Doctoral Thesis

Realization of silicon based ultrasound micro-systems

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Realization of Silicon Based Ultrasound Micro-Systems

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Abstract

This thesis describes the conception, design, implementation and experimental characterization of ultrasound micro-systems. Such systems are intended for ultrasound control applications in industrial automation, which so far relied on costly and bulky piezo-ceramic transducers. Two examples are discussed - an ultrasound barrier detector micro-system and a miniaturized range finder. Both application make use of micromachined silicon membrane transducers, which are fabricated in a standard bipolar, CMOS or BiCMOS circuit technology. The membranes are about 1mm by 1mm of size and can be excited thermally to emit ultrasound power at 80-100 kHz. They can also be stimulated by incident ultrasound waves and the resulting membrane vibrations are detected by a piezo-resistive Wheatstone bridge.

The barrier detector micro-system is intended for the presence detection of liquid, transparent or light sensitive objects, especially at places where space is limited. It consists of an ultrasound transmitter/receiver pair facing each other at a distance up to 10 cm. No wired connection exists between transmitter and receiver in order to allow high flexibility within the application. The challenge in the design of the barrier detector is how to secure reliable operation of the system despite the non-idealities of the membrane transducers. Among others these are in particular the temperature and process dependence of the narrow bandwidth (quality factor Q=100) mechanical resonance and the weak sensor signals. An elegant solution is found to provide best performance of the transmitter. By embedding the membrane into the loop of an electro-mechanical oscillator, the latter's oscillation frequency is always determined by the membrane resonance, where the emitted ultrasound power is maximum. For the receiver a simple and efficient self-calibration concept is developed, which is used to maximize the receiver sensitivity. It consists of a high gain (70 dB), ac-coupled amplification stage to boost the extremely weak sensor signals to a sufficient level. A subsequent tuning/tracking circuit automatically tunes
the membrane resonator in order to find and track the highest possible sensitivity. Objects can be detected within 10 ms when they interrupt the sound-path between transmitter and receiver.

The miniaturized range finder system uses the silicon membranes as transmitter/receiver for contact-less short range distance measurement in air. It can determine the distance from an object to the membrane surface up to 10 cm with a 5 mm precision and features an extremely small blind space of 1 cm, where the measurement is not possible. For the micro-system a novel quasi continuous-wave measurement procedure called Δf phase shift method is developed to enable the short range measurement despite the slow thermal excitation of the silicon membrane transducers. The new method, however, requires a coherent beam of ultrasound waves. This is performed by applying the concept of a phase locked loop circuit (PLL) to the tunable membrane resonator. The resulting electro-mechanical PLL is able to lock the emitted ultrasound waves to the electrical signal and hence generates the required coherent beam. The same unique concept is used for the self-calibrating receiver to guarantee zero degree electrical phase shift and at the same time maximum sensitivity. Experimental and theoretical investigations show, that the accuracy is limited by inadvertent cross-coupling effects between transmitter and receiver and multi-path reflections. Both effects are heavily influenced by the micro-system package and their suppression is a demanding task, especially for small package dimensions. An optimized package is presented, with which these disturbing effects can be substantially reduced and the resolution of 5 mm can be achieved.

Both systems are implemented in technologies which allow the co-integration of the silicon transducers and the attached interface electronics (monolithic solution). The system architecture is chosen to allow full integration of all electronic components and - as transmitter and receiver share most of the required building blocks - only one combined transmitter/receiver circuit has to be designed. A monolithic micro-system with such an architecture is a versatile and cost efficient solution and an attractive alternative to the existing piezo ceramic devices.
Kurzfassung

In dieser Dissertation wird die Konzeption, der Entwurf, die Realisierung und Charakterisierung von Ultraschall-Mikrosystemen behandelt, welche für Überwachungsaufgaben in der industriellen Fertigung eingesetzt werden können. Heutzutage werden dafür größtenteils piezokeramische Ultraschallwandler eingesetzt, die oft teuer und voluminös sind. Zwei typische Anwendungen werden vorgestellt — eine miniaturisierte, selbst-kalibrierende Ultraschallschranke und ein Distanzmessermikrosystem. Als Ultraschallwandler dienen mikromechanisch hergestellte, 1mm x 1mm große Siliziummembranen, hergestellt in eine Standard-CMOS, Bipolar- oder BiCMOS Schaltungs technologie. Die Membranen können thermisch angeregt werden und strahlen Ultraschallwellen mit einer Frequenz von 80-100 kHz ab. Eintreffende Schallwellen können die Membranen ebenfalls stimulieren, was mit Hilfe einer am Rand angebrachten piezo-resistiven Wheatstonebrücke detektiert werden kann.

Die Ultraschallschranke ist vorgesehen zur Detektion von transparenten oder lichtempfindlichen Objekten sowie von Flüssigkeiten (Tropfen). Sender und Empfänger des Systems stehen sich in einem Abstand von bis zu 10 cm gegenüber, und die Objekte werden innerhalb von 10 ms detektiert, wenn sie den Schallstrahl dazwischen unterbrechen. Damit die Schranke flexibel eingesetzt werden kann, darf keine elektrische Verbindung zwischen Sender und Empfänger bestehen, ähnlich wie bei einer Lichtschranke. Die Herausforderung bei der Realisierung dieses Mikrosystems ist, dessen zuverlässige Arbeitsweise trotz der zum Teil schwerwiegenden Nachteile der Membranwandler zu garantieren. Diese sind insbesondere die temperatur- und prozessabhängige, schmalbandige Resonanz (Qualitätsfaktor $Q=100$) und die sehr schwachen Sensorsignale. Eine elegante Lösung lässt sich für den Sender finden: Durch Einbettung der Membran in die Rückführung eines elektromechanischen Oszillators wird automatisch sichergestellt, dass die thermische Anregungsfrequenz immer mit der Membranresonanz identisch ist. Die abgegebene Schallleistung ist maximal, unabhängig von Tem-
peratureinflüssen oder Fabrikationstoleranzen. Der selbst-kalibrierende Empfänger besteht aus einem ac-gekoppelten 70 dB-Eingangsverstärker, welcher die winzigen Sensorsignale auf einen genügend hohen Pegel anhebt. Die nachfolgende raffinierte und effiziente elektronische Regelung gleicht die Empfangsmembran immer auf die Frequenz maximaler Empfindlichkeit ab.


Beide Mikrosysteme werden in Standardtechnologien hergestellt, was die gemeinsame Integration der Siliziummembranen mit der benötigten Ansteuerelektronik zulässt (monolithisches Mikrosystem). Die Systemarchitektur ist in beiden Fällen so gewählt, dass alle Funktionen vollständig integriert werden können und dass insbesondere eine effiziente Kombination von Sender und Empfänger im selben Chip ermöglicht wird. Ein Mikrosystem mit einer derartigen Architektur wird dann zu einer kostengünstigen, flexiblen und daher attraktiven Konkurrenz für viele bestehende piezo-keramische Lösungen.
Chapter 1

Introduction

1.1 Micro-system technology

The miniaturization of technical components and systems has always been a topic of great interest in scientific and industrial applications. The example maybe most famous for the illustration of this trend is the development of the computer, which led from room-sized, slow and unreliable machines for simple calculation tasks to today's ultrafast, lightweight laptops for all kinds of applications. In the particular field of sensors and actuators as the main objects of interest in this thesis, a similar strong tendency towards miniaturized products exists for various reasons. A reduction in size of a sensor or actuator can lead to an increase in its applicability through:

- lower weight (greater portability)
- lower manufacturing cost (less material)
- increased functionality (sensor system)

Many different technologies have been proposed for the fabrication of miniaturized sensors and actuators, also called transducers, or even combined devices incorporating both a sensor and an actuator.
The miniaturized devices are called *micro-sensors*, *micro-actuators* or *micro-transducers*, accordingly. However, one of the most promising approaches towards miniaturized sensors and actuators is the use of silicon wafers, normally used for the fabrication of electronic circuits, as the basic sensor/actuator material. Monocrystalline silicon is known to have many promising physical properties suitable for various types of sensors and actuators [1, 2]. These can be fabricated using similar processing steps as used in the manufacturing of integrated electronic circuits (ICs). As a very useful consequence, batch fabrication of the sensors/actuators is possible as well as the co-integration of sensing elements and electronic signal conditioning circuits on the same piece of silicon. The combination of micro-sensors/actuators together with integrated electronics is called a *micro-system* or *smart-sensor/actuator* and has lead to several successful commercial products like pressure sensors or acceleration meters [3, 4].

A notable trend in this context, which is also present in this thesis, is the growing effort to combine micro-sensors/actuators and integrated circuits using the same IC-technology to form smart sensors and sensor systems (*monolithic micro-system*). The benefits of realizing monolithic micro-systems using an IC-technology is not only to combine the sensors and the circuits on the same die, but also to tap the strong manufacturing base of these technologies, which have perfected themselves in low-cost mass production. While sensors and sensor systems entirely based on silicon may not always boast better sensitivity or further range, etc. than those implemented in specialized sensor technologies and hybrid circuit configurations, the advantages of monolithic sensor systems lie in that they can be manufactured inexpensively, and their performance is enhanced by on-chip conditioning circuits.

In a great variety of today's existing micro-systems [5], the sensor or actuator part is based on *micro-machined* structures. This can be e.g. the seismic mass of an accelerometer or sensitive membranes, beams and bridges. Thin silicon membranes for example can be obtained by processing silicon wafers with an electro-chemical etching step (for micro-machining techniques see e.g. [2, 6]). As an illustration, Fig. 1.1 shows the cross-section of four typical micro-systems with increasing degree of integration from hybrid (a) to monolithic (d). Fig. 1.1a shows the cross-section of a hybrid accelerometer. The sensor head consists of a sandwich structure of several micro-machined silicon layers which are
1.2. Smart sensors and smart actuators

 bonded together. The read-out or interface-electronics is implemented in a separate chip and connected to the sensor head by short bonding wires. In Fig. 1.1b the sensor is fabricated using a silicon technology which is optimized for a particular application by adding special sensitive layers or structures on the sensor area. This specialized sensor technology is not intended to implement electronic circuits, so that these have to be again realized in an additional chip. Fig. 1.1c and d show monolithic micro-systems. The difference between the two is only the additional sensitive layer/structures which cover the sensor area in Fig. 1.1c whereas in Fig. 1.1d a standard IC technology is used.

1.2 Smart sensors and smart actuators

The downsizing of existing sensors/actuators using silicon technologies not only leads to the desired miniaturized products, but unfortunately sometimes is also accompanied by a substantial reduction of performance. Micro-sensors often suffer from weak sensitivity to the measurand (physical quantity to measure), offset, bad linearity or undesired temperature dependence.

On the other hand, in a micro-system or smart sensor/actuator, these imperfections can partially be made up by an integrated circuit which is attached to the micro-sensor/actuator [7]. It performs e.g. low
noise and high gain amplification of the weak sensor signals or achieves a temperature compensation of the measurand. Moreover, the IC can include additional (high level) features like the interface to a superior system or a built-in self-test of the whole micro-system. Fig. 1.2 shows a simplified block diagram of a typical smart sensor/actuator with all the relevant functional parts. A smart sensor (Fig. 1.2a) consists of the following:

a) Smart sensor

b) Smart actuator

**Figure 1.2: Smart sensors and smart actuators**

- **Package**: Protects the micro-system from environmental influences like dust, humidity or light. On the other hand, it serves as a guide for the measurand to reach the sensitive silicon surface. The two functions are sometimes quite conflicting and the sensor performance can be heavily be diminished by a poor package design. In a micro-system, the package design has to be considered carefully in order to get maximum performance.
1.2. Smart sensors and smart actuators

- **Micro-sensor**: Micro-machined silicon sensor, which translates the physical quantity (measurand) into an electrical signal. This can be either voltage, current or charge. According to their operating principle, micro-sensors are divided into two types [6]: *Modulating* or *active* sensors need an additional energy source to generate an electrical output signal (e.g. sensors with a Wheatstone bridge), whereas *self-generating* or *passive* sensors operate on its own like the well known thermocouple. In active sensors, the physical and electrical properties can be influenced by the additional energy source(s), which enables automatic sensor adjustment or correction of non-idealities (see below).

- **Front-end amplifier**: Low noise and high-gain amplifier directly connected to the micro-sensor terminals. Besides the performance of the micro-sensor, the signal level and quality of the whole micro-system are basically determined by this amplifier.

- **Interface circuit**: Provides the connection of the sensor to a superior system. This means further processing of the amplified sensor signal to achieve a signal form which is acceptable for the system. Often analog-to-digital (A/D) conversion is performed for an easy link of the smart sensor to a digital bus system as well as subsequent digital processing by a CPU.

- **Conditioning circuit**: The conditioning circuit is used to compensate for the physical and electrical imperfections of the micro-sensor or to adjust the latter's operating range. The imperfections can be e.g. offset, nonlinearity, temperature dependence or the package influence mentioned above. Compensation is performed by feeding a compensating signal (e.g. from a temperature sensing device) to the active sensor, where it cancels the undesired physical or electrical effect. With the feedback loop shown in Fig. 1.2, even more advanced features like offset compensation or linearity improvement can be realized. Some typical examples for micro-systems with conditioning circuits can be found in [8].

A smart actuator, shown in Fig. 1.2b, can be described by a block diagram similar to that of a smart sensor. The main difference is only the direction of the signal flow, which is from the electrical input to the physical output rather than the opposite direction. The description of the individual blocks is almost identical to the ones described in
the smart sensor above and is therefore not discussed in much detail. Instead of the front-end amplifier in the smart sensor, an *excitation circuit* builds the interface between the electronic circuit and the micro-actuator. The task of this block is to translate the internal signal into electric energy, which is delivered to the actuator in a form needed to excite it.

### 1.3 Design method for micro-systems

The design of a micro-system for a particular application requires careful design planning. The main reason for that comes from the fabrication procedure of miniaturized silicon structures, both sensors/actuators and electronic circuits. Apart from costly trimming it is (almost) impossible to modify such devices after the micro-structures are completed. In a poor design, the performance of the micro-system therefore can hardly be improved, even by additional external circuits or devices. If feedback loops between sensor/actuator and the electronics exist, like shown in Fig. 1.2, the situation is even more delicate, because poorly designed closed loop systems can become unstable and the micro-system then is useless. However, to successfully develop a 'first-time-right' micro-system nonetheless, a safe step-by-step design flow should be followed.

Such a step-by-step design flow is proposed and illustrated in Fig. 1.3. The sequence shows the required design steps necessary for the development of a monolithic smart sensor micro-system. Again, only minor changes have to be considered, if the design of a smart actuator is the object of interest.

In the present research, several examples of micro-systems have been successfully developed by consequently following the design guidelines illustrated in Fig. 1.3. For two ultrasound micro-system applications described in this work, integrated circuits for both smart sensors and smart actuators have been designed, fabricated and experimentally characterized.
1.3. Design method for micro-systems

- System specifications
  - Characterization and modelling of micro-sensor: Definition of a physical or empirical model which takes into account the influence of the package.
  - Micro-system measurement concept: Depending on the previously derived model, an IC-compatible measurement concept is chosen.
- Feasibility study:
  - Implementation of the micro-system using discrete components and the micro-sensor.
  - Characterization: Determine system performance.
  - IC-specification: Extraction of parameters for the integrated circuit using results from feasibility study.
  - IC Implementation: Design of the IC for a hybrid micro-system.
  - Characterization of hybrid micro-system.
- Monolithic micro-system:
  - Combine micro-sensor and IC on the same die.
- Characterization of monolithic micro-system.

Smart sensor

Figure 1.3: Design flow of a smart sensor.
1.4 Ultrasound transducers and applications

The focus of the present research is on the design of silicon micro-systems for industrial applications using ultrasound. The generation and detection of the ultrasound waves is performed by miniaturized silicon transducers which have recently been realized and which are promising in terms of monolithic fabrication and cost effectiveness. Besides the already well established micro-system products like pressure transducers or accelerometers, the miniaturized ultrasound sensors and systems may gain much attention in the future, since ultrasound can be used for many different applications.

Today's ultrasound technology covers an enormous range of measuring, diagnosing and other applications. From non-destructive testing to medical scanning, from surveillance in process plants to surface acoustic wave (SAW) filters. According to the broad spectrum of the ultrasound applications, the operating principle of the ultrasound sensors, actuators and transducers also show up in many different forms [9]. The most important transducer type, however, is the piezoelectric transducer which works at operating frequencies from around 100 kilohertz up to 50 megahertz. Piezoelectric ultrasound transducers are typically used for surveillance tasks in industrial automation or process plants by performing the following sensor functions among many others:

- Detect the level of a fluid.
- **Barrier detector.** Used to detect and count pieces.
- Ultrasound *sonar* or range finder.
- Measuring the density of a fluid or a gas.
- Flow-meter.

Since the device generates reasonable ultrasound power, it can be used as broad-band transducer and exhibits enough sensitivity to incident ultrasound waves. As a remarkable drawback, however, these transducers are rather bulky devices and in addition quite expensive to fabricate. Consequently, a considerable motivation towards miniaturized ultrasound devices exists, in particular for the ultrasound applications mentioned above. Much research has been done in the field of
ultrasound micro-transducers and many different solutions have been presented up to now. Fig. 1.4 shows three examples of different approaches providing miniaturized transducers for ultrasound in air.

![Capacitive ultrasound transducer based on a BiCMOS process](image)

![Micromachined capacitive ultrasound transducer](image)

![Thermal excited ultrasound transducer](image)

**Figure 1.4:** Three examples of micro-transducers for ultrasound in air. a) and b) Electrostatic driven transducers. c) Electro-thermal operating transducer used in this research.

The two devices depicted in Fig. 1.4a and b [10, 11] are electrostatic driven (capacitive) transducers. They generate or detect ultrasound in air around 10 MHz and need high biasing voltages of 20...100V for proper operation. Because of this excessive voltage, both types can hardly be used as part of a hybrid or monolithic micro-system. The voltage level is too high for standard VLSI (very large scale integration) circuits, which operate around 5 V. On the other hand, the microtransducer shown in Fig. 1.4c [12] is a very promising device for an ultrasound micro-system - for two reasons. First, it is fabricated using a standard CMOS (complementary metal oxide semiconductor) or bipolar technology which makes it suitable for an inexpensive monolithic
Chapter 1. Introduction

micro-system solution. Second, it operates at voltage and current levels which can be handled easily by integrated circuits. This transducer operates in a frequency range which is two orders of magnitude lower than the other two examples (around 100 kHz) which makes it well suited for ultrasound applications in air or gases. The physical properties and the fabrication of the device has been investigated intensively in two PhD studies [13, 14] and it is used in this work to develop the two ultrasound applications for industrial automation.

1.5 Organization of the thesis

In this thesis, the realization of silicon micro-systems is discussed by means of examples of ultrasound applications using miniaturized ultrasound transducers. Consequently, the organization of the thesis corresponds to the systematic design flow of micro-systems presented in the previous section (Fig. 1.3).

In Chapter 2 the operation principle of the silicon ultrasound transducers is explained. The focus hereby is on how the devices can be used as part of an ultrasound micro-system rather than explaining the detailed physical and acoustic properties.

Chapters 3 to 5 present applications to the transducers. The chapters also cover more detailed insight into the transducer characteristics, whenever required for the particular application.

Chapter 3 describes the design of a miniaturized barrier detector micro-system which extensively relies on dedicated front-end and conditioning electronics to overcome the imperfections of the silicon membrane transducers.

Chapter 4 presents a feasibility study of an ultrasound sonar for short range applications (1...10 cm). First, measurement techniques are devised, which are suitable for the special characteristics of the miniaturized transducers. Then circuit concepts are developed, which enable the silicon transducers to be used as ultrasound transmitter/receiver of the sonar and which are compatible to a subsequent integration in an IC technology.
1.5. Organization of the thesis

Chapter 5 finally covers the implementation of a CMOS mixed analog/digital front-end ASIC (application specific integrated circuit) for the range finder application based on the specification of the feasibility study of chapter 4.

The results and achievements of the experimental work during the studies are discussed in a concluding chapter 6, which also includes suggestions, approaches and applications for future work in the field of ultrasound micro-systems.
Chapter 2

Thermally excited micro-transducer

2.1 Operating principle

The micro-transducer considered in the present research is a thermally excited silicon membrane resonator [12, 13, 14], whose simplified cross-section and top view schematic are shown in Figs. 1.4c and 2.1, respectively. It consists of a 1 mm by 1 mm silicon membrane of 5-6 μm thickness, which is obtained by etching away the underlying bulk silicon of a finished wafer of standard bipolar or CMOS technology. The silicon membrane is released by using an electro-chemical etching technique [15, 6], which results in a uniform and reproducible membrane thickness. In the middle of the membrane, low ohmic resistors are diffused for the thermal excitation of the transducer as an ultrasound actuator. As a sensor, the piezoresistive Wheatstone bridge placed at the border of the membrane picks up the mechanical vibrations of the latter and translates them into an electrical signal \( V_{\text{out}} \).

The device can be used to generate or detect ultrasound waves at frequencies around 100 kHz. As an ultrasound actuator or transmitter, the membrane is excited thermally to vibrate and emit ultrasound by applying an ac voltage \( V_{\text{ac}} \cdot \cos(\omega t) \) superimposed on a dc voltage \( V_{\text{dc}} \).
to the heating resistor $R_h$.

![Schematic top-view of the membrane resonator](image)

**Figure 2.1:** Schematic top-view of the membrane resonator (see also Fig. 1.4c).

The resulting thermal power $P$ generated at the resistor $R_h$ is

$$P = \frac{1}{R_h} \left[ V_{dc}^2 + \frac{1}{2} V_{ac}^2 + 2 V_{dc} V_{ac} \cos \omega t + \frac{1}{2} V_{ac}^2 \cos 2\omega t \right]$$

$$= P_{\text{stat}} + P_{\text{dyn1}} \cos \omega t + P_{\text{dyn2}} \cos 2\omega t \quad (2.1)$$

The membrane resonator can be stimulated by both dynamic heating power, $P_{\text{dyn1}} \cos \omega t$ and the harmonic $P_{\text{dyn2}} \cos 2\omega t$. The heating power causes a dynamic temperature profile around the heating resistor $R_h$, which results in an oscillating bending moment due to the different thermal expansion coefficients of the layers incorporated in the membrane’s sandwich structure. If the excitation frequency $\omega$ or $2\omega$ match the mechanical resonant frequency of the membrane, its oscillation amplitude becomes very large and the device emits ultrasound waves with sufficient pressure. In a first order approximation, the level of the emitted ultrasound pressure is directly proportional to the applied dynamic heating power.
2.2. Transducer efficiency

On the other hand, the static heating power \( P_{\text{stat}} \) causes a constant mechanical stress in the membrane, which leads to a slight buckling of the latter. The buckling effect on his part substantially influences the membrane's resonant frequency, up to \( \pm 7\% \) of the initial value. This effect can be used to tune the resonant frequency of the membrane by simply applying different dc heating voltages \( V_{\text{dc}} \), which turns out to be very useful feature for the efficient usage of the silicon micro-transducer.

When the transducer is used as an ultrasound detector or receiver, the membrane acts as a modulating or active sensor (according to the sensor classification mentioned in chapter 1.2). The membrane deflections are measured by a piezo-resistive strain gauge (Wheatstone bridge) and proportionally translated into an electrical signal. Incident ultrasound waves around the resonant frequency can excite the membrane large enough to produce measurable electrical signals with an amplitude of 10 \( \mu \text{V} \) to 250 \( \mu \text{V} \).

2.2 Transducer efficiency

A major goal in the design of a micro-system is the optimization of the micro-sensor/actuator performance. For the described ultrasound membrane transducers this means maximizing both the efficiency of the ultrasound generation for high ultrasound pressure and the sensitivity to incident ultrasound waves. These optimization tasks are not only a question of how the micro-sensor/actuator is designed and fabricated, but also how properly it is used in conjunction with the attached interface electronic:

The amount of ultrasound pressure delivered by the vibrating membrane transducer directly depends on the dynamic heating power \( P_{\text{dyn1}} \) or \( P_{\text{dyn2}} \), generated by the resistor \( R_h \). To achieve maximum ultrasound pressure, the dynamic heating power therefore has to be as high as possible. For a given supply voltage \( V_{dd} \), the excitation voltage \( V_h = V_{ac} \cos \omega t + V_{dc} \) across the heating resistor may look like shown in Fig. 2.2a. Since the maximum and the minimum voltage for \( V_h \) is limited by the two supply rails, operating conditions for \( V_{ac} \) and \( V_{dc} \) can be derived and lead to the dark shaded operating range in Fig. 2.2b. For \( V_{ac} = V_{dc} = V_{dd} / 2 \), both quantities \( P_{\text{dyn1}} \) and \( P_{\text{dyn2}} \) achieve the absolute maximum according to the Eq. 2.1.
Figure 2.2: a) Excitation voltage \( V_h = V_{ac}\cos\omega t + V_{dc} \) and b) operating range for a given supply voltage \( V_{dd} \).

In Eq. 2.2 the heating power \( P_{dyn1} \) is four times larger than \( P_{dyn2} \). It is therefore much more efficient to use the term \( P_{dyn1} \cos \omega t \) for the thermal excitation of the membrane than trying to stimulate it with the harmonic \( P_{dyn2} \cos 2\omega t \). Besides, the emitted ultrasound then has the same frequency as the electrical excitation signal which is convenient for most applications, especially for those described in this thesis.

The dynamic heating power to excite the membrane becomes highest, if a rectangular signal \( V_h = V_{dc} + V_{ac} \text{rect}(\omega t) \) \(^1\) is applied to the heating resistor, as indicated with the dotted line in Fig. 2.2b. The power established at the resistor \( R_h \) is

\[
max(P_{dyn1}) = \frac{1}{2} \cdot \frac{V_{dd}^2}{R_h}, \quad max(P_{dyn2}) = \frac{1}{8} \cdot \frac{V_{dd}^2}{R_h} \quad (2.2)
\]

\(^1\text{rect}(\omega t) = \frac{4}{\pi} [\cos \omega t - \cos \frac{3\omega t}{3} + \cos \frac{5\omega t}{5} - ...] \)
2.3. Characterization and modeling

$$P = \frac{1}{R_h}[V_{dc}^2 + V_{ac}^2 + 2V_{dc}V_{ac} \text{rect}(\omega t)]$$

$$= V_{dc}^2 + V_{ac}^2 + \frac{8}{\pi} V_{dc}V_{ac} \left[ \cos \omega t - \frac{\cos 3\omega t}{3} + ... \right]$$

$$= P_{\text{stat}} + P_{\text{dyn1}} \cos \omega t + P_{\text{dyn3}} \cos 3\omega t + ...$$

where the fundamental component $P_{\text{dyn1}} \cos \omega t$ again is the strongest contribution of the heating power and therefore preferred for the excitation of the membrane. Maximum power is achieved for $V_{dc} = V_{ac} = V_{dd}/2$ and is calculated to be

$$\max(P_{\text{dyn1}}) = \frac{2}{\pi} \cdot \frac{V_{dd}^2}{R_h}$$

which is 27% higher than in case of the sinusoidal signals investigated in the initial contributions on the silicon membrane transducers [12]. Rectangular excitation therefore is the preferred concept for achieving best ultrasound efficiency.

2.3 Characterization and modeling

It is crucial for the implementation of a micro-system to have comprehensive knowledge about the physical and electrical properties of the micro-sensor/actuator. Well understood and characterized micro-sensors /actuators enable both the physical performance optimization of the latter as well as a successful design plan of the whole micro-system (see also Fig. 1.3).

2.3.1 Transfer function

To investigate the (non-linear) electrical-thermal-mechanical behavior of the membrane transducer and to derive a practical model for the application, it is convenient to consider the transducer as a linearized, electrical two-port device with a characteristic, frequency dependent transfer
Chapter 2. Thermally excited micro-transducer

a) Ultrasound

\[ V_{in} = V_{ac}(j\omega) + V_{dc} \]

\[ V_{out} = V_{sig}(j\omega) + V_{off} \]

b) Transfer Function of Membrane for \( V_{dc}=3.45V, V_{ac}=100mV \) or \( P_{stat}=85mW, P_{dyn1}=5mW \) @ \( R_{h}=140\Omega \)

\[ f_0 = 91.4 \text{ kHz} \]

\[ f_{3dB} = 160 \text{ Hz} \]

Figure 2.3: Small signal membrane transfer function \( G_M(j\omega) = \frac{V_{sig}(j\omega)}{V_{ac}(j\omega)} \), \( \omega = 2\pi f \) from \( f=10 \text{ Hz} \) to \( 400 \text{ kHz} \). The membrane area is \( 0.84 \text{ mm} \times 0.84 \text{ mm} \) and the device is fabricated in a \( 2\mu m \) CMOS technology from AMS (Austria Mikro Systeme).

The transfer function can be divided into two significant regions, one around the membrane resonant frequency \( f_0 \) and the other between
2.3. Characterization and modeling

dc up to about 3 kHz. At low frequencies the membrane acts like an inverting first order low-pass filter with a cut-off frequency of 160 Hz. Slow varying input signals quasi-statically alter the heating power $P_{\text{stat}}$, which is directly translated into different buckling height and converted back to voltage by the Wheatstone bridge. The roll-off at 160 Hz is due to the large thermal time constant of the membrane, which is characteristic for a given technology and membrane design. For the generation of ultrasound pressure, this low frequency behavior has negligible significance. It is important, however, when the membrane tuning capabilities are investigated. The resonant frequency, which depends on $P_{\text{stat}}$, can only be changed as fast as allowed by the time constant of the low pass mentioned above.

Around $f=f_0=91.4\, \text{kHz}$ the transfer function is that of a high $Q$ ($Q \approx 100$) bandpass resonator with zero degrees phase shift at the resonance. There, the membrane emits maximum ultrasound pressure as an ultrasound transmitter, and exhibits best sensitivity to incident ultrasound waves when used as a receiver. Outside the narrow 3 dB bandwidth of the bandpass characteristic, the excitation is attenuated significantly and the gain of $G_M(j\omega)$ falls rapidly to levels 30 dB below the peak at $f_0$. The emitted ultrasound in this frequency range is tiny and the already small sensitivity virtually becomes zero. Consequently, it is the most important aspect in the design of a micro-system using these silicon micro-transducers to ensure best matching of the excitation frequency to the mechanical resonance. This turns out to be a rather challenging task, because the mechanical resonant frequency $f_0$ is not constant but strongly depends on technology spread, temperature effects and the package influence.

Fig. 2.4 shows a close up view of $G_M(j\omega)$ in the frequency range 80 kHz ... 110 kHz, for increasing dc heating voltages $V_{dc}$ or static heating power $P_{\text{stat}}$, respectively. As expected by previous consideration, the resonant frequency $f_0$ is different for each $V_{dc}$. Moreover, it also depends on the amplitude of the ac excitation $V_{ac}$ as can be seen in the measurements of Fig. 2.4. The reason for this behavior is that a strong amplitude $V_{ac}$ also contributes to static heating power $P_{\text{stat}}$ according to the basic equation 2.1. The two sets of measurement curves in Fig. 2.4 for $V_{ac}=100\, \text{mV}$ and $V_{ac}=-2\, \text{V}$ reflect the two basic operating modes of the membrane: transmitter and receiver. For weak excitation signals ($V_{ac} \ll V_{dc}$) the electrical transfer function $G_M(j\omega)$ exactly re-
Figure 2.4: The transfer function $G_M(j\omega)$ for different dc heating voltages $V_{dc}$. Weak excitation ($V_{ac}=100 \text{ mV}$) does not give the same transfer function as strong excitation ($V_{ac}=2 \text{ V}$)!

Reflects the frequency dependent sensitivity to incident ultrasound, which is the receiving mode of the membrane. On the other hand, strong excitation signals are needed when the membrane is used as transmitter. The characteristic transfer function $G_M(j\omega)$ then may look like the one for $V_{ac}=2 \text{ V}$ and all resonant frequencies are slightly larger compared to weak excitation.

In Fig. 2.5 this important relationship $f_0$ vs. $V_{dc}$ for weak and strong excitation again is illustrated. It is slightly non-linear but monotonic and the gradient $K_f = \Delta f_0/\Delta V_{dc}$ for weak excitation achieves values 9...9.4 kHz/V in the interval $V_{dc}=2.5...4.5 \text{ V}$. $K_f$ turns out to be one of the critical design parameters for the ultrasound micro-systems and will be described in more detail when the focus is on the application (chapter 3-5).
2.3. Characterization and modeling

Resonant Frequency vs. DC Heating Voltage

![Graph showing the relationship between resonant frequency and DC heating voltage.](image)

Figure 2.5: Measured resonant frequency as a function of the heating voltage $V_{dc}$ for weak and strong excitation.

2.3.2 Membrane offset

As already mentioned before, the output of the transducers $V_{out}$ is composed of an ac signal $V_{sig}(j\omega)$ and a dc component $V_{off}$. While $V_{sig}(j\omega)$ reflects the frequency dependent excitation of the membrane, the term $V_{off}$ represents a (undesired) constant offset voltage which appears at the terminals of the Wheatstone bridge. The offset voltage is caused by two different effects. The first is the static buckling of the membrane, which is induced by both initial compressive stress, and the static power $P_{stat}$ applied to the heating resistor. As the piezo-resistive Wheatstone bridge measures the bending of the membrane, this buckling can translate into a substantial dc voltage. The second contribution to offset voltage is due to the technology dependent mismatch between the diffused piezo-resistors of the Wheatstone bridge. Both effects can lead to offset voltages as high as 80 mV, which in case of an ultrasound receiver is up to 80 dB (!) higher than the sensed ultrasound signal. Careful
consideration have to be accomplished in the design of subsequent front-end amplifiers, which should amplify the sensed ac signal rather than be saturated by applied dc offset (see Chapter 3).

### 2.3.3 Small signal model

![Small signal band pass model of the membrane resonator including parasitic coupling effects.](image)

**Figure 2.6**: *Small signal band pass model of the membrane resonator including parasitic coupling effects.*

In the vicinity of the resonant frequency, the membrane small signal behavior can be characterized accurately by an empirical second order band pass model. The differential equations which describe the mechanical membrane resonator, the electrical-thermal-mechanical excitation and the behavior of the piezo-resistive Wheatstone bridge can all be represented by a lumped, electrical RLC resonator as depicted in Fig. 2.6. The advantage of using electrical network elements for the modeling of a mixed electrical - non-electrical object like the present membrane resonator is that most of the relevant membrane characteristics can then be studied easily. In addition, the model can be used effectively during the design and verification of a complete micro-system. Thanks to its pure electrical nature, the model is convenient and fast to analyze together with the interface electronics using numerical network simulators like e.g. SPICE.
2.3. Characterization and modeling

In the membrane model illustrated in Fig. 2.6, resistor $R_h$ and controlled voltage source $E_1$ describe the electro-thermal excitation, whereas $E'_2$ models incident ultrasound. $E_2, E_3$ and $R_o$ model the piezo-resistive bridge, and capacitors $C_1$ and $C_2$ represent parasitics associated with layout and the packaging. The model parameters are identified by a set of appropriate measurements. The quantities for $R_h$ and $R_o$ can be obtained by a simple static impedance measurement. To determine the intrinsic resonator parameters $R, L$ and $C$ and the gain of the membrane $G_o$, the membrane model characteristics as a function of the described parameters has to be analyzed first. Thereby, it is reasonable to consider the parasitics $C_1$ and $C_2$ to be negligible initially.

In the frequency domain, the band pass transfer function of the membrane model is

$$G_M(s) = \frac{V_o}{V_i} = \frac{G_o}{RLC} \cdot \frac{sL}{s^2 + \frac{s}{RC} + \frac{1}{LC}} = K \cdot \frac{\omega_o s}{s^2 + \frac{\omega_o s}{Q} + \omega_o^2}$$

(2.5)

where $\omega_o = 2\pi f_o = \sqrt{\frac{1}{LC}}$ is the resonant frequency, $Q = R\sqrt{\frac{C}{L}}$ the quality factor and $K = \omega_o \frac{L}{R} G_o$ the gain. The quantities for the model parameters $R, L, C$ and $G_o$ can be extracted by a frequency domain measurement using a network analyzer. Unfortunately, this simple and ideal band pass model does not explicitly describe the small signal behavior of the membrane resonator. Parasitic capacitive cross coupling from input to output due to non-symmetrical membrane layout and package can have significant influence on the resonator performance. To take into account the parasitic capacitive effects, the capacitors $C_1$ and $C_2$ are included in the model as shown by the dashed network in Fig. 2.6. The improved transfer function $G'_M(s)$ of the membrane then results in

$$G'_M(s) = \frac{s \Delta \omega (s^2 + \frac{\omega_o^2}{Q_1} s + \omega_o^2)}{(s^2 + \frac{\omega_o^2}{Q_2} s + \omega_o^2)(s + \omega_1)(s + \omega_2)}$$

(2.6)

where
\[ \omega_1 = \frac{1}{R_o C_1}, \quad \omega_2 = \frac{1}{R_o C_2}, \quad \Delta \omega = \omega_2 - \omega_1 \]  \hspace{1cm} (2.7)

and

\[ \omega_z = \omega_0 \sqrt{1 + \frac{K}{\omega_o} \cdot \frac{\omega_1 \omega_2}{\Delta \omega}}, \quad Q_z = \frac{\omega_z}{\omega_o} \cdot \frac{Q}{1 + Q \frac{K}{2} \frac{\omega_1 + \omega_2}{\Delta \omega}} \]  \hspace{1cm} (2.8)

As can be seen in Eq. 2.6, the parasitic capacitors \( C_1 \) and \( C_2 \) generate two additional poles at \( s = -\omega_1 \) and \( s = -\omega_2 \) as well as a (complex) pair of zeros with the characteristic frequency \( \omega_z \). The two additional poles are typically far away from the resonant frequency, since the parasitic capacitances are small (as an example: \( R_o = 2 \, k\Omega, C_1, C_2 \approx 5 \, pF \rightarrow \omega_1, \omega_2 \approx 2\pi \cdot 16 \, MHz \)). They do not influence the transfer function within the critical frequency range around the membrane resonant frequency. On the other hand, the pair of zeros can have a significant impact on the membrane characteristics, especially when the parasitics \( C_1 \) and \( C_2 \) have pretty different values. The effect can be demonstrated easiest when the transfer function \( G'_M \) is plotted for different capacitance ratios as shown in Fig. 2.7 where \( C_1/C_2 = 1 + m, \, m=0.8\% \ldots 0.8\% \).

The zeros cause an additional 180° phase step at the characteristic frequency \( \omega_z \), which can be either lower than the resonant frequency \( (m<0) \) or higher \( (m>0) \). For frequencies higher \( \omega_z, \omega_o \) the transfer function converges to a high pass characteristic which reflects the capacitive feed forward path in the model. The described effects of the parasitic capacitive coupling turns out to be quite disturbing for most membrane application (see e.g. Chapter 3) since it aggravates the reliable control of the membrane at resonance. It is therefore an important task in the design of membranes and micro-systems to reduce these effects as much as possible.

The approach is easy to understand, when we again look at the set of transfer functions in Fig. 2.7. It can be noticed, that the location of the disturbing zeros heavily depends on the ratio \( C_1/C_2 \) and moves far away from the resonant frequency when the two capacitors are nearly equal. Moreover, it can be shown that for \( C_1 = C_2 \) \( (m=0) \) the zeros
2.3. Characterization and modeling

Transfer function for different capacitance ratios $C_1/C_2=1+m$; $m=-0.8\%...0.8\%$

![Transfer function graph](image)

**Figure 2.7:** Numerical evaluation of the transfer function including parasitic capacitive effects. To illustrate the cross coupling effects, the transfer function is plotted for different ratios $C_1/C_2$.

disappear completely and the transfer function $G'_M$ looks like the original simplified transfer function $G_M$, shaped by an uncritical low-pass function with a high frequency pole at $\omega_p = \omega_1 = \omega_2$ (see Eq. 2.9).

$$G'_M(s)|_{\omega_1=\omega_2=\omega_p} = K \cdot \frac{\omega_p s}{s^2 + \frac{\omega_p}{Q} + \omega_p^2} \cdot \frac{\omega_p}{s + \omega_p}$$  (2.9)

To suppress the unwanted zeros induced by capacitive cross coupling, the value matching of the parasitic capacitors $C_1$ and $C_2$ has to be as good as possible. This can be achieved by a careful symmetrical membrane layout as well as a symmetrical package. Best matching performance is achieved, when the membrane and the attached interface electronics is integrated on the same die. In such a monolithic solution (like shown in Fig. 1.1d) the capacitance values are well controlled and
the matching properties are usually very good. In the present research, different membranes in different technologies and layout have been used. The minimization of parasitic cross coupling has been an important aspect in the design of membrane resonators and several solutions for low parasitic resonators have been proposed [14].

Fig. 2.8 shows the comparison of the measured transfer function of an early membrane resonator and the appropriate small signal model. For the extracted model parameters and the coupling capacitors $C_1=5.22 \text{ pF}$ and $C_2=5 \text{ pF}$ the matching between the band pass model and the measurement is excellent.
Chapter 3

Miniaturized ultrasound barrier detector

3.1 Introduction to the application

An example of an ultrasound silicon micro-system is presented in this chapter, which relies extensively on circuit functions to make up the imperfections of the micro-sensor/actuator [16]. The goal is to develop a miniaturized ultrasound barrier (or presence) detection system using silicon membrane resonators as the ultrasound transducers. Ultrasound barrier detection systems are widely used in industrial automation to detect liquid, transparent or light sensitive objects [17, 18]. Dusty or smoky surroundings are also places where ultrasound is used for presence detection instead of light.

In a barrier detector micro-system, as shown in Fig. 3.1, well focused ultrasound is generated by a high Q silicon membrane resonator and detected by another membrane which has identical dimensions to the transmitter. The silicon membranes are connected to a transmitter/receiver circuit to provide both the ultrasound generating and detecting functions. To allow flexibility within the system, there should not be any hard-wired connection between transmitter and receiver. The ultrasound detector micro-system then operates like the commonly used
light barrier found in many industrial applications. Unlike the electronics of a conventional ultrasound barrier based on piez-ceramic transducers and discrete circuits [18], the use of silicon membrane transducers requires more complex circuit structures. The low sensitivity of the silicon membranes means that the signal received by the piezoresistive Wheatstone bridge is very weak, which requires amplification by more than 1000. Mismatch of the membrane resonance frequencies due to fabrication tolerances or temperature difference has to be compensated by a self-calibration procedure. Combining all these functions in a single, fully integrated IC not only allows the system to be miniaturized, but also reduces cost and improves reliability.

For the realization of the mixed signal IC the 2 μm BiCMOS technology (combined bipolar and CMOS) of Mietec Alcatel is chosen, which not only enhances the design freedom and performance of the analog building blocks, but also enables the co-integration of the silicon membranes and the electronic circuits on the same die. The epitaxial layer of the BiCMOS technology, used as anode in the electro-chemical etching procedure [15], has a thickness of 6 μm [20] and ideally can be used as bottom layer of the silicon membrane (see cross section Fig. 1.4).
3.2 Self-calibrating ultrasound transmitter/receiver

As already found out during the investigation of the membrane transducer described in chapter 2.3, the sensitivity and the emitted ultrasound are rather small outside the narrow 3 dB-bandwidth, which is due to the membrane's high selectivity. When used in an ultrasound barrier application, this narrow band behavior may cause several problems.

The first is the matching of the signal frequency of the electrical driving source to the resonant frequency of the transmitter membrane. Due to the narrow bandwidth of the membrane even a small mismatch between these two frequencies leads to drastically reduced emitted ultrasound power and will limit the overall system performance.

The second problem is the matching of the receiver resonant frequency to the incident ultrasound, generated by the transmitter. Even though the design geometries of transmitter and receiver membranes are identical, their resonant frequencies differ up to 5% due to process variations or temperature differences. This mismatch can cause substantial loss of sensitivity of the receiver and again reduces the range of operation.

In order to achieve reliable operation of the system, the loss in sensitivity or emitted ultrasound power has to be prevented by tuning schemes which automatically match the transducer resonant frequency to the required frequency. As the resonant frequency depends on the static heating power applied to the heating resistor, a dc voltage across the latter will be an ideal controlling parameter for automatic tuning.

For the ultrasound transmitter, however, such a tuning procedure may not be the best way to solve the matching problems. A simpler and more elegant way is to embed the membrane into the loop of an electro-mechanical oscillator as depicted in the block-diagram of the chip (Fig. 3.2) with the toggle switches in transmit-mode (T) position. In this oscillator the membrane resonance determines the oscillation frequency while two series-connected 35 dB amplifiers deliver the electrical amplification necessary for an overall loop gain of more than one.
In the receiver it is more difficult to match the membrane resonance to the frequency of the incident ultrasound. Since a wired connection between transmitter and receiver is missing, a direct link to the transmitting frequency does not exist and information on the latter can only be extracted from the sensor signal itself. The problem has been solved by using an incremental tuning/tracking scheme which automatically looks for maximum sensitivity by tuning the dc voltage across the heating resistor. Fig. 3.2 with toggle switches in receive-mode (R) position shows a schematic representation of the algorithm and more details will be explained in section 3.4.

As illustrated in Fig. 3.2, both transmitter and receiver circuits share most of the building blocks. It is therefore efficient to integrate both transmitter and receiver functions on the same die. This will not only lower the cost of the die-fabrication but also reduce cost of chip assembly and packaging since then transmitter and receiver are exactly the same.

Figure 3.2: Block-diagram of transceiver IC with transmit-mode (T) and receive-mode (R).
3.3 The ultrasound source

The task of the ultrasound transmitter is the generation of a focused ultrasound beam of sufficient power. The wavelength of ultrasound at 80 kHz is about 4 mm. Holes or grids in the transmitter package with similar dimensions therefore can be used to bundle the ultrasound waves. An appropriately designed package can thus focus the ultrasound into a beam and achieve up to 30% more sound pressure [21]. Apart from this physical amplification or beam-forming of ultrasound waves, the optimization of ultrasound power by the electrical excitation of the ultrasound transducer is of great importance (see also Chapter 2!).

Maximum emitted ultrasound power is obtained by matching the excitation frequency to the membrane resonance. This can be achieved by embedding the membrane into the loop of an electro-mechanical oscillator as is described in [22]. Considering the membrane as an electrical two port device, the overall transfer-gain has been measured to be around -50 dB, depending on the excitation amplitude.

Fig. 3.3 shows a schematic representation of the electro-mechanical oscillator with the membrane as the resonator. The weak electrical output signal of the membrane with transfer function $G'_M$ is amplified and positively fed back to the electro-thermal excitation at the input summing node. More than $A=60$ dB of amplification is needed to secure an overall loop gain of more than one and to fulfill the oscillation condition. This is needed to make up for the poor conversion gain of the membrane’s electro-thermal-mechanical excitation, on the one hand, and the low sensitivity of the Wheatstone bridge on the other.

However, pure amplification in the electro-mechanical circuit may not be sufficient for a reliable oscillation. The additional feed forward
zeros of $G'_M$, induced by parasitic capacitive cross-coupling can prevent
the oscillation condition, as is illustrated by the root locus curve in
Fig. 3.4. On the right hand side of the picture, the closed loop poles
of the system in Fig. 3.3 are plotted for different gains $A$ around the
dominant pole of the resonator. The root locus curve is evaluated for
$C_1 = C_2$ and for $C_1 \neq C_2$, where $C_1$ and $C_2$ are the parasitics in the
model of Fig. 2.6. For $C_1 = C_2$ the resulting root locus curve is a
straight line, which moves towards the imaginary axis for increasing
gains $A > 0$. The oscillation condition is fulfilled, when the curve
hits the imaginary axis which in the present example is achieved for
$A=50$ dB. On the other hand, for $C_1 \neq C_2$, feed forward zeros appear
near the dominant poles of the resonator. If the zeros are very close
to the poles as shown in the figure, the resulting root locus curve is
"attracted" by them and may not reach the imaginary axis, even for
very high gain. The oscillation condition then cannot be fulfilled and
the electro-mechanical circuit never starts oscillating.

![Location of Poles (X) and Zeros (O) of Openloop Transfer Function](image)

![Root Locus Curve for $|A|=20...54$dB](image)

**Figure 3.4:** *Root locus curve of electro-mechanical oscillator.*

In the above example, the two parasitic capacitors have values of
3.3. The ultrasound source

5 pF and 7 pF, respectively. The difference between the two is not much, however, the resulting feed forward zeros inhibit reliable oscillation. This situation has to be avoided and it can be done with a dedicated membrane layout, as already mentioned in the previous chapter. To achieve parasitics with matched values in order to move the unwanted zeros far away from the dominant pole, the design of a symmetrical membrane layout and package is mandatory. The same considerations are valid, if a hybrid system is targeted using a printed circuit board (PCB). The PCB has also to be designed with a symmetrical layout for the interconnection network between the membrane and the electronics.

3.3.1 AC coupling

The amplification stage in the electro-mechanical oscillator of Fig. 3.3 is the actual front-end circuit of the ultrasound source and the microsystem, respectively. It provides the gain necessary to sustain oscillation and provides differential to single ended conversion for subsequent processing on the chip. As this front-end circuit is a high gain amplifier, special attention must be paid to the input dc offset. Due to different expansion coefficients of the layers incorporated in the sandwich structure of the silicon membrane, the latter is slightly buckled. Because the piezo resistive Wheatstone bridge measures the bending of the membrane, it sees an electrical offset which can be as high as 80 mV as has been discussed already in chapter 2. AC coupling of the differential input is necessary in order to avoid saturation of the amplifiers. This combination of high gain at a relatively low frequency of 80 kHz and ac coupling requires a very large time constant for an integrated circuit. If two op amp (operational amplifier) based differentiators are used in cascade, as shown in Fig. 3.5, the time constant $\tau_1 = R \cdot C_1$ (see Fig. 3.5) of each differentiator will be on the order of 220 $\mu$s. Even for a 15 pF coupling capacitance, 15 M$\Omega$ of resistance is required. Since it would take too much silicon area to implement such resistors with polysilicon or diffusion, an artificial floating resistor using MOS transistors is devised to emulate the feedback resistors in Fig. 3.5.
Artificial floating resistor

Artificial resistors are widely used in various integrated circuits like active RC filters or continuous time signal processing in general. Many solutions for the emulation of resistors using MOS transistor have been suggested, most of them intended for highly linear applications [23, 24, 25]. For the present ac coupling function, however, the simpler, but nonlinear circuit configuration shown in Fig. 3.6 is an attractive alternative. High linearity is not really required for the ac coupling function and additionally the presented circuit structure is inherently offset free, which is not the case in most of the floating resistors found.
3.3. The ultrasound source

Figure 3.6: Artificial floating resistor using an NMOS and PMOS biased in triode region.

in literature.

The basic idea is to use long channel transistors biased in the triode region. A single transistor implementation has the problem that, as the terminal connected to the op amp output is expected to swing symmetrically with respect to the terminal connected to the inverting input of the op amp (virtual ground), the former terminal becomes the source of the transistor during either the positive or the negative half-swing of the oscillation. As the equivalent resistance during this half swing decreases rapidly as the amplitude increases, the resulting oscillation will become very asymmetric and distorted. An NMOS-PMOS combination between the op amp output and input results in the I-V characteristic shown in Fig. 3.7, which can be designed to be symmetric with a well controlled linear range. The biasing circuitry in Fig. 3.6 is designed to ensure that the sources of both the NMOS and the PMOS transistors are biased at analog ground.

For the analysis of the circuit and to derive design considerations for a symmetrical I-V characteristic, the well known quadratic-law MOS transistor model of Shichman and Hodges [26] is used. It can be shown, that the equivalent resistance $R_o$ of the proposed artificial resistor at
zero voltage $V_R$ (see Fig. 3.7) is the sum of the inverse of the drain-to-source transconductance of the two main transistors $M1$ and $M2$, which operate in triode region:

$$R_o = \frac{1}{g_{DS1}} + \frac{1}{g_{DS2}} = \frac{1}{\beta_1} \sqrt{\frac{\beta_1}{2I_{b2}}} + \frac{1}{\beta_2} \sqrt{\frac{\beta_3}{2I_{b1}}} \quad (3.1)$$

here $\beta = \frac{W}{L} \mu_s C_{ox}$ is the transistors intrinsic transconductance and the index at the parameters denote the individual devices $M1$...$M4$ in the above schematic. For a high voltage $V_R$ across the artificial resistor, either $M1$ or $M2$ enter the saturation region, where the current becomes independent of the applied voltage. To achieve a symmetrical I-V characteristic as is depicted in Fig. 3.7, these two limiting (saturation) currents $I_{s1}$ and $I_{s2}$ have to be equal. Let’s first consider the voltage $V_R < 0$ and the terminal $R_1$ connected to a virtual ground at the potential of analog ground. The gate voltage $V_{GS2}$ of $M2$ is supplied by the diode connected $M3$ and remains constant. The maximum drain current of $M2$ therefore is limited by its saturation. On the other hand, the gate voltage $V_{GS1}$ of $M1$ is not constant and gets larger when $V_R$ becomes more negative, which makes the channel of $M1$ more and more low impedance. The influence of $M1$ on the overall I-V characteristic then disappears and is completely dominated by transistor $M2$. The limiting current $I_{s2}$ then becomes equal to the drain current of $M2$ in saturation:
3.3. The ultrasound source

\[ I_{s2} = \frac{\beta_2}{\beta_3} \cdot I_{b1} \quad \text{(Current mirror M2, M3)} \quad (3.2) \]

which can be interpreted as a current mirror composed of transistors M2 and M3. For \( V_R > 0 \) the situation is slightly different. Transistors M1 and M4 also build a current mirror, whose current limitation \( I_{s1} \) is due to the saturation of transistor M1. However, this current is scaled down by a source degeneration effect of M1, whose source terminal is connected to M2 biased as resistor in triode region. The exact nonlinear calculation of \( I_{s1} \) as a function of the transistor parameters leads to messy and incomprehensible formulas. However, if M2 is treated as linear resistor for small values \( V_{DS2} \), a reasonable approximation for \( I_{s1} \) can be found:

\[
I_{s1} \approx 2 I_{b1} \frac{\beta_2^2}{\beta_1 \beta_3} (B + 1 - \sqrt{2B + 1}), \quad B := \frac{\beta_1}{\beta_2} \sqrt{\frac{I_{b2} \beta_3}{I_{b1} \beta_4}} \quad (3.3)
\]

Setting equal the equations for the two limiting currents \( I_{s1} := I_{s2} \) (Eq.3.2 and Eq. 3.3) defines the condition for a symmetrical I-V characteristic of the artificial floating resistor, out of which guidelines for the latter’s design can be extracted. Solving this equation for the ratio \( \beta_1 / \beta_2 \) results in:

\[
\frac{\beta_1}{\beta_2} = \frac{W_1}{L_1} \mu_{on} \approx \frac{1}{\left( \sqrt{I_{b2} \beta_3} - \frac{1}{2} \right)^2} \quad (3.4)
\]

Where \( \mu_{on} \) and \( \mu_{op} \) are the mobilities of NMOS and PMOS, respectively, and \( W_1, L_1, W_2, L_2 \) are the transistor dimensions. The design guideline represented by Eq.3.4 has been considered at the implementation of a tunable high impedance artificial resistor, which was done in order to verify the proposed principle and to use as feedback resistor in Fig. 3.5. Fig. 3.8 shows the measured I-V characteristics. The resistor tuning range is from less than 14 M\( \Omega \) up to 40 M\( \Omega \).
Front-end characteristics

To minimize the cost of the front-end op amp in terms of silicon and power, the op amp’s unity gain bandwidth is designed to be 8–9 MHz, just high enough to provide 30–35 dB gain for each differentiator at the frequency of interest. The resulting characteristics are that of a bandpass filter with high peak gain $A_c$ (see Fig. 3.5). Small feedback capacitors are provided so that a small flat gain region exists around the oscillation frequency, between the corner frequencies of the highpass and lowpass characteristics.

In this work, the front-end op amp is implemented as an operational transconductance amplifier (OTA). Assuming a simplified one pole transfer function for the op amp with high dc gain $A_o$ and unity gain bandwidth $\omega_b$,

$$G_{OTA}(s) = A_o \frac{\omega_b}{s + \frac{\omega_b}{A_o}}$$ (3.5)
3.3. The ultrasound source

the transfer function $G_F(s)$ of one differentiating amplification stage results in the second order bandpass

$$G_F(s) = \frac{\omega_b}{\omega_c} \cdot \frac{s\omega_c}{s^2 + s \frac{C_1}{\omega_c} + \omega_c^2},$$

(3.6)

under the reasonable assumptions $C_1 \gg C_2$ (high peak gain) and $A_0 \gg 1$. The center frequency $\omega_c$, the quality factor $Q_c$ and the peak gain $A_c$ can be obtained easily by assuming $C_1 \gg C_2$ and $\tau_2 = RC_2 \gg 1/\omega_b$, respectively:

$$\omega_c = \sqrt{\frac{\omega_b}{\tau_1}}, \quad Q_c = \sqrt{\frac{1}{\omega_b\tau_1} \cdot \frac{C_1}{C_2}}, \quad A_c = \frac{C_1}{C_2}$$

(3.7)

At 80 kHz, the cascade of two differentiators has approximately 70 dB gain and zero degree phase shift. The measured transfer characteristics of one stage are shown in Fig. 3.9. The measurement shows that the region around the center frequency ($\omega_c$) is not as flat as expected from an ideal bandpass characteristic. This is due to the distributed lowpass behavior of the high ohmic feedback resistor which results in a small peaking of the transfer function at 50 kHz. However, any in-accuracies of the phase characteristic in the differentiator chain due to peaking and process variations will be compensated by the high $Q$ of the mechanical resonator so that oscillation will be guaranteed and close to the peak of the membrane resonance.

3.3.2 Limiter

In many oscillator circuits, a control mechanism is required to set the oscillation amplitude to a predefined level. The stabilization of the amplitude can be achieved in different ways. One simple solution is to use a non-linear element like a limiter clamping the output to a defined level (which can be the power supply). On the other hand, if the oscillator should provide signals with low distortion, an automatic gain control (AGC) can be applied.
In order to achieve maximum ultrasound power, the amplitude of the oscillation should be as large as possible. Rather than simply using the (natural) clipping behavior of the amplification stages to the power supply, a MOS limiter as shown in Fig. 3.10 is used to generate quasi rectangular signals of 50 % duty-cycle which are then applied to the output driver. The reason for using signals smaller than what can be delivered by the output of the previous stage is to prevent the bipolar power output stage from operating in deep saturation (see also section 3.3.3).

The limiter-cell is applied to the output of an OTA (see Fig. 3.2). For the description of the circuit, the inspection of one half of the circuit in Fig. 3.10 is sufficient since the structure is symmetrical. Transistor $M_1$ is biased by $M_1'$ and two current sources in such a way that its gate potential is $V_{gs} - V_{lim}$ relative to analog ground. Assume a decreasing voltage signal, supplied by the output of the OTA in the oscillator, applied to the limiter. As long as the voltage is larger than $-V_{lim}$, the transistor $M_1$ is turned off and the input impedance of the limiter cell is

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**Figure 3.9:** Measured transfer-function of one amplification stage.
3.3. The ultrasound source

Figure 3.10: The limiter.

high. When the input reaches $-V_{lim}$, transistor $M1$ enters strong inversion region, providing a low ohmic node to the output of the OTA. The gain of the OTA is then reduced drastically and the resulting oscillation amplitude is limited to an amplitude close to $V_{lim}$.

Fig. 3.11 shows the measured voltage transient across the heating resistor for different voltage limits $V_{lim}$. Note that the oscillation frequency depends on the amplitude of the driving signal. This is because larger amplitudes generate more dc power on the heating resistor and therefore change the resonant frequency (equation 2.11).

3.3.3 The output driver

The last important building block of the transmitter is the buffer amplifier which closes the loop of the oscillator. It drives the rather low ohmic heating resistor (130-200 $\Omega$) of the membrane. A BiCMOS implementation of the buffer amplifier is shown in Fig. 3.12.

It is a two-stage amplifier whose second stage is composed of a bipolar power transistor and the transducer heating resistor $R_h$ as the load, which maximizes the power efficiency of the driver because all power current flows into the load.
Chapter 3. Miniaturized barrier detector

Figure 3.11: Transient measurement of voltage across the heating resistor for different oscillation amplitudes, which leads to different oscillation frequencies.

A drawback of the circuit in Fig. 3.12 is the potential deep saturation of the bipolar transistor $Q_1$ for an output swing close to the negative power-supply. Although an output swing close to the supply-rails is preferred for the high ultrasound power generated, a saturated power transistor can cause severe phase shift in the feedback loop of the oscillator. If this phase shift exceeds a critical value the oscillation eventually dies.

Fig. 3.13 shows the measured amplitude of the first amplification stage, which is proportional to the vibration amplitude of the membrane and therefore a measure for the ultrasound pressure. The magnitude increases proportional to the power amplifier output voltage until the latter reaches the saturation voltage of the bipolar transistor $Q_1$ where it suddenly collapses.

Saturation can be avoided if the amplitude of the input signal to the output driver is limited so that $Q_1$ does not operate in saturation as described in the previous section. The output buffer amplifier has a closed loop gain of +2. The input common-mode potential for the full output swing is then reduced to one half of what it would be for a unity
3.4. The ultrasound receiver

The task of the receiver is to detect incident ultrasound and to translate it into an electrical signal of sufficient level and signal quality. While the signal quality is basically determined by the front-end amplifier interfacing to the Wheatstone bridge, sufficient signal amplitude can only be obtained if the ultrasound frequency matches the resonance of the receiver membrane, as described in section 3.2.

An automatic tuning scheme which performs this task continuously requires reference information about the transmitted ultrasound signal like phase, frequency or amplitude. However, as the barrier detection micro-system should be general purpose, a wired connection between

Figure 3.12: BiCMOS output amplifier including the resistive load in the second stage.
transmitter and receiver is not available which means that the information on the ultrasound signal can only be extracted by the receiver membrane itself.

One solution to this problem is to use a simple tuning scheme to look automatically for the highest possible sensitivity within a range of voltage applied to the heating resistor. Such a simple tuning and tracking algorithm has been developed in this work. Its operating principle is illustrated in Fig. 3.14.

When a slowly varying voltage ramp is applied to the heating resistor of the receiving membrane, the amplitude observed at the membrane Wheatstone bridge describes the shape shown in Fig. 3.14. The maximum is achieved when the membrane's resonant frequency, tuned by the applied heating power, exactly matches the frequency of the incident ultrasound. At this point the derivative of the membrane voltage with respect to the heating voltage changes the sign from positive to negative which can be used as a detection flag for a tracking algorithm.

The tracking algorithm starts at one end of the tuning interval by applying $V_1$ at the heating resistor. Then, the heating voltage $V_h$ is increased by a $\Delta V_h$ and the change in the amplitude of the bridge signal is observed. If the slope is positive, the track direction was correct.
3.4. The ultrasound receiver

3.4.1 Tracker implementation

After an amplification of 70 dB, the signal from the bridge is rectified and low-pass filtered in order to obtain the amplitude. It is then fed to the slope-detector, which generates a digital signal that is high for a
positive slope and low for a negative gradient. The slope-detector is an OTA which operates either as a unity-gain buffer or as a comparator. Fig. 3.15 shows the schematic.

\[ V_{dd} \]

Figure 3.15: The slope detector.

In the sampling phase, the switch is closed (\( \Phi = V_{dd} \)) and the sampling capacitor \( C_c \) acts as the compensation for the OTA. The sampled rectifier voltage plus the offset of the OTA arc stored on the capacitor.

If the switch is open, the OTA then behaves like an auto-zero comparator, which detects if the actual input signal is higher or lower than the stored value on the capacitor.

Care has to be taken of the OTA unity gain bandwidth and the dimensions of the sampling capacitor and the switches, respectively. The non-zero impedance \( R_{sw} \) of the closed switch together with \( C_c \) causes an additional pole at \( 1/R_{sw}C_c \) in the open loop transfer function of the OTA which may affect the stability during sampling operation.
3.4. The ultrasound receiver

$R_{sw}$ can be made small by a large $W/L$-ratio of the switch transistors in order to move this pole to high frequencies. On the other hand, large area switches are not preferable to achieve low charge injection on the sampling capacitor. The choice of appropriate values for $C_c$ and the $W/L$-ratio is therefore a tradeoff between accuracy (low charge injection) and speed. In the present design, the speed requirement on the slope detector are relaxed so it is not critical to fulfill above design considerations. The sampling capacitor $C_c$ is 10 pF, $R_{sw}=4500 \ \Omega$ and the bandwidth of the OTA is designed to be 450 kHz in order to achieve a phase margin of 85° (see Fig. 3.15).

![Diagram of a simple EXOR gate and up/down counter](image)

**Figure 3.16:** The tracking algorithm with a simple EXOR as decision logic.

The slope detector builds the analog/digital interface to the digital portion of the tracking algorithm, whose central part is a simple EXOR gate plus an up/down counter as shown in Fig. 3.16. In the first phase
(Φ₁) the counter is enabled to generate a new incremental data output, which is then translated to the heating voltage by a 6 bit D/A converter followed by a buffer amplifier. As explained in the previous paragraph, this causes a change in amplitude of the received signal which is monitored by the slope detector during phases Φ₂ and Φ₃, respectively. The subsequent EXOR gate compares the actual slope with the previous one stored in the flip-flop and determines the new tracking direction (up or down counting) at the end of phase Φ₃.

The phases Φ₁, Φ₂, Φ₃ are derived from a low frequency on-chip oscillator as depicted in the block diagram of Fig. 3.2. It is a simple relaxation type oscillator [27] whose oscillating frequency depends on the bias current and an on-chip capacitor. The oscillating frequency may vary up to 30% due to process tolerances, which cannot be tolerated in many sampled system application. However, in the present micro-system the self-calibrating (tracking) process operates slowly and an accurate time base for the system clock is not required. A fully integrated oscillator is therefore easy to implement and the preferred solution in the presence detector micro-system.

### 3.4.2 The front-end amplifier

When the transceiver IC operates as a self-calibrating receiver, the ultrasound signal supplied by the Wheatstone bridge is amplified by the same ac-coupled stages as the transmitter (see Fig. 3.2, toggle switches in R position). Although for the transmit-mode the non-idealities of this amplification stage like nonlinearity, gain inaccuracy or noise, are not of primary interest, some of them have to be considered carefully when the IC is used as a receiver. The small voltage signals obtained from the transducer membrane are in the range of 10 μV to 250 μV depending on the distance between the transmitter and the receiver, which makes a closer consideration of the amplification stages necessary in terms of noise. Since the gain of each stage in the amplification chain is 35 dB at 80 kHz, the noise is dominated by the first stage and the examination of one ac-coupled stage is sufficient.

There are basically two noise sources which contribute to the overall noise performance, the input equivalent noise of the OTA and the thermal noise due to the large feedback resistor R (see Fig. 3.5). However,
3.4. The ultrasound receiver

due to the transimpedance structure of the amplifier, the noise contribution of $R$ is not amplified [28] and it can be shown that the output power spectral density $\frac{\overline{v}^2_{or}}{\Delta f}$ due to $R$ alone at the center frequency $\omega_c$ (for $C_1 \gg C_2$ and $RC_2 \gg \frac{1}{\omega_b}$) is

$$\frac{\overline{v}^2_{or}(\omega_c)}{\Delta f} = \frac{\overline{v}^2_{r}}{\Delta f}Q_c^2$$  \hspace{1cm} (3.8)$$

where $\overline{v}^2_{r} = 4kTR\Delta f$ is the voltage noise generator of the resistor $R$ and $Q_c$ is the quality factor of the bandpass characteristics of the amplification stage (see Fig. 3.9). $\omega_b$ is the unity gain bandwidth of the OTA and $Q_c$ is calculated using Eq. 3.7. To achieve a flat phase characteristic around the center frequency $\omega_c$, the quality factor $Q_c$ is designed to be less than one. This has the consequence that the noise contribution generated by $R$ alone is not amplified according to Eq. (3.8). On the other hand, the noise $\overline{v}^2_{d}$ generated by the input stage of the OTA is enlarged by $A_c = \frac{C_1}{C_2}$ at $\omega_c$ which is the same amplification factor as in the signal path.

It can be seen from the above consideration that at high peak amplification $A_c \gg 1$, the noise contribution of the OTA dominates ($\overline{v}^2_{oa} \geq 10 \cdot \overline{v}^2_{or}$) the overall noise performance, even if large feedback resistors $R$ are applied.

For the OTA used in this design, the following specification are met: $\omega_b = 2\pi \cdot 9$ MHz, equivalent input noise of OTA = 13.5 nV/√Hz, closed loop amplification $A_c$=37.5 dB @ $\omega_c$=2π•80 kHz. For $\overline{v}^2_{oa} \approx 10 \cdot \overline{v}^2_{or}$ the maximum allowed noise due to $R$ is 500 nV/√Hz which translates into an $R$ of 15 MΩ and hence a coupling capacitor $C_1$ of 15 pF. Fig. 3.17 shows the measured output noise power spectral density (PSD) of one amplification stage. To illustrate the individual noise contribution of the amplifier and the feedback resistor, ideal PSD curves derived from a first order calculation are included in the figure.

3.4.3 The rectifier/envelope detector

The rectifier/envelope-detector is a combination of a full-wave rectifier and a low pass filter to extract the amplitude of the applied signal.
Figure 3.17: Measured noise PSD of one amplification stage. For comparison, the ideal noise contribution of the resistor and the OTA (dashed line) as well as the total output referred noise of the stage are included.

To minimize off-chip components, the first order Gm/C-type low-pass filter required to suppress the ripple of the rectifier is fully integrated (see Fig. 3.18).

The ac input voltage is first translated into the input current $I_1$ by $G_{m1}$ and then rectified by the bipolar transistors $Q1$ and $Q2$. If the sign of $I_1$ is negative, the transistor $Q1$ is turned on, providing a low-ohmic impedance node for $I_1$. At the collector terminal of $Q1$, a copy of $I_1$ appears which is equivalent to the negative half-wave of the applied ac current. For the positive part of $I_1$, transistor $Q2$ is turned on in the same way, delivering the positive half-wave current at its collector terminal. This current is then copied by the mirror $M3, M4$ and added to the negative half-wave at the summing node $A$ of the output current mirror $M1, M2$. A full-wave rectified current thus results from the drain-terminal of $M2$. 
The output stage of the rectifier consists of a small $G_m2$ and a filter capacitor $C$. The task of this Gm/C-type low-pass filter is twofold. First, it transforms the drain current of $M2$ into a voltage and second, it extracts the dc component of the rectified signal by suppressing the higher order harmonics of the ripple. If 1% ripple is required at 80 kHz, the time constant for a first order low pass filter is 75 $\mu$s, which translates into a small $G_m2$ of 200 $\mu$A/V for $C=15$ pF. As this small $G_m$ provides high gain at the output, the current driving $G_m2$ has to be scaled down properly in order to not saturate the output. This is done by the 100:1 current mirror $M1, M2$. A detailed schematic of $G_m1$ and $G_m2$ is shown in Fig. 3.19.

The transconductance is obtained by biasing a MOS transistor in the triode region at constant drain source voltage $V_{DS}$ [29]. The $G_m$ therefore is proportional to the clamping voltage $V_{DS}$ which is defined by $R_1 \times I_1$ (see Fig. 3.19, $G_m1$). The output transconductance $G_m2$ is slightly different.

Since $G_m2$ is very small, the bias current flowing through the input MOS-pair $M3, M4$ cannot be too high in order to not require excessively
large drain-to-source voltage $V_{DS}$ and hence limit the output swing. However, a reasonably large current is needed for the bipolar cascode transistors to form a low impedance node. Since these two requirements are conflicting, an additional current source has been inserted as shown in Fig. 3.19, $G_{m2}$. The drain current of the input transistors and the emitter-current of the bipolar cascode are thus independent.

Fig. 3.20 a) shows measurements of the input and output voltages of the rectifier/envelope detector for a 100 Hz test signal and b) shows the transfer characteristics at 100 kHz.

Filters using very small $G_m$ generally suffer from degraded noise performance. However, if the equivalent noise bandwidth is small and noise requirements are relaxed like in the described application, the usage of small $G_m$'s for a fully integrated system is a viable alternative to expensive off-chip solutions.
3.5 System measurements

To facilitate the characterization of the silicon membrane transducers and the electronic circuit, the two parts have been integrated on two separate chips. A die photograph of the ASIC is shown in Fig. 3.21 where the analog and digital partitions of the system are marked with white boxes. Fig. 3.22 shows two prototypes of a complete ultrasound barrier detector system with packaged transmitter/receiver micro-systems. In the upper one the membrane, the IC and a few discrete components for biasing and power supply decoupling are mounted on a ceramic sub-
strate. The substrate is enclosed in a plastic housing with a 8 mm opening for the ultrasound waves. To protect the membrane against dust, the opening is covered with a fine grain sieve [14]. The lower part of the pictures shows a more advanced prototype, where the micro-system is based on a small PCB and put in an industrial package [14].

![Die photograph of the transceiver ASIC. The die size is 3.4 x 2.8 mm^2 including pads.](image)

**Figure 3.21:** Die photograph of the transceiver ASIC. The die size is 3.4 x 2.8 mm^2 including pads.

In order to measure the performance of the transmitter/receiver IC in the application, the upper prototype system of Fig. 3.22 has been set up by facing transmitter and receiver at a distance of D=8 cm. In the first measurement, the tuning range of the receiver has been determined (see Fig. 3.23) which is the range of the ultrasound frequency for which the receiver circuit is able to find and track the maximum sensitivity. For a 5 V supply the measured range is 77.4 kHz up to 84.3 kHz, corresponding to a 7% variation of the transmitted ultrasound frequency. The maximum sensitivity is only achieved if the membrane's resonant frequency is matched to the transmitted ultrasound. Since this is the case for the measured dots in Fig. 3.23, the latter also shows the (nonlinear) relationship between the membrane resonant frequency and the required heating voltage, supplied by the self-calibration circuit.
Figure 3.22: The complete ultrasound barrier detector system. The assembling and the packaging of the ultrasound micro-system is provided by the Physical Electronics Lab. (PEL), ETH Zurich, and Baumer Electric AG, Frauenfeld. (Photograph PEL)
Due to the 130 Ω heating resistor as the load of the output amplifier, the power consumption of the whole barrier system is quite substantial. In transmit-mode, the measured average power for maximum output amplitude is 93 mW, where 78% are delivered to the heating resistor. In receive-mode, a system power of 96 mW . . . 165 mW is measured, depending on the frequency of the transmitted ultrasound.

\[
\begin{array}{c|c|c|c|c|c|c|c|c|c}
\hline
\text{Ultrasound Frequency [kHz]} & 77 & 78 & 79 & 80 & 81 & 82 & 83 & 84 & 85 \\
\hline
\text{Heating Voltage [V]} & 2.4 & 2.8 & 3.2 & 3.6 & 4 & 4.4 \\
\hline
\end{array}
\]

**Figure 3.23:** Tuning range of the self-calibrating receiver.

The measurement in Fig. 3.24 shows the startup sequence of the tracking circuit for a strong (D<4 cm) and a weak (D>10 cm) signal. As the slope of the receiver amplitude with respect to time for a weak signal is smaller than in the case of a strong one, noise makes it more difficult for the tuning algorithm to converge to the maximum. The startup period therefore lasts longer.

In Fig. 3.25 b) the behavior of the barrier detector micro-system is shown when an obstacle interrupts the acoustic path. The received ultrasound signal disappears and triggers the detector output. During this period, the tracking circuit is disabled automatically and the heating voltage remains constant. This prevents the voltage across the heating resistor from drifting away from the optimum during the absence of incident ultrasound.
3.6 Conclusions

The realization of a barrier detector system based on a BiCMOS technology has been discussed in the context of a micro-system. Sensor-
actuator and read-out electronics can be fabricated using the same industrial IC-process. The miniaturized ultrasound membrane transducers can be modeled accurately by a RLC-resonator and therefore be included for the design and verification of the micro-system. On-chip ac-coupling and an on-chip rectifier are part of a design realizing large time constants of $100 - 200 \, \mu s$. The presented ultrasound barrier detector micro-system is a typical example for a smart sensor and also for a smart actuator, where physical imperfections and process tolerances are compensated by dedicated conditioning circuits (see Fig. 1.2). Frequency matching problems between the transmitter and the receiver of the barrier system are solved by embedding the transmitting membrane in the loop of an electro-mechanical oscillator as well as applying a simple tracking algorithm to the receiver. The miniaturization of ultrasound sensors or actuators and dedicated electronics with a minimum of external components can be an interesting solution for many applications of industrial automation.
Chapter 4

Miniaturized ultrasound range finder system

4.1 Introduction to the application

In this chapter, a demonstrator of a miniaturized ultrasound range finder system for short distances (10—100 mm) is described, which uses silicon membranes to generate and detect ultrasound waves [30] at 90-100 kHz. The main intention is the investigation of measurement methods and system architectures for the range finder based on the silicon membranes. In this context not only the special characteristics of the transducers is considered but also the possibility of a fully integrated solution in a CMOS technology. Existing measurement procedures known from piezo-ceramic ultrasound transducers and laser range finders are reviewed with respect to their applicability to the silicon transducer and a novel technique is derived, which is suitable for the new devices.

Ultrasound range finders (sonar, distance sensor) are used in a great variety of applications. Ultrasound is employed in liquid or gas at frequencies from above 20 kHz to several megahertz to locate objects like shoals of fish, measuring the depths of a tank, determine the level of liquid or e.g is used as parking aid for un-talented car drivers. One large category of range finders applied in industrial automation is employed
for contact-less distance measurement in air [18]. Most of the range finder systems in this category are based on piezo-ceramic transducers, which generate or detect short ultrasound pulses used for time-of-flight (TOF) [9] based measurement methods.

The measurement range (scanning range) of such range finders lies in the cm to m range, with a resolution of 0.5 mm to 3 mm [18] equivalent to about 10 bits. The response time is distance dependent and can easily achieve tens of milliseconds up to 250 ms [18]. Although range finders of the described type usually have good performance and are successfully used in many applications, they suffer from a few significant drawbacks. The piezo-ceramic transducers are rather expensive and bulky, which makes them difficult to use in applications where only little space is available. In addition, they need high excitation voltages (40-100 V) to achieve sufficient ultrasound power for the full scanning range and accuracy.

On the other hand, the use of micro-machined silicon membranes as the ultrasound transducers of a range finder system can reduce the latter's size and cost, as has already been discussed in chapter 3. The thermally excited silicon membrane transducers are only 1 mm by 1 mm in size and can be fabricated inexpensively by an electro-chemical etching technique using a standard CMOS integrated circuit technology. Other than in the barrier application of chapter 3, the membrane thickness is determined by the depth of the implanted n-well and e.g. reaches about 5-6 μm in a 0.8 μm CMOS technology [31, 14].

The advantage of the silicon membrane transducer in terms of price and dimensions is obvious compared to the existing piezo-ceramic transducers, but, the physical-electrical characteristics are rather non-ideal. Due to the small membrane area the generated ultrasound pressure is very weak (0.25 Pa measured at 50 mm distance [32]) and limits the practical distance range to about 100 mm. In addition, the receiver sensitivity to incident ultrasound is low. At 5 V operation, the signals obtained from the piezo-resistive Wheatstone bridge hence are rather small (10 μV to 250 μV) and noisy. High-gain, ac-coupled amplification and self-calibration is mandatory, as has been described already in chapter 3. To generate the short ultrasound pulses required in time-of-flight measurements, the applied transducers have to be reasonable broad band. The bandwidth of the silicon transducers, however, is only 900 to 1000 Hz.
due to the high Q of the resonant structure, which makes it extremely difficult to use them for this widely used TOF principle. Other measurement techniques like the continuous wave (CW) method, also known as phase shift method, may be more suitable in terms of bandwidth requirement, but they suffer from a very limited measurement range and distance dependent distortion. Section 4.2 reviews all relevant measurement techniques including the TOF and the CW method.

To overcome the drawbacks of both the TOF as well as the CW and related techniques, a novel measurement technique based on the CW method is presented in this research, which exhibits a 10 to 100 mm measurement range. It takes into account the limited bandwidth of the silicon transducer, as well as the low sensitivity of the thermal-electromechanical excitation and detection. Mismatch between the transmitter and receiver membranes as well as excitation and resonant frequencies are solved by self-calibrating front-end and driving circuits, similar to the ultrasound barrier in chapter 3. The origin of distance dependent distortion mentioned above is explained and techniques are presented to substantially reduce it.

The implementation of such a range finder as a fully integrated CMOS micro-system operating from a single 5 V supply is then an attractive alternative to today's expensive and large scale solutions.

4.2 Measurement methods

4.2.1 The time-of-flight (TOF) method

In a TOF measurement, also called pulse-echo method, a transmitter (T) emits a packet of ultrasound waves towards the surface of an object, where the waves are reflected. The reflected waves are detected by a receiver (R) and the time interval between emission and detection is evaluated as measure for the distance $X$. Fig. 4.1 shows the principle.

The distance $X$ is determined straightforward by equation Eq.4.1

$$X = t_x \cdot \frac{c}{2}$$ (4.1)
where $t_x$ is the time-of-flight mentioned above and $c$ is the sound velocity (in air).

The basic arrangement in Fig. 4.1a) is an expensive solution since it requires two identical transducer elements and, as a consequence, also a costly assembling. A combined receiver/transmitter (R/T) solution as shown in Fig. 4.1b) is more practical, since it is based only on one transducer element and therefore is more simple and cost effective. After emitting the ultrasound burst pulses, the transducer is switched to the receive mode in order to detect the echo from the object.

The combination of transmitter and receiver using a single piezoceramic transducer is a simple and inexpensive solution, however, it may be unsuitable for short distances $X$. If the echo from the object surface hits the transducer before the previous excitation envelope declines to zero, the echo is blanked out completely and a detection of the echo is impossible. A minimum distance exist, called blind space, below which it is impossible to detect any echo. The blind space of today’s state of the art range finders is about 3 to 5 cm, which is too much for those applications where space is a critical parameter.

On the other hand, the blind space is made much smaller when the costly variant of Fig. 4.1a) is applied. Unlike in the single transducer variant of Fig. 4.1b ), the echo from the object can be clearly separated.
from the excitation, which is preferable for a small blind space. Due to today’s expensive transducers, however, not many commercial range finders based on two transducing elements exist. By the use of batch fabricated silicon transducers instead of expensive piezo ceramic devices, the cost of a complete range finder system no longer is dominated by the transducer elements, even if separate transmitter and receiver are applied. A miniaturized range finder based on separate silicon transducers therefore is an attractive candidate for short range applications. In the next paragraph, an experimental range finder is investigated, which is based on the TOF principle and two silicon membrane transducers.

4.2.2 TOF and silicon transducers

In this first experiment, an electro-mechanical start-stop oscillator has been built based on the ultrasound source of chapter 3.3 to generate the ultrasound burst pulses required by the TOF method. The receiver is manually tuned to get maximum sensitivity. Fig. 4.2 shows the block diagram. Transmitter and receiver face each other at a distance of a few cm.

Figure 4.2: Electro-mechanical start-stop oscillator and receiver.
Figure 4.3: The time-of-flight measurement. a) Simplified 100 kHz transducer with 25 kHz bandwidth. Transmitter and receiver face each other at 8 cm (numerical experiment). b) The measurement with silicon transducers. The bursts have a frequency of 80 kHz and a pulse repetition rate of 200 Hz. Note that the a) and b) have the same time-scale.

To achieve packets of ultrasound waves, the excitation of the transducer is modulated by a rectangular signal by simply turning on and off the supply of the transmitter Wheatstone bridge (see Fig. 4.2). The dc heating voltages $V_{dc1}, V_{dc2}$ of both transmitter and receiver are manually tuned in order to get matched resonance frequencies. The resulting
4.2. Measurement methods

measurement is shown in Fig. 4.3 b). The upper trace (T) is the excitation voltage across the heating resistor whereas the lower trace shows the amplified signal from the receiver membrane (R). For comparison, a numerical experiment using an idealized 100 kHz transducer with 25 kHz bandwidth is included as Fig. 4.3 a).

Due to the slow thermal excitation principle ($\tau_{\text{heat}}=0.5\ldots1\text{ms}$) and the high Q resonant structure of the silicon membrane, the transient response of both transmitter and receiver is expected to be a rather slow process. It can be seen in the measurement of Fig. 4.3b) that the initial time interval between the edge of the transmitter excitation and the detected signal rising from the system noise floor is already 0.5 ms, which corresponds to an unacceptable large 16.5 cm offset distance. In addition, the noise level in the receiver is high in general and the steepness of the received pulses is more than an order of magnitude away from what can be delivered by a broad band transducer (Fig. 4.3a). All these characteristics make it extremely difficult to accurately determine the time-of-flight as the desired measurement output and it is desirable to consider different measurement techniques.

4.2.3 Continuous wave (CW) and related methods

A measurement method which is absolutely uncritical in terms of transducer bandwidth is the continuous wave (CW) method [9, 33, 34]. The transmitter emits a beam of ultrasound waves at one fixed frequency $f_c$ which is reflected at the object and detected by the receiver (setup as Fig. 4.1a). The phase shift $\varphi_x$ between the electrical excitation (T) and the received signal (R) is evaluated as a measure for the distance X. Fig. 4.4 shows the principle.

The phase shift $\varphi_x$ can be determined directly from Fig. 4.4 to be

$$\varphi_x = \frac{t_x}{T_c} \cdot 2\pi + \varphi_0 = 4\pi \cdot \frac{f_c}{c} + \varphi_0 = 4\pi \cdot \frac{X}{\lambda_c} + \varphi_0$$  (4.2)

Where $t_x$ is the time between the two zero crossover point of T and R, respectively. $T_c$ is the period of the emitted wave, $\lambda_c$ is the wavelength and $\varphi_0$ is an initial electrical phase shift. This unintentional initial
Figure 4.4: Continuous wave measurement (see text).

phase shift is originated at the imperfect electrical-mechanical transfer characteristics of the ultrasound transducers on the one hand and also due to the non-ideal acoustic impedance matching of the transducer to the air (see next section).

The major disadvantage of the described CW method, however, is the limited measurement range. As the phase shift \( \varphi_x \) can only be measured uniquely in the interval \( 0 \ldots 2\pi \), the resulting distance measurement range is restricted to half a wavelength, which is only about 1.65 mm at 100kHz.

The restriction of maximum measurement range maybe can be over¬come by combining or modifying both the TOF and the CW method. Apart from the two basic methods various related measurement tech¬niques both for ultrasound and laser range finders are proposed in lit¬erature. Amplitude modulation (AM) is very popular for laser range finders, Chirp modulation uses frequency modulated (FM) pulses like used by some bats at echo location or the frequency modulated con¬tinuous wave (FMCW) method which is similar to the chirp method [9, 35, 36, 34, 33]. Unfortunately, all these techniques also require a cer¬tain amount of bandwidth, which can hardly be provided by the silicon membrane transducers.

Another approach is the resonance CW technique, also mentioned in [9]. It is a variation of the phase shift technique but avoids the limitation of the measurement range due to the phase shift, which can only be uniquely measured in in the range \( 0 \ldots 2\pi \). It does two determinations of phase shift at slightly different frequencies \( f_1 \) and \( f_2 \) and calculates the phase difference \( \Delta \varphi_x \) as a measure of the distance. It can be shown
(see next section) that the distance $X$ is proportional to $\Delta \varphi_x$ as is stated by Eq. 4.3.

$$X = \frac{c}{4\pi} \cdot \frac{\Delta \varphi_x}{\Delta f}, \quad X < \frac{c}{2\Delta f} \tag{4.3}$$

where $\Delta \varphi_x$ denotes the above phase difference and $\Delta f = f_2 - f_1$.

![Figure 4.5](image)

**Figure 4.5**: Resonance CW technique. a) Frequency $f_1$ and $f_2$ applied at the same time. b) Alternate excitation between $f_1$ and $f_2$ and simultaneous adapted membrane transfer function $G_M$ (see text).

The transmitter can be excited at the two frequencies either simultaneously as shown in Fig. 4.5a) or sequentially (Fig. 4.5b)). When the transducer is excited at the two frequencies at the same time, the required bandwidth of the transducer has to be at least as much as the separation $\Delta f = f_2 - f_1$ of the applied signals. For the silicon membranes this turns out to be rather difficult. Due to the extremely narrow bandwidth both the transmitted ultrasound power as well as the receiver sensitivity are substantially reduced at the two frequencies, as shown in Fig. 4.5a).

On the other hand, the resonance CW technique can also be applied
when the transducers are excited by periodically switching back and forth between frequencies \( f_1 \) and \( f_2 \), as illustrated in Fig. 4.5b). This technique is much more convenient for the silicon membrane transducers, since for both frequencies \( f_1 \) and \( f_2 \) the actual membrane resonance can be adjusted by appropriate dc heating power (see Fig. 2.4), even if the frequency separation \( \Delta f \) is large.

Using the resonance CW method by periodically alternating two slightly different excitation frequencies is the basic concept for the novel range finding technique developed in the present research and is called \( \Delta f \) phase shift method.

### 4.3 The \( \Delta f \) phase shift method

![Figure 4.6: The \( \Delta f \) phase shift measurement principle. For schematic illustration, the transmitted and reflected waves from the setup of Fig. 4.1a) are drawn in one plane and the same direction.](image)

In Fig. 4.6 a simplified schematic of the principle is shown. It depicts the two ultrasound waves according to slightly different frequencies \( f_1 \) and \( f_2 \) and illustrates the linear, distance dependent phase difference \( \Delta \phi_x \). For convenience, the transmitted and the reflected wave from the
4.3. The $\Delta f$ phase shift method

setup in Fig. 4.1a) are drawn in the same plane and direction. Note that at the object surface no phase discontinuity occurs and the system can be treated as if transmitter and receiver face each other at a distance $2X$.

In the measurement procedure, the transmitter first emits a continuous ultrasound beam of frequency $f_1$ (wavelength $\lambda_1$). At the receiver, the ultrasound is detected and the phase $\varphi_1$ between the received ultrasound and the electrical signal driving the transmitter is evaluated. Then the transmitter is switched to a slightly different frequency $f_2$ (wavelength $\lambda_2$) and again the phase $(\varphi_2)$ at the receiver is determined. For the two frequencies, using a $2\pi$ phase detector, the phase-versus-distance diagram in Fig. 4.7 results.

![Figure 4.7: The phase diagram for the two frequencies $f_1$, $f_2$.](image)

The shape of the phase plotted against the distance is that of a sawtooth function with a period equal to half a wavelength $\lambda_1$ and $\lambda_2$, respectively. At $X = 0$ an arbitrary initial phase $\varphi_0$ is present which can be e.g. due to the non-ideal electrical transfer characteristics of the transducers (see also Eq. 4.2).

It can be shown that the phase difference $\Delta \varphi = \varphi_2 - \varphi_1$ is a unique representation of the distance $X$, despite the periodicity of the phase and the (unknown) initial phase shift $\varphi_0$. The distance dependent phase difference can be calculated if the following two conditions are fulfilled. Firstly:
\[ \varphi_1(X = 0) = \varphi_2(X = 0) := \varphi_0 \quad (4.4) \]

Which means, that the phase at \( X = 0 \) is the same for both frequencies \( f_1 \) and \( f_2 \) but can have an arbitrary value \( \varphi_0 \). And secondly, the unique measurement range of the distance is limited to

\[ X < \frac{\lambda_1 \lambda_2}{2(\lambda_1 - \lambda_2)} = \frac{c}{2\Delta f} \quad (4.5) \]

Under the conditions mentioned above, the phase difference \( \Delta \varphi \) can have two different values, according to Fig. 4.7.

**Interval 1:** \( n \cdot \frac{\lambda_2}{2} + X_2 < X \leq n \cdot \frac{\lambda_1}{2} + X_1 \), \( n \) = number of half wavelengths within \( X \).

\[ \Delta \varphi = \varphi_2 - \varphi_1 = 4\pi X \left( \frac{1}{\lambda_2} - \frac{1}{\lambda_1} \right) - 2\pi \leq 0 \quad (4.6) \]

**Interval 2:** \( n \cdot \frac{\lambda_1}{2} + X_1 < X \leq (n + 1) \cdot \frac{\lambda_2}{2} + X_2 \)

\[ \Delta \varphi = 4\pi X \left( \frac{1}{\lambda_2} - \frac{1}{\lambda_1} \right) > 0 \quad (4.7) \]

By adding \( 2\pi \) if the phase difference \( \Delta \varphi \leq 0 \), a unique expression for \( \Delta \varphi \) can be derived:

\[ \Delta \varphi = 4\pi X \left( \frac{1}{\lambda_2} - \frac{1}{\lambda_1} \right) = 4\pi X \frac{\Delta f}{c} \quad (4.8) \]

Eq. 4.8 shows the linear relationship between \( \Delta \varphi \) and the distance to measure.
4.4 Electro-acoustic model of the transducer

For the desired distance range in the present application \( X = [0...10 \text{ cm}] \) and a sound velocity of \( c = 340 \text{ m/s} \) the frequency interval \( \Delta f \) has to be 1.7 kHz according to the above equation. The resulting \( \Delta \varphi \) then is zero for the minimum distance at the transmitter and \( 2\pi \) for the maximum 10 cm distance.

### 4.4 Electro-acoustic model of the transducer

An important assumption of the described measurement method is that the phase of the received signal for zero distance \( X = 0 \) is independent of the ultrasound frequency (Eq. 4.4). Using ideal transducers with frequency independent phase characteristics would certainly be ideal to satisfy this condition, however, it is much more difficult to achieve the same with the silicon membrane transducers, whose transfer function are heavily frequency dependent. To get a close examination of the frequency characteristics, especially the frequency dependence of phase, a complete electro-acoustic model of the measurement setup of Fig. 4.6 is derived.

In this model not only the electrical-mechanical transfer functions are considered, but also the influence of acoustic effects due to the package and the channel. The physical constants and transfer functions which appear in Fig. 4.8 of the transmitter are the following:

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>( e )</td>
<td>electrical excitation</td>
</tr>
<tr>
<td>( a_T )</td>
<td>membrane deflection of transmitter</td>
</tr>
<tr>
<td>( p_{0T} )</td>
<td>ultrasound pressure at membrane surface</td>
</tr>
<tr>
<td>( p_{1T} )</td>
<td>ultrasound pressure at outlet</td>
</tr>
<tr>
<td>( f_T )</td>
<td>mechanical force to excite membrane</td>
</tr>
<tr>
<td>( o_T )</td>
<td>electrical output (Wheatstone bridge)</td>
</tr>
<tr>
<td>( K_{EF} )</td>
<td>electro-thermal-mechanical excitation</td>
</tr>
<tr>
<td>( G_M^* )</td>
<td>intrinsic mechanical resonator</td>
</tr>
<tr>
<td>( G_M )</td>
<td>electro-mechanical transfer function</td>
</tr>
<tr>
<td>( Z_{PT} )</td>
<td>acoustic input impedance of package</td>
</tr>
<tr>
<td>( G_{PT} )</td>
<td>pressure transfer function of package</td>
</tr>
</tbody>
</table>

Whereas the table below depicts the same for the receiver.
The generation of ultrasound pressure waves is performed at the transmitter by applying an electrical signal $e$. Due to the electro-thermal-mechanical effect $K_{EF}$, the electrical signal is translated into a force $f_T$ which is the actual mechanical excitation of the membrane. The intrinsic high Q mechanical resonant structure of the membrane is modeled by $G_{M*}$, which converts the applied force $f_T$ in a membrane deflection $a_T$. In the mechanical resonator $G_{M*}$, also the influence of the package has to be considered, especially the packaging of the membrane into a closed housing [21]. Additional resonances can appear,
4.4. Electro-acoustic model of the transducer

which significantly influence the response of the mechanical resonator. The vibrating membrane builds the mechanical-acoustic interface and generates the ultrasound pressure waves.

The relationship between the membrane deflection $a_T$ and the ultrasound pressure generated by the latter can be described by the use of acoustic impedances $Z_a := p/v$ which describe the package influenced propagation of sound waves by the local pressure $p$ and the local phase velocity or particle velocity $v$ (see chapter 4.8.2). A differentiator translates the membrane deflections $a_T$ into the surface velocity of the latter, which is required as the input of the acoustic impedance of the package $Z_{PT}$. The output then is the ultrasound pressure generated just at the surface of the membrane.

Before the ultrasound pressure waves are emitted into the free air (channel $G_{CH}$), they have to pass the transducer package, which also can significantly influence the pressure waves. As the consequence, a pressure transfer function $G_{PT}$ is defined in the present model, which takes into account the acoustic properties of the package. Note, that the acoustic impedance $Z_{PT}$ from the paragraph above, and the pressure transfer function $G_{PT}$ underly the same physical equations, therefore always have to be considered together.

Due to the presence of the Wheatstone bridge, the membrane deflections can be measured through the membrane sensitivity constant $K_{AV}$ and are available as an electrical output $o_T$. The combination $K_{EF} \cdot G_M \cdot K_{AV}$ then is identical to the membrane transfer function $G_M$ of Fig. 2.3 and can be measured easily. $G_M$ hence is a scaled version of the intrinsic resonator $G_M*$ and therefore can be used to qualitatively describes the membrane behavior (which was already done in chapter 2).

At the receiver, the ultrasound pressure waves again pass the package ($G_{PR}$) before reaching the surface of the receiving membrane. There, the ultrasound pressure is translated to a force $f_R$ which is the input to the same intrinsic mechanical resonator $G_M*$, as described above. The resulting receiver membrane deflection $a_R$ is converted into voltage by the sensitivity constant $K_{AV}$ to build the overall output of the measurement setup $o_R$.

According to the $\Delta f$ phase shift method, the overall phase shift in the present model should only depend on the propagation of the pres-
sure waves in the channel $G_{CH}$ and not on other electrical or mechanical effects. However, the frequency dependent phase behavior of the above model is not only determined by the sound propagation, but also by the transducer characteristics, namely the package of the transducer and especially the high Q intrinsic mechanical resonator (see measurement of $G_M$ in Fig. 2.4). As we know from previous considerations on the membrane transfer function, these phase shift contributions are additionally not constant but rather depend on temperature and process variations.

The question therefore arises, how can the phase shifts of the transducers be controlled in terms of frequency independence? Can the $\Delta f$ phase shift method still be applied? Both questions can be answered positively as will be shown in the following sections. In the present research, both solutions to the control of the electrical as well as the acoustic phase shift have been found. The package-induced phase shift and other acoustic problems are solved by appropriately designed broad band packages (see section 4.8) whereas the problem of frequency dependent phase shift due to $G_M*$ is solved by the use of phase locked loop (PLL) circuits, as will be presented in the following section.

### 4.5 Phase locked ultrasound

To lock the periodic membrane deflection to a fixed, frequency independent phase relation to the electrical excitation, the silicon membrane transducer is embedded into the loop of an electro-mechanical phase-locked-loop (PLL) circuit as shown in Fig. 4.9.

The concept is derived from the well known PLL-circuits widely used in telecommunication applications [37, 38] and consists of a phase detector, a loop-filter (LF) and a voltage controlled oscillator (VCO). In case of the present PLL, the VCO is the electro-mechanical oscillator of Fig. 3.3 with the dc heating voltage as the controlling voltage. The advantage of the circuit is not only that the output phase, which reflects the actual membrane deflections, is locked to the electrical input signal. The electro-mechanical VCO also provides that the membrane operates always at its resonance (see chapter 3) where the emitted sound pressure is maximum.

For the analysis of the PLL normally a linear model is used, which
4.5. Phase locked ultrasound

![Diagram: Electro-mechanical phase locked loop circuit with the circuit of Fig. 3.3 as VCO.]

Figure 4.9: Electro-mechanical phase locked loop circuit with the circuit of Fig. 3.3 as VCO.

describes the system in the phase domain. Fig. 4.10 depicts the corresponding model for the above circuit.

![Diagram: Small signal model of the electro-mechanical PLL circuit of Fig. 4.9.]

Figure 4.10: Small signal model of the electro-mechanical PLL circuit of Fig. 4.9.

The analysis of electronic PLL using a small signal model in the phase domain has been studied extensively in literature and only the relevant considerations are repeated once again in here. In an electronic PLL, the frequency response of the overall phase transfer function $\Phi_{out}/\Phi_{in}$ normally is that of a second order system, whose characteristics are dominated by the transfer function of the loop filter $G_{LF}(s)$. The residual phase difference $\varphi_e = \varphi_{in} - \varphi_{out}$ is calculated to be
\[ \varphi_e = \frac{\Delta \omega}{K}, \quad K = K_{VCO} \cdot K_{PD} \]

where \( \Delta \omega = 2\pi \Delta f \) is a frequency step at the input, \( K_{VCO} \) is the gain of the VCO and \( K_{PD} \) reflects the gain of the phase detector. The gain of the phase detector \( K_{PD} \) depends on the latter’s type. An EXOR phase detector exhibits a \( K_{PD} \) of \( V_{dd}/\pi \) where \( V_{dd} \) is the supply voltage. For a digital \( 2\pi \) phase phase detector, \( K_{PD} \) becomes \( V_{dd}/2\pi \).

What is different in the electro-mechanical PLL of Fig. 4.10 compared to a pure electronic type, however, is the special characteristics of the electro-mechanical VCO. Other than in a typical electronic VCO, the response to a changing input voltage is not followed by an immediate change of the oscillating frequency. The change of oscillating frequency can only be as fast as allowed by the large time constant of the electro-thermal excitation (see chapter 2.3) because the mechanical parameters responsible for the resonant frequency (buckling) only alter as fast as the thermal stimulation. In the above circuit, the electro-mechanical VCO is therefore modeled as an ideal VCO \( \left( \frac{K_{VCO}}{s} \right) \) which is controlled by a large time constant low-pass \( G_{ML}(s) \) with gain equal to one. As we will see later in this chapter, \( G_{ML}(s) \) corresponds to the low frequency part of the membrane transfer function \( G_M(s) \) shown in Fig. 2.3.

To verify above consideration and to determine the VCO characteristics \( (K_{VCO}, G_{ML}(s)) \), a measurement is performed with the setup shown in Fig. 4.11.

![Diagram](image_url)

**Figure 4.11:** Measurement setup to extract the VCO characteristics.

The electro-mechanical VCO is attached to the REF output of a network analyzer which generates an ac signal whose frequency is swept.
between 10 Hz and 1 kHz. The ac signal is superimposed with a dc source \( V_{dc} \) in order to bias the VCO at a free running frequency \( f_0 \) of 93.8 kHz and 100.5 kHz, respectively. The output of the VCO is fed to a frequency demodulator (a broad band PLL) which detects the instantaneous VCO frequency and converts it into a voltage \( V_{FM} \). Feeding back the signal \( V_{FM} \) to the R input of the network analyzer enables the measurement of the frequency response \( V_{FM}/V_{in} \) which is equal to \( G_{ML}(s) \) when normalized to 0 dB for \( s = 0 \). Fig. 4.12 shows the measurement result, for the two different frequencies mentioned above. For both frequencies, the measured frequency response is that of a first order low-pass with a 3 dB bandwidth of 150 Hz. The measurement hence shows the accurate agreement between the low frequency part of the membrane transfer function \( G_{M}(s) \) (Fig. 2.3) and \( G_{ML}(s) \). The gain \( K_{VCO} \) of the VCO has been measured to be \( 2\pi \cdot 8.5 \text{ kHz/V} \) which is, not surprisingly, equal to the slope of the resonant frequency vs. heating voltage measurement in Fig. 2.5 scaled by a factor \( 2\pi \).

**Figure 4.12:** Measurement of \( G_{ML}(s) \) of a membrane fabricated in a 2μm CMOS technology.

A prototype circuit has been built to verify the PLL principle of
Fig. 4.9 and the idea of an electro-mechanical PLL was confirmed. However, the original concept turned out to be not best suited for the range finder application, due to the following reasons.

The residual phase $\varphi_e$ depends on the applied frequency difference $\Delta f$ and the overall loop gain $K$. For high gain $K$ the residual phase $\varphi_e$ becomes independent of the frequency, as is required by the first condition (Eq. 4.4) of the $\Delta f$ phase shift method. However, evaluating $K$ for the electro-mechanical PLL using a $2\pi$ phase detector operating at a 5 V supply leads to $K = K_{VCO} \cdot K_{PD} = 2\pi \cdot 8.5 \text{ kHz/V} \cdot 5/2\pi \text{ V/rad} = 4.25 \cdot 10^4 \text{ rad/s}$. The residual phase due to the applied frequency step of 1.7 kHz in the range finder application then is $\varphi_e = 0.25 \text{ rad}$ or $14.4^\circ$. As this phase difference cause an intolerable error in the overall measurement result, the basic circuit structure of Fig. 4.9 cannot be used for accurate measurements based on the proposed phase shift method.

In addition, the design of the loop filter required to suppress the high frequency components at the output of the phase detector is aggravated by the presence of the additional low frequency pole (150 Hz) due to the membrane ($G_m(s)$). The closed loop system of Fig. 4.10 then is a higher order system which can easily become unstable if the pole of the low-pass type loop filter is close to 150 Hz.

Luckily, both of the difficulties mentioned above can be overcome by using a modified PLL structure, as shown in Fig. 4.13.

Instead of using an electro-mechanical VCO in the loop of a PLL to control the periodic membrane deflections, the same task is performed by a feed forward structure and an additional integrator building block in the modified circuit. As we will see in the next paragraph, this combination (shaded part of Fig. 4.13) then behaves like a VCO in a limited range for the membrane heating voltage.

The membrane is stimulated at a fixed input frequency which generates a membrane response according to the transfer function $G_m(s)$ of Fig. 2.3. Depending on the actual resonant frequency, the phase can be everything between $-90^\circ$ and $+90^\circ$ as expected in a second order bandpass system. On the other hand, if the input frequency matches the resonant frequency, the phase shift is zero degree and the amplitude response is maximum at the same time. This operating point is exactly the same as when the membrane is used as an electro-mechanical VCO.
4.5. Phase locked ultrasound

Figure 4.13: Improved electro-mechanical PLL: Better stability properties and negligible residual phase $\phi_e$. The shaded part behaves like a VCO in a limited range for $V_{dc}$.

and hence is highly suitable for the $\Delta f$ phase shift method. To find this dedicated operating point, the membrane resonant frequency has to be tuned until above matching is reached and this can be done ideally with the dc part $V_{dc}$ of the heating voltage.

The improved PLL circuit in Fig. 4.13 provides the control loop required to achieve an automatic matching of the resonant and the input frequency. The output phase is compared to the input reference and leads to a phase error $\phi_e$. As in the previous PLL, a subsequent LF suppresses high frequency signals of the phase detector. After the phase error $\phi_e$ is translated by the LF into a voltage at node $A$, it is compared to a reference voltage $V_{ref}$ and the difference is integrated to build the dc part of the membrane heating voltage. After settling of the control loop, the voltage at node $A$, which represents the phase error $\phi_e$, becomes equal to $V_{ref}$ due to the (infinite) high dc gain of the integrator. The reference voltage $V_{ref}$ hence is set to a level so that the phase between $REF_T$ and the output of the membrane is zero degrees and at the same time the amplitude of the emitted ultrasound is maximum.

On the other hand, for an EXOR type phase detector, the ideal operating point is achieved, when the input signals have around $90^\circ$ phase shift, rather than if they are complete in phase. The output voltage after the LF then is $V_{dd}/2$ with $V_{dd}$ equals the supply voltage of the EXOR.
To maintain correct operation of the above circuit using an EXOR phase detector nevertheless, the reference voltage is set to \( V_{\text{ref}} = V_{dd}/2 \) and the required additional 90° offset phase shift is added in the feedback loop of the PLL.

An important advantage of this modified PLL structure is that the LF after the phase detector is not really required. The integrator block in front of the membrane not only secures zero degree phase error \( \varphi_e \), but also sufficiently suppresses higher order harmonics of the phase detector.

### 4.5.1 Circuit analysis

For the analysis of the electro-mechanical circuit, especially for stability and settling consideration, a closer examination of the membrane characteristic has to be performed again. The most critical characteristic according to the above paragraph, is the phase and the amplitude response of the membrane transfer function \( G_M \), this time not as function of the frequency but rather as a function of the dc heating voltage \( V_{dc} \). Fig. 4.14 shows the corresponding measurement, for three different excitation frequencies.

The measurement of Fig. 4.14 shows that the set of phase responses \( \varphi_{GM}(V_{dc}) \) are non-linear, but monotonic functions. The steepness is highest at \( \varphi_{GM} = 0^\circ \) and at the same operating point, the amplitude response has the maximum. Due to the monotonicity of the phase transfer-function, the correct large signal behavior of the modified electro-mechanical PLL is guaranteed. This means, that the circuit will always find the required operating point at zero degrees phase shift, during start-up as well as at a switching from one to another input frequency.

On the other hand, to have insight in the stability and settling behavior of the circuit, a small signal model is derived which uses a linearization of the transfer-function around the operation point. The corresponding critical parameter is defined as phase gain \( K_\varphi = \Delta \varphi_{GM}/\Delta V_{dc} \) and describes how strong the phase varies when the applied heating voltage changes. \( K_\varphi \) depends on the applied excitation frequency and is measured to be \( K_\varphi = 10.7...16.5 \text{ rad/V} \) for excitation frequencies between 88 kHz and 96 kHz.

The resulting small signal model is depicted in Fig. 4.15 and rep-
4.5. Phase locked ultrasound

Figure 4.14: Measured membrane transfer characteristics as a function of the dc heating voltage $V_{dc}$. Excitation amplitude = 1.9 V.

represents the system in the phase domain, similar to the PLL previously described. As the function of the LF is not really necessary according to the consideration mentioned above, the LF is omitted in the following.

The model consists of a phase detector with gain $K_{PD}$, an integrator $K_{IN}$ with $K_{IN} = 1/T_i$, $T_i$ = time constant of integrator, and a dedicated model for the membrane phase response. As is explained in the above paragraph, the phase gain $K_{\phi}$ as part of this model describes how the signal after the integrator is translated into the output phase of the membrane. Due to the same slow electro-thermal mechanism described earlier in this section, the phase response to a changing $V_{dc}$ is again limited in speed. As we know already, this effect can be accurately be modeled by a first order low-pass transfer function $G_{ML}(s)$ (see Fig. 4.12) which is included in the above model after $K_{\phi}$. To complete the small signal model, the feed forward path is included in the phase summing node in front of the low-pass transfer function $G_{ML}(s)$. 
The resulting solution for the phase error \( \Phi_e \) in the Laplace domain is calculated by using a first order low-pass function \( G_{ML}(s) = \frac{\omega_{ML}}{s + \omega_{ML}} \)

\[
\Phi_e = \Phi_{in} \cdot G_e(s) = \Phi_{in} \cdot \frac{s^2}{s^2 + s\frac{\omega_{PL}}{Q_{PL}} + \omega_{PL}^2}
\]  

(4.10)

where

\[
\omega_{PL} = \sqrt{K \cdot \omega_{ML}} \quad \text{and} \quad Q_{PL} = \frac{\omega_{PL}}{\omega_{ML}}
\]  

(4.11)

\( K = K_{PD} \cdot K_{IN} \cdot K_p \) is the overall loop gain of the PLL. The phase error transfer function \( G_e(s) = \Phi_e / \Phi_{in} \) is a second order high pass function. For a frequency step \( \Phi_{in} = \Delta \omega / s^2 \), the resulting phase error after settling becomes

\[
\varphi_e(t \rightarrow \infty) = \lim_{s \rightarrow 0} s \cdot \frac{\Delta \omega}{s^2} \frac{s^2}{s^2 + s\frac{\omega_{PL}}{Q_{PL}} + \omega_{PL}^2} = 0
\]  

(4.12)

as expected.
4.6. Self-calibrating receiver

The characteristic frequency $\omega_{PL}$ of the system depends on the low frequency pole $\omega_{ML}$ and the overall loop gain $K$. $\omega_{PL}$ as well as $Q_{PL}$ can be set by the gain $K_{IN} = 1/T_i$ of the integrator, so that the PLL is stable, despite variations of the membrane parameter $K_v$ (see measurement in Fig. 4.12).

Based on the above design consideration, a discrete prototype PLL has been implemented as part of the complete range finder system (see chapter 4.7). The integrator has a time constant of $T_i = 9$ ms and the EXOR type phase detector exhibits a $K_{PD}$ of $3/\pi$ V/ rad. The required 90° degrees phase shift in the feedback loop for the EXOR is included in the whole front-end amplification chain by a slight mismatched between center frequency and signal frequencies around 100 kHz. The resulting characteristic frequency $\omega_{PL}$ becomes around $2\pi \cdot 190$ Hz according to Eq. 4.10 and $Q_{PL}=1.27$. As a representative example of the PLL, Fig. 4.16 shows the start-up transient measurement for one particular frequency.

The system is settled 25 ms after the (digital) 98 kHz input signal is applied. In the first 15 ms, the integrator output changes linearly with time, which is a typical large signal phenomena in many control loops and is called slewing. Slewing limits the speed for large input changes, however, for small changes of the input frequency, as present in the $\Delta f$ phase shift method, the slewing effect is negligible.

With the present PLL structure, the periodic membrane deflection can be controlled reliably. Independent of the frequency, the phase shift is zero degree and the emitted ultrasound power is maximum. The time needed to settle between two different frequencies depends on the characteristic frequency $\omega_{PL}$ and the corresponding quality factor $Q_{PL}$, and lies between 10 to 20 ms.

4.6 Self-calibrating receiver

The task of the ultrasound receiver is to translate the incident ultrasound waves, which are reflected by the object, into an electrical signal of sufficient quality and level. Moreover, to fulfill the phase requirement of the $\Delta f$ phase shift method, no additional frequency dependent phase shift should be added due to the membrane or the receiver electronics.
The quality (noise, offset, linearity) of the received signal is mainly determined by a high gain, ac-coupled front-end amplifier as has already been discussed in the previous chapter of the thesis. Apart from the discussion about signal quality which basically is interesting for the resolution of the ultrasound range finder, the behavior of the receiver in terms of phase shift is of primary interest for the system.

According to the receiver model in Fig. 4.8, the frequency dependent phase behavior is dominated again by the intrinsic membrane resonator $G_M^*$ and therefore exhibits zero degree phase shift and maximum sensitivity only at the latter's resonant frequency. A self-calibrating procedure is required to secure action at this particular operating point, independent of the frequency. To perform the resulting automatic tuning of the membrane resonant frequency, the concept of using an electro-
mechanical PLL again is ideal, this time not to achieve maximum emitted ultrasound, but rather maximum sensitivity with zero degree phase shift. Fig. 4.17 shows the schematic representation of the circuit, which is implemented as a mixed analog/digital signal solution.

The circuit consists of a phase detector, a (digital) discrete time integrator followed by two registers (f1,f2) and a digital-to-analog (D/A) converter. The switch S1 chooses whether the output of one of the registers or the integrator is connected to the D/A-converter, where the dc heating voltage is generated and added to the input signal. The output of the membrane (Wheatstone bridge) is amplified, quantized and fed back to the phase detector.

During the calibration cycle, the switches S1, S2 are closed and the resulting circuit becomes identical to the transmitter PLL of Fig. 4.13. The membrane is excited by an electrical input signal with small amplitude and frequency $f_1$. After settling of the loop, a dc heating voltage is established at the heating resistor of the membrane, for which the membrane operates at maximum sensitivity. At the same time, the output of the integrator represents the digital version of the dc heating voltage for frequency $f_1$ which can easily be stored as a calibration value in register f1. Then, the same procedure is done for the second frequency ($f_2$) or any further frequencies.

After calibration, the small electrical input signal is switched off (S2)
and the individual dc heating voltage for one of the frequencies is simply selected by setting the switch S1 to read the corresponding register. The receiver circuit then is adjusted to maximum sensitivity for a incident ultrasound wave with that particular frequency and the amplified and quantized ultrasound signal is obtained at the terminal \( OUT_R \) or the phase of the latter with respect to \( REF_R \) can be read after the phase detector.

The choice of a mixed analog/digital architecture for the self-calibrating receiver is due to two reasons, both with respect to a future realization of the system in an IC technology (see chapter 5). The first reason, which also affects the design of the transmitter, is the large time constant of the integrator. To implement continuous time integrators with time constants in the millisecond range, very large capacitors are needed, which is too expensive for a cost effective integrated solution. As the consequence, for very low frequency applications, the use of digital signal processing often is the only alternative.

The second reason concerns the storage of the calibration value (dc heating voltage). As these values have to be constant over time, it is convenient to use a digital register for the storage of the corresponding calibration value. In the above self-calibrating receiver, both requirements - low frequency signal processing as well as storage capabilities, could be combined ideally by the mixed analog/digital version of the electro-mechanical PLL circuit.

### 4.6.1 Mixed signal analysis

For a mixed analog/digital system like the self-calibrating receiver, where continuous time signal processing is replaced by sampled and digital circuits, two questions arise immediately and have to be answered before analyzing the functionality of the system:

How much is the sampling frequency and how many bits are required to represent the desired analog quantities. To explain the situation for the present self-calibrating circuit, the critical interfaces are depicted in Fig. 4.18, showing the mixed signal replacement of the integrator of Fig. 4.13.

The digital phase detector (PD) builds the interface from the analog continuous time (\( \sim \)) phase domain to its sampled digital (\( # \)) representa-
4.6. Self-calibrating receiver

The conversion from continuous time to sampled signals certainly underlies the well known sampling theorem, which requires a sampling frequency $f_s$ which is at least twice as the bandwidth of the continuous signal spectrum. Due to the narrow bandwidth (150 Hz) of the membrane transfer function ($G_{ML}$), the output of the membrane as the input of the phase detector is limited in bandwidth already as well as the reference $REF_R$ which is slowly switched between two single frequencies. The sampling condition can be fulfilled easily without an additional anti-aliasing filter for sampling frequencies well above 320 Hz. To achieve acceptable noise performance in the system, however, the sampling frequency $f_s$ is chosen not to that lower limit but rather as high as reasonable for the digital signal processing and the overall power consumption.

Fig. 4.19 shows a detailed schematic of the digital 8 bit $2\pi$ phase detector. A quantization level of 8 bit is chosen because the overall range finder system is targeted to have 8 bit resolution over the full (0...10 cm) measurement range, which directly translates to 8 bit resolution in phase accuracy due to the linear relationship in Eq.4.8.

The synchronously clocked phase detector consists of two edge detectors which generate a single positive pulse for every positive edge of the two input signals $REF$ and $IN$, respectively, followed by a JK-flip-flop. The length of the resulting pulse at the output of the flip-flop ($CE$) is measured by a resetable counter, whose output $CNT$ is stored in a register at the end of the period of the reference signal. Every end of the reference period, a new digitized phase value $\Phi_k$ is present at the output of the phase detector. Fig. 4.20 shows the corresponding timing diagram of the individual signals.
To achieve an 8 bit resolution for input signals around 100 kHz, the clock frequency has to be at least 256 times the frequency of the input signal, which translates to $f_{CLK} = 25.6$ MHz. Rather than using an independent clock frequency for the digital phase detector, it is much more convenient to use a clock signal with exactly and always 256 times the frequency of the signal $REF$ (coherent sampling). The scheme not only simplifies the design of the overall digital system but also, which is more important, enables the digital phase detector to measure the phase independent of the signal frequency. For a (continuous time) phase $\varphi$ between $REF$ and $IN$, the integer $\Phi_k$ then becomes always $\Phi_k = \text{round}(\varphi / 2\pi \cdot 2^8)$, independent of the frequency of $REF$. 
4.6. Self-calibrating receiver

Furthermore, it's reasonable to use the signal frequency $f_s$ as sampling frequency for the complete mixed signal receiver of Fig. 4.17 since at every end of the reference period a new digital phase sample is generated and ready to be processed further. With a frequency of about 100 kHz, $f_s$ is well suited to fulfill both sampling condition and at the same time is the highest possible sampling frequency for the above simple digital phase detector.

The digital-to-analog (D/A) converter at the end of the mixed/signal integrator translates the sampled digital values back to the analog signals within the loop of the electro-mechanical PLL. Other than in case of the digital phase detector considered above, the decision of how many bits or dynamic range the converter should have is more difficult here. The reason is the non-linear phase transfer function $\varphi_{GM}(V_{dc})$ of the subsequent membrane, which has been discussed already in section 4.5.1 and whose measurement is shown in Fig. 4.14. To achieve an overall 8 bit resolution along the signal path of the PLL despite the above non-linear transfer function, the maximum allowed quantization step $\Delta V_{dc}$ of the D/A converter (operating at the maximum steepness at resonance) is $\Delta V_{dc} = \Delta \varphi_{GM}/K_{\varphi} = 2\pi/2^8/K_{\varphi}$. As the point of maximum steepness, equal to the operating point of the circuit, can be anywhere within the interval $V_{dc} = V_1...V_2$, the above maximum allowed quantization step has to be held also within this interval. This leads to a minimum dynamic range of the converter of $V_{DA} \cdot K_{\varphi} \cdot 2^8/2\pi$ where $V_{DA} = V_2 - V_1$ is the required output voltage range of the D/A converter. For the membrane measured in Fig. 4.14, the minimum dynamic range is calculated to 11...12 bits, which is much higher than the overall 8 bit phase accuracy.

After knowing the details about sampling frequency and dynamic range (number of bits) the remaining part to explain in the mixed signal receiver is the sampled digital integrator and the functionality of the overall analog-digital-analog system. Fig. 4.21 shows the implementation of the first order integrator using a 16 bit register plus a 16 bit adder/subtractor.

The input of the digital integrator is laid out for the 8 bit values delivered by the digital phase detector whereas the output is 12 bit wide to provide samples for the subsequent 12 bit D/A converter. The present simple integrator as part of the mixed signal receiver PLL not only acts to force the input samples $\Phi_{in}$ to a reference $\Phi_{ref}$ like the analog counter.
part in Fig. 4.13, but also provides an elegant interpolation of the 8 bit quantized input samples to the 12 bit output.

What is interesting and deciding for the overall mixed signal PLL, however, is how the digital sampled integrator, together with the phase detector and the D/A converter behaves as a quasi continuous time solution within the system. The analysis can be performed easily by using the relationship between the discrete time $z$-transform to the continuous time Laplace-transform, that is $z = e^{sT_s}$ with $T_s = 1/f_s$ equals the sampling interval. The discrete time transfer function $G_{IN}(z)$ of the integrator in Fig. 4.21 then can be transformed approximately for frequencies well below the sampling frequency or $sT_s \ll 1$ to

$$G_{IN}(z) = \frac{1}{a} \cdot \frac{1}{1 - z^{-1}}$$

which corresponds to a continuous time integrator with a time constant $T_i = a \cdot T_s$. This continuous time model described by Eq. 4.13 can now be included in the PLL analysis performed in the previous sections. The overall gain of the mixed signal PLL then becomes $K = K_{PD#} \cdot K_{IN#} \cdot K_{DA} \cdot K_\varphi$ where $K_{PD#} = 2^8/2\pi$, $K_{IN#} = 1/a/T_s$ and $K_{DA} = V_{DA}/2^{12}$. For the same membrane used in the transmitter
4.7. The range finder system

PLL, a sampling frequency of 100 kHz, a=16 and $K_{DA}=2.5V/2^{12}$, the characteristic frequency of the mixed signal PLL then becomes $\omega_{PL}=2\pi \cdot 230$ Hz and the $Q_{PL}=1.53$.

The electro-mechanical receiver PLL has been realized as part of the whole range finder system using discrete components for the front-end amplifier chain, the comparator at the input of the digital phase detector and the power amplifier to drive the heating resistor. The digital part is implemented in this feasibility study using field programmable gate arrays (FPGA) which are best suited for prototyping purpose.

In the present mixed signal scheme for an ultrasound receiver, the concept of a PLL circuit can be elegantly used as self-calibrating procedure, which tunes the membrane to have zero degrees phase shift and maximum sensitivity for the particular ultrasound frequencies used for the $\Delta f$ phase shift measurement. Together with the transmitter PLL considered in the previous section it builds the prerequisite for the successful application of the silicon membrane for this measurement method. In the next section, the overall control system is explained, which relies on the above electro-mechanical transmitter PLL and self-calibrating receiver as improved ultrasound transducers.

4.7 The range finder system

In Fig. 4.22 a block diagram of the complete range finder system is depicted.

It consists of the above transmitter and receiver circuits, a digital controller and a frequency synthesizer, which uses a crystal oscillator as precise frequency reference. The frequency synthesizer generates the system clock in the 25 to 30 MHz range, from which the accurate reference signal $REF$ at frequency $f_1$ or $f_2$ is derived by dividing the clock signal by 256. For the desired distance range of $X=[0...10 \text{ cm}]$, $f_2$ is 101.7 kHz and $f_1=100$ kHz (see example in section 4.3), which translates into clock frequencies of 26.035 MHz and 25.6 MHz, respectively. The two frequencies are selected by the controller by changing the variable ratio divider of the frequency synthesizer ($F$ signal).

The heart of the digital controller, which covers all the relevant blocks
needed for the $\Delta f$ phase shift measurement, is depicted in Fig. 4.23. According to the measurement principle described in section 4.3, it performs the measurement of the phase at the two frequencies $f_1$ and $f_2$ and calculates the phase difference as digital 8 bit value $\Delta \Phi$.

Rather than calculating the phase difference after the phases for both frequencies are determined, the same task is performed with a different scheme using two phase detectors ($\varphi, \varphi'$) and a programmable delay circuit $T$ (see Fig. 4.23).

The idea is that after the measurement of either $\varphi_1(f_1)$ or $\varphi_2(f_2)$ a phase shift $-\varphi_1(f_1)$ or $-\varphi_2(f_2)$ is added to the path of the input signal by the delay circuit, so that the resulting output phase $\Phi$ becomes zero degree. During the next measurement interval, the phase difference $\Delta \Phi$ then directly appears at the output, either with a positive sign ($f_1 \rightarrow f_2$), or as negative ($f_2 \rightarrow f_1$) value. The additional calibration value for the programmable delay circuit is provided by an auxiliary phase detector $\varphi'$, whose inverted output is stored in a register at the end of each measurement interval (signal $CAL$). To obtain a new, positive $\Delta \Phi$ for every transition $f_1 \rightarrow f_2$ and $f_2 \rightarrow f_1$, a buffer is added, which simply multiplies by one or minus one, depending on which of the two transition is present.
4.7. The range finder system

![Diagram of the range finder system](image)

**Figure 4.23:** The heart of the range finder system.

The advantage of the circuit described above is that the modulo $2\pi$ limitation, which makes necessary the consideration of phase in different intervals (see Fig. 4.7), of the main phase detector $\varphi$ is circumvented. After the calibration value for the digital delay circuit is applied, the linear output range of $\varphi$ is a full $2\pi$ interval, and the phase difference can be measured directly.

In Fig. 4.24 a timing diagram of the system is depicted, which gives detailed insight in the operation of the above circuit as well as illustrates the self-calibrating process of the receiver during start-up.

The phase detector $\varphi$ is the same as used for the self-calibration of the receiver. During this procedure, the delay circuit gets a fixed value $\Phi_0$, selected by the multiplexer and $TEN$. The reason for this fixed delay (or phase shift) is again due to the operating point of the $2\pi$ digital phase detector, which is ideally at $180^\circ$ phase shift (see also the discussion for the EXOR phase detector used in the transmitter PLL). To achieve zero degrees phase shift between reference and the output of the membrane as well as to operate the digital phase detector at this particular operating point, the value for $\Phi_0$ is set to $128$. 
### 4.7.1 Noise improved phase detector

Apart from the modulo $2\pi$ limitation and the operating point considerations, the digital phase detector as interface between the continuous time membrane input signal and the sampled digital controller requires another important discussion: noise. As the signals obtained at the receiver membrane are very small, especially for higher distances, the general noise level in the front-end electronics and the quantization stage becomes an important issue. The origin of noise in the receiver electronics is basically due to the thermal noise of the resistors of the Wheatstone bridge and thermal (and also *Flicker*) noise from the front-end amplifier (see also chapter 3).

The input phase as well as the noise contribution is sampled every 10 $\mu$s and leads to a quantized phase plus a digital noise signal with variance $\sigma_1$, which can be well above the targeted 8 bit quantization level. Fig. 4.25 shows the phase transfer function ($\varphi \to 8$ bit code) of the original phase detector of Fig. 4.19, if a noisy signal is applied.

![Timing diagram of the range finder system.](image)

**Figure 4.24:** *Timing diagram of the range finder system.*
4.7. The range finder system

Figure 4.25: Phase transfer function \( \varphi \rightarrow 8 \) bit code including statistical code variation due to noise.

The 8 bit code is composed of the phase value according to the ideal (noiseless) phase transfer function and a statistical deviation represented by a \( 2\sigma \)-band around the latter. At \( \varphi = 0 \) and \( \varphi = 2\pi \) the output of the phase detector even jumps between low numbers around the minimum (0) and the maximum (255), which is due to the modulo \( 2\pi \) limitation of a \( 2\pi \) phase detector.

As the noise contribution in the range finder system turned out to be much too high when using the original phase detector, a modified digital phase detector has been developed which has an improved noise performance.

The reduction of noise in a system can be achieved by limiting the latter’s bandwidth, which is performed by low-pass (band-pass) filtering the signal of interest together with the superimposed noise contribution. The minimum applicable bandwidth thereby is given by the bandwidth of the signal. As the settling time of the electro-mechanical PLL is in the order of 10 ms, plenty of time is available for low-pass filtering the digital phase values during every measurement interval and therefore limiting the latter’s phase signal and noise bandwidth. In Fig. 4.26 two possibilities are illustrated, with which the low-pass filtering can be performed.

One possibility is the integration of the signal \( CE \) of the original phase detector (Fig. 4.19) over \( N \) reference periods instead of only one,
Figure 4.26: a) Phase detector circuits with reduced noise bandwidth. b) Phase transfer function of the circuits from a). $f_{bw}$ is the remaining noise bandwidth (equal to the bandwidth of the low-pass filter).

Another solution is the filtering with a subsequent first order low-pass filter. Both solutions lead to the phase transfer function of Fig. 4.26b, where the digital noise signal exhibits a reduced variance $\sigma_2 = \sigma_1 \sqrt{2f_{bw}} f_s$. $f_{bw}$ denotes the remaining noise bandwidth, which in case of the low-pass filter is equal to the latter's bandwidth.

Around the particular phases $\varphi = 0$ and $\varphi = 2\pi$ in the diagram Fig. 4.26b, the filtering (averaging) of the statistical values leads to a smooth transition from 255 to 0. For $\varphi = 2\pi$ a number around $\Phi = 128$ is generated, which is the same value as if $\varphi = \pi$. Due to this averaging effect, the mapping between the input phase $\varphi$ and the digital output $\Phi$ hence no longer is a one-to-one relationship within the interval $\varphi = 0...2\pi$. As the $\Delta f$ phase shift requires this one-to-one mapping in order to continuously compose the phase difference from the phase intervals in Fig. 4.7 and to achieve the desired linear continuous relationship of Eq. 4.8, the effect described above generates errors. At every transition ($2\pi \rightarrow 0$) a wrong phase $\Delta \varphi$ is generated, which is unacceptable.

The problem can be solved by using a modified structure for the
4.7. The range finder system

In terms of digital signal processing (low-pass filtering), the structure is equivalent to the IIR low-pass filter depicted in Fig. 4.26. However, instead of feeding back the output of the integrator to the summing node after the phase detector, it is fed to the programming input of a digital delay circuit (same as in Fig. 4.23), which performs an additional phase shift of the input signal prior to the phase detector. Due to the integrator in the feedback loop of the circuit, the output of the phase detector is forced to a predefined level \( \Phi_r = 128 \). The advantage of biasing the phase detector output at a fixed level in the middle of the operating range is that the unintentional averaging of the statistical jumps at \( \varphi = 0 \) and \( 2\pi \) no longer occur. The phase transfer function then becomes as shown in Fig. 4.27. The digital noise signal has the same reduced variance \( \sigma_2 = \sigma_1 \sqrt{\frac{2f_{bw}}{f_s}} \) as found already for the low-pass previously described and the transition at \( \varphi = 2\pi \) no longer is a smooth curve but rather a sharp edge.

The noise improved phase detector of Fig. 4.27 is well suited for the range finder system, since it converts the phase to the correct 8 bit number with subsequent low-pass filtering within the full \( 2\pi \) input range. The bandwidth \( f_{bw} \) of the low-pass filter is set by the time constant of the integrator. A high time constant strongly reduces the influence of noise on the output accuracy but at the same time increases the settling time of the phase detector. In the present design, a bandwidth of \( f_{bw} = 250 \text{ Hz} \) was chosen, which sufficiently suppresses the noise contri-
bution and at the same time has a settling time of 3.5 ms well below the time of the measurement interval (10-20 ms).

With the range finder system of Fig. 4.22 all important distance measurements of the present thesis have been performed - to prove to principle, to investigate the ultrasound acoustic and to derive the specifications for a subsequent integration of the system on a chip.

As an example of what signals are obtained in the system, Fig. 4.28 shows the measured receiver and the heating voltage $V_{dc}$, when the reference frequency is switched from $f_1=98$ kHz to $f_2=99.555$ kHz. To simplify the design of the frequency synthesizer, the frequency difference $\Delta f = f_2 - f_1=1.555$ kHz is somewhat slower as the required 1.7 kHz, which gives a slightly larger measurement range of $X=[0...11$ cm].

Although it would be logical, the actual measurement of the phase difference $\Delta \Phi$ over the full distance range, provided by the described digital controller, is not presented at this stage, but rather in section 4.9. The reason for this is that one very critical element in the overall range finder system is not considered in much detail so far, although it is
4.8 Transducer package

The task of a miniaturized package of an ultrasound micro-system is twofold. On the one hand, it is needed to protect the sensitive transducers and also the electronics against environmental influences like humidity, dust or light. On the other hand, it builds a dedicated interface between the ultrasound generation/detection on the membrane surface and the free air. The two tasks are often conflicting and the optimization of both requirements leads to the tradeoff: good protection properties versus good sensitivity. The protection of the silicon transducers used in the present thesis has been investigated by M. Hornung [14] who uses thin layers of porous material in front of the membranes, which are permeable to air but not to dust or water. The materials can fulfill the required protection functions, however, one has to accept a 50% loss of ultrasound power or sensitivity [21], which is a significant number in view of the already small ultrasound signals.

Apart from power/sensitivity reduction problems due to protecting layers, the design of a miniaturized package as acoustic interface between membrane and air turns out to be a challenge, first of all due to the small dimension of the transducers compared to the ultrasound wavelength.

A naked membrane (without package), whose 1 mm-diameter is much smaller than the ultrasound wavelength $\lambda=3.4$ mm at 100 kHz, i.e. emits spherical ultrasound waves, rather than a sharp ultrasound beam with a small cross-section. As such a beam is preferable for undistorted ultrasound distance measurements over spherical waves, a physical transformation (beam-forming, focusing) of the raw ultrasound waves is necessary. The focusing not only increases the ultrasound pressure per area and therefore increases the operating range of the system, but also suppresses parasitic reflections and acoustic cross coupling from transmitter to receiver (see below). This physical transformation is provided by a dedicated sensor package design, which can be a rather difficult task, especially when the targeted dimensions of the miniaturized package are
Chapter 4. Miniaturized range finder system

in the same order of magnitude than the ultrasound wavelength where diffraction effects play an important role.

In this section, the influence of the package on the performance of the range finder system is investigated and guidelines for a dedicated sensor package are derived. One type of package is then presented and analyzed, which is promising in many aspects. Due to the significance of the package for the performance of the system, various distance measurements using different sensor packages are shown in the following section in order to verify the theoretical considerations.

There are three ways of how the package can influence the accuracy of the applied $\Delta f$ phase shift measurement (and also all other phase related measurement).

- Influence of the package on the propagation of the ultrasound waves from the transmitter to the receiver. Cross-coupling and multi-path reflections lead to a distance dependent distortion of $\Delta \varphi(X)$.

- Frequency dependent phase shift due to the acoustic transfer functions $Z_{PT}$, $G_{PT}$ and $G_{PR}$ (see Fig. 4.8), which can lead to a phase offset.

- Very small object distances to the package surface can provoke parasitic resonances which alter the acoustic impedance $Z_{PT}$. This can also lead to a frequency dependent and distance dependent phase shift.

4.8.1 Distance dependent distortion

The first of the three above effects turned out to be a serious problem for the miniaturized range finder system since it can drastically limit the overall measurement performance. The reason for the distance dependent distortion of the phase difference $\Delta \varphi$ is parasitic acoustic reflections as well as cross coupling effects between transmitter and receiver. As a result this leads to a non-linear distortion of the phase-distance relationship and therefore directly affects the measurement result.
Figure 4.29: The four main sources of distance dependent distortion. The thick black arrow indicates the desired fundamental path for the ultrasound waves to travel from transmitter to receiver.

In Fig. 4.29 the four main sources for undesired parasitic acoustic cross-coupling or reflection can be identified, if a bad package design is applied:

1. Mechanical direct coupling via package.
2. Direct coupling via air.
3. Reflection at the environment.
4. Multi-path reflection at the object (standing waves).

The non-linear distortion of the phase-distance relationship arises when a portion of the ultrasound waves, emitted by the transmitter, does not travel along the fundamental path (thick black arrow in Fig. 4.29) towards the object and back to the receiver, but rather take one of the four parasitic ways with a different path-length $D$. At the receiver, the parasitic waves are superimposed with the fundamental one and influence amplitude and phase of the received signal. For the situations (1)...(3) the following simplified model can be derived with which the described influence can be investigated. Considering a fundamental ultrasound pressure wave $p_0(t, x) = P_0 \cos(\omega t + \nu x)$ with $\nu = 2\pi f/c$ ($\nu$ is the wave number) and a parasitic damped wave $p_1(t, x) = \alpha \cdot p_0(t, x)$,
(α ≪ 1) which are superimposed at the receiver. The voltage obtained at the output of the ultrasound receiver then becomes

\[ v_R = V_R \cos(\omega t + 2\nu X + \varphi_0) + \alpha \cdot V_R \cos(\omega t + \nu D + \varphi_0) \quad (4.14) \]

where \( \varphi_0 \) is an initial electrical phase, but which can be considered to be zero without restrictions for the following discussion. The superposition of the two sinusoidal components with the same frequency \( \omega = 2\pi f \) again leads to a sinusoidal voltage but with a variation in amplitude and phase:

\[ v_R = V'_R \cos(\omega t + \varphi), \quad V'_R = V'_R(f, X), \quad \varphi = \varphi(f, X) \quad (4.15) \]

The variation of the amplitude can be calculated to

\[ V'_R = V_R \sqrt{1 + \alpha^2 + 2\alpha \cos \nu(2X - D)} \quad (4.16) \]

whereas the variation in phase, which is more interesting, becomes

\[ \varphi = \varphi_x + d\varphi = \arctan \left[ \frac{\sin(2\nu X) + \alpha \sin(\nu D)}{\cos(2\nu X) + \alpha \cos(\nu D)} \right] \quad (4.17) \]

where \( \varphi_x = 2\nu X \) is the undistorted, linear distance dependent phase and \( d\varphi \) is the deviation from the latter due to the influence of the parasitic wave. \( d\varphi \) is calculated to

\[ d\varphi = \arctan \left[ \frac{\alpha \sin \nu(D - 2X)}{1 + \alpha \cos \nu(D - 2X)} \right] \approx \alpha \sin(D - 2X) \quad (4.18) \]

and can be approximated by the right hand part of Eq. 4.18 for \( \alpha \ll 1 \). Finally, for the result of the \( \Delta f \) measurement \( \Delta \varphi = \varphi_2(f_2) - \varphi_1(f_1) \) using the above approximation we get
\[ \Delta \varphi = \Delta \varphi_x + \Delta \varphi_e \approx \Delta \varphi_x + 2\alpha \cos \left[ \frac{2\pi}{c} (D - 2X) \bar{f} \right] \sin \left[ \frac{\pi}{c} (D - 2X) \Delta f \right] \] (4.19)

which is the non distorted-phase difference \( \Delta \varphi_x = 4\pi X \Delta f / c \) plus an error phase \( \Delta \varphi_e \). \( \Delta f = f_2 - f_1 \) is the frequency difference whereas \( \bar{f} = (f_2 + f_1) / 2 \) denotes the average between the two frequencies.

In Fig. 4.30 a numerical example to the above equation is depicted for a parasitic path-length of \( D=4 \) cm and a strength of \( \alpha=15\% \) of the fundamental wave.

![Diagram](image)

**Figure 4.30:** Distance dependent distortion due to cross-coupling. The parasitic cross-coupling wave has a path-length of 4 cm and a strength of 15\% of the fundamental wave.

At \( X = D/2=2 \) cm the phase error \( \Delta \varphi_e \) has a minimum whereas at \( X = c/4/\Delta f + D/2=7 \) cm it achieves the absolute maximum \( 2\alpha \). For a 15\% cross-coupling effect, the maximum phase error \( \Delta \varphi_e(\text{max}) \) is 17.2° which translates to distance error of 1.8 mm.
Standing waves

The last source for non-linear distorted distance measurement according to Fig. 4.29 is multi-path reflection (4) or *standing waves* between the surface of the range finder sensor and the object. Standing waves are formed when the ultrasound beam from the transmitter is reflected not only once at the object, but oscillates many times between the latter and the sensor surface which is in parallel to the object surface. The chain of ultrasound waves are superimposed the same way as described in the previous paragraph and also lead to a distance dependent distortion.

Again, a simple model helps to understand, how the multi-path reflection affect the accuracy of the distance dependent phase. Assuming an object which ideally reflects the ultrasound waves and a sensor surface which returns a fraction $r$ (*reflection coefficient*) of the incoming ultrasound echo. The superposition of the resulting multiple echos between object and sensor then leads to a receiver voltage

$$v_R = V_R' \cos(\omega t + \varphi) = V_R \sum_{i=0}^{\infty} r^i \cos(\omega t + 2\nu(i + 1)X)$$

(4.20)

which again shows up with distance dependent variation of the amplitude as well as a non-linear phase. As the reflection coefficient $r$ should be much smaller than one for a low distortion due to standing waves, the above formula can be simplified by only considering the first parasitic reflection $rV_r \cos(\omega t + 4\nu X)$. All higher order terms which are scaled with $r^2, r^3, \ldots$ then are negligible. The result for the phase error can be derived directly from Eq. 4.19 by setting $D := 4X$ and $\alpha := r$ leading to

$$\Delta \varphi \approx \Delta \varphi_x + 2r \cos \left[ \frac{4\pi}{c} X f \right] \sin \left[ \frac{2\pi}{c} X \Delta f \right]$$

(4.21)

The shape of the distance dependent variation of the phase error is the same as discussed for parasitic cross-coupling (Fig. 4.30) with the maximum deviation $\Delta \varphi_c = 2r$ in the middle of the measurement range ($X = c/4/\Delta f=5$ cm).
4.8. Transducer package

As we now know how the ultrasound propagation from transmitter to receiver is responsible for either linear distance dependent phase shift but also undesirable non-linear distortion of the latter, the question arises how a dedicated package can influence the sound propagation and improve the performance of the system, respectively. According to the above consideration, the goal is to minimize the parasitic cross-coupling waves (\( \alpha \rightarrow 0 \)) and at the same time suppress the standing waves as much as possible (\( r \rightarrow 0 \)). For the four situations illustrated in Fig. 4.29 this leads to the following design guidelines for the sensor package.

- The mechanical coupling via package (1) can be suppressed by a damping spacer between transmitter and receiver, which on the one hand provides the mechanical connection of the two parts, but also acoustically isolates them.

- Direct coupling via air (2) and reflection at the environment (3) can be circumvented when a sharp and guided ultrasound beam is applied. One way of achieving this is to increase the equivalent emitting area of the transmitter using a housing with an outlet much wider than the ultrasound wavelength. The shape of the emitted ultrasound waves then no longer resembles a sphere, but rather is bundled to form a club.

  Similar package consideration can be applied for the receiver housing to obtain a certain directivity. The directivity of the receiver is preferable, since it then can be made only sensitive to incident waves from the object rather than to parasitic waves aside.

- Finally, the standing waves (4) can be significantly reduced by using an acoustic absorbing surface of the range finder as well as minimizing the overall sensor area facing the object.

Fig. 4.31 shows an improved package according to the above guidelines. The sensor surface is covered by an acoustic absorbing foam which lowers the reflection coefficient and hence reduces standing waves. The bundling of the transmitter waves and the directivity of the receiver is achieved by an exponential horn [39, 19] for both transmitter and receiver. The exponential horn is a very promising device, not only as it fulfills shaping and directivity purpose, but also due to its broadband matching properties due to the acoustic impedance \( Z_{PT} \). With
the package of Fig. 4.31 the best performance of the feasibility study could be achieved, orders of magnitude better than if a simple package as shown in Fig. 4.29 is applied (see measurement of next section).

In the following, the acoustic properties of the exponential horn as well as its impact on the performance of the system are investigated.

4.8.2 Exponential horn

An exponential horn is a cylindrical pipe, whose cross-section area \( S(x) = S_1 \cdot e^{cx} \) is an exponential function of the length \( x \), where \( c \) is the flare rate of the horn and \( S_1 \) is the initial cross-section at the throat. Fig. 4.32 shows a schematic.

Due to this particular shape, the horn acts as a loss-less acoustic impedance transformer, which linearly translates the low pressure at the surface of the membrane to a high pressure in the far field (\( x \gg a_2 \)) of the outlet (mouth). For frequencies well above a certain cut-off frequency the transformation is ideal, in particular no non-linear phase shift of the pressure waves occurs. How can this be achieved?

The area of the mouth at \( x = L \) is significantly enlarged due to the exponential function \( S(x) \), which on the one hand leads to the desired bundling of the ultrasound waves but also improves the bandwidth of the acoustic matching to the air. Other than in case of the simpler
4.8. Transducer package

**Figure 4.32:** Cross-section of an exponential horn. The characteristic properties important for the range finder system are the radius of the mouth \( a_2 \), the acoustic input impedance \( Z_{PT} \) and the pressure amplification gain (see text).

Conical horn [39] the equivalent emitting ultrasound area can be made large, in particular also for small horn length \( L \), which is preferable for the targeted miniaturized package dimensions.

The matching and ultrasound emission of the membrane plus horn can be described with the acoustic impedance \( Z_{PT} = \frac{p_1}{v_1} \) which is the relationship between the local pressure \( p_1 \) and the particle velocity \( v_1 \) at the throat of the horn. The acoustic impedance \( Z_{PT} \) translates the surface velocity of the vibrating membrane into pressure and leads to the following frequency dependent and complex function, assuming that the emission into the air is free of reflections at the mouth [39].

\[
Z_{PT} = Z_0 \left( \sqrt{1 - \left( \frac{\varepsilon c}{2\omega} \right)^2} + j \frac{\varepsilon c}{2\omega} \right) \tag{4.22}
\]

\( Z_0 \) is the characteristic acoustic impedance of air (similar to the wave impedance of an electrical cable). The real part of \( Z_{PT} \), responsible for the transmission of ultrasound power, is a high pass function and for
frequencies $\omega > \omega_{HP} = \varepsilon c/2$ ($\omega_{HP}$ = horn cutoff frequency for propagation) the impedance rapidly converges to $Z_0$. For frequencies close to $\omega_{HP}$, the generated ultrasound power is small and the phase between ultrasound pressure and membrane velocity strongly and non-linear depends on the frequency.

The pressure transfer function from the membrane surface ($p_1$) to the mouth ($p_2$) for the transmitter can be calculated by the following expression.

\[
P_{PT}(j\omega) = \frac{p_2}{p_1} = e^{-\frac{\varepsilon L}{2}} e^{-j\frac{L\omega}{c}} \sqrt{\left(\frac{\omega}{c}\right)^2 - \left(\frac{\varepsilon}{2}\right)^2} \\
\approx e^{-\frac{\varepsilon L}{2}} e^{-j\frac{L\omega}{c}}, \quad \omega \gg \omega_{HP}
\]

The above equation shows that the pressure at the mouth of the horn is lowered by a factor $e^{\frac{\varepsilon L}{2}}$. However, this reduction does not stand for a lowered pressure due to a power dissipation in the horn — the power through every cross-section $S(x)$ remains constant and is calculated to $P_{sound} = S(x) \cdot v(x) \cdot p(x) = const.$

The frequency dependent phase shift $\varphi(\omega) \approx -L \cdot \omega/c$ of $G_{PT}$ for high frequencies is the same as if the pressure is measured at the distance $L$ without horn, which means that for a phase shift measurement no additional phase shifts have to considered.

The lowering of the ultrasound pressure at the mouth of the horn is certainly not what is expected from a good package design. However, the effect of bundling the waves by an increased emitting area makes up this reduction and even increase the sound pressure, as we will see in the following paragraph.

The bundling effect of the horn can be investigated easiest by considering the planar pressure waves at the mouth as a piston diaphragm with membrane radius $a$ equal to the radius of the mouth. For vibrating piston diaphragms the sound propagation in the far field ($r \gg a$) can be determined and leads to the description of the sound pressure as a function of the distance $r$ and the angle $\theta$, which is the deviation from the piston axis (polar representation):
4.8. Transducer package

\[ p(r, \theta) = \frac{2J_1(\nu a \sin \theta)}{\nu a \sin \theta} j \frac{Z_0 \nu}{2r} a^2 v e^{j(\omega t - \nu r)}, \quad r \gg a \]

(4.24)

\[ = \Gamma(\theta) \cdot j \frac{Z_0 \nu}{2r} a^2 v e^{j(\omega t - \nu r)} \]

\[ J_1 \] is the Bessel-function of the first kind and \( v \) is the local particle velocity. Eq. 4.24 describes the emission of ultrasound as spherical waves, whose pressure amplitude is inversely proportional to the distance \( r \) and shaped by a distance and angle dependent function called directional gain \( \Gamma(\theta) \).

\[ \Gamma(\Theta) \]

\[ \Phi = 35^\circ \]

\[ \text{exponential horn} \]

\[ \text{8 mm mouth} \]

\[ f=100 \text{ kHz} \]

\[ 1 \text{ mm naked membrane} \]

\[ -90^\circ \]

\[ -60^\circ \]

\[ -30^\circ \]

\[ 0^\circ \]

\[ 30^\circ \]

\[ 60^\circ \]

\[ 90^\circ \]

Figure 4.33: Directional gain \( \Gamma(\theta) \) for the 1 mm membrane and for the exponential horn with a diameter of 8 mm at 100 kHz operating frequency.

In Fig. 4.33 this directional gain \( \Gamma(\theta) \) is plotted for the original membrane \((a=0.5 \text{ mm})\) and for an exponential horn with a mouth radius of \( a=4 \text{ mm} \). The curve for the naked membrane is almost a semicircle whereas for the increased diameter of the horn the directional gain decreases rapidly for angles \( \theta \) larger than plus/minus a few degrees.

The directional gain function and hence the pressure amplitude in the far field achieves the absolute maximum for \( \theta=0^\circ \) and it is certainly an issue of great interest to compare the maximum sound pressure with and without the exponential horn.
For high frequencies, the acoustic impedance both at the throat and the mouth of the horn becomes \( Z_0 \), leading to \( v_1 = p_1/Z_0 \) and \( v_2 = p_2/Z_0 \) (see Fig. 4.32). Due to the previously mentioned power conservation in the horn, the following must be satisfied: \( P_1 := P_2 \rightarrow S_1 v_1 p_1 = S_2 v_2 p_2 \), where \( S_1, S_2 \) are the cross-section at the throat and mouth, respectively.

Using these equations and the exponential behavior of the cross-section and the pressure leads to a local velocity at the mouth of the horn

\[
v_2 = v_1 \frac{S_1 p_1}{S_2 p_2} = v_1 \cdot e^{-\varepsilon L} \cdot e^{\varepsilon L/2} = v_1 \cdot e^{-\varepsilon L/2} = v_1 \cdot \sqrt{\frac{S_1}{S_2}} \quad (4.25)
\]

which now can be applied in Eq. 4.24 to compare the maximum sound pressure \( p_1(r, \theta) \) generated by a naked membrane (radius \( a_1 \)) and \( p_2(r, \theta) \), generated by a membrane with an exponential horn with mouth radius \( a_2 = a_1 e^{\varepsilon L/2} \):

\[
A_p(\theta) = \frac{p_2(r, \theta)}{p_1(r, \theta)} = \frac{\Gamma_2(\theta)}{\Gamma_1(\theta)} \cdot e^{\varepsilon L/2}, \quad A_p(0^\circ) = e^{\varepsilon L/2} = \sqrt{\frac{S_2}{S_1}} \quad (4.26)
\]

For \( \theta = 0^\circ \) the resulting pressure gain \( A_p \) (turn ratio) becomes the maximum \( \max(A_p) = \sqrt{S_2/S_1} = a_2/a_1 \), which can be much higher than one. For the present horn with \( a_1 = 0.5 \text{ mm} \) and \( a_2 = 4 \text{ mm} \) the gain \( \max(A_p) \) is calculated to 18 dB.

### Design tradeoffs

In order to get a large diameter of the mouth, preferable for a large pressure gain and also a large bandwidth of acoustic matching, the flare rate \( \varepsilon \) of the horn has to be large for a given length \( L \). On the other hand, a large \( \varepsilon \) increases the lower frequency corner \( \omega_{HP} \), which limits the frequency bandwidth for proper operation.

In addition, a second frequency corner can be derived from the previous assumption, that the acoustic impedance at the mouth is real \( (Z_0) \) so that no reflections occur at that particular place. It can be shown that this is only the case for frequencies well above the horn cut-off frequency for radiation \( \omega_{HR} = 2c/a_2 = 2c/a_1 \cdot e^{-\varepsilon L/2} \).
4.9. Measurements

In this work, careful consideration has been taken to the above design tradeoffs in order to get a large pressure gain and bandwidth and at the same time reasonable package dimensions \((L,a_2)\). The horn has the following parameters: \(\varepsilon = 1180 \text{ m}^{-1}\), \(L=3.5 \text{ mm}\), \(a_1=0.5 \text{ mm}\), \(a_2=4 \text{ mm}\). The resulting pressure gain is 18 dB and the two cut-off frequencies are calculated to \(\omega_{HP} = 2\pi \cdot 32.86 \text{ kHz}\) and \(\omega_{HR} = 2\pi \cdot 27.85 \text{ kHz}\) which are well below the operating frequencies 90-100 kHz. Fig. 4.33 shows the corresponding directional gain, including the characteristic beam-width \(\Phi=35^\circ\), which is twice the angle \(\theta\) for the 3 dB corner of the ultrasound power in the beam.

As a receiver (microphone), the characteristics of the horn is expected to be reversal to the transmitter. The directivity, which can be used to select reflected waves only from a particular direction, has the same shape as in case of the transmitter. Ultrasound pressure waves arriving at the large mouth of the receiver horn are amplified by the same exponential factor \(G_{PR} = e^{\varepsilon L/2}\), as they are damped in case of the transmitter (see Eq. 4.23). By using an exponential horn as package of the receiver, not only the directivity can be improved but also the sensitivity is increased by a significant factor.

One can conclude that due to the use of two identical exponential horns as package of the range finder system, the overall sensitivity can be significantly be improved. In addition the ultrasound waves, emitted by the transmitter, are bundled to form a sharp ultrasound beam with an increased pressure amplitude along the axis of the horn. The selectivity of the receiver is enhanced by the same mechanism that improves the directivity of the transmitter. Both the packaged transmitter and receiver membrane therefore reduce the disturbing cross-coupling effects (see Fig. 4.29) and are an important prerequisite for the \(\Delta f\) phase shift method and any other phase related measurements.

4.9 Measurements

The experimental range finder system of Fig. 4.22 has been implemented in order to perform the overall system measurements (static measurement of \(\Delta \varphi\) as a function of the distance) and to investigate the influence of different package types. Fig. 4.34 shows a photography of the measurement setup. The object is a 2.5 cm disc, mounted on an automatic
position unit in order to be moved precisely within the complete measurement range. The environment around the disc is covered with the same acoustic absorbing foam as used to suppress standing waves at the surface of the sensor.

![Experimental range finder setup](image)

**Figure 4.34**: Experimental range finder setup.

Early measurements have been performed by using a simple transducer package, without appropriate bundling properties and without absorbing sensor surface. The resulting phase difference at the output of the system was so distorted, that a correlation between distance and output phase could hardly be observed. The first measurements where a certain connection from phase difference output to the distance could be seen at all, was obtained by using exponential horns to suppress cross-coupling effects.

Fig. 4.35 shows the measurement result. High fluctuations of the digital phase output are observed, with an amplitude almost within the complete 360° phase range. However, one can also recognize the correct linear tendency for distances $X > 10$ mm — the solid line is the calculated phase difference for a frequency difference $\Delta f = 1550$ Hz and
4.9. Measurements

Package with exponential horns, separation 8mm

Exponential horn for T and R, d=8 mm

Figure 4.35: First experiment using a package with exponential horns.

A sound velocity $c=343$ m/s at room temperature. For small distances $X < 10$ mm the effect of a longer sound path due to the 8 mm separated transducers has to be taken into account which leads to the flattened curve around $X=0$ mm.

To improve the unacceptable distortion of the phase in the above measurement, the sensor surface as well as every part of the setup that can cause multi-path reflections have been covered by an acoustic absorbing material. The resulting measurement is shown in Fig. 4.36. The fluctuations are significantly reduced but still much too high for the targeted accurate distance measurement.

The shape of the distorted measurement in Fig. 4.36 strongly resembles the theoretical predictions by the derived model for cross-coupling between transmitter and receiver (see section 4.8.1). By assuming that the remaining distance dependent distortion is due to these cross-coupling effects, the amount $\alpha$ of parasitic acoustic pressure coupling from transmitter to receiver can be estimated. For the maximum phase error $\Delta \varphi_e = 2\alpha$ observed in the measurement of Fig. 4.36, $\alpha$ becomes 35%, which is a rather high value.

If more than one third of the ultrasound pressure waves travels directly to the receiver instead towards the object to measure, the question
arises, are the transmitter and receiver of the range finder maybe two close?

In the measurement of Fig. 4.37 the two membranes including the exponential horn are separated by \( d = 26 \text{ mm} \) instead of the minimum distance \( d = 8 \text{ mm} \). With this configuration, the best performance of the range finder prototype could be achieved. Due to the remaining influence of multi-path reflection and cross-coupling, the accuracy is limited to about 5 mm within the complete measurement range. Note, that at higher distances \( X = 80-100 \text{ mm} \) the accuracy is improved, which can be explained by a reduced effect of standing waves, lowered due to one-over-distance behavior of the pressure amplitude.

### 4.10 Conclusion

In this chapter, a feasibility study of a miniaturized range finder system was presented. The goal of the study was to derive a dedicated measurement method for the range finder, which can be applied to the miniaturized ultrasound transducers. In addition, careful consideration...
4.10. Conclusion

Figure 4.37: Best achieved measurement.

A novel continuous wave measurement called $\Delta f$ phase shift technique has been developed and implemented. In this measurement, the phase difference induced by two continuous waves of different frequencies is evaluated as absolute measure of the distance. Prerequisite to this method is the full control of the phase between the electrical signals and the ultrasound wave, which is a challenging task since the phase depends on many electrical and physical influences.

By the use of an electro-mechanical PLL as a transmitter, one important step to that phase control could be achieved: the coherence of the
electrical excitation to the ultrasound pressure at the membrane surface. The same PLL structure could be used as self-calibrating concept in the receiver. For the two frequencies required in the above measurement, the phase shift between incident ultrasound and the sensed electrical signals then is zero degrees and at the same time the sensitivity to these frequencies is maximized.

When the transmitter PLL and the self-calibrated receiver operate as part of a miniaturized range finder system, another aspect becomes a critical issue: the transducer package. In this context not only the additional phase shift due to the propagation of sound in the package has to be considered, but also - more important - the appearance of parasitic cross-coupling and multi-path reflections. By the use of an exponential horn as package for both the transmitter and the receiver membrane, a significant improvement of the overall sensor sensitivity as well as the reduction of cross-coupling effects could be obtained. The reduction of inadvertent standing waves between sensor surface and object could be reduced by covering the sensor surface with acoustic absorbing material.

The best performance could be achieved using identical horns for the two membranes, separated by a 26 mm gap and enclosed with an acoustic absorbing foam. The accuracy is distance dependent and achieves 1.5 mm to 5 mm within the full measurement range. The remaining distortion, which limits the accuracy of the system to about 5%, is basically due to residual multi-path reflections at the surface of the sensor.

Although 5 mm overall system accuracy is not good enough for precise distance measurements (e.g. liquid level detector), there exist many other applications, where the accuracy requirements are relaxed. A miniaturized range finder system using silicon membrane transducers and the concepts developed in this chapter then could be an attractive alternative to existing piezo-ceramic solutions. How the miniaturization of the electronics is carried out in detail is subject of the following chapter of the thesis.
Chapter 5

Mixed analog/digital IC implementation

5.1 Design concept: transmitter/receiver circuit

The aim of this final chapter 5 is to describe the implementation of a complete front-end IC suitable for the miniaturized range finder based on the $\Delta f$ phase shift measurement. The IC operates on a single 5 V supply and features all relevant building blocks to perform receiver and transmitter functions using silicon membrane transducers as well as to control the measurement procedure.

The targeted technology of both the silicon membrane transducers and the IC is a standard 0.8 $\mu$m process, which allows the co-integration of the silicon transducers and the electronics on the same die (monolithic micro-system). The specifications of the IC, in terms of functionality and performance of the micro-system, are derived from the feasibility study described in the previous chapter. To achieve a high level of miniaturization, the design concepts are chosen in a way to allow full integration of the complete electronics on a single VLSI chip.

Fig. 5.1 shows the outline of the IC (shaded blocks) as part of the
complete range finder system. It can be noticed that the block diagram of the system is almost identical to the one of the previously described discrete prototype (Fig. 4.22).

![Block Diagram of the Miniaturized Range Finder System](image)

**Figure 5.1:** The miniaturized range finder system. The shaded parts (transmitter or receiver electronics and frequency synthesizer) are combined in one single mixed analog/digital front-end IC.

What is different, however, is the transmitter. Other than in case of the discrete type implementation, the transmitter of the micro-system is implemented as a mixed analog/digital PLL, the same way as the self-calibrating receiver. As has already been concluded in the description of the mixed signal receiver, this is is a convenient way to circumvent the need of large time constant continuous time integrators. Transmitter and receiver of the range finder system then are identical blocks as indicated in the figure. The desired functionality (transmitter or receiver) is defined by the digital signals applied by the controller, which is basically the 12 bit bus \((DA_{T/R})\) of the D/A converter and the digital excitation \((REF_{T/R})\).

As due to this architecture the transmitter and the receiver of the system can share most of the critical building blocks, it is efficient to design the front-end electronics as a combined transmitter/receiver IC, whose individual function can be configured externally. Fig. 5.2 shows the block-diagram of the chip.
To drive and read-out the silicon membrane transducers as well as to accomplish the measurement procedure, the IC basically consists of three parts.

a) Signal amplification and quantization (receiving function).
b) Excitation of the silicon membrane transducers (transmitting function).
c) Frequency synthesizer (accurate timing reference for the system).
In receive mode, high gain, ac-coupled and low-noise amplification of the weak sensor signals (10 μV - 250 μV) is performed to achieve sufficient signal level and quality. After an amplification of 72 dB at 100 kHz the signal is fed to a self-calibrating comparator with less than 800 μV offset voltage.

As a transmitter, the chip has to excite the silicon membrane transducer. This is performed by adding a rectangular ac-signal with an amplitude of 1.5 V to a dc voltage and apply the sum via a power amplifier to the 100 Ω heating resistor. The dc voltage is derived from a monotonic 12 bit D/A converter. All reference voltages and currents are generated inside the chip, based on a bandgap voltage cell.

The frequency synthesizer is used to accurately generate the 25.6 MHz system clock and the two different measurement frequencies around 100 kHz needed for the Δf phase shift measurement. Input to the synthesizer is a crystal oscillator, which needs the only external device of the whole system, that is the Quartz crystal.

In the following sections, the three main parts of the IC are explained in more detail, especially in terms of how the specifications derived from the feasibility study are translated into fully integrated, CMOS compatible circuits. The performance of selected critical building blocks are presented as well as measurement traces of the complete range finder system based on the above IC.

5.2 The front-end circuits

5.2.1 Amplification and quantization

Fig. 5.3 shows the complete sensor amplification chain, including a self-calibrating (offset calibrating) comparator for the required (1 bit) quantization prior to the digital controller. A low offset comparator is really needed to make the phase shift between the amplified sensor signal and the resulting digital output independent of the former's amplitude and frequency.

The basic structure of the high gain, ac-coupled amplification stage is almost equal to the one used in the barrier detector application of chap-
5.2. The front-end circuits

ter 3. What is slightly different, however, is that the second ac-coupled amplification stage is included in the structure of the self-calibrating comparator. By this scheme, the intrinsic offset of both the OTA and the subsequent comparator can be suppressed and leads to the required low offset structure.

Figure 5.3: The complete amplification chain: ac-coupled low noise amplifiers (two sections) and a self-calibrating comparator.

In the calibration phase, the switches S1 and S3 are closed, S2 is open. The OTA of the second stage operates in unity gain configuration and the offset appears as input signal to the comparator. The offset of this comparator can be adjusted in a limited range by tuning the bias current in the branches of the input differential pair (see below). A 6 bit D/A converter, applied to a 6 bit counter, performs the automatic adjustment of the offset. After a system reset (signal RB=0), the systematic offset of the comparator is set to the lower bound of the D/A converter (about -25 mV), and the resulting output of the comparator is zero, independent of any random offsets. The systematic offset then is linearly increased by the counter, until the output of the comparator changes from zero to one. This transition is monitored by an edge detector circuit (signal CI), immediately stops the counter from further increasing and at the same time terminates the calibration period. The switches S1 and S3 then are open whereas S2 is closed to again enable operation of the OTA as ac-coupled second amplification stage. Due to this calibration technique, the residual offset of the whole amplification chain can be limited to one LSB (least significant bit) of the 6 bit D/A converter, which is about 800 μV.
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**Noise**

The design of the amplification stages is determined not only by the need of ac-coupled, high gain (> 72 dB) amplification of the weak sensor signals, but also by a certain amount of signal quality. At such low levels of signal strength the influence of noise, both due to the sensor itself and due to the electronic amplification can limit the overall system accuracy.

![Diagram of sensor amplification quantization phase measurement]

Fig. 5.4 depicts how the noise sources, basically due to the sensor resistance (Wheatstone bridge) and the front-end amplifier contribute to the signal-to-noise $S/N$ ratio SNR of the system at different stages. The figure also illustrates how the variation of amplitude in the signal (amplitude noise $v^2_{oa}$) translates into the corresponding fluctuation of the output phase (phase noise $\varphi_1^2$). As can be derived from the simplified model showed in the lower part of Fig. 5.4, the variance of the phase noise $\sigma_{\varphi 1} = \sqrt{\varphi_1^2}$ is direct proportional to the variance $\sigma_{oa} = \sqrt{\varphi_{oa}^2}$ of the amplitude noise at the output of the amplifier. The resulting SNR of the phase equal $\varphi_1^2/(2\pi)^2$ hereby remains the same as the SNR $v^2_{oa}/V_a^2$ prior to the comparator/phase detector.

To increase the quality of the phase signal at the output of the phase detector in order to achieve the required 8 bit Neq. (number of equiv-
5.2. The front-end circuits

alien bits), a digital 250 Hz low pass filter is added, which limits the noise bandwidth of the output (see also section 4.7.1 “Noise improved phase detector”).

According to the above considerations, the variance or rms value of the amplitude noise at the output of the amplifier should be less than 10 mV, which defines gain and noise requirements of the complete amplification stage. For a coupling capacitor $C_1=50$ pF and a feedback resistor $R=3.2$ MΩ (Fig. 5.3) used in the identical ac amplification stages, the specifications for the low-noise OTA are the following.

<p>| | |</p>
<table>
<thead>
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<tr>
<td>Gain bandwidth GBW</td>
<td>10 MHz</td>
</tr>
<tr>
<td>Thermal input noise</td>
<td>$&lt; 5$ nV/$\sqrt{Hz}$</td>
</tr>
<tr>
<td>Flicker noise corner$^1$</td>
<td>$&lt; 1$ kHz</td>
</tr>
</tbody>
</table>

The above numbers, especially thermal noise and flicker noise specifications, are not always easy to achieve, in particular when the power consumption and the area on the chip are limited. For the present design, a folded cascode [40] OTA has been chosen, due to its potential advantage in terms of low frequency 1/f noise. Fig. 5.5 shows the schematic.

Other than in a ‘normal’ current mirror based OTA the noise of the load transistors $M3, M4$ of the folded cascode OTA can be controlled in a wide range without affecting bandwidth or stability of the latter. In particular the flicker noise contribution, which is inversely proportional to the area of the transistors, can be minimized by designing large area transistors for $M3, M4$. To achieve the highest possible $Gm$ of the OTA for a moderate bias current (120 μA) the input differential pair operates in weak inversion.

Self-calibrating comparator

For the comparator in Fig. 5.3 as interface between continuous time analog sensor signals and the sampled digital output a regenerative latched comparator [41] has been chosen. The circuit features high-speed and good precision and occupies only a small area. Fig. 5.6 shows

$^1$The flicker noise corner is the frequency where low frequency 1/f noise hits the thermal noise floor.
Two non-overlapping clock phases $\phi, \bar{\phi}$ operating at the system frequency (25-30 MHz) are needed to enable the correct sampled comparator function. In the reset phase ($\phi = V_{dd}$), the current provided by
the input differential pair $M_1, M_2$ flows through transistors $M_3, M_4$ and partially through the switch transistor $M_{11}$. The voltage across the closed $M_{11}$ depends on the input differential voltage of the comparator and builds the initial state of the latched comparator for the subsequent regenerative (decision) phase $\Phi$. In this second phase, the logic output state is established based on the initial voltage described above. The decision speed (and also the sensitivity to input signals) hereby typically is very high compared to other comparator structures, which is an intrinsic property of structures with positive feedback like the present latch ($M_3, M_5, M_7, M_8$).

The desired calibration or adjustment of random comparator offset as well as offset contributions due to the OTA in advance can be performed by applying additional bias current ($I_c + \Delta I_c, I_c - \Delta I_c$) to the transistors $M_1$ and $M_2$ of the input differential pair. By this scheme, the imbalance of the branch currents due to mismatch in the transistors or an applied small voltage can be made up in order to force the overall offset to zero. For the automatic offset calibration described here the additional bias currents are provided by a 6 bit D/A converter as is illustrated in the block diagram of Fig. 5.3.

A convenient choice for the 6 bit calibration D/A converter is a resistor based $R-2R$ ladder converter [42] because the analog output signal of this converter type usually is current. Moreover, the $R-2R$ ladder principle can also be applied if MOS transistors are used to emulate the resistors, which makes the technique well suited for an implementation in pure CMOS technology [43].

Fig. 5.7 depicts such a MOS based R-2R D/A converter with a differential current output ($I_{OUT}, I^{OUT}_{OUT}$) for the offset calibration described above. The circuit consists of a MOS ladder out of 6 identical ladder slices ($M_1, M_2, M_3, M_4$) and a symmetrical cascode summing stage. Each slice generates a binary weighted current which is summed at one of the two low ohmic impedance nodes $A$ and $B$, respectively. The resulting branch currents $I_A$ and $I_B$ are the complementary output currents of the ladder and the sum $I_A + I_B$ is always equal to the input ladder current $I_{LAD}$. The low ohmic impedance node $A$ (and $B$, respectively) is provided by a regulated cascode [44] stage which consists of the cascode transistor $M_5$ with a gain transistor $M_6$ and two bias sources $I_1, I_2$. Due to this regulated cascode structure, the impedance at the source of $M_5$ becomes extremely low which is preferable for a
good linearity of the ladder output current $I_A$ [43]. $I_A$ is folded to the drain current of the cascode transistor and builds, via the current mirror $M7, M8$ the output of the D/A converter.

Figure 5.7: MOSFET-only 6 bit D/A converter to generate the offset compensation currents $I_{OUT}$ and $I_{OUT'}$.

The described current mode D/A converter generates quantized output currents between 5 µA and 20 µA. When applied to the input differential pair of the previously described comparator, an offset compensation of +/- 25 mV of the latter can be achieved. This is certainly a sufficient range to consider both the random offset of the comparator as well as the offset of the applied OTA of the second amplification stage.

5.2.2 Excitation circuit

12 bit monotonic D/A converter

The membrane excitation circuit, responsible for the electro-thermal excitation of the membrane as well as the setting of the latter’s operating point, consists of a monotonic by design 12 bit D/A converter [42] and a power amplifier driving the 100 Ω heating resistor of the membrane.
5.2. The front-end circuits

Fig. 5.8 shows the corresponding block diagram.

**12 Bit Monotonic DA Converter**

![Diagram of 12 Bit Monotonic DA Converter](image)

**Power Amplifier**

![Diagram of Power Amplifier](image)

Figure 5.8: Monotonic 12 bit D/A converter and power amplifier as interface circuit between the digital control circuits and the excitation of the membrane.

A 12 bit monotonic D/A converter is a convenient choice for this mixed signal interface circuit due to the requirements of the electro-mechanical PLL described in section 4.6. Although the quantization level of the output voltage across the heating resistor has to be high in order to achieve the required phase accuracy of the PLL, an absolute accuracy (linearity) of 12 bits of the D/A converter is not really required. For the correct and accurate operation of the control loop of the PLL it is sufficient to keep the converter output characteristic monotonic. The
design of such a 12 bit monotonic D/A converter is much easier than if a true 12 bit (full 12 bit accuracy) type has to be constructed, and is performed with a coarse-fine architecture as shown in Fig. 5.8.

The converter consists of two resistor strings, each comprising 64 unit resistors $R$. One of the two resistor strings provides the course quantization of the output voltage (string voltage) whereas the other is used for the fine division of each voltage drop provided by the course resistor string. The individual voltage levels are selected by MOS switches applied to each node of the resistor string and the corresponding gate voltages for these MOS switches are provided by two digital encoding blocks ('MSB Segment Selection', 'LSB Tab Selection'). The arrangement of the switches and an additional analog multiplexer hereby has been chosen in a way so that the digital encoding logic, which usually occupies most of the converter area, becomes simple [42]. When the switches are set correctly as a function of the input digital code, the analog output signal obtained after the analog multiplexer is 12 bit quantized and intrinsically monotonic.

Reference concept

Every D/A converter requires an accurate reference voltage $V_{ref}$ (or reference current) from which the converted analog output quantity is derived. For the present 12 bit converter as part of the excitation circuit, this reference voltage has to be considered in the context of the power output amplifier driving the low ohmic heating resistor of the membrane.

The power amplifier is a similar circuit as has already been used in the barrier detector application of chapter 3 (Fig. 3.12) - a two stage Miller compensated amplifier whose second stage is composed of an NMOS power transistor directly applied to the 100 $\Omega$ load. A high W/L-ratio (3000 $\mu$m/0.8 $\mu$m) NMOS is used as gain transistor and performs a gain of 24 dB for the second stage, despite the low ohmic 100 $\Omega$ heating resistor. The resulting output voltage $V_h$ appears across the heating resistor, between the positive power supply $V_{dd}$ and the drain of the NMOS. Consequently, the required reference voltage of the D/A converter has also to be referred to $V_{dd}$, in order to prevent the output voltage $V_h$ from being power supply dependent.

In the above excitation circuit, the converter reference voltage is
5.2. The front-end circuits

generated by applying a current source $I_{DA} = V_{BG}/40R$, proportional to an accurate on-chip voltage reference $V_{BG} \approx 1.25 \text{ V}$ (bandgap voltage), to the coarse resistor string. $I_{DA}$ creates the upper and lower bounds of the coarse string voltage $V_1, V_2$ as well as a $V_{ref}$ between $V_1$ and $V_2$ as reference potential. $V_{ref}$ is buffered by a 50 pF coupling capacitor and then applied to the positive terminal of the inverting power amplifier, which features a gain of -2. The potential for $V_1$ is chosen as high as possible, but well below the supply voltage to allow the usage of NMOS-only switches. $V_2$ is chosen as low as possible, but high enough to provide sufficient headroom to the saturation voltage of the current source $I_{DA}$. $V_{ref}$, finally, is set to an appropriate level between $V_1$ and $V_2$, so that the resulting output voltage $V_h$ exhibits a maximum range between the two supply rails $V_{dd}$ and $V_{ss}$.

All resistors are an integer number of series connected unit resistors $R$. By this scheme, the resistors are matched and the output voltage $V_h$ becomes independent of absolute resistor values, since it is always composed of a sum of unit resistors $R$, multiplied with the reference current $I_{DA} = V_{BG}/40R$. With the resistors depicted in the schematic of Fig. 5.8 the output voltage range between the lowest and the highest digital input number becomes $V_h = V_{BG}[0.35 \ldots 3.55]$, which is calculated to $[0.44 \ldots 4.44 \text{ V}]$ for a band-gap voltage of 1.25 V.

Periodic membrane excitation

The rectangular membrane excitation at 100 kHz is performed by applying an alternating current source to the artificial ground of the inverting power amplifier. The current is multiplied by the feedback resistor $(80R)$ and directly translates to the rectangular oscillating output voltage. Again, the current amplitude is set proportional to the band-gap voltage $V_{BG}$, like for the reference current of the D/A converter, which makes the amplitude of the oscillating heating voltage an accurate derivative of $V_{BG}$.

Reference circuit

The task of the reference circuit is to deliver all reference voltages and reference currents needed in the complete mixed/signal front-end IC. In particular this means the generation of the low impedance analog ground of the IC as well as to provide the reference current $I_{DA}$ needed
in the monotonic D/A converter described above.

![Reference Circuit Diagram]

**Figure 5.9:** Reference circuit providing a buffered bandgap circuit for the analog ground and the reference current for the D/A converter.

Fig. 5.9 shows the schematic of a reference block which performs the two tasks simultaneously. The generation of the analog ground voltage around 2.5 V and the reference current $I_{DA}$ is accomplished by a non-inverting, buffered amplifier, which amplifies the applied bandgap voltage by a factor of 2. The current flowing in the source follower $M1$ hereby is given by the voltage across the two feedback resistors, which is already the required quantity for $I_{DA}$, namely $V_{BG}/40R$. A current mirror ($M2, M3$) monitors the drain current of $M1$ and leads it to the output terminal for $I_{DA}$ by an auxiliary current mirror $M4, M5$. The low impedance analog ground is provided by a replica circuit ($M1', M2', M6$) which copies the buffered output voltage from the source of $M1$. This source terminal certainly can also directly be used as the analog ground node, however, the use of a replica circuit shows up with two significant advantages.

The first is that the load of the analog ground is decoupled from the feedback loop of the buffer amplifier. It then can be designed independent of the (unknown) complex load of the analog ground and always remains stable. The second advantage is that due to the decoupling any interference on the analog ground like charge packets from switching transients do not influence the reference current $I_{DA}$ for the D/A
5.2. The front-end circuits

With the present reference circuit the analog ground voltage as well as the output of the 12 bit D/A converter becomes independent of supply voltage fluctuations, which is an important prerequisite for reliable and accurate operation of the system. The analog ground as overall reference potential in the IC is derived by an on-chip band-gap cell and is powered by a decoupled, low ohmic replica circuit. If necessary, the voltage can be buffered by a large (1..10 nF) external capacitor $C_{ext}$ which lowers the ac-impedance of the analog ground.

5.2.3 Frequency synthesizer

The last of the three main parts of the front-end IC builds the frequency reference for the whole system, that is a versatile fractional-N frequency synthesizer [38, 45]. Fig. 5.10 shows the block diagram.

![Block diagram of the frequency synthesizer](image)

**Figure 5.10:** Frequency synthesizer with fractional N divider.

With a fractional-N synthesizer accurate frequencies can be generated in a predefined, digitally controlled frequency interval. In case of the present synthesizer the output frequency, which is used as clock frequency of the overall system (see also section 4.7), can be set digitally.
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in unit steps between 24.576 MHz and 26.112 MHz.

The frequency synthesizer consists of a phase locked loop (PLL), whose VCO output is fed back to the phase detector via a fractional-N divider [45]. This digital circuit divides the VCO frequency by a variable ratio $k = 16 + \text{FREQ}/2^8$ where FREQ is the 8 bit digital frequency input. When the PLL is locked to the input reference frequency at 1.536 MHz, the VCO generates a system clock signal of $k$-times the input reference, which leads to the above mentioned frequency range. The above 1.536 MHz reference frequency of the synthesizer is provided by an on-chip crystal oscillator operating at a standardized frequency of 12.288 MHz, which is divided by 8 before applied to the phase detector.

Frequency synthesizer like the above circuit are widely used in measurement equipment and telecommunication applications to provide high quality (modulation) signals with an accurate and stable frequency. Critical design issues of such synthesizers is the phase noise of the output signal, basically determined by the performance of the VCO and the loop filter (LF). In the present application, however, the requirements on frequency stability and phase noise is not a critical question, since only 8 bit of phase accuracy is targeted. This relaxed requirements, however, enables the frequency synthesizer to be fully integrated in the front-end IC since then a low performance VCO as well as an on-chip loop filter can be used.

Fig. 5.11 shows an on-chip VCO which is an all-NMOS three stage differential ring oscillator [46]. The circuit consists of three identical differential inverters which are configured in a unity gain feedback loop. Each stage features a gain of $A_o = 5$ and a cut-off frequency of $f_o = 14.43$ MHz. It can be shown, that the oscillation condition for this structure is fulfilled, when the gain of each stage exceeds a minimum of 2. The oscillating frequency is calculated to be $\sqrt{3}$ times the pole frequency $f_o$, which gives a free running frequency of the VCO of $f_{osc} = 25$ MHz.

The pole frequency $f_o$ of each stage is process dependent and basically is determined by the intrinsic cut-off frequency of the long channel load transistors $M_3$ and $M_4$, that is $f_o \approx f_t = G_{m3}/C_{gs3}$. No additional capacitors are needed. Tuning of the oscillation frequency is performed by changing the bias current through $M_3$ and $M_4$, which changes the latter’s $Gm$ and then alters the pole frequency, respectively. The gain
5.2. The front-end circuits

Three Stage Differential Ring Oscillator

Figure 5.11: VCO as a three stage differential ring oscillator.

\[ A_0 = \frac{G_{m1}}{G_{m3}} \] of the stage, however, is not altered by the tuning since both the \( G_{m1} \) of the input differential pair and the \( G_{m3} \) of the load are modified in the same direction (tracking).

An additional differential pair (\( M5, M6 \)) provides the tuning current for the oscillator, which finally enables the circuit to be used as a voltage controlled oscillator. The VCO frequency can be set between 18 MHz and 30 MHz, if a differential voltage signal of +/-200 mV is applied to the tuning input of the VCO. The frequency range of the VCO is large enough to cover the frequency span of the range finder application despite the expected typical +/-20% variation of the free running frequency \( f_{osc} \) due to process variations.

The ring oscillator generates differential voltage signals with an amplitude of about 500 mV. Before the output of the VCO can be used as a digital clock signal for the system as well as for the fractional N divider, these small amplitude signals have to be amplified to the full 0 to 5 V digital swing. A differential to single ended stage as shown in Fig. 5.11 therefore is added to the output of one of the three oscillator stages.

The complete mixed/signal front-end IC has been implemented in pure 0.8 \( \mu \text{m} \) CMOS, which is the same technology as used for the real-
ization of the silicon membranes. Fig. 5.12 shows the chip micro-graph of the implemented chip, where the building blocks described above are outlined with white boxes. The core area is $1.8 \text{ mm} \times 1.67 \text{ mm} = 3 \text{ mm}^2$, from which a significant part is occupied by the 12 bit monotonic D/A converter.

**Figure 5.12:** Chip micro-graph of the mixed/signal front-end ASIC. The core area (chip without pads) is $1.8 \text{ mm} \times 1.67 \text{ mm}$.
5.3 Measurements

In order to verify the functionality and the performance of the prototype ASIC as part of the miniaturized range finder, a complete transmitter/receiver system has been built, similar to the experimental setup of Fig. 4.34. Fig. 5.13 shows a photography of this setup, where the following two parts can be identified: The bottom PCB contains the programmable FPGA logic for the digital controller whereas the perpendicular PCB carries the transmitter and the receiver IC as well as the packaged membrane transducers (backside view). Both the silicon membrane transducers and the ASIC are fabricated in the same 0.8 μm CMOS technology.

![Image of range finder setup](image)

**Figure 5.13:** Range finder setup using the transmitter/receiver IC. Both the silicon membrane transducers and the IC are fabricated in the same 0.8 μm CMOS technology.
Transmitter and receiver are separated by a 1 mm slot in the PCB in order to minimize mechanical sound coupling via the common substrate. The complete measurement setup is powered by a single 5 V supply and contains a two-digit hexadecimal display as well as a parallel interface for the acquisition of the digital phase values.

With the present setup all of the following measurement traces have been performed. The measurements are focussed on the most important characteristics of the range finder system, in particular they complement system aspects which have not been covered by the discrete prototype described in chapter 4.

![Amplified Receiver Voltage @ f=93.515 kHz](image)

Figure 5.14: Signal strength over distance from sensor surface (horn mouth) to 150 mm distance. The object is a 8 x 8 cm plastic box.

5.3.1 System measurements

In a first experiment the amplitude of the receiver signal has been investigated as a function of the object distance $X$. The measurements have
been accomplished both for a sensor covered with acoustic absorbing foam and also - to again demonstrate the influence of multi-path reflection - without any absorbing material at the sensor surface. Fig. 5.14 shows the amplitude of the 72 dB amplified receiver signal, when the transmitter membrane is excited by a 93.515 kHz rectangular heating voltage with 3 Vpp amplitude.

The amplitude achieves an absolute maximum at $X=13$ mm for both cases and then declines to zero by a $1/X$ behavior typical for the sound propagation of piston transducers (see Eq.4.24). Periodic amplitude fluctuations can be observed which are induced by multi-path reflections between the sensor surface and the object. In case of the uncovered sensor, these amplitude variations can be more than a factor of two whereas for the covered version they are limited to less than 20% within the distance range of interest. Moreover, the amplitude level in general is much higher for the covered sensor which is preferable for a good signal-to-noise ratio of the overall measurement system.

![Figure 5.15: The most critical signals of the system. Distance $X=5$ cm.](image)

Fig. 5.15 shows a transient measurement of the most critical electrical signals involved in the range finder measurement system. The signals are captured for a fixed distance of $X=5$ cm and a reference fre-
frequency of 93.515 kHz. The heating voltage of the transmitter is locked
to the reference signal by the electro-mechanical PLL and exhibits an
amplitude of 3 Vpp. The resulting amplified receiver signal is a pure
sine-wave with an amplitude of 500 mV corresponding to a 125 µV
signal delivered by the membrane Wheatstone bridge.

In Fig. 5.16 another measurement of the above signals (transmitter
heating voltage, receiver signal) is depicted, this time for a complete
$\Delta f$ phase shift measurement cycle and at a different time scale.

![TeK Run: 500ks/s Hi Res
Ch1 Zoom: 5.0X Vert 0.01X Horz]

Figure 5.16: $\Delta f$ measurement cycle. Distance $X=5$ cm.

The measurement illustrates how the electro-mechanical PLL adapts
the average heating voltage at the transmitter membrane when the ex-
citation frequency is switched from $f_2=95.000$ kHz to $f_1=93.515$ kHz.
For the CMOS membrane applied in this measurement, the time needed
for the settling of the PLL is 20 ms. The level of the amplified receiver
voltage is about 500 mV for both $f_1$ and $f_2$ and corresponds to the max-
imum sensitivity for each of the two frequencies. It can be concluded
from the above measurement that the automatic tuning/calibration of
both transmitter and receiver membrane works well for excitation fre-
5.3. Measurements

Fig. 5.17 finally shows the digital $\Delta \Phi$ output of the system for the same excitation frequencies as used in the previous measurement.

![Digital $\Delta \Phi$ - Measurement @ X=5 cm](image)

**Figure 5.17:** *Digital $\Delta \Phi$ output of the range finder system.*

The sampling frequency for the above transient measurement is equal to the excitation frequency of the two measurement intervals, as has been discussed in section 4.6.1. 4096 phase samples are acquired for each $\Delta f$ measurement interval, which results in an interval time of 44 ms. About 20 ms after the frequency is switched from 95.000 kHz to 93.515 kHz or visa versa the digital phase values are settled to within an LSB of the 8 bit digital phase detector. For both transients ($f_1 \rightarrow f_2$ and $f_2 \rightarrow f_1$) the digital phase value $\Delta \Phi$ converge to $165^\circ$ which correspond to an object distance of $X=5.2$ cm.
5.3.2 Phase locked loop measurement

The electro-mechanical PLL as self-calibrating transmitter/receiver concept is the most important part of the complete range finder system, since it enables the generation of coherent ultrasound waves and at the same time maximum sensitivity/efficiency. It is therefore important to know how far the PLL is able to fulfill this task and what are the limiting factors, respectively.

A characteristic measurement for the PLL is the input frequency range for which the PLL is able to lock the output phase accurate and stable. Fig. 5.18 shows this relationship between applied input frequency and the resulting average voltage across the heating resistor within the stable range of the PLL. The measurement has been evaluated for both transmitter (excitation=3 Vpp) and receiver (excitation=200 mVpp) PLL.

![Tuning (Locking) Range PLL](image)

**Figure 5.18: PLL Measurement for transmitter and receiver.**

The transmitter PLL can be tuned between 93 kHz and 104 kHz whereas the receiver PLL exhibits a wider range of 85.5 kHz ... 109 kHz. The reason for the lower tuning range of the transmitter is simply the
smaller headroom for the dc part of the heating voltage within the supply voltage, since the amplitude of the excitation voltage already takes 3 Vpp. The relationship between the applied frequency and the measured dc heating voltage can be expected to be the same curve as has been measured for the resonant frequency vs. dc heating voltage shown in Fig. 2.5, since the PLL operates always at the resonant frequency of the membrane. For the receiver the resulting measurement curve is monotonic within the entire PLL operating range. However, in case of the transmitter PLL the measurement exhibits a substantial peak around 100 kHz. The reason for this peaking is a local package resonance, induced by the membrane backside package of the measurement setup (see Fig. 5.13). For the strongly excited transmitter membrane, the sound pressure at the package resonance may alter the membrane characteristics, which is then observable in the above measurement. Apart from the package resonance between 93 kHz and 98 kHz, no peaking occurs and defines a safe and reliable area for ultrasound generation - preferable for the range finder measurement.

5.3.3 Signal quality

The last measurement important for the performance characterization of the range finder prototype is the measurement of the ultrasound signal quality, basically determined by the noise in the system. As has already been discussed earlier, the primary source for noise in the system is due to the resistors of the receiver Wheatstone bridge and the limited noise performance of the high gain front-end amplifier. In order to determine the noise contribution of the amplifier and the sensor, the power spectral density (PSD) of the amplified receiver signal for an average distance of \( X = 5 \) cm has been measured (Fig. 5.19). The measured amplitude of the receiver signal at a frequency of 93.515 kHz is 470 mV. The noise PSD of the signal is band limited and leads to a total integrated voltage noise contribution of 12 mVrms. The resulting signal-to-noise ratio becomes \( S/N = 28.8 \) dB, which is good enough for the subsequent 8 bit phase detection according to Fig. 5.4.

With the PSD measurement of Fig. 5.19 also the distortion of the receiver signal can be investigated. Although the receiver output is expected to be a pure sinusoidal signal due the twofold high Q bandpass filtering of transmitter and receiver membrane, it is somewhat distorted
by the non-linear amplification of the front-end amplifier. However, it can be calculated from the above PSD measurement that the total harmonic distortion (THD) of the amplified receiver signal is less than 1.3% despite the use of non-linear feedback resistors in the ac-coupled amplification stages.

5.4 Conclusion

In this chapter the implementation of a fully integrated mixed analog/digital front-end IC for the miniaturized range finder system has been demonstrated. The system architecture and the specifications are derived from the feasibility study previously developed. The architecture hereby has been chosen in a way to allow both transmitter and receiver function, only with a small amount of area overhead.
A prototype range finder system has been built using the above front-end IC and the silicon membrane transducers, both implemented in the same 0.8 μm CMOS technology. The resulting measurements confirm that the prototype fulfills the tight specifications of the discrete feasibility study, in particular the function of the electro-mechanical PLL and the low noise amplification of the tiny membrane signals.

The complete range finder system is quite complex - in case of the discrete version more than 100 components are used for the transmitter and receiver circuits. To achieve a miniaturized range finder solution despite these complex front-end electronics, a fully integrated system as presented in this chapter is really required. An integrated system not only is preferable because of the small size, but also as it typically requires much less power than a discrete type implementation. For the present prototype micro-system more than 95% of the overall electrical power is spent for the electro-thermal excitation in the heating resistor whereas in case of the feasibility study a considerable amount of power is wasted in the discrete electronics.

What remains to be done, however, is the combination of the described IC, the digital controller and the silicon membrane transducer on the same die, which is an easy step after the successful verification of functionality and performance by the above prototype. The resulting micro-system then is either the self-calibrated transmitter or receiver of the range finder system, ready to be used for the Δf phase shift method or any other phase shift measurement technique.
Chapter 6

Conclusion

In this dissertation the realization of two silicon based ultrasound micro-
systems is discussed. Both micro-systems - an ultrasound barrier de-
tector and a miniaturized range finder - are intended for industrial au-
tomation applications. They are based on thermally excited silicon
membrane transducers for ultrasound generation and detection, fabri-
cated in a standard bipolar, BiCMOS or CMOS circuit technology. The
purpose of these micro-systems is the fabrication of a versatile and in-
expensive monolithic sensor/actuator solution by the combination of
silicon based ultrasound transducers and attached interface electronics
in the same VLSI technology.

The present work describes a systematic approach for the realization
of ultrasound micro-systems, from the idea/specifications to a working
prototype. In particular it includes the most important design steps
required for a successful implementation which are the following tasks:

- Characterization and modeling of the silicon transducer.
- Development of a dedicated measurement principle.
- Feasibility study using circuit concepts compatible to integrated
  version.
- Package consideration.
• Implementation as a fully integrated circuit.

In a first step the development of an empirical model of the silicon sensor/actuator is performed which reflects the latter’s operating principle. With this model the behavior of the sensor/actuator can be studied, in particular in the context with the subsequent measurement concepts and self-calibration techniques. To develop a dedicated model for the silicon membrane transducers, the physical mechanism for the generation and detection of ultrasound waves has been studied and the relevant model parameters have been extracted. The most important parameters among all the mostly non-linear physical characteristics are the resonant frequency, the sensitivity (gain) and the quality factor of the resonant mechanical structure. Model parameters which reflect the tuning capabilities of the membrane are the voltage versus resonant frequency relationship $K_f$, the phase gain $K_p$ and the thermal tuning bandwidth $\omega_{ML}$. Non-idealities like electrical offset or parasitic capacitances have been investigated and included in the model when necessary. The physical description of the sensor/actuator including the extracted parameters (differential equations) have been translated to a simplified model using electrical network elements. Such an empirical model, represented by pure electrical network elements, has the advantage that it can be used in the numerical verification of the complete micro-system using a network simulator like SPICE.

As already be mentioned in the above paragraph, the characterization and modeling of the sensor/actuator is an important prerequisite to the second step, that is the selection of a dedicated measurement method. In case of the barrier detector micro-system this is straightforward. An ultrasound source generates a continuous beam of ultrasound power whose presence is detected by a receiver. For the range finder micro-system, however, a unique measurement method called $\Delta f$ phase shift measurement has been developed, since the thermally excited membrane transducers are not well suited for any of the classical measurement principles like TOF etc..

A prototype system has been built using discrete components and the silicon micro-transducer. The idea of the prototype implementation is to investigate the system behavior, to verify the measurement principle and to determine the system specifications for a subsequent implementation on an IC. Consequently, the electronic circuit hereby are also
designed for the later IC development, which means that one foregoes the use of non-integrable components like inductors, large and precise capacitors or resistors. In this context also self-calibration is an important issue, since both the electronics and the silicon sensor/actuator have statistical variations of their characteristics or exhibit a temperature dependence.

An important aspect in the complete micro-system design flow is the realization of the package. There are two items which have to be considered: the influence of the package on the electronics on the one hand, and the direct impact of the package on the physical quantity (to measure) on the other. Both effects are rather critical for the performance of the micro-system and have to be considered carefully. Non-symmetrical layout of the packaged silicon membranes may invoke inadvertent feed-forward zeroes in the resonator model, which may prevent the ultrasound transmitter from the correct operation in both applications. Furthermore, the package has a strong influence on the acoustics of the system and limits the performance of the range finder system. One can conclude in general that the package of a micro-system is one of the most critical issues in the design and its potential influence has to be taken into account from the very beginning. In particular, it should be considered also during the characterization and modeling of the silicon transducer as it can be strongly linked to the physical characteristics of the sensor/actuator.

The implementation of the complete measurement system as a fully integrated solution is the final challenge in the design flow of a micro-system. Fully integrated means no external discrete components or only a minimum amount of non-integrable devices like a quartz crystal, for the benefit of reduced space, fabrication cost and also for increased reliability. To achieve a fully integrated front-end IC, a mixed analog/digital system architecture has been chosen in both applications. High gain, ac-coupled and low noise amplification is provided to boost the weak sensor signals. The ac-coupling hereby is performed by a combination of moderate size on-chip capacitors and a novel artificial resistor using MOS transistors in the triode region. The electro-thermal excitation of the membrane is accomplished by an efficient two stage power amplifier whose second amplification stage includes the low ohmic heating resistor as the load. Conditioning electronics (see smart sensor/actuator block diagram in Fig. 1.2) is included providing a self-calibrating feedback
loop from the amplified sensor signal to the thermal excitation. In case of the barrier detector micro-system this consists of the simple EXOR based tune and track circuit whereas for the range finder system it is the electro-mechanical PLL. On-chip reference circuits like the analog ground generation are also included to make the system a single supply solution.

In case of the two ultrasound micro-systems of this thesis, another feature is interesting in the context of the system architecture. For both systems, the architecture can be designed efficiently as a combined transmitter/ receiver solution, since transmitter and receiver share most of the electronic building blocks. Such a scheme simplifies the design of the transmitter/ receiver system in general and also doubles the count of mass-fabricated dies, preferable for a low cost system.

Both ultrasound micro-systems can fulfill the desired tasks for their intended industrial applications. In case of the barrier detector application a self-calibrated transmitter/ receiver system has been built which is as flexible as the commonly used light barrier but offers the advantage of ultrasound as carrier: It is robust against a dusty or smoky environment and it can also be applied for objects which normally are transparent to light like bottles or drops of solution. An object can be detected within less than 10 ms in a distance range up to 10 cm.

The range finder micro-system performs a contact-less distance measurement in a range of 1 cm to 10 cm with a precision of 5 mm and a measurement speed of 10...20 ms. Limitation to the accuracy of the system are acoustic cross-coupling or multi-path reflection effects. However, a well designed package can be used to concentrate the power of the ultrasound waves and focus the latter to form a narrow beam, which is preferable for low cross-coupling from transmitter to receiver.

Although much effort has been invested in the optimization of both dedicated sensor electronics and package design, it can be concluded from the range finder prototype that the silicon membrane transducers are not best suited for fast and precise distance measurements.

However, the use of these devices is not limited to range finders and barrier detector micro-systems like the one described here. As already be mentioned in the introduction (chapter 1), one can think also of other ideas for ultrasound applications, where the silicon membranes can be used as effective replacement of the existing expensive (piezo-ceramic)
ultrasound transducers.

\[ \Delta \phi = 4 \pi f \frac{\Delta X}{c} \]

\[ \Delta \phi = -2 \pi f \frac{\Delta \theta k \theta X_0}{c^2} \]

\[ \Delta \phi = 4 \pi f \frac{v X_0 \cos \theta}{c^2} \]

**Figure 6.1:** Alternative applications for the miniaturized ultrasound transducers. a) Paper stack control. b) Instantaneous air (gas) temperature measurement. c) Flow meter d) Position control.

In Fig. 6.1 four examples are presented which benefit from the potential advantages of the silicon membrane transducers and where disturbing effects like standing waves or cross-coupling play an inferior role. The first three examples are applications where the phase is evaluated at one fixed frequency as a function of a small distance difference (a), the temperature of a gas (b) or the gas flow velocity (c). Application (d) makes use of the small dimensions of the silicon membranes as position control of textile material (*fork barrier*), where the commonly used light barrier is less qualified due to its sensitivity to dust, smoky environment or humidity.

The range finder of example a) can be used for the paper stack
control in industrial or desktop printing applications. It monitors the simple but important task of whether one or two sheets of paper are fed into the printing machine. As the miniaturized range finder of this thesis, operating at one fixed frequency, can measure the distance accurately within one half wavelength ($\lambda/2=1.7$ mm @ $f=100$ kHz), it is well suited to distinguish whether one or two 80 $\mu$m thick papers are stacked on each of the other.

In the second example the instantaneous temperature of a gas is measured by evaluating the temperature dependence of the sound velocity $c = c_0 + k_0 \Delta \vartheta$ ($c_0$ is the sound velocity at room temperature, $k_0$ is the first order temperature coefficient of the latter and $\Delta \vartheta$ is a temperature step). Again, the phase is an ideal parameter to evaluate this dependence, since it is pretty sensitive to temperature fluctuations. According to the formula given in Fig. 6.1b the sensitivity at a fixed distance $X_0=5$ cm becomes $\Delta \varphi/\Delta \vartheta = 9^\circ/\degree{}C$. Measurement samples can be obtained in the millisecond to ten-millisecond range, which is orders of magnitude faster than existing temperature meters for air or gas.

The third application in Fig. 6.1 is a gas flow meter based on an measurement setup found in [9]. It uses a pair of combined ultrasound transmitter/receiver which are arranged obliquely across the gas stream, flowing with an average velocity $u$. The phase difference induced by the different effective sound velocities upstream and downstream can be evaluated and lead to the equation given in Fig. 6.1c. The phase difference $\Delta \varphi$ can be uniquely be determined within one full $2\pi$ interval, which limits the range of operation: For a typical arrangement ($f=100$ kHz, $X_0=5$ cm, $\theta=45^\circ$, $u=343$ m/s), the flow velocity measurement range then is $v=0...16.6$ m/s.

Finally, the example shown in Fig. 6.1d is worth mentioning, since it is a very simple, but effective application of the silicon membrane transducers. It is an ultrasound fork barrier for the correct positioning of textile material. The position accuracy is expected to be in the same order as the dimension of the membrane, that is 1 mm. The ultrasound principle makes it well suited for dusty or smoky environment and the elevated temperature of the membrane surface is superior for insensitivity to humidity. Other than in case of the barrier detector described in chapter 3, a wired connection between transmitter and receiver is possible in this case, which relaxes the requirements to a self-calibrating receiver. The system specifications then are quite similar to the ones of
the range finder and, in fact, the following conclusion can be drawn:

For all four applications described above the same mixed signal front-end IC as described in chapter 5 can be used, only with minor changes on the system level. Self-calibration procedures and coherent ultrasound generation remains the same for all application shown in Fig. 6.1, what is different, however, is the evaluation of the digital phase values or the amplitude of the received ultrasound signal (example d). The ultrasound micro-system as a monolithic combination of a silicon transducer and attached front-end electronics is then a versatile measurement subsystem, attractive for mass-fabrication and with the potential advantages against existing ultrasound solutions.

The demand for miniaturized technical components and systems today is irresistible. In particular this trend is also present in control applications, where more and more sensors and actuators are used to perform advanced surveillance tasks. Typical examples can be found in automotive applications, in industrial automation, in bio-medical and air/space applications. The miniaturization of sensors and actuators, however, is often accompanied by a significant reduction of signal quality, both in terms of signal level and also linearity. The tendency especially is observed in micro-systems using standard IC technologies like CMOS, BiCMOS or bipolar processes, since these technologies are not optimized for a particular physical quantity to measure. On the other hand, advances in the same technologies in terms of miniaturization and mass-fabrication enable the efficient integration of sophisticated read-out electronics as presented in this work to compensate for the typical non-idealities mentioned above. Micro-systems based on such an approach thus can be an attractive alternative to many of today's existing bulky and expensive sensor/actuator solution.

The design of micro-systems will more and more be supported by modern computer aided design and engineering (CAD/CAE) tools, in particular also for a combined development of micro-sensor/actuator and integrated electronics. This helps a micro-system engineer, who has to dispose of excellent multi-disciplinary design skills, to simplify the design flow of a micro-system project, especially the modeling of a sensor and to develop a dedicated measurement concept.

The growing demand for miniaturized sensor/actuator solutions then can be satisfied faster and the products will exhibit advanced complexity, similar to the general tendency in today's micro-electronics.
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