis current	[A]
R_{hst}	Resistor for I_{hst} programming	[Ω]
V_{in}	Schmitt trigger input test voltage	[V]
V_{sw}	Schmitt trigger schematic output voltage	[V]
f_{twv}	TWV signal frequency	[Hz]
P_{trig}	Schmitt trigger power consumption	[W]
$\hat{\sigma}_L$	Estimated liquid conductivity	[S.m ⁻¹]
$\hat{\varepsilon}_L$	Estimated liquid permittivity	[—]

APPENDIX B

List of Notations: chapter 3

Magnetic laws

$\mathbf{B}$	Magnetic field	[T]
I_w	Straight wire current	[A]
μ_0	Absolute magnetic permeability	[H.m ⁻¹]
R	Distance from straight wire	[m]
$\mathbf{F}$	Lorentz displacement force	[N]
q	Elementary charge	[C]
$\mathbf{v}$	Charge velocity	[m.s ⁻¹]
$\mathbf{E}$	Electric field	[V.m ⁻¹]

Inductor geometric parameters

W_w	Track width	[m]
W_{sp}	Track spacing	[m]
H_w	Track thickness	[m]
N_s	Turn number	[–]
L_c	Inner width	[m]
H_{box}	Box thickness	[m]
A_{eff}	Device area	[m ²]

Inductor electrical model

ε_{ind}	Net induced voltage	[V]
L_S	Serial inductance	[H]
R_S	Serial resistance	[Ω]
C_P	Parallel distributed capacitance	[F]
Z_L	Inductive device load impedance	[Ω]
V_{ind}	Inductive device induced voltage	[V]
Φ_e	Captured magnetic flux	[Wb]
σ_m	Metal conductivity	[S.m ⁻¹]
ε_{box}	Box relative permittivity	[–]

Inductor FEM model

ω	Input signal frequency	[rad.s ⁻¹]
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σ	Material conductivity	[S.m ⁻¹]
$\varepsilon_0\varepsilon_r$	Material permittivity	[F.m ⁻¹]
$\mu_0\mu_r$	Material permeability	[H.m ⁻¹]
V	Electric potential	[V]
$\mathbf{A}$	Magnetic potential	[T.m]
I_{inw}	Inductor stimulation current	[A]
σ_{metal}	Metal track conductivity	[S.m ⁻¹]
$\mathbf{H}_{ext}$	External applied magnetic field	[A.m ⁻¹]
$\mathbf{H}$	Magnetic field notation	[A.m ⁻¹]
$\mathbf{H}_0$	Enforced boundary field notation	[A.m ⁻¹]
$\mathbf{n}$	Normal surface unity vector	[—]
d_{avg}	Average device width	[m]
ρ	Wheeler formula factor	[—]
B_z	Sensor plane normal external magnetic field	[T]
δ	Skin depth	[m]
l_{av}	Wire average length	[m]
Q	Device quality factor	[—]

Inductor sizing methodology parameters

d_{min}	Technology metal track resolution	[m]
A_{max}	Device maximal area	[m ²]
ε_{min}	Minimal net induced voltage	[V]

Inductor interface methodology notations

V_{osc}	Saw-tooth oscillating voltage	[V]
I_L	Oscillator control current	[A]
C_L	Oscillator load capacitance	[F]
V_{reset}	Feedback trigger digital reset voltage	[V]
v_+	Schmitt trigger higher level	[V]
v_-	Schmitt trigger lower level	[V]
T_0	Initial saw-tooth oscillation period	[s]
V_{dd}	Circuit supply voltage	[V]
V_{ref}	Input diode reference voltage	[V]
V_{hst}	Schmitt trigger hysteresis voltage	[V]
I_{d0}	Diode initial bias current	[A]
T_{out}	Observed output saw-tooth period	[s]
β	MOS transistor current factor	[F.V ⁻¹ .s ⁻¹]
V_{th}	MOS transistor threshold voltage	[V]
n	MOS transistor substrate factor	[—]
U_t	Thermal voltage	[V]
I_s	Weak inversion current factor	[A]
μ_n	Electron mobility	[m ² .V ⁻¹ .s ⁻¹]

C_{ox}	Gate oxide surfacic capacitance	[F.m ⁻²]
$I_{d0,SI}$	Diode bias current in strong inversion	[A]
$I_{d0,WI}$	Diode bias current in weak inversion	[A]
$I_{d,SI}$	Diode observed current in strong inversion	[A]
$I_{d,WI}$	Diode observed current in weak inversion	[A]
f_e	Sine induced voltage frequency	[Hz]
T_e	Sine induced voltage period	[s]
Ω_{out}	Circuit output characteristic	[—]
$\Omega_{out,SI}$	Circuit output characteristic in strong inversion	[—]
$\Omega_{out,WI}$	Circuit output characteristic in weak inversion	[—]
$V_{osc,SS}$	Steady-state oscillation voltage	[V]
$v_{osc,AC}$	Small-signal response voltage	[V]
V_G	MOS gate voltage	[V]
g_{in}	Circuit input conductance	[S]
v_{in}	Circuit input voltage notation	[V]
A_Φ	Magnetic flux output transfer characteristic	[V.Wb ⁻¹]
$g_{in,SI}$	Circuit strong inversion conductance	[S]
$g_{in,WI}$	Circuit weak inversion conductance	[S]

Inductor discrete interface

r_w	Power line metal radius	[m]
d_i	Power line insulating material thickness	[m]
M_s	Mutual inductance	[H]
f_w	Power wire signal frequency	[Hz]
I_{arc}	Arc flash current	[A]

Inductor CMOS interface

ω_w	Power wire signal angular frequency	[rad.s ⁻¹]
IC	CMOS inversion factor	[—]
g_m	MOS gate transconductance	[S]
g_d	MOS output conductance	[S]
g_{mb}	MOS substrate conductance	[S]
P_{tot}	Circuit inteface power consumption	[W]
f_{osc}	Saw-tooth oscillating frequency	[Hz]

APPENDIX C

Calculation details of liquid sensing system

C.1 IDE maximum phase and sizing equations

Starting from the general form of the IDE structure impedance equation (C.1) and considering $s = j\omega$, the real and imaginary parts are isolated in expressions (C.2) and (C.3).

$$Z_{IDE} = \left[\frac{Z_0}{j\omega} \right] \left[\frac{j\omega + \omega_z}{j\omega + \omega_p} \right] \quad (C.1)$$

$$\Re(Z_{IDE}) = \frac{\omega Z_0 (\omega_p - \omega_z)}{\omega^2 + \omega_p^2} \quad (C.2)$$

$$\Im(Z_{IDE}) = -\frac{Z_0 (\omega^2 + \omega_z \omega_p)}{\omega \omega^2 + \omega \omega_p^2} \quad (C.3)$$

The phase expression is therefore given by expression (C.4).

$$\Phi_{Z_{IDE}} = \arctan \left(\frac{\Im(Z_{IDE})}{\Re(Z_{IDE})} \right) = \arctan \left(\frac{\omega^2 + \omega_z \omega_p}{\omega \omega_z - \omega \omega_p} \right) \quad (C.4)$$

The position of the phase maximum ω_{max} is next extracted by solving equation (C.5), resulting in expression (C.6).

$$\left. \frac{d\Phi_{Z_{IDE}}}{d\omega} \right|_{\omega=\omega_{max}} = \frac{\omega_{max}^2 (\omega_z - \omega_p) + \omega_p^2 \omega_z - \omega_z^2 \omega_p}{\omega_{max}^4 + \omega_{max}^2 (\omega_z^2 + \omega_p^2) + \omega_p^2 \omega_z^2} = 0 \quad (C.5)$$

$$\omega_{max} = \sqrt{\frac{\omega_z^2 \omega_p - \omega_z \omega_p^2}{\omega_z - \omega_p}} \quad (C.6)$$

The value of this maximum is thus provided by (C.7).

$$M_\Phi = \max(\Phi_{Z_{IDE}}) = \arctan \left(\frac{2\sqrt{\omega_z \omega_p}}{\omega_z - \omega_p} \right) \quad (C.7)$$

The ω_z and ω_p expressions, being given by (C.8) and (C.9), are injected in (C.7) in order to form a second order equation (C.10) where C_A is the unknown value.

$$\omega_z = \frac{1}{R_L (C_A + C_L + C_B)} \quad (\text{C.8})$$

$$\omega_p = \frac{1}{R_L (C_L + C_B)} \quad (\text{C.9})$$

$$C_A^2 \tan^2 M_\Phi - 4C_A(C_B + C_L) - 4(C_B + C_L)^2 = 0 \quad (\text{C.10})$$

The solution of this last equation is finally given by (C.11); M_Φ being the criterion that has to be fixed for the IDE sensor sizing.

$$C_A = 2(C_L + C_B) \frac{1 + \sqrt{1 + \tan^2 M_\Phi}}{\tan^2 M_\Phi} \quad (\text{C.11})$$

C.2 Calculation of the voltage response segments

Assuming that the output wave (2.37) is probed on a rising segment after M half-periods, all the hard-step functions $u(t - nT)$ for $n < M$ are elapsed and equal to 1:

$$\begin{aligned} v_a(t, M) &= \beta t + \zeta - \frac{\zeta}{2} e^{-\frac{t}{\tau}} + \frac{3\zeta}{2} e^{-\frac{(t-T)}{\tau}} \\ &- \frac{(2M+1)\beta T}{2} \\ &- 2\zeta \sum_{n=2}^M \left[(-1)^n e^{-\frac{(t-nT)}{\tau}} \right] \end{aligned} \quad (\text{C.12})$$

A time translation is then applied following the variable change (C.13):

$$t' = t - MT \quad (\text{C.13})$$

The expression (C.12) can be rewritten in (C.14):

$$\begin{aligned} v_a(t', M) &= \beta t' + \zeta - \frac{\zeta}{2} e^{-\frac{(t'+MT)}{\tau}} \\ &+ \frac{3\zeta}{2} e^{-\frac{(t'+(M-1)T)}{\tau}} \\ &- \frac{\beta T}{2} - 2\zeta e^{-\frac{t'}{\tau}} \sum_{n=0}^{M-2} \left[(-1)^n e^{-\frac{nT}{\tau}} \right] \end{aligned} \quad (\text{C.14})$$

If we consider the square current signal as an infinite sum of hard-step functions, the expression (C.14) can be computed for $M \rightarrow \infty$. The two exponential terms of respective coefficients $-1/2$ and $3/2$ thus disappear; a last serial progression has to be computed in the resulting equation (C.15):

$$v_a(t', \infty) = \beta t' + \zeta - \frac{\beta T}{2} - 2\zeta e^{-\frac{t'}{\tau}} \sum_{n=0}^{\infty} [(-1)^n e^{-\frac{nT}{\tau}}] \quad (\text{C.15})$$

The last term of equation (C.15) corresponds to the well known geometrical progression (C.16) where s is the sum value, a is a scale factor and r is the common ratio. The progression converges only if the common ratio r modulus is smaller than unity. This is always verified in the present case because the parameters T and τ are always greater than zero.

$$s = \sum_{n=0}^{\infty} ar^n = \frac{a}{1-r} \quad (\text{C.16})$$

The final segmented expression of the output voltage wave is thus given by expression (C.17); the equation has been vertically shifted to set the starting voltage at zero.

$$v_u(t) = \beta t + \frac{2\zeta}{1 + e^{-\frac{T}{\tau}}} (1 - e^{-\frac{t}{\tau}}) \quad (\text{C.17})$$

APPENDIX D

Calculation details of magnetic sensing system

D.1 Calculation detail of FEM model equations

The development of the FEM model constitutive equations starts from the Maxwell-Ampère's law (D.1) and use the definition of the magnetic and electric fields given by the expressions (D.2), (D.3), (D.4), and (D.5).

$$\nabla \times \mathbf{H} = \mathbf{J} = \sigma (\mathbf{E} + \mathbf{v} \times \mathbf{B}) + j\omega \mathbf{D} + \mathbf{J}^e \quad (\text{D.1})$$

$$\mathbf{B} = \nabla \times \mathbf{A} \quad (\text{D.2})$$

$$\mathbf{E} = -\nabla V - j\omega \mathbf{A} \quad (\text{D.3})$$

$$\mathbf{B} = \mu_0 (\mathbf{H} + \mathbf{M}) \quad (\text{D.4})$$

$$\mathbf{D} = \varepsilon_0 \mathbf{E} + \mathbf{P} \quad (\text{D.5})$$

In the present work, the variables $\mathbf{M}$, $\mathbf{P}$, $\mathbf{J}^e$, and $\mathbf{v}$ are unused since the inductive system does not include any magnetic remanent field, electric polarization, external current density, and carriers motion problems. Combining all these expressions allows for rewriting Ampere's law as (D.6); this constitutes the first FEM equation. The second equation (D.7) is obtained by taking the divergence of the Ampere's law.

$$(j\omega\sigma - \omega^2\varepsilon_0\varepsilon_r) \mathbf{A} + \nabla \times \left(\frac{1}{\mu_0\mu_r} \nabla \times \mathbf{A} \right) + (\sigma + j\omega\varepsilon_0\varepsilon_r) \nabla V = 0 \quad (\text{D.6})$$

$$-\nabla \cdot \left[(j\omega\sigma - \omega^2\varepsilon_0\varepsilon_r) \mathbf{A} + (\sigma + j\omega\varepsilon_0\varepsilon_r) \nabla V \right] = 0 \quad (\text{D.7})$$

D.2 Cascoded CMOS diode current expression

The calculation of the cascoded CMOS diode current (Fig.D.1) starts from the classical expressions (D.8) and (D.9) of any MOS transistor in strong and weak inversion operation; I_s being the current coefficient given by equation (D.10).

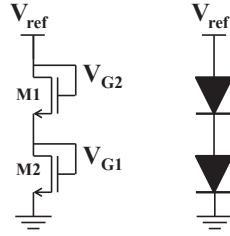


Figure D.1: CMOS cascoded diode configuration

$$I_{dsat,SI} = \frac{\beta}{2n} [V_{GS} - V_{th}]^2 \quad (D.8)$$

$$I_{dsat,WI} = I_s e^{\frac{V_{GS}}{nU_t}} \left[1 - e^{\frac{-V_{DS}}{U_t}} \right] \quad (D.9)$$

$$I_s = \mu_n C_{ox} \frac{W}{L} (n-1) U_t^2 e^{\frac{-V_{th}}{nU_t}} \quad (D.10)$$

Supposing the two transistors perfectly indetical and their substrate connected to the ground, $V_{D2} = V_{G2} = V_{ref}$ and $V_{DS1} = V_{GS1} = \frac{V_{ref}}{2}$. The injection of those equalities in the current equations yields the current expressions (D.11) and (D.12).

$$I_{d0,SI} = \frac{\beta}{8n} [V_{ref} - 2V_{th}]^2 \quad (D.11)$$

$$I_{d0,WI} = I_s e^{\frac{V_{ref}}{2nU_t}} \left[1 - e^{\frac{-V_{ref}}{2U_t}} \right] \quad (D.12)$$

D.3 Large-signal characteristic calculations

We first rewrite all the common expressions used for both strong and weak inversion modes. Expressions (D.13) and (D.14) are given for the saw-tooth

oscillating periods T_0 in the absence of any magnetic phenomenon and T_{out} when a magnetic perturbation is sensed; I_{d0} and $I_d(t)$ being respectively the currents flowing in the mos diodes in the absence and presence of any induced voltage. The output transfer characteristic $\Omega_{out,SI}$ is evaluated using (D.15).

$$T_0 = \frac{C_L(v_+ - v_-)}{I_{d0}} = \frac{C_L V_{hst}}{I_{d0}} \quad (D.13)$$

$$\bar{T}_{out} = \frac{C_L V_{hst}}{\bar{I}_d(t)} \quad (D.14)$$

$$\Omega_{out} = \sqrt{\left| \frac{T_0}{\bar{T}_{out}} - 1 \right|} \quad (D.15)$$

The induced voltage V_{ind} is supposed as a pure sinusoid (D.16).

$$V_{ind}(t) = |V_{ind}| \cos(2\pi f_e t) = 2\pi f_e \Phi_e \cos(2\pi f_e t) \quad (D.16)$$

D.3.1 Strong inversion mode

The expressions of the diode current in strong inversion mode are first recalled as (D.17) in the absence of any magnetic perturbation and (D.18) when a sinusoidal induced voltage (D.16) is measured.

$$I_{d0,SI} = \frac{\beta}{8n} [V_{ref} - 2V_{th}]^2 \quad (D.17)$$

$$I_{d,SI}(t) = \frac{\beta}{8n} [V_{ref} + V_{ind}(t) - 2V_{th}]^2 \quad (D.18)$$

The mean of the diode current is calculated using (D.19) on one oscillating period T_e of the induced voltage; resulting in expression (D.20).

$$\bar{I}_{d,SI} = \frac{1}{T_e} \int_0^{T_e} I_{d,SI}(t) dt \quad (D.19)$$

$$\bar{I}_{d,SI} = \frac{\beta}{8n} (V_{ref} - 2V_{th})^2 + \frac{\beta}{16n} |V_{ind}|^2 \quad (D.20)$$

Equations (D.17) and (D.20) are next injected in (D.13) and (D.14) to give the saw-tooth periods expressions. The usage of formula (D.15) thus allows for giving the large-signal output characteristic in strong inversion mode (D.21).

$$\Omega_{out,SI} = \frac{|V_{ind}|}{\sqrt{2} (V_{ref} - 2V_{th})} \quad (D.21)$$

D.3.2 Weak inversion mode

Similarly to the strong inversion mode, the expressions of the weak inversion mode currents are recalled as (D.22) and (D.23), $V_{ind}(t)$ being the contribution of the induced voltage (D.16) and I_s the current coefficient given by (D.24).

$$I_{d0,WI} = I_s e^{\frac{V_{ref}}{2nU_t}} \left[1 - e^{\frac{-V_{ref}}{2U_t}} \right] \quad (D.22)$$

$$I_{d,WI}(t) = I_s e^{\frac{V_{ref} + V_{ind}(t)}{2nU_t}} \left[1 - e^{\frac{-V_{ref} - V_{ind}(t)}{2U_t}} \right] \quad (D.23)$$

$$I_s = \mu_n C_{ox} \frac{W}{L} (n-1) U_t^2 e^{\frac{-V_{th}}{nU_t}} \quad (D.24)$$

The mean of the diode current is calculated using (D.25) on one oscillating period T_e of the induced voltage; resulting in expression (D.26). The calculation of the current mean has been performed according to equation (D.27); the function $I_\nu(x)$ being the modified Bessel function of the first kind.

$$\bar{I}_{d,WI} = \frac{1}{T_e} \int_0^{T_e} I_{d,WI}(t) dt \quad (D.25)$$

$$\bar{I}_{d,WI} = I_s e^{\frac{V_{ref}}{2nU_t}} I_{\nu=0} \left(\frac{|V_{ind}|}{2nU_t} \right) - I_s e^{\frac{V_{ref}(1-n)}{2nU_t}} I_{\nu=0} \left(\frac{|V_{ind}(1-n)|}{2nU_t} \right) \quad (D.26)$$

$$I_\nu(x) = \frac{1}{\pi} \int_0^\pi \cos(\nu\theta) \exp(x \cos \theta) d\theta \quad (D.27)$$

The equations (D.22) and (D.26) are finally injected into (D.13), (D.14), and (D.15) in order to give the large-signal output characteristic (D.28) in the weak inversion mode.

$$\Omega_{out,WI} = \sqrt{\left| \frac{I_{\nu=0} \left(\frac{|V_{ind}|}{2nU_t} \right) - 1 + e^{\frac{-V_{ref}}{2U_t}} \left[1 - I_{\nu=0} \left(\frac{|V_{ind}(1-n)|}{2nU_t} \right) \right]}{1 - e^{\frac{-V_{ref}}{2U_t}}} \right|} \quad (D.28)$$

D.4 Small-signal circuit input conductance

The following expressions provide the small-signal input conductance of the circuit. This parameter results from the equivalent conductance g_{in} of the cascoded CMOS diodes as shown in Fig. D.2. The general expression of the input conductance is given by (D.29).

$$g_{in} = \frac{\partial I_{d0}}{\partial V_{ref}} \quad (D.29)$$

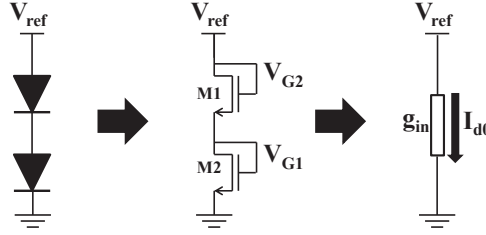


Figure D.2: CMOS cascoded diode input conductance representation

D.4.1 Strong inversion mode

The conductance expression in strong inversion results from the differentiation of the strong inversion current expression (D.11). The input conductance in strong inversion mode is thus given by (D.30).

$$g_{in,SI} = \frac{2I_{d0,SI}}{V_{ref} - 2V_{th}} \quad (D.30)$$

D.4.2 Weak inversion mode

The conductance expression in weak inversion results from the differentiation of the weak inversion current expression (D.12). The input conductance in weak inversion mode is thus given by (D.31).

$$g_{in,WI} = \frac{I_{d0,WI}}{2nU_t} \left(1 + \frac{ne^{\frac{-V_{ref}}{2U_t}}}{1 - e^{\frac{-V_{ref}}{2U_t}}} \right) \cong \frac{I_{d0,WI}}{2nU_t} \quad (D.31)$$

D.5 Magnetic flux formula

The following development explains in detail the magnetic flux calculation. The start point is the expression of the magnetic field generated by a straight

conductor (D.32) and the formula to calculate the magnetic flux (D.33). Following the geometrical configuration shown by fig. D.3, the magnetic flux is calculated.

$$B = \frac{\mu_0 I_w}{2\pi r} \quad (\text{D.32})$$

$$\Phi = \iint B dS \quad (\text{D.33})$$

The third dimension needs no computation, the resulting flux must only be multiplied by L_c . The turn number N_s has also to be considered and has to multiply the flux. The wire thickness is supposed negligible in comparison with the wire radius.

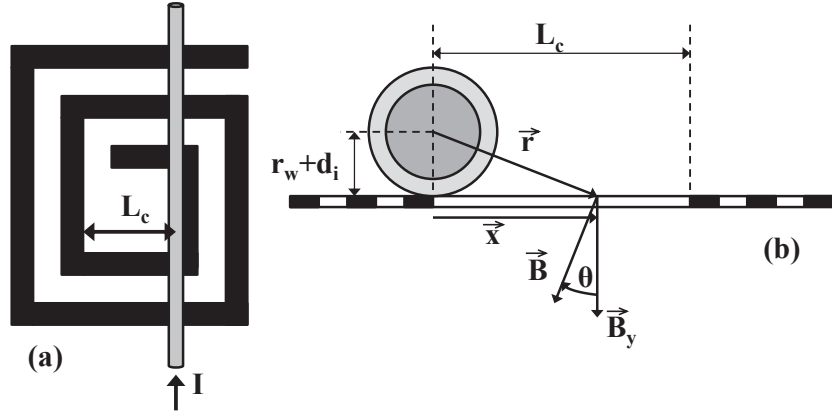


Figure D.3: Sensing device and current line positions

The integral to compute is given by the following expression:

$$\Phi_b = N_s L_c \int_0^{L_c} B_y dx \quad (\text{D.34})$$

The B_y component is the projection of the magnetic field B on the vertical axis:

$$B_y = B \cos \theta = \frac{\mu_0 I_w}{2\pi r} \frac{x}{r} = \frac{\mu_0 I_w}{2\pi} \frac{x}{x^2 + (r_w + d_i)^2} \quad (\text{D.35})$$

The final expression to integrate is:

$$\Phi_b = \frac{N_s L_c \mu_0 I_w}{2\pi} \int_0^{L_c} \frac{x \cdot dx}{x^2 + (r_w + d_i)^2} \quad (\text{D.36})$$

The following variable change provides an elementary expression:

$$u = x^2 + (r_w + d_i)^2 \quad (\text{D.37})$$

$$du = 2x dx \quad (\text{D.38})$$

The transformed expression after adaptating the new integration limits yields:

$$\Phi_b = \frac{N_s L_c \mu_0 I_w}{4\pi} \int_{(r_w + d_i)^2}^{(r_w + d_i)^2 + L_c^2} \frac{du}{u} \quad (\text{D.39})$$

The magnetic flux after the integral calculation is:

$$\Phi_b = \frac{N_s L_c \mu_0 I_w}{4\pi} \ln\left(\frac{L_c^2 + (r_w + d_i)^2}{(r_w + d_i)^2}\right) \quad (\text{D.40})$$

APPENDIX E

Layout views of integrated CMOS circuits

E.1 Schmitt Trigger architecture and layout

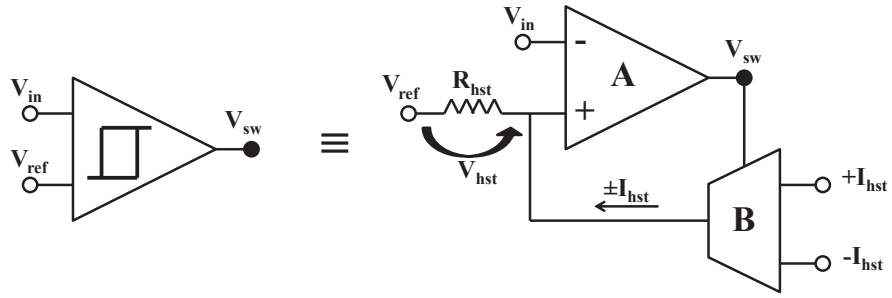


Figure E.1: Schmitt trigger block diagram.

Transistor	Type	W [μm]	L [μm]	V_{th} [V]
M1	NMOS	0.5	1	0.48
M2	NMOS	0.5	1	0.48
M3	PMOS	0.5	1	-0.6
M4	PMOS	0.5	1	-0.6
M5	NMOS	1	1	0.48
M6	NMOS	5	1	0.48
M7	PMOS	4.5	0.35	-0.6
M8	NMOS	1.5	0.35	0.48

Table E.1: Transistors sizes and threshold voltages of Fig. 2.47.

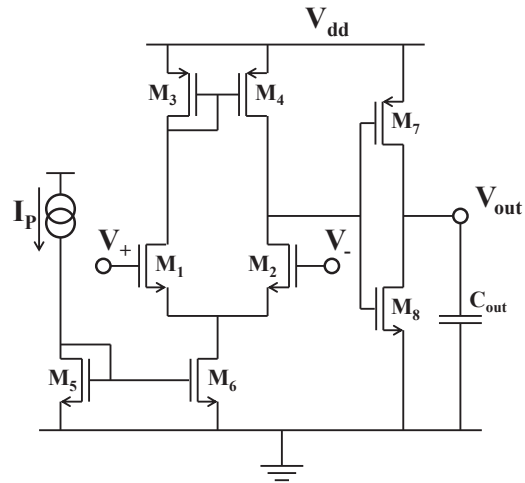


Figure E.2: Input comparator stage.

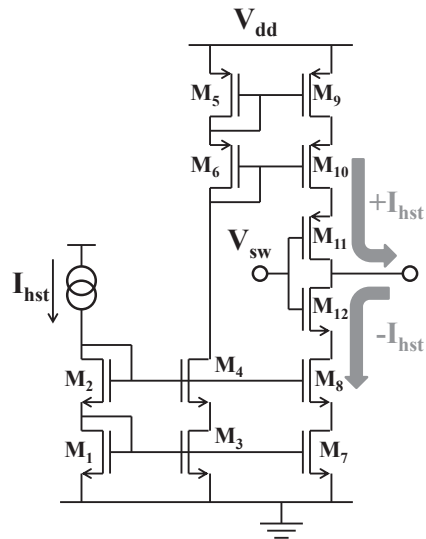


Figure E.3: Hysteresis stage.

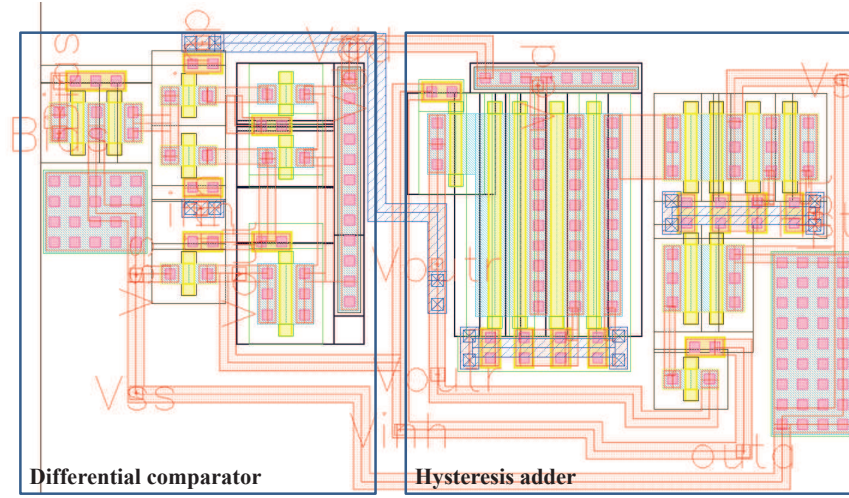


Figure E.4: Schmitt trigger layout view.

Transistor	Type	W [μm]	L [μm]	V_{th} [V]
M1	NMOS	0.3	0.6	0.175
M2	NMOS	0.3	0.6	0.175
M3	NMOS	0.3	0.6	0.175
M4	NMOS	0.3	0.6	0.175
M5	PMOS	0.3	0.6	-0.29
M6	PMOS	0.3	0.6	-0.29
M7	PMOS	0.3	0.6	-0.29
M8	PMOS	0.3	0.6	-0.29
M9	PMOS	3	0.6	-0.29
M10	PMOS	3	0.6	-0.29
MSP	PMOS	3	0.6	-0.29
MSN	NMOS	3	0.6	-0.29

Table E.2: Transistors sizes and threshold voltages of Fig. 2.48.

E.2 Switched current mirrors for IDE sensor interfacing

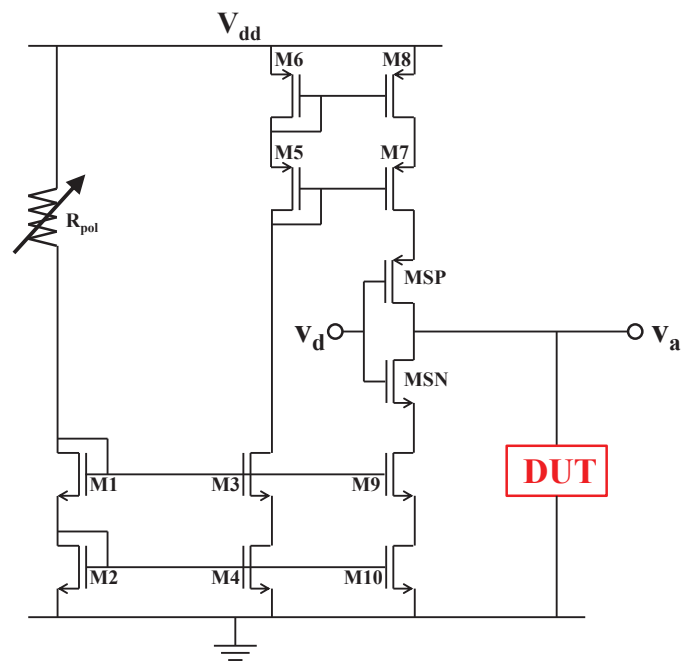


Figure E.5: CMOS Switched current source architecture.

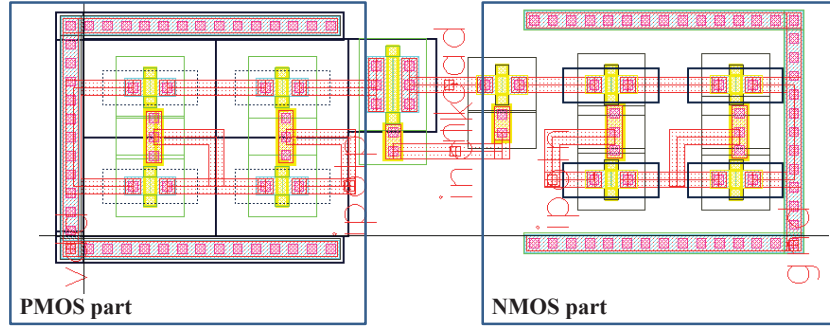


Figure E.6: Interdigitated sensor interface mirrors layout.

Transistor	Type	W [μm]	L [μm]	V_{th} [V]
M1	NMOS	0.3	0.6	0.175
M2	NMOS	0.3	0.6	0.175
M3	NMOS	0.3	0.6	0.175
M4	NMOS	0.3	0.6	0.175
M5	PMOS	0.3	0.6	-0.29
M6	PMOS	0.3	0.6	-0.29
M7	PMOS	0.3	0.6	-0.29
M8	PMOS	0.3	0.6	-0.29
M9	NMOS	0.3	0.6	0.175
M10	NMOS	0.3	0.6	0.175
MSP	PMOS	0.9	0.6	-0.6
MSN	NMOS	0.3	0.6	0.48

Table E.3: Transistors sizes and threshold voltages of Fig. 2.45.

E.3 Current mirrors for inductive interfacing

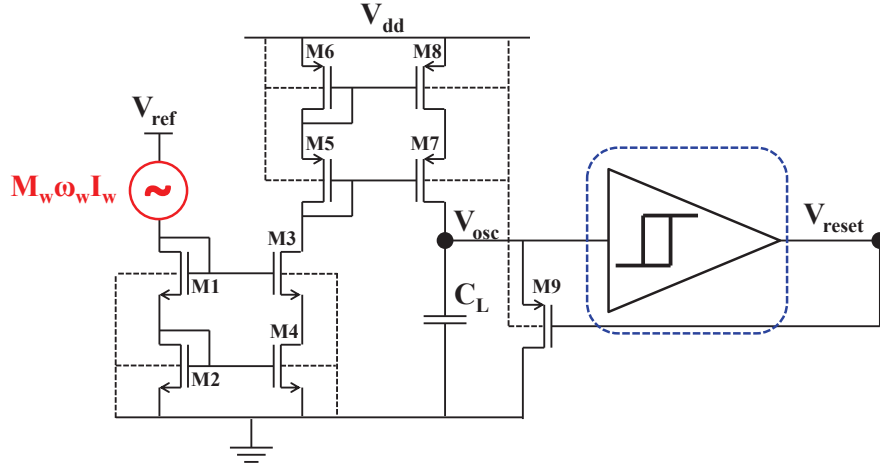


Figure E.7: Current source mirrors CMOS architecture, the inductive device is represented by the red voltage source and the MOS devices bulk connections are drawn with discontinuous lines.

Transistor	Type	W [μm]	L [μm]	V_{th} [V]
M1	NMOS	0.3	0.6	0.175
M2	NMOS	0.3	0.6	0.175
M3	NMOS	0.3	0.6	0.175
M4	NMOS	0.3	0.6	0.175
M5	PMOS	3	0.6	-0.29
M6	PMOS	3	0.6	-0.29
M7	PMOS	3	0.6	-0.29
M8	PMOS	3	0.6	-0.29
M9	PMOS	3	0.6	-0.29

Table E.4: Transistors sizes and threshold voltages of Fig. E.7.

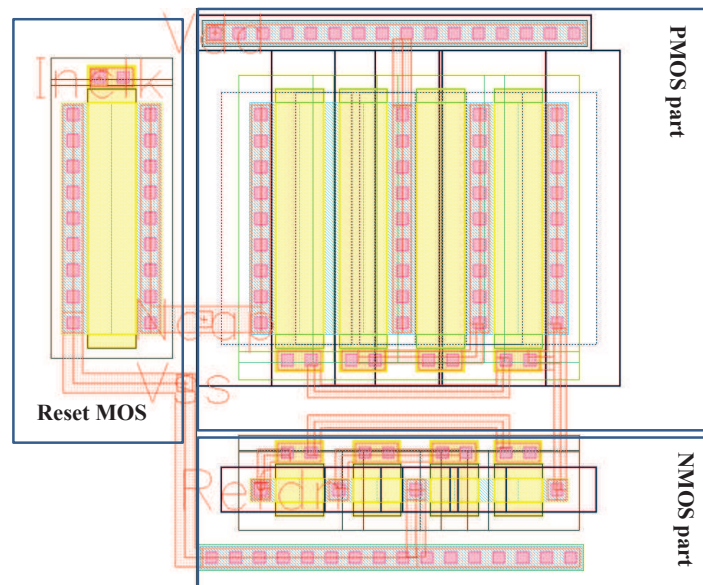


Figure E.8: Magnetic sensor interface mirrors layout.