Piezoelectric and ferroelectric
device technologies for
microwave oscillators

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Cover:
Detail of oscillator based on ferroelectric varactors.

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Abstract

The purpose of this thesis is to investigate piezoelectric and ferroelectric thin film device technologies for application in microwave oscillators.

Thin film varactors based on ferroelectric materials are considered. Experimental development of practical varactors based on paraelectric phase Ba$_x$Sr$_{1-x}$TiO$_3$, in terms of layout design and model extraction, is presented in the thesis. Experimental results of voltage-controlled oscillators based on ferroelectric varactors operating at 16 GHz and 19 GHz are also presented. The ferroelectric device technology is furthermore compared to traditional varactor technologies, and discussed from the perspective of oscillator applications.

Thin film bulk acoustic resonators based on piezoelectric materials and biased electrostrictive materials are considered. Specifically, fixed-frequency resonators based on AlN and tunable resonators based on paraelectric phase Ba$_x$Sr$_{1-x}$TiO$_3$ are investigated in the thesis. An integration concept is developed for AlN resonators, and experimentally demonstrated by 2 GHz oscillators. Additionally, modelling and measurement techniques for resonators based on AlN and Ba$_x$Sr$_{1-x}$TiO$_3$ are developed. The investigated technologies are compared to traditional planar resonator technologies.

**Keywords:** Thin film devices, ferroelectric varactors, thin film bulk acoustic resonators, microwave oscillators, piezoelectrics, ferroelectrics, electrostrictive materials
List of publications

Appended papers

The thesis is based on the following papers.


**Other papers**

The papers listed below in chronological order are overlapped by the appended papers or not relevant to the work, and are thus not appended to the thesis.


Notations and abbreviations

Notations

c elastic coefficient
C capacitance
$C_{CW}$ Curie-Weiss constant
$C_m$ motional capacitance
$C_{\text{max}}$ maximum (zero-bias) capacitance
$C_v$ varactor capacitance
$\tilde{C}_v$ varactor capacitance per unit area
E electric field
$f_{\text{off}}, \omega_{\text{off}}$ offset (angular) frequency
h piezoelectric coefficient
$k_t^2, k_{t,\text{eff}}^2$ (effective) piezoelectric coupling coefficient
$\mathcal{L}$ phase-noise relative to carrier
$L_m$ motional inductance
m second-order electrostriction coefficient
P electric polarisation or power
q electrostrictive coefficient
$Q_m$ acoustic (mechanical) quality factor
$Q_p$ parallel resonance quality factor
$Q_s$ series resonance quality factor
$Q_v$ varactor quality factor
$R_m$ motional resistance
<table>
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<tr>
<th>Symbol</th>
<th>Description</th>
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<tr>
<td>$S$</td>
<td>strain</td>
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<tr>
<td>$T_C$</td>
<td>capacitance tunability</td>
</tr>
<tr>
<td>$T_{CW}$</td>
<td>Curie-Weiss temperature</td>
</tr>
<tr>
<td>$T_f$</td>
<td>frequency tunability</td>
</tr>
<tr>
<td>$V$</td>
<td>voltage</td>
</tr>
<tr>
<td>$V_{1/2}$</td>
<td>half-capacitance voltage</td>
</tr>
<tr>
<td>$\beta$</td>
<td>impermeability</td>
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<tr>
<td>$\epsilon$</td>
<td>permittivity</td>
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<tr>
<td>$\epsilon_b$</td>
<td>background permittivity</td>
</tr>
<tr>
<td>$\epsilon_r$</td>
<td>relative permittivity</td>
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<tr>
<td>$\phi$</td>
<td>acoustic phase</td>
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<tr>
<td>$\Phi$</td>
<td>charge</td>
</tr>
<tr>
<td>$\chi$</td>
<td>susceptibility</td>
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**Abbreviations**

- **BAW** bulk acoustic wave
- **BJT** bipolar junction transistor
- **BSTO** $\text{Ba}_x\text{Sr}_{1-x}\text{TiO}_3$
- **BTO** $\text{BaTiO}_3$
- **CQF** commutation quality factor
- **EM** electromagnetic
- **FET** field effect transistor
- **FOM** figure-of-merit
- **HBT** heterojunction bipolar transistor
- **HRS** high-resistivity silicon
- **IC** integrated circuit
- **IF** intermediate frequency
- **LO** local oscillator
- **MBVD** modified Butterworth-Van Dyke model
- **MEMS** microelectromechanical system
- **PLL** phase-locked loop
- **MOS** metal-oxide-semiconductor
- **PZT** $\text{PbZr}_x\text{Ti}_{1-x}\text{O}_3$
- **RF** radio frequency
<table>
<thead>
<tr>
<th>Acronym</th>
<th>Full Form</th>
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<tr>
<td>SAW</td>
<td>surface acoustic wave</td>
</tr>
<tr>
<td>SEM</td>
<td>scanning electron microscopy</td>
</tr>
<tr>
<td>STO</td>
<td>SrTiO$_3$</td>
</tr>
<tr>
<td>TCF</td>
<td>temperature coefficient of frequency</td>
</tr>
<tr>
<td>TFBAR</td>
<td>thin film bulk acoustic resonator</td>
</tr>
<tr>
<td>VCO</td>
<td>voltage-controlled oscillator</td>
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Chapter 1

Introduction

This chapter serves as a brief introduction to the work presented in this thesis, by providing motivation and presenting the investigated solutions. It also describes the organisation of the thesis.

1.1 Motivation

Devices used in today’s microwave systems for communication and sensing are subject to continuously increasing requirements on performance and cost. Traditional device technologies are pushed toward the limits of their capabilities, and there is a need for better alternatives. This process creates opportunities for new, emerging technologies which are now being studied and scrutinised by the research community. New materials and fabrication techniques are explored in pursuit of better performance and improved functionalities.

This work is an attempt to contribute to this development by evaluating some emerging technologies for use in microwave systems. Although these technologies rely on material properties that are well-known, they have not yet been extensively used in microwave circuits. The motivation of this work is that these technologies must be scientifically investigated, scrutinised, and developed before they, potentially, can be of commercial interest. If these devices indeed are successfully commercialised, they will support better and cheaper microwave systems in the future.

1.2 Investigated solutions

This work investigates thin film devices based on ferroelectric and piezoelectric materials, and their application in compact microwave oscillators. It follows two parallel tracks as illustrated by Fig. 1.1. In the first track, ferroelectric varactors based on Ba$_x$Sr$_{1-x}$TiO$_3$ are investigated and developed. These devices represent an alternative to traditionally
used semiconductor varactors of various kinds. Microwave tunable oscillators based on ferroelectric varactors are designed to investigate and demonstrate the device technology experimentally.

The second track considers electroacoustic thin film resonators based on piezoelectric materials. Nontunable resonators based on piezoelectric AlN, and tunable resonators based on bias induced piezoelectricity in Ba$_x$Sr$_{1-x}$TiO$_3$ are investigated and developed. Also, low-noise reference oscillators based on AlN resonators are designed to demonstrate integration techniques developed in the work.

### 1.3 Thesis organisation

The scientific core of this work is presented in chapters 2, 3, and 4. Chapters 2 and 3 present ferroelectric varactors and thin-film resonators, respectively. Underlying physical properties, device design (paper [A]), and modelling (paper [B] and [C]) are discussed, and the investigated device technologies are compared with traditional technologies. Microwave oscillators based on the previously introduced device technologies are discussed in chapter 4. Demonstrator circuits (paper [D], [E] are presented, and [F]), and the capability and potential of the devices for this type of application is discussed. The thesis is then summarised in chapter 5.
Chapter 2

Ferroelectric varactors

This chapter presents ferroelectric varactors – tunable capacitors based on ferroelectric materials. After introducing the basic properties of ferroelectric materials, these devices are described and discussed in terms of performance characteristics and design optimisation, and compared with competing device technologies.

2.1 Ferroelectric materials

Ferroelectric materials in the ferroelectric phase belong to the family of pyroelectrics, meaning that they demonstrate spontaneous polarisation under changing temperature. Furthermore, for ferroelectric materials this spontaneous polarisation can by definition be switched by an external electric field. The relationship between electric field and polarisation is, in other words, nonlinear and hysteretic as illustrated by Fig. 2.1a. The spontaneous, or zero-bias, polarisation typically vanishes with increasing temperature, as the ferroelectric material undergoes a phase-transition and enters the so-called paraelectric phase. A ferroelectric material in the paraelectric phase is still highly nonlinear, as schematically shown in Fig. 2.1b, i.e. the permittivity of the material is strongly field-dependent. This in contrast to common dielectric materials, Fig. 2.1c, which can be considered linear also for strong external fields. Moreover, the permittivity of a ferroelectric material is generally very high, typically $\epsilon_r = 100$–1000, and temperature-dependent.

The ferroelectric devices presented in this work are all based on Ba$_x$Sr$_{1-x}$TiO$_3$ (BSTO) which belongs to a common class of ferroelectric materials known as the Perovskites, Fig. 2.2. BSTO is a solid solution of BaTiO$_3$ (BTO) and SrTiO$_3$ (STO), the former is paraelectric above 390 K while the latter is paraelectric at least down to about 10 K (an *incipient* ferroelectric). By controlling the Ba/Sr composition ratio, it is possible to tailor the ferroelectric-to-paraelectric phase-transition temperature of BSTO. For instance, Ba$_{0.5}$Sr$_{0.5}$TiO$_3$ is paraelectric at
about 250 K and above. For tuning applications, paraelectric phase materials are preferred in order to avoid the hysteresis effect that is associated with the ferroelectric phase, and the transition temperature is hence normally chosen to be below the operating temperature (room temperature, commonly). On the other hand, by operating close to the transition point, the parameters of permittivity, tunability and loss all increase. This may be beneficial for some applications where high tunability is more important than low loss. The Ba/Sr composition ratio thereby represents some flexibility when designing devices based on BSTO.

### 2.2 Thin film varactors

One way of making tunable capacitors, or varactors\(^1\), is to combine a ferroelectric material with a pair of electrodes. A bias voltage controls the permittivity of the device which from a small-signal perspective appears as a tunable capacitor. There are two basic configurations of pla-

\(^1\)Varactor is short for variable reactor.
thin film capacitors; parallel-plate, Fig. 2.3a–e, and coplanar-plate capacitors, Fig. 2.3f–g. Coplanar-plate capacitors are simple to fabricate but require higher tuning voltages than parallel-plate capacitors. This thesis is focused on the parallel-plate design.

The capacitance $C_v$ of a typical parallel-plate varactor (in the paraelectric phase) measured versus bias voltage is shown in Fig. 2.4. The measured varactor is of circular layout, Fig. 2.3e, and is characterised in a cryo-chamber by an LCR-meter [1]. The capacitance tuning curve is practically symmetric with respect to bias voltage. The capacitive tunability is defined as

$$T_C = \frac{C_{v,max} - C_{v,min}}{C_{v,max}}$$

(2.1)

where $C_{v,max/min}$ is the max/min capacitance over a certain tuning range. For paraelectric phase BSTO, $T_C$ is typically 20–80% (depending on composition, temperature, and film quality), for bias fields of about 30–60 V/μm. The thickness of the ferroelectric film in a parallel-plate capacitor is usually about 0.1–1 μm; meaning that the necessary bias voltages range from a few volts up to tens of volts. By increasing the film thickness, linearity is improved at the cost of a higher required bias voltage. Fig. 2.4 also clearly illustrates the temperature dependence of a ferroelectric material. The zero-bias permittivity goes up as the temperature is decreased toward the phase-transition temperature which for this particular Ba/Sr composition should be around 120 K. The tunability is increasing with permittivity, meaning that the capacitive tuning range available at e.g. 300 K is also available at 200 K, albeit at a somewhat higher tuning voltage.

The tuning speed of ferroelectric films is less than 10 ns [2], and the capacitance is practically frequency independent (in the microwave regime). Fig. 2.5 shows the capacitance extracted from the reflection coefficient of a device measured using a network analyser (from paper [A]). The device has a design similar the layout shown in Fig. 2.3b.

The capacitance of a varactor in the paraelectric phase can be modelled [3] as

$$C_v(V) = \frac{C_{max}}{2 \cosh \left[ \frac{2}{3} \sinh^{-1} \left( \frac{2V}{V_{1/2}} \right) \right] - 1}$$

(2.2)

where $C_{max} = C_v(0)$, and $V_{1/2}$ is the half-capacitance voltage; $C_v(V_{1/2}) = C_{max}/2$. Despite having only two parameters, (2.2) typically fits measured capacitance of parallel-plate varactors well, although for small parallel-plate varactors or coplanar varactors the model may need some correction for fringing-field nontunable capacitance. The model is readily scaled with device area $A$ and thickness $t$ as $C_{max} \propto A/t$ and $V_{1/2} \propto t$, and it is conveniently used in circuit design work. For large-signal simulations the varactor charge $\Phi$ can be derived by integration.
Fig. 2.3: Design of thin film capacitors, (a) cross-section and (b–e) layouts of parallel-plate capacitors, (f) cross-section and (g) layout of coplanar-plate capacitor.
of (2.2), as

$$\Phi(V) = \frac{3}{2} C_{\text{max}} V_{1/2} \sinh \left[ \frac{1}{3} \sinh^{-1} \left( \frac{2V}{V_{1/2}} \right) \right].$$  \hfill (2.3)

In the paraelectric phase the temperature dependence of the zero-bias permittivity can be modelled by the Curie-Weiss law as

$$\epsilon = \frac{C_{\text{CW}}}{T - T_{\text{CW}}} + \epsilon_L$$  \hfill (2.4)

where $C_{\text{CW}}$ and $T_{\text{CW}}$ are the Curie-Weiss constant and temperature, respectively. The latter is always smaller than the phase-transition temperature. Moreover, $\epsilon_L$ is the high-temperature permittivity of the order of 50 for typical Perovskites [4]. The minimum permittivity approached at high bias voltages is temperature independent, as seen in Fig. 2.4.
2.3 Quality factors

Apart from the capacitance tuning characteristic, the device quality factor is the most important parameter for varactors used in oscillators. It is defined as

$$Q_v = \frac{-X_v}{R_v}$$

where $R_v$ and $X_v$ are the real and imaginary parts, respectively, of the device impedance $Z_v$. When discussing device impedance in this work, it is generally assumed to include not only the overlap area of the varactor but also any short interconnects necessary to integrate the device in a practical circuit. The device quality factor should as such represent something that is readily available for a circuit designer. Moreover, it is important to note that $Q_v$ is not a resonator quality factor. The varactor is not a resonator, but a lossy reactor. However, if the varactor is combined with an inductive element of device quality factor $Q_L$ to form a resonator, the total resonator quality factor is

$$Q_{resonator}^{-1} = Q_v^{-1} + Q_L^{-1}.$$ 

For integrated technologies at microwave frequencies, $Q_v$ is usually lower or comparable to $Q_L$, meaning that the performance of the resonator largely depends on the performance of the varactor. When used in an oscillator, having a high varactor quality factor is therefore crucial in reaching a low phase-noise, as elaborated in chapter 4.

While the capacitance of a ferroelectric varactor is fairly easy to measure, it is more difficult to measure the quality factor accurately. To determine the quality factor of a varactor over a range of microwave frequency, the device is usually measured using a network analyser. This can be done either in a direct reflection oneport structure, Fig. 2.6a, or in a resonant transmission structure, Fig. 2.6b. The first method of direct reflection (impedance) measurement is straightforward and provides the device quality factor over an arbitrary continuous frequency range. However, such measurements of devices of high quality factors may suffer significantly from measurement errors [5], and should hence always be verified against some other type of measurement. The second type of measurement based on transmission structures, also known as Deloach structures [6], provides a more reliable method of determining the varactor quality factor. The transmission line is loaded by a resonator consisting of the varactor combined with an inductive element of a known device quality factor (which can be determined from measurements of long through lines). The resonator quality factor can be accurately determined from the 3-dB transmission bandwidth of the structure, and the varactor quality factor can then be calculated given the inductor quality factor. The disadvantage with this method is that it only provides the varactor quality factor at one discrete frequency.
point, specifically at the resonance frequency of the structure. Direct reflection measurements of the quality factor for a typical varactor are presented in Fig. 2.7. In paper [A] these measurements show reasonable agreement with resonant measurements, at least below 20 GHz.

Generally the quality factor $Q_v$ is expected to increase with increasing bias voltage, since $C_v$ is decreasing and $R_v$ is fairly constant with bias, as explained below. However, as seen in Fig. 2.7a the quality factor at 12 V bias is significantly degraded at some periodically spaced frequency points. In between, it tends to go up as could be expected. This is further illustrated in Fig. 2.7b showing that the quality factor goes down with increasing bias at 10 GHz in the vicinity of a resonance point, while it goes up at 20 GHz in between resonance points. This phenomenon of spurious resonance points is caused by the strong electrostrictive effect of the material; piezoelectricity is induced in the material by the applied voltage, and the varactor behaves as an electroacoustic resonator [7, 8]. While this mechanism can be used for resonator applications (see section 3.6), it is highly undesired for varactors. Through this bias induced piezoelectricity, energy is coupled and lost to the acoustic domain. The effect degrades the quality factor more or less for all frequencies, but at the periodically spaced resonance points it is further enhanced as acoustic standing waves appear in the device. The phenomenon is typically seen in devices with rather thin films with thin electrodes, and one way of suppressing at least the resonant behaviour is to use thick electrodes which provide acoustic loading [9]. The resonances may also be eliminated over a chosen frequency range by proper
Fig. 2.7: Measured varactor quality factor (a) versus frequency, and (b) versus bias ($\text{Ba}_{0.25}\text{Sr}_{0.75}\text{TiO}_3$, 300 nm, paper [A]).

Aside from solving the problem of electroacoustic behaviour appearing at high bias voltages, it is necessary to understand where other significant loss contributions are located in order to maximise the device quality factor. For parallel-plate varactors prominent loss contributions are ohmic loss of electrodes and interconnect strips, and film loss (due to various mechanisms in the actual film). To analyse this, let's consider the simple varactor model in Fig. 2.8, where ohmic loss is represented by $R_s$, and film loss is represented by $G_p = \omega C_v \tan \delta$. The total varactor
impedance is
\[ Z_v = R_s + \frac{1}{G_p + j\omega C_v}, \]  
(2.7)
and the real and imaginary parts of (2.7) are
\[ R_v = R_s + \frac{1}{\omega C_v} \frac{\tan \delta}{1 + \tan^2 \delta} \]  
\[ \simeq R_s + \frac{1}{\omega C_v} \tan \delta \]  
(2.8)
and
\[ X_v = -\frac{1}{\omega C_v} \frac{1}{1 + \tan^2 \delta} \]  
\[ \simeq -\frac{1}{\omega C_v} \]  
(2.9)
where the approximations hold if \( \tan^2 \delta \ll 1 \). The quality factor is consequently
\[ Q_v = \frac{1}{\omega R_s C_v (1 + \tan^2 \delta) + \tan \delta} \]  
\[ \simeq \frac{1}{\omega R_s C_v + \tan \delta}. \]  
(2.10)
It is generally difficult to distinguish between ohmic loss and film loss given the impedance (vs. frequency) of a single device. The loss tangent is frequency dependent, and among several loss mechanisms included in \( \tan \delta \), loss due to charged defects is \( \tan \delta_{ch} \propto \omega C_v \) at high frequencies [12, 13]. The contribution to \( R_v \) from that particular loss mechanism is constant with frequency and bias voltage making it inseparable from \( R_s \). At high frequencies \( R_s \) will have a frequency dependence due to the skin effect, but it is still difficult to accurately separate ohmic loss and film loss from the measured \( R_v \) of a single device. However, studies using more elaborated measurement techniques (e.g. measurements of multiple devices of varying geometry) seem to conclude that the loss tangent of BSTO thin films is quite low, and should translate to \( Q_v \) of several hundreds at tens of gigahertz if the ohmic loss could be eliminated. For
instance, in [14] a special test structure is used to determine the loss tangent of Ba$_{0.25}$Sr$_{0.75}$TiO$_3$ to about 0.005 at 10 GHz, i.e. the quality factor should be about 200 without ohmic loss. Practical varactors based on films of similar quality tend to have $Q_v < 50$ at 10 GHz, indicating that ohmic loss represents a significant loss contribution. Fig. 2.9 shows the product of quality factor and frequency for a particular sample. The product is essentially a cut-off frequency serving as a FOM, and it provides a good illustration of the frequency dependence in $Q_v$. For this sample, the product is fairly constant from about 10 GHz to 20 GHz, indicating that ohmic loss (and possibly some small contribution from tan $\delta_{ch}$) is dominating the quality factor over this frequency range. At lower frequencies, $\omega R_s C_v < \tan \delta$ and the product is increasing with frequency. At very high frequencies, the product tends to decrease as higher order frequency components in tan $\delta$ grow large. Additionally, the skin effect comes into play. Knowing the role of various loss contributions is important when developing these components. For practical ferroelectric varactors operated at a few gigahertz or higher, it is crucial to have good electrodes (high conductivity, sufficient thickness), and proper device layout (see section 2.4) in order to keep the ohmic loss to a minimum.

Tunability can be traded for quality factor by combining varactors of moderate quality factor with nontunable capacitors of higher quality factor. It is however not possible to do the opposite, i.e. to increase the tunability beyond what is offered by the varactors. Thus, for a circuit designer it may be more convenient to have a device with a high tunability and a somewhat moderate quality factor, rather than the other way around. When using BSTO, the film loss can be traded for tunability by controlling the the Ba/Sr composition ratio (see section 2.1), and it may consequently be advantageous to optimise the Ba/Sr composition for a large tunability rather than for a low film loss. Furthermore, since
the quality factor of a practical varactor is largely determined by ohmic
loss, an increase in film loss will merely have a limited effect on the
quality factor.

Since tunability and quality factor can be traded against one an-
other, by various means, it is reasonable to consider a varactor figure-
of-merit which includes both tunability and quality factor. For a device
tuned from 0 to \( V_{\text{max}} \), the commutation quality factor (CQF) [15] is
defined as

\[
\text{CQF} = \frac{(n - 1)^2}{n} \cdot Q_v(0) \cdot Q_v(V_{\text{max}})
\]  

(2.11)

where \( n = C(0)/C(V_{\text{max}}) \). In [16] it is reported that the highest CQF
for BSTO film (not a practical varactor) is given by a Ba/Sr composition
of 30/70.

### 2.4 Layout optimisation

A parallel-plate capacitor of simple layout as shown in Fig. 2.3b is vul-
nerable to any small fabrication misalignment of the top and bottom
electrodes. At high frequencies typical capacitance values are small,
implying small varactor dimensions particularly since the device per-
mitivity is very high (\( \epsilon_r = 100–1000 \), typically). The aligning precision
of a common mask aligner used in processing of these devices may be
about 0.1–1 \( \mu \text{m} \), which for a simple capacitor with an overlap area of
say \( 5 \times 5 \ \mu \text{m}^2 \) can translate to an area variation of 2–20% (or more, if
there is a misalignment in both dimensions). One solution to this issue
is to define the overlap area by a window in an intermediate oxide layer
[9, 13]. This kind of intermediate layer would also serve as protection of
the ferroelectric layer during subsequent processing steps. However, this
solution is not suitable for high-permittivity varactors due to their small
size. It is difficult to make the intermediate layer thick for small window
dimensions, and the parasitic capacitance associated with the interme-
diate layer overlap may easily account for 10–20% of the total capac-
itance, thereby reducing the overall tunability substantially. Another
way of solving the misalignment problem without using an intermediate
layer is to arrange the electrodes such that misalignment errors do not
change the overlap area. The layouts in Fig. 2.3c–d are examples of this
approach. In paper [A], varactors of layout similar to the ones shown
in Fig. 2.3b–d are analysed and characterised by direct measurements
and resonant measurements. Variation in shape factor (width/length
ratio of overlap) is studied in order to maximise the device quality fac-
tor of each layout type. It is demonstrated that the simple type of
layout, Fig. 2.3b, really do suffer from small misalignments in that the
capacitance is not constant with varying shape factor (the overlap area
does not remain constant). As expected, varactors of layouts shown in
Fig. 2.3c–d demonstrate constant capacitance, however at the cost of
a slightly lower tunability. These layouts incorporate a larger bottom electrode periphery, and the ferroelectric film covering the step of the bottom electrode has a lower tunability. It is furthermore shown that the split-area layout, Fig. 2.3d, offers the highest quality factor.

For practical ferroelectric varactors (having ohmic connect lines, etc.), the quality factor generally varies with the device geometry, or say with the area $A$ and shape factor $\phi$. This is evident just by considering (2.10); the $R_s C_v$ term is generally not constant with area while $\tan \delta$ is. In paper [A] the varactor resistance associated with the layouts shown in Fig. 2.3b–c is modelled as

$$R_v = R_W \frac{1}{\sqrt{A\phi}} + R_L \sqrt{\frac{\phi}{A}} + R_A \frac{1}{A} \quad (2.12)$$

where $R_W$, $R_L$, and $R_A$ are constants determined by the contributions of ohmic and film loss in the device. By measuring devices of varying $A$ and $\phi$, the model parameters $R_W$, $R_L$, and $R_A$ can be determined, and an optimal shape factor (minimum $R_v$) can be calculated as $\phi_{opt} = R_W / R_L$. Note that (2.12) is a design tool that only attempts to model the variation with area and shape factor. There is no distinction made between contributions from $R_s$ and $\tan \delta$ (cf. (2.8)), and the parameters should furthermore be considered frequency dependent. However, using the expression it can be qualitatively concluded that for a certain shape factor the quality factor scales with area as

$$\frac{1}{Q_v} = \omega C_v \left( \alpha \frac{1}{\sqrt{A}} + \beta \frac{1}{A} \right) \quad (2.13)$$

where $\alpha$ and $\beta$ are coefficients related to the device periphery and the device area, respectively. Furthermore, by expressing the area as $A = C_v / \tilde{C}_v$ where $\tilde{C}_v$ is the capacitance per unit area, $Q_v$ is recast as

$$\frac{1}{Q_v} = \omega \left( \alpha \sqrt{\tilde{C}_v} C_v + \beta \tilde{C}_v \right) \quad (2.14)$$

It is seen that the quality factor at a fixed capacitance $C_v$ will be increased by reducing $\tilde{C}_v$. In addition to simply increasing the film thickness, there are at least two ways this can be done. First, the Ba/Sr composition ratio may be tuned to tailor the permittivity. A reduction of the permittivity will however inevitably also lower the tunability as discussed in section 2.1. The second alternative is to use composite films, in which the ferroelectric material is combined with a fraction of a low permittivity, nontunable material [17, 18] or amorphous ferroelectric material of lower permittivity [19]. As long as the permittivity of the non-tunable material is much lower than the permittivity of the ferroelectric film, the effective permittivity of the composite can be significantly lowered without reducing the tunability too much. This is however only true for columnar composites, i.e. when the low and high permittivity materials are effectively in parallel.
2.5 Technology comparison

For microwave applications the main competitors to ferroelectric varactors are variations of traditional semiconductor varactors (pn and Schottky junctions, MOS structures) using interdigital electrodes, e.g. [20, 21, 22, 23, 24, 25, 26]. A layout of a semiconductor varactor is shown in Fig. 2.10a, next to a layout of a ferroelectric varactor, Fig. 2.10b. The layouts are schematic, but the difference in size for a given capacitance value is illustrated. At high frequencies, say above 10 GHz, ohmic loss limits the performance of both technologies, although in terms of loss contributions mesa-structured semiconductor varactors are somewhat different from parallel-plate ferroelectric varactors. First of all, the active region of a semiconductor varactor (e.g. the pn junction in Fig. 2.11a) has a doping profile such that the doping density is higher near the junction and decreases away from the junction, in order to accommodate a high capacitive tunability [27]. The undepleted section of this volume constitutes an ohmic part of the varactor electrodes, ap-
pearing as a loss contribution that is proportional to the inverse of the of the device area. Combined with contact resistance, this is represented by $R_d \propto 1/A$ in Fig. 2.11b. Secondly, in mesa-structured semiconductor varactors, semiconducting sublayers partly constitute the current path to the active region. Granted these sublayers are highly doped, they still represent a higher resistance than metallic electrodes, and result in a spreading resistance approximately proportional to the inverse of the device periphery, say $R_{\text{sub}} \propto 1/\sqrt{A}$ in Fig. 2.11b.

Ferroelectric parallel-plate varactors, on the other hand, suffer from ohmic loss in narrow interconnect strips as discussed in section 2.4 and in paper [A]. The effective permittivity of typical ferroelectric varactors is one or two orders of magnitude higher than the substrate permittivity, so the varactor dimensions are much smaller than the dimensions of other passive elements, e.g. 50 Ω transmission lines. Because of this, interconnect strips must be narrow and consequently contribute with substantial ohmic loss. Ferroelectric varactors are also much smaller than semiconductor varactors, rendering interdigital electrode layouts impractical and difficult to fabricate. Due to these periphery related loss contributions, the quality factors of ferroelectric varactors vary with area as described by (2.13). In principle, this also applies for semiconductor varactors due to the contribution of $R_{\text{sub}}$, but as mentioned above, interdigital layouts have large periphery-to-area ratio rendering $R_{\text{sub}}$ small compared to $R_d$. The quality factors of such semiconductor varactors are hence fairly constant with area.

Beyond processing practicalities there is no limitation on metal thickness of the electrodes in ferroelectric or semiconductor varactors – the respective tuning mechanisms do not depend on the metallisation. The thickness should of course be sufficient (in relation to skin depth), and the conductivity should be high. It is however difficult to conclude what technology can offer the highest quality factor at a given metallisation. Going toward small capacitances, the quality factor of ferroelectric varactors is increasing, as illustrated by (2.14), approaching a limit set by the film loss. This contribution is probably lower than the area dependent ohmic loss seen in the active region of semiconductor varactors ($R_d$ in Fig. 2.11b), meaning that for small capacitances ferroelectric varactors should offer higher quality factors. For large capacitances, on the other hand, interdigital semiconductor varactors probably offer higher quality factors.

All in all, ferroelectric varactors and semiconductor varactors are comparable in terms of sheer performance (quality factor, tunability). From a fabrication point-of-view, regarding processing and integrability, there are however some differences. Ferroelectric varactors require less sophisticated fabrication equipment, compared to what is necessary for semiconductor technologies. Ferroelectric varactors can be tailored for a particular application, the film composition can for instance be adjusted to trade loss against tunability, and the film thickness can be adjusted
to trade linearity against tunability. Semiconductor varactors are on the other hand normally part of a bigger process technology and can not be easily modified since they share layers with active components. Being part of integrated processes is of course an intrinsic advantage when selecting varactors for IC designs. The comparison is further elaborated in section 4.2.2.

In this brief comparison of varactor technologies, some words should also be said about *microelectromechanical system* (MEMS). Just like the ferroelectric device technology, MEMS is an emerging technology which may be used in future microwave circuits [28, 29], and some examples of MEMS varactors are presented in [30, 31, 32, 33, 34]. Typical MEMS varactors have tuning ratios $C_{\text{max}}/C_{\text{min}}$ of about 1.5–4, comparable to what ferroelectric and semiconductor varactors can offer. Reported quality factors seem to be around 10–100 at 1 GHz. MEMS varactors are however quite slow with tuning speeds of about 1–100 $\mu$s, i.e. several orders of magnitude slower than ferroelectric and semiconductor varactors. This disqualifies MEMS for many VCO applications. Additionally, the technology requires sophisticated hermetic packaging. The technology is likely better suited for switchable devices used in high-power applications, than for continuously tuned oscillator applications.
Chapter 3

Thin film bulk acoustic resonators

This chapter presents thin film bulk acoustic resonators. The underlying material properties are described, and the devices are discussed in terms of design, characteristics and modelling. In the end, the technology is compared with other resonator technologies.

3.1 Piezoelectric and electrostrictive materials

3.1.1 Piezoelectricity

Dielectric crystals can be categorised in 32 crystal point groups. Out of these, 20 have the property of piezoelectricity, by definition meaning that the crystals are electrically polarised when put under mechanical stress. The effect is reversible, under applied electric field the crystals are mechanically deformed. The relationships of direct piezoelectricity (mechanical stress generating electric polarisation) and converse piezoelectricity (electric field generating mechanical strain) are linear by definition. Mathematically, piezoelectricity couples a pair of mechanical variables, typically strain $S$ and stress $T$, to a pair of electrical variables, electric field $E$ and either polarisation $P$ or electric flux density $D$. It can for instance be formulated as,

\[
T = c^D S - hD \tag{3.1}
\]

\[
E = -hS + \beta^S D \tag{3.2}
\]

where $c^D$ is the elastic coefficient (under constant $D$), $h$ is the piezoelectric coefficient, and $\beta^S$ is the dielectric impermeability (under constant $S$). The coupling coefficient, or coupling factor, is defined as

\[
k_t^2 = \frac{h^2}{c^D \beta^S}. \tag{3.3}
\]
The most well-known piezoelectric material is quartz (SiO$_2$) – bulk electroacoustic resonators based on crystalline quartz have been used for almost a century. More recent gigahertz thin film resonators have mostly been based on ZnO and AlN. In contrast to ZnO, AlN is compatible with CMOS processing and seems today to be the material of choice for commercial thin film devices [35]. Materials from the Perovskite family, e.g. PbZr$_x$Ti$_{1-x}$O$_3$ (PZT) and BaTiO$_3$, have also been considered for resonator applications because of their tunable characteristics. Low acoustic loss, high coupling coefficient, processability, and temperature stability are some important material parameters characterising good piezoelectric materials.

3.1.2 Electrostriction

Electrostriction is yet another physical property which is of importance for the devices studied by this thesis. The phenomenon, which is found in all dielectric materials, is a second-order relationship between strain and electric polarisation. It can be expressed as

$$S = \delta P^2$$  \hspace{1cm} (3.4)

where $S$, $\delta$ and $P$ are the strain, electrostriction coefficient and polarisation, respectively. In contrast to piezoelectricity, this relationship between polarisation and strain is nonlinear, and there exists no reverse relationship – polarisation cannot be induced by an applied mechanical stress.

The electrostrictive effect is interesting because it can be used to make tunable electroacoustic devices. When a material is put under a bias voltage, it is due to the electrostrictive effect deformed and a lattice asymmetry appears. Thereby, the material becomes piezoelectric as described in the previous section. In other words, the electrostrictive effect can be used to induce a small-signal piezoelectric effect in any material having a sufficiently large electrostriction coefficient. The mechanism can be mathematically described by considering the polarisation to be composed by a bias component $P_{dc}$ and a small-signal component $P_{ac}$,

$$P = P_{dc} + P_{ac},$$  \hspace{1cm} (3.5)

and by expanding the expression for the strain, (3.4),

$$S = \delta P_{dc}^2 + \delta P_{ac} \frac{b(P_{dc})}{b(P_{dc})} P_{ac} + \delta P_{ac}^2.$$  \hspace{1cm} (3.6)

A linear relationship between the small-signal parts of the strain and polarisation can then be identified as

$$S_{ac} = b(P_{dc})P_{ac}$$  \hspace{1cm} (3.7)
where \( b(P_{dc}) \) effectively is an induced, bias dependent piezoelectric coefficient. In most materials the electrostrictive effect is too weak to be useful for practical applications – the bias induced piezoelectric effect is negligible also at high bias voltages. However, ferroelectric materials tend to have large electrostriction coefficients, related to the fact that these materials have high permittivities. These materials have consequently been considered for tunable electroacoustic resonators.

### 3.2 Thin film resonators

The thin film bulk acoustic resonator (TFBAR) is fundamentally a pair of electrodes sandwiching a piezoelectric thin film in which acoustic waves can be induced by applied electric signals through the converse piezoelectric effect. Resonance occurs when the thickness of the active layer equals an integer of half acoustic wavelengths, at which point a standing wave appears, Fig. 3.1. Due to the direct piezoelectric effect, this wave induces a polarisation opposing the electrically generated polarisation. The resonance consequently appears electrically as a parallel resonance (zero current). At a slightly lower frequency, the acoustically generated polarisation is in phase with the electrically generated polarisation and a series resonance appears (zero voltage). The measured frequency response of a typical TFBAR is shown in Fig. 3.2. TFBARs belong to the family of bulk acoustic wave (BAW) devices, since the resonance frequencies are determined by the thickness dimension, independent of lateral dimensions. The maximum resonance frequency of TFBARs is limited by how thin active layers of sufficient quality can be deposited. TFBARs have been demonstrated at 20 GHz, e.g. [36], although most devices are designed for various wireless standards around 1–5 GHz.

To support a high quality factor of the acoustic resonances in the active layer, the resonator volume must properly be isolated (acoustically) from the lossy substrate underneath the bottom electrode. TFBARs are categorised as solidly mounted or membrane mounted depending
on how this isolation is provided\textsuperscript{1}. Fig. 3.3a shows the first type of TFBAR which is based on an acoustic reflector, a Bragg reflector, comprising $\lambda/4$-layers having alternating high and low acoustic impedance. Typically two or three pairs of low/high layers are necessary to provide a sufficiently good isolation. Ideally, the acoustic loading below the bottom electrode is zero at the centre frequency of the reflector. The second type, shown in Fig. 3.3b, relies on an air cavity formed beneath the bottom electrode (by micromachining of the substrate, or by formation of an air bridge). Fig. 3.3c shows a schematic layout where the top electrode is asymmetric (apodised) in order to suppress lateral modes of acoustic resonance.

TFBARs and other electroacoustic devices are generally compact, since their dimensions are determined by acoustic wavelength. Acoustic propagation speed in solid materials is of order $10^3$–$10^4$ m/s, while electromagnetic propagation speed is about $10^7$–$10^8$ m/s. Acoustical wavelength is thus four to five orders of magnitude lower than electrical wavelength. This makes TFBARs much smaller than electromagnetic resonators based on transmission line segments, for instance. The overlap area of typical AlN-TFBAR in a 50 $\Omega$ system is about $(100 \ \mu m)^2$, which is smaller than the area occupied by a typical lumped $LC$ resonator with an integrated spiral inductor. Furthermore, for common piezoelectric materials such as ZnO and AlN the acoustic loss is fairly low up and including gigahertz range. For instance, quality factors of 280 at 20 GHz for AlN-TFBARs have been demonstrated\textsuperscript{36}. Typical TFBARs offer quality factors of more than 1000 at 2 GHz.

\textsuperscript{1}Solidly mounted TFBARs are commonly known as \textit{solidly mounted resonators} (SMR), while membrane mounted TFBARs are known as \textit{free-standing bulk acoustic resonator} (FBAR). In this work, TFBAR is used for all types of thin film BAW resonators.
Fig. 3.3: Design of TFBARs, (a) cross-section of reflector type resonator, (b) cross-section of membrane type resonator, and (c) layout of oneport resonator with apodised top electrode.

### 3.3 Characteristics

The main TFBAR characteristics relevant for oscillators are the resonator quality factors and the associated resonance frequencies. Indirectly, the coupling coefficient is also important in that the quality factors, and more generally the impedance response, depends on it.

The quality factor of a resonator device is defined as

\[
Q = 2\pi \frac{E_{\text{max}}}{\Delta E}
\]  

(3.8)

where \(E_{\text{max}}\) is the maximum, or peak energy stored in the resonator, and \(\Delta E\) is the energy dissipated in its lossy parts during one resonance period. For simple resonators, \(Q\) is commonly determined from the 3-dB bandwidth of the impedance or admittance response, but this fails for devices such as TFBARs having multiple resonances in close proximity. Instead, the quality factor can be determined by the phase of the device impedance \(Z\), by evaluating

\[
Q(\omega) = \frac{\omega}{2} \left| \frac{\partial \arg Z}{\partial \omega} \right|
\]  

(3.9)

in the points of resonance, \(Q_{s,p} = Q(\omega_{s,p})\). This method is easy to implement and is used in the papers appended to this thesis. It is
Table 3.1: Resonance frequency definitions [38].

<table>
<thead>
<tr>
<th>Series resonance</th>
<th>Parallel resonance</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_m$</td>
<td>max. abs($Y$)</td>
</tr>
<tr>
<td>$f_s$</td>
<td>max. re($Y$)</td>
</tr>
<tr>
<td>$f_r$</td>
<td>im($Y$) = 0</td>
</tr>
<tr>
<td>$f_n$</td>
<td>max. abs($Z$)</td>
</tr>
<tr>
<td>$f_p$</td>
<td>max. re($Z$)</td>
</tr>
<tr>
<td></td>
<td>$f_a$ im($Z$) = 0</td>
</tr>
</tbody>
</table>

However, a bit problematic to use numerically since it is based on a derivative of the impedance-phase that changes abruptly around the resonance points. In order to get reliable quality factors using (3.9), it may be necessary to smooth the measured data before determining the derivative and then the quality factor. Otherwise, a small measurement fluctuation may cause a significant shift in the calculated quality factor. An alternative method of determining the quality factor [37] is

$$Q(\omega) = \omega \cdot \tau(\omega) \cdot \frac{|\Gamma|}{1 - |\Gamma|^2}$$

(3.10)

where $\Gamma$ is the reflection coefficient of the device, and $\tau$ is the group delay defined as

$$\tau(\omega) = -\frac{\partial \text{arg} \Gamma}{\partial \omega}.$$  

(3.11)

This method relies on the derivative of the reflection-phase which varies more uniformly with frequency, and its derivative is hence easier to evaluate without curve smoothing.

In relation to determining the quality factors, a few words need to be said about the resonance frequencies, and how they can be defined. To reiterate, the TFBAR has a series resonance and a parallel resonance, Fig. 3.2. Normally, a device is said to be resonating when its susceptance or reactance vanishes. There are however in fact at least three set of definitions for the points of series and parallel resonance of lossy devices [38], as listed in Table 3.1. For devices of high quality factors and coupling coefficients, the above definitions tend to coincide, but for devices of low quality factors and coupling coefficients there is some variation between the parameters. For instance, tunable devices described in section 3.6 may not be resonant in reaching zero reactance for low bias voltages, rendering the definition based on zero susceptance or reactance void. In this work, the second set of definitions given in Table 3.1 is used, unless otherwise is stated. Quality factors $Q_s$ and $Q_p$ are evaluated using (3.9) in the points of $f_s$ and $f_p$. One may additionally consider the so-called acoustic or mechanic quality factor $Q_m$ which is calculated from the model in Fig. 3.8 described in the next section, as $Q_m = \sqrt{L_m/C_m/R_m}$. It is a measure solely of acoustic loss, while $Q_s$ and $Q_p$ additionally depend on ohmic electrode loss and dielectric loss, respectively.

24
Once the resonance frequencies are established, the effective coupling coefficient can be calculated [39] as

$$k_{t,\text{eff}}^2 = \frac{\pi}{2} \frac{f_s}{f_p} \frac{1}{\tan \left( \frac{\pi f_s}{2 f_p} \right)}.$$  \hspace{1cm} (3.12)

It is determined by the material coupling coefficient of the active layer, as defined by (3.3), and to a less extent by the device geometry. This is an important measure when designing filters based on TFBARs, since the $f_s/f_p$ ratio relates to the filter bandwidth. Wide bandwidths require high effective coupling coefficients, meaning materials of high material coupling coefficients. For oscillator design, the coupling coefficient is less important since an oscillator only makes use of one resonance point, either $f_s$ or $f_p$, and the resonance separation is usually irrelevant. The coupling coefficient is however implicitly important in that $Q_s$ and $Q_p$ depends on it. A material with a low acoustic loss is useless from an electrical point-of-view, if the coupling coefficient is too small.

In paper [F], the variation in resonance frequencies and quality factors of AlN-TFBARs distributed over a four-inch wafer of test resonators is measured and investigated. The test wafer is shown in Fig. 3.4. In this work, the AlN active layer is deposited using a reactive sputtering process [40], and as a result the film thickness $t$ is of Gaussian shape, $t \propto \exp \left( -d^2 \right)$, where $d$ is a normalised parameter ranging from 0 (centre of wafer) to 1 (edge of wafer),

$$d = \frac{\text{distance from centre of wafer}}{\text{maximum distance}}.$$  \hspace{1cm} (3.13)

Variation in $f_p$ and $Q_m$ with $d$ is shown in Fig. 3.5. The resonance frequency is increasing when approaching the wafer edge, as the film becomes thinner. Simultaneously the quality factor drops, mostly because the resonance point goes away from the centre of the reflection.
Fig. 3.5: Variation in (a) parallel resonance frequency, and (b) acoustic quality factor with wafer position (paper [F]).

band of the Bragg reflector. The large spread in the quality factor is not only due to actual variation in the film quality, but also because of inaccuracies in the measurement and extraction procedures. For instance, variations in the contact resistance of the probes will have a significant influence when measuring devices with very high quality factors. There are also numerical issues when estimating the quality factor using (3.9) as discussed above, in this case curve fitting was used to find a reasonable phase-derivative. The average measure is reliable, however.

This kind of wafer variation needs to be battled to allow high-yield commercial fabrication of TFBARs. Practically, it is indeed hard to achieve high uniformity of AlN films over large areas of 4–8 inch wafers [41, 42]. In paper [F] the main goal was not to solve this problem, but to demonstrate oscillator circuits implemented in a novel integration concept (described in section 4.3.1). The variation was considered by a resonator model parameterised by $d$, which allowed the subsequently
3.4 Electroacoustic modelling

The electrical response of a TFBAR is a function not only of the dielectric response, but also of the acoustic nature of the device. Generally, a one-dimensional electroacoustic device may be represented by a three-port model, Fig. 3.6. One port is electrical, corresponding to the electrodes of the resonator, while the other two are mechanical ports to which mechanical loads above and below the resonator volume, $Z_L$ and $Z_R$, are connected. By solving the piezoelectric equations in the one-dimensional case (ignoring lateral effects), an expression of the input impedance of the electrical port is given as

$$Z = \frac{1}{j\omega C_0} \left( 1 - k_{t,\text{eff}}^2 \tan \phi \frac{z_m}{z_m} \right)$$  \hspace{1cm} (3.14)

where $C_0$ is the dielectric parallel-plate capacitance, $k_{t,\text{eff}}^2$ is the effective coupling coefficient, $\phi$ is the acoustic phase over the active layer, and

$$z_m = \frac{(z_r + z_l) \cos^2 \phi + j \sin 2\phi}{(z_r + z_l) \cos 2\phi + j(z_r z_l + 1) \sin 2\phi}$$  \hspace{1cm} (3.15)

describes the mechanical loading of the resonator [39]. In (3.15) the mechanical loads are normalised by the characteristic acoustic impedance of the active layer, $Z_P$, as $z_l = Z_L/Z_P$ and $z_r = Z_R/Z_P$. The expressions are valid for thickness longitudinal modes of resonance, such as fundamental resonances in TFBARs, under so-called L-effect coupling (depolarisation field parallel to acoustic propagation vector) and ideal boundary conditions.

Mason’s circuit model [43] shown in Fig. 3.7 is essentially (3.14) expressed in circuit elements, and may as such be more suitable for implementation in circuit simulators. For instance, in case of layered structures, the layers may be represented by acoustic transmission lines of
Fig. 3.7: Mason’s model of a piezoelectric resonator.

\[
Z_A = jZ_P \tan \phi \\
Z_C = -jZ_P / \sin 2\phi \\
n^2 = 2\phi/(k_{\text{eff}}^2 \omega C_0 Z_P)
\]

certain lengths, propagation constants, and characteristic impedances. The mechanical loading above and below the resonator volume, \(Z_L\) and \(Z_R\), are then implemented by cascading these transmission lines accordingly. This is an intuitive and simple way of experimenting and developing layered TFBAR structures. Elements representing ohmic and dielectric loss are readily added to the electrical port of the model, and acoustic loss in active layer and reflector layers (due to viscosity) can be introduced by complex propagation constants.

Another commonly used model is the modified Butterworth-Van Dyke (MBVD) model [44], of which two variants are shown in Fig. 3.8. The parallel-plate dielectric capacitance is represented by \(C_0\), and the acoustic resonance is modelled by capacitance \(C_m\), inductance \(L_m\) and acoustic loss \(R_m\). Ohmic loss of the electrodes is represented by \(R_s\), while dielectric loss is represented either by a series element \(R_0\) as in Fig. 3.8a, or by a parallel element \(G_0\) as in Fig. 3.8b (of different frequency dependencies). Usually the former alternative is used, but in paper [B] the latter is preferred to model BSTO-TFBARs. It better reflects the additive nature of loss tangents due to different dielectric loss mechanisms, especially toward zero frequency.
Mason’s model is generally used to predict the response of a device, before fabrication. It may for instance be used to optimise the layers of a Bragg reflector. The MBVD model, on the other hand, is typically extracted from measurements and used as a scalable model in circuit simulations. A circuit designer may scale the model with device area $A$ and thickness $t$, assuming that material parameters such as coupling coefficients and quality factors are invariant with geometry. The dielectric capacitance $C_0$ is proportional to $A/t$, and resonance frequencies $f_s$ and $f_p$ are proportional to $1/t$. From this, $L_m$ and $C_m$ can be determined. Loss elements $R_m$ and $G_0$ are determined by quality factors. This works well as long as the device area is sufficiently big relative to the thickness, otherwise the parallel-plate approximation and one-dimensional acoustic modelling will fail.

The remaining element, $R_s$, may be a bit more difficult to scale appropriately with device area. For peculiar layouts, e.g. tapered interconnect lines and apodised top electrodes, it may be necessary to use an electromagnetic (EM) simulator to calculate the ohmic loss. Fortunately, one can use internal ports to connect the acoustic branch of the MBVD model to the EM simulated electrical network [45]. This technique can also be used in combination with Mason’s model. Fig. 3.9 shows an example simulation where the ohmic loss is EM simulated by ADS Momentum$^2$, and combined with the acoustic part of Mason’s model. The top electrode metal thickness is varied. A larger thickness will reduce the ohmic loss, say $R_s$, but it may also degrade the device acoustically if the acoustic quality factor of the metal is poor. As seen in this example, the quality factors initially increase with metal thickness. However, with growing thickness the top electrode will constitute a gradually larger fraction of the active medium (the piezoelectric medium combined with electrodes, given that interfaces below and above electrodes are ideal). The overall quality factors, $Q_s$ and $Q_p$, will hence increasingly depend on the quality of the top electrode. As seen in the simulation, $Q_s$ and $Q_p$ degrade for large metal thickness.

### 3.5 Roughness modelling

Another aspect of TFBAR modelling is the impact of film roughness. Rough film interfaces may cause scattering of incident acoustic waves which in turn result in extra loss. Moreover, the active layer of TFBARs comprises thousands of grain columns, of varying thickness and area, as seen in Fig. 3.10 showing a scanning electron microscopy (SEM) image of a BSTO film. Naturally, processing is aimed at having a film of uniform thickness and quality, but there is always some variation due to bottom electrode roughness and deposition conditions. There will be a variation in the column thickness across the surface of the device, and

---

$^2$ADS Momentum is a trademark of Agilent Technologies Inc.
Fig. 3.9: TFBAR quality factors at (a) series and (b) parallel resonances, simulated using ADS Momentum (ohmic and dielectric loss) combined with Mason’s model (acoustic loss). Parameter $\kappa$ is the acoustic quality factor of the electrodes normalised by the acoustic quality factor of the active layer.

Fig. 3.10: Typical SEM cross section image of a BSTO-TFBAR, average column thickness $t$ is about 350 nm.
this will have an impact on the TFBAR performance, as considered in paper [C]. As it turns out, this effect seem to be quite substantial.

The simplest way of modelling this kind of variation, or roughness, is to consider the columns as parallel-connected resonators of varying resonance frequency and characteristic admittance. The column thicknesses are assumed to follow a normal distribution characterised by a nominal (average) thickness \( t_{\text{nom}} \), and a standard variation \( \sigma_t \), Fig. 3.11. Columns over a range of thickness, \( t_{\text{nom}} \pm \Delta t \), are considered. The range should be chosen wide, say \( \Delta t = 3\sigma_t \) including about 99.7% of all columns. The range is then discretized into a large number \( N \) intervals; \( t_i \pm \delta t \), where

\[
\delta t = \frac{\Delta t}{N - 1} \quad (3.16)
\]

and \( i = 1, \ldots, N \). If \( \delta t \) is sufficiently small, columns within the \( i \)th interval can be considered to have the same thickness,

\[
t_i = t_{\text{nom}} - \Delta t + 2\Delta t \frac{i - 1}{N - 1}. \quad (3.17)
\]

The equivalent area for columns within the \( i \)th interval is then determined by the statistical distribution as

\[
A_i = A_{\text{total}} \left( F(t_i + \delta t) - F(t_i - \delta t) \right) \quad (3.18)
\]

where \( A_{\text{total}} \) is the total device area and \( F(\cdot) \) is the cumulative normal distribution function. The admittance for the \( i \)th column group, \( Y_i \), is a function of the area \( A_i \), thickness \( t_i \) and, if the TFBAR material is tunable, by the bias field \( E_i = V_i/t_i \). The total admittance of the TFBAR is then calculated as

\[
Y_{\text{tot}} = \sum_{i=1}^{N} Y_i. \quad (3.19)
\]
From the calculated admittance, quality factors $Q_s$ and $Q_p$ are determined using (3.9). Fig. 3.12 presents the quality factors against $\sigma_t$ (i.e. roughness), from a simulation of a TFBAR based on a Bragg reflector. For this qualitative analysis, the acoustic quality factors of all columns and all layers in the reflector are set to 1000, and dielectric loss is ignored. Remaining material parameters are calculated for BSTO, as described in section 3.6.

Modest standard deviations of just a few nm seem to degrade the series resonance quality factor significantly, and also to a less degree the parallel resonance quality factor. This is remarkable since it corresponds to a roughness variation of just a few percent of the acoustic wavelength. However, if the individual columns represent high quality factors also small shifts in resonance frequency should dramatically reduce the quality factors. Parts of the assumptions made in the analysis may be questionable, for instance whether the columns really can be modelled as parallel-connected networks, or if there is some mutual acoustic coupling between the columns that needs to be considered. It seems reasonable to conclude, nevertheless, that having smooth interfaces and accurately grown films is crucial in order to reach optimal quality factors.

3.6 Tunable resonators

Traditional TFBARs are generally not tunable; their electroacoustic properties do not change under a bias field. However, as described in section 3.1.2, a small-signal piezoelectric effect can be induced by a bias voltage in materials having large electrostriction coefficients. This bias dependent effect enables tunable TFBARs.

In this thesis, tunable TFBARs based on paraelectric phase BSTO films are considered. Fig. 3.13 shows the bias dependent parameters ex-
tracted from measurements of a typical device. The effective coupling coefficient, Fig. 3.13a, is vanishing for zero bias voltage, since unbiased paraelectric phase BSTO is centrosymmetric and thus not piezoelectric. As the bias voltage is increased, the lattice becomes increasingly asymmetric (due to electrostriction) and the coupling coefficient grows. Since the material is in the paraelectric phase, no hysteresis is observed in the extracted parameters as the voltage is swept from 0 V to 18 V and then back to 0 V. A device based on ferroelectric phase material would be hysteretic.

This effect of bias induced piezoelectricity has been modelled ([46] and paper [c]) as

$$h = h_0 + 2qP_{dc}$$  \hspace{1cm} (3.20)

where $h$ is the effective piezoelectric coefficient, $h_0$ is the zero-bias piezoelectric coefficient (vanishing for a paraelectric material), $P_{dc}$ is the bias polarisation, and $q$ is the electrostriction coefficient. Moreover, the elastic coefficient is bias dependent as

$$c^D = c^0 - mP_{dc}^2 - \frac{h^2\epsilon_b}{1 + \epsilon_b/\chi_f}$$  \hspace{1cm} (3.21)

where $c^0$ is the zero-bias elastic coefficient, $m$ is a higher-order electrostriction coefficient, $\epsilon_b$ is the background permittivity, and $\chi_f = \epsilon - \epsilon_b$ is the ferroelectric contribution to the susceptibility. The latter is bias dependent since the material is ferroelectric. The resonance frequencies, Fig. 3.13b, consequently change with bias voltage, and the result is a tunable TFBAR.

The acoustic branch quality factor $Q_m$ is believed to be independent of bias voltage, Fig. 3.13c. For an intrinsic device, a condition for resonance (zero reactive part) is

$$Q_m k_{i, eff}^2 > \frac{\pi^2}{4}$$  \hspace{1cm} (3.22)

assuming $Q_m \gg 1$ and $C_0 \gg C_m$. In this case $Q_m$ is about 120, implying that the condition is true for an effective coupling coefficient of at least 2%, corresponding to a bias voltage of about 7 V. The series resonance quality factor $Q_s$ grows with increasing $k_{i, eff}^2$ until (3.22) is true, at which point it saturates at

$$Q_s \simeq \frac{Z_c}{R_m + R_s}.$$  \hspace{1cm} (3.23)

If $R_s$ is small compared to $R_m$, then $Q_s \simeq Q_m$. For this example the ohmic loss of $R_s$ limits the device performance and $Q_s$ saturates at a level below $Q_m$.

Fig. 3.14a shows the measured reflection coefficient for a typical BSTO-TFBAR biased at 4 and 18 V. The resonance loop grows with
Fig. 3.13: Bias dependent parameters extracted from measurements of a BSTO-TFBAR, paper [B], (a) the effective coupling coefficient, (b) series and parallel resonance frequencies, and (c) mechanical and series quality factors. The bias voltage is swept $0 \rightarrow 18 \rightarrow 0$ V.
bias voltage, and as predicted by (3.22), the device becomes resonant at a bias voltage of about 7 V. For comparison, the measured reflection coefficient for a typical AlN-TFBAR is shown in Fig. 3.14b. Although the device sizes and resonance frequencies are different, the plots qualitatively illustrate the difference between a tunable BSTO-TFBAR and a traditional AlN-TFBAR. The difference in resonance loop size is obvious even though the particular AlN-TFBAR shown is not state-of-the-art. Especially, the parallel resonance resistance of the AlN-TFBAR is much larger than the counterpart of the BSTO-TFBAR. For given capacitance and resonance frequencies, the size of the resonance loop is governed by the acoustic quality factor $Q_m$ and the effective coupling coefficient $k_{t,\text{eff}}^2$. In this example the $Q_m$ of the AlN-TFBAR is about ten times higher than the $Q_m$ of the BSTO-TFBAR. The bias independent $k_{t,\text{eff}}^2$ of the AlN-TFBAR is similar to the $k_{t,\text{eff}}^2$ of the BSTO-TFBAR biased at 18 V.

Fixed-frequency TFBARs based on e.g. AlN tend to be rather easy to characterise using the MBVD model for which numerical routines have been developed [47]. On the other hand, tunable TFBARs are because of their inherently low coupling factor (at low bias voltages) and currently lower quality factors more difficult to extract accurately. For instance, approximations such as $f_s = f_m$ hold well for devices of high quality factors, but are not adequate for BSTO-TFBARs. In response to this problem, paper [B] suggests a robust routine for extraction of tunable TFBARs. The proposed method is a step-wise fitting procedure to determine all parameters per bias point. First, all non-acoustical parameters ($R_s$, $G_0$ and $C_0$ in Fig. 3.8b, contingently also a series inductance $L_s$) are extracted from a simple impedance fitting away from the acoustic resonances. One could expect it to be trivial to determine
the remaining elements of the acoustic branch \((R_m, C_m, L_m, \text{Fig. 3.8b})\) simply by removing all non-acoustical parameters from the measured impedance. However, when the coupling coefficient is low, a small error in the non-acoustical parameters will cause a large error in the (vanishing) acoustic branch. Instead the method relies on an alternative way of determining the resonance of the acoustic branch, \(\omega_m = (L_mC_m)^{-1/2}\). In short, \(\omega_m\) is defined from a minimum in the frequency-derivative of the intrinsic susceptance \(B_i\). The intrinsic part of the model is simply everything but the series elements \(R_s\) and \(L_s\) which can be determined with high accuracy since they are not bias dependent. The minimum in \(\frac{\partial B_i}{\partial \omega}\) is not a function of \(C_0\) nor \(G_0\), and is hence not affected by any associated extraction errors. In the resonance point defined by \(\omega_m\) any susceptance in the measured impedance can be attributed to the dielectric capacitance \(C_0\), and the acoustic branch can then be robustly determined from a narrow frequency range around \(\omega_m\). Paper [B] also suggests a way of having lower and upper estimations of the acoustic branch characteristic impedance \(Z_c = (L_mC_m)^{1/2}\) and the quality factor \(Q_m = Z_c/R_m\). These estimations provide an indication of whether or not the model and extraction procedure are accurate. Parameter \(Q_m\) presented in Fig. 3.13c has been extracted using the suggested method, and the curve represents the average of the lower and upper estimations of \(Q_m\).

### 3.7 Technology comparison

During recent years TFBARs of different materials and configurations have demonstrated quality factors above 2000 at a few gigahertz. Avago reportedly have devices of \(Q \simeq 5000\) at 2 GHz [48], implying \(Q \times f\) products comparable to those of low-frequency quartz resonators. TriQuint, EPCOS, and Skyworks are other examples of companies offering TFBAR products. TFBAR duplexer filters for mobile phones in the frequency range 1–3 GHz have been successfully commercialised.

A selection of planar resonator technologies are listed in Table 3.2, and compared in terms of frequency, tunability (if any), quality factor and temperature coefficient of frequency (TCF) [35, 49, 50, 51]. Applications based on surface acoustic wave (SAW) technology, mostly filters, have been used since the 1970s, and during the 1980s the technology was developed for microwave electronics. Before the advent of the TFBAR technology by the early 2000s, SAW had a unique position as a planar high quality factor technology. It is a mature and cheap technology adapt for mass fabrication, not as vulnerable as TFBAR to issues of film uniformity as discussed in section 3.3. However, SAW relies on accurate photo-lithography which sets the upper frequency limit to about 5 GHz, and the technology requires special substrate materials (commonly LiNbO\(_3\) or LiTaO\(_3\)), while TFBARs can be processed on almost
Table 3.2: Typical parameters of some planar resonator technologies.

<table>
<thead>
<tr>
<th>Technology</th>
<th>Frequency [GHz]</th>
<th>Tunability</th>
<th>$Q \times f$ [GHz]</th>
<th>TCF [−ppm/K]</th>
</tr>
</thead>
<tbody>
<tr>
<td>AlN-TFBAR</td>
<td>&lt; 20</td>
<td>-</td>
<td>2000–5000</td>
<td>&lt; 30</td>
</tr>
<tr>
<td>BSTO-TFBAR</td>
<td>&lt; 20</td>
<td>&lt; 3%</td>
<td>100–500</td>
<td>≃ 50</td>
</tr>
<tr>
<td>SAW</td>
<td>&lt; 5</td>
<td>-</td>
<td>1000–2000</td>
<td>&lt; 40</td>
</tr>
<tr>
<td>Integrated stub</td>
<td></td>
<td>with varactor</td>
<td>300–3000$^2$</td>
<td></td>
</tr>
<tr>
<td>Integrated LC</td>
<td></td>
<td>with varactor</td>
<td>30–300$^2$</td>
<td></td>
</tr>
</tbody>
</table>

$^1$Estimated for room-temperature $\text{Ba}_{0.25}\text{Sr}_{0.75}\text{TiO}_3$, at $f_s$.

anything, for instance silicon. This is a great advantage since processing methods and infrastructure developed for standard Si processes can be used. The power-handling capability of SAW is also generally lower than that of TFBARs.

Lumped $LC$ resonators or distributed elements (e.g. stub resonators) fabricated in some kind of integrated process are examples of planar electromagnetic resonators. The quality factor of such solutions is determined by substrate and metallisation. Stub resonators can reach high quality factors on low-loss substrates, but they are large compared to TFBARs, since electromagnetic wavelength is about four order of magnitude longer than acoustic wavelength. Discrete $LC$ resonators are smaller than stub elements, although integrated inductor coils normally occupy a large chip area. Typical quality factors of $LC$ resonators are however quite poor.

Tunable TFBARs based on electrostriction represent the only intrinsically tunable technology of Table 3.2. Varactors can be used to tune the other technologies, but this adds cost and lowers the quality factor. Since 2007, tunable TFBARs based on BSTO have been published by a number of research groups (including Chalmers). As summarised in [52], TFBARs of reflector type, [46, 53, 54, 55] and paper [c], and membrane type, [56, 57, 58] have been demonstrated for frequencies from about 1 to 7 GHz. Membrane type TFBARs tend to have a slightly better performance in terms of coupling coefficient and quality factor, as is the case for traditional TFBAR technologies based on AlN or ZnO. Typical quality factors are about 100–200 at a few gigahertz, and tunabilities less than 3%. Usually the presented quality factors are the maximum values over the tuning range, at the highest bias voltage. Furthermore, the tunabilities are typically somehow calculated from zero bias to maximum bias, and thus include a range where true resonances do not exist (the condition defined by (3.22) is not satisfied over the full tuning range). The practical tunability available for say a microwave oscillator is then roughly half of the presented tunabilities.
The published results of tunable TFBARs are so far quite modest, and the technology requires substantial development to be of practical use for oscillators. Being based on electrostriction, these devices will inevitably have vanishing resonances (low quality factors) at low bias voltages. Although the electrostrictive effect is strong, it will eventually saturate at some level of bias field, and the useful tuning range is thus limited. It may be more interesting to use the devices as switchable resonators, rather than as continuously tuned resonators.

However, most of the research groups that have initiated the development of tunable TFBARs have backgrounds in the field of ferroelectric varactors, rather than in the field of acoustic devices. The published tunable TFBARs are thus of simple, rudimentary designs. On the other hand, the past decade’s dramatic development of traditional TFBAR technology has been driven by companies and other research groups with backgrounds in acoustic technologies. Industry state-of-the-art AlN-TFBARs likely involve numerous tweaks of proprietary technology, much of which is yet to be exploited by BSTO-TFBARs. One example is the importance of having films with smooth interfaces and well-defined column structure, as discussed in section 3.5 and in paper [C]. Films of modest roughness may work fine for ferroelectric varactors, but are terrible for electroacoustic applications. Improvements here should give the tunable TFBARs a significant performance boost.
Chapter 4

Oscillators based on thin film devices

This chapter begins with fundamental oscillator theory, before presenting experimental results of oscillators based on ferroelectric varactors and TFBAR devices. The device technologies are then discussed from the perspective of oscillator applications.

4.1 Oscillator theory

A harmonic oscillator is essentially a resonator combined with an active device. The purpose of the active device is to cancel any loss in the resonator, thereby supporting a constant oscillation amplitude at a frequency defined by the resonator. The performance of the oscillator naturally depends on the characteristics of both parts, in addition to how the parts are combined (circuit topology). This work is focused on development of thin film devices which can be used in the resonator part of microwave oscillators in order to improve their performance. The most important performance measures for oscillators studied in this thesis are frequency tunability and phase-noise. The former is obviously a function of the resonator’s tunability, and in turn a function of the capacitive tunability of the varactor if one is used. The frequency tunability of the oscillator is always reduced by parasitic, nontunable, reactive elements in the circuit. In this work, frequency tunability is defined as $T_f = \Delta f / f_c$ where $\Delta f$ is the tuning range and $f_c$ is the centre frequency.

Compared to tunability, phase-noise is vastly more complicated to discuss and measure. Consider an oscillator output signal given as

$$v(t) = A(t) \cos[\omega_0 t + \phi(t)].$$

(4.1)

Amplitude $A(t)$ and phase $\phi(t)$ are ideally time-invariant, but in practice both fluctuate over time due to noise sources external and internal to
the oscillator. Examples of internal noise sources are flicker noise of the active device and thermal noise in lossy circuit elements. Bias supply noise leaking through the bias network is an example of an external noise source. Amplitude-noise and phase-noise are defined as short term stochastic variations in $A(t)$ and $\phi(t)$ (time scales of order fundamental oscillation period). The former is limited by the amplitude-restoring mechanism inherent to the oscillation steady-state. Phase-noise is on the other hand persistent, and is consequently the main contributor to oscillator noise manifested as broadening of the oscillation spectrum. For a small-valued $\phi(t)$, (4.1) is approximately

$$v(t) \simeq A \cos(\omega_0 t) + A' \phi(t) \sin(\omega_0 t)$$

(4.2)

where the fluctuations in the amplitude are disregarded and $A' = A/\text{rad}$. In the single-sideband frequency domain, the power spectrum of the oscillation is

$$S_v(\omega) = A \delta(\omega - \omega_0) + A' S_\phi(\omega) * \delta(\omega - \omega_0)$$

(4.3)

where $*$ denotes convolution, and $\delta(\cdot)$ is Dirac’s delta function. The power spectrum of the phase-noise, denoted $S_\phi$, appears as sidebands around the centre frequency of the output signal which is said to be phase-modulated. This frequency broadening has a negative impact on communication systems; for many applications it is crucial to keep the phase-noise to a minimum. For instance, a local oscillator (LO) with a high degree of phase-noise may cause channel interference when used in a down-converting receiver stage, as illustrated in Fig. 4.1.

Phase-noise is usually expressed relative to carrier, as the ratio of the sideband power at offset frequency $\omega_{\text{off}}$ to the carrier power at frequency $\omega_0$. The sideband power is typically calculated by integrating the power spectrum over the unit bandwidth (1 Hz), and the phase-noise is then expressed as

$$L(\omega_{\text{off}}) = 10 \log \frac{S_v(\omega_0 + \omega_{\text{off}}, 1 \text{ Hz})}{S_v(\omega_0)}.$$
Although the above measure is traditionally referred to as phase-noise, it also includes a minor contribution of amplitude-noise. Leeson’s phase-noise formula [59],

$$L(\omega_{\text{off}}) = 10 \log \left\{ \frac{2FkT}{P_s} \cdot \left[ 1 + \left( \frac{\omega^2}{\omega_{\text{off}}} \right)^2 \right] \cdot \left( 1 + \frac{\omega_3}{\omega_{\text{off}}} \right) \right\}, \quad (4.5)$$

is commonly used as a qualitative model of phase-noise spectra. Here, $F$ is a fitting parameter related to the noise factor of the active device, and $P_s$ is the power dissipated in the resonator tank. The expression predicts three distinct spectrum regions of $\omega^{-3}$, $\omega^{-2}$ and, $\omega^0$ slopes, as illustrated in Fig. 4.2. These regions are defined by corner frequencies $\omega_2 = \omega_0/(2Q)$, and $\omega_3$. The latter is related to the corner-frequency of the active device flicker noise but is generally considered to be a fitting parameter [60]. Corner frequency $\omega_2$ is determined by a quality factor $Q$ which commonly is assumed to be the loaded quality factor of the resonator which in fact is merely approximately correct. It has been shown [61] that this spectrum-shaping quality factor also depends on the characteristics of the active device, and should be expressed as

$$Q = \frac{\omega_0}{2} \left| \frac{d}{d\omega} \ln \Psi \right| = \frac{\omega_0}{2} \sqrt{ \left( \frac{d}{d\omega} \ln \text{abs } \Psi \right)^2 + \left( \frac{d}{d\omega} \arg \Psi \right)^2 } \quad (4.6)$$

where $\Psi$ is the corresponding net-immittance of the circuit, including not only the resonator but also the active device. The traditional interpretation of $Q$ as the loaded quality factor is problematic since the active device in an oscillator will cancel the energy lost in the resonator. Steady-state oscillation hence implies having a loaded quality factor approaching infinity. A high resonator quality factor is nevertheless crucial to limit the phase-noise. For practical and well-designed oscillators, the amplitude variation in $\Psi$ around the resonance is rather small, and (4.6) is reduced to

$$Q = \frac{\omega_0}{2} \frac{d}{d\omega} \arg \Psi \quad (4.7)$$
where most of the phase variation in $\Psi$ is due to the fast impedance variation of the resonator tank near its resonance frequency. The phase-noise at small offset frequencies (say, below 10–100 kHz) is partly determined by the noise properties of the active device, but at higher offsets the thermal noise converted to $\omega^{-2}$-noise is dominant and active device noise is no longer particularly important. The quality factor of the resonator sets the overall phase-noise level at all offset frequencies below $\omega_2$.

In addition to frequency tunability and phase-noise characteristics, oscillators are commonly compared in terms of a figure-of-merit defined as

$$FOM = -\mathcal{L}(\omega_{\text{off}}) + 10 \log \left( \frac{\omega_0}{\omega_{\text{off}}} \right)^2 - 10 \log \left( \frac{P_{\text{dc}}}{1 \, \text{mW}} \right). \quad (4.8)$$

This FOM comprises the phase-noise (scaled by centre and offset frequencies), and the power consumption $P_{\text{dc}}$. As such it is suitable for comparison of low-power reference oscillators.

### 4.2 Oscillators tuned by ferroelectric varactors

As discussed in chapter 2, varactors used in integrated microwave circuits typically suffer from poor quality factors. At high frequencies, say above 10 GHz, the varactor quality factor is usually lower than the quality factor of available inductor elements. The phase-noise performance of oscillators based on integrated resonators is consequently limited by the traditionally used varactor technologies. In this context ferroelectric varactors represent an interesting alternative, not only because they may offer higher quality factors but also due to their flexibility in that film composition and thickness can be tailored to a certain application with only minor processing adjustments. A number of oscillators using ferroelectric varactors as tuning element have been published, as listed in Table 4.1. Most of them are megahertz oscillators [63, 64, 65, 67, 68, 71], but there are also a few examples of gigahertz oscillators. In [66], a 1.6 GHz VCO using an AlGaN/GaN HFET as active device is reported to have a phase-noise of $-81.4$ dBC/Hz at 100 kHz offset and a frequency tunability of 3%. In [62] a cryogenic GaAs pHEMT VCO is reported operating at 43 K to reach a high tunability of the varactors based on STO. The varactors are formed on a LaAlO$_3$-substrate and the VCO operates at 16.7 GHz with a 3% tuning, phase-noise is not reported.

At the time for publication, paper [D] reported the highest operating frequencies for VCOs based on room-temperature ferroelectric varactors, with centre frequencies of 16.5 GHz and 19.6 GHz. It was also the first design having the ferroelectric devices integrated on carrier substrates of silicon, specifically p-type high-resistivity silicon (HRS). The
Table 4.1: Summary of oscillators tuned by ferroelectric varactors.

<table>
<thead>
<tr>
<th>Year</th>
<th>Ref.</th>
<th>Varactor</th>
<th>Technology</th>
<th>( f_0 ) [GHz]</th>
<th>( T_f ) [%]</th>
<th>( P_{\text{out}} ) [dBm]</th>
<th>( P_{\text{dc}} ) [mW]</th>
<th>( L(f_{\text{off}}) ) [dBc/Hz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1998</td>
<td>[62]</td>
<td>( \text{SrTiO}_3 )</td>
<td>GaAs pHEMT</td>
<td>16.7</td>
<td>3</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2003</td>
<td>[63]</td>
<td>( \text{Ba}<em>{0.95}\text{Ca}</em>{0.05}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.0024</td>
<td>8</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2004</td>
<td>[64]</td>
<td>( \text{Ba}<em>x\text{Sr}</em>{1-x}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.04</td>
<td>28</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2006</td>
<td>[65]</td>
<td>( \text{Ba}<em>x\text{Sr}</em>{1-x}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.05</td>
<td>11</td>
<td></td>
<td></td>
<td>-120 @ 100 kHz</td>
</tr>
<tr>
<td></td>
<td>[66]</td>
<td>( \text{Ba}<em>{0.6}\text{Sr}</em>{0.4}\text{TiO}_3 )</td>
<td>AlGaN/GaN HFET</td>
<td>1.6</td>
<td>3</td>
<td>32</td>
<td></td>
<td>-81.4 @ 100 kHz</td>
</tr>
<tr>
<td>2007</td>
<td>[67]</td>
<td>( \text{Ba}<em>{0.5}\text{Sr}</em>{0.5}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.21</td>
<td>5</td>
<td></td>
<td></td>
<td>-90 @ 10 kHz</td>
</tr>
<tr>
<td></td>
<td>[D]</td>
<td>( \text{Ba}<em>{0.25}\text{Sr}</em>{0.75}\text{TiO}_3 )</td>
<td>SiGe HBT</td>
<td>16.5</td>
<td>6.7</td>
<td>3</td>
<td></td>
<td>-95 @ 100 kHz</td>
</tr>
<tr>
<td></td>
<td>[D]</td>
<td>( \text{Ba}<em>{0.5}\text{Sr}</em>{0.5}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.59</td>
<td>4</td>
<td></td>
<td></td>
<td>-102 @ 100 kHz</td>
</tr>
<tr>
<td>2007</td>
<td>[68]</td>
<td>( \text{Ba}<em>{0.5}\text{Sr}</em>{0.5}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.21</td>
<td>5</td>
<td></td>
<td></td>
<td>-90 @ 10 kHz</td>
</tr>
<tr>
<td></td>
<td>[D]</td>
<td>( \text{Ba}<em>{0.5}\text{Sr}</em>{0.5}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.59</td>
<td>4</td>
<td></td>
<td></td>
<td>-102 @ 100 kHz</td>
</tr>
<tr>
<td>2008</td>
<td>[69]</td>
<td>( \text{Ba}<em>{0.3}\text{Sr}</em>{0.7}\text{TiO}_3 )</td>
<td>Gunn diode</td>
<td>17</td>
<td>0.8</td>
<td>9</td>
<td></td>
<td>-95 @ 100 kHz</td>
</tr>
<tr>
<td></td>
<td>[69]</td>
<td>( \text{Ba}<em>{0.5}\text{Sr}</em>{0.5}\text{TiO}_3 )</td>
<td>BJT</td>
<td>0.3</td>
<td>6</td>
<td></td>
<td></td>
<td>-118 @ 100 kHz</td>
</tr>
</tbody>
</table>

1. operated at 43 K
2. 25 GHz version
3. 16 GHz version
4. 28 GHz version
5. 19 GHz version
Table 4.2: Experimental results, paper [D].

<table>
<thead>
<tr>
<th>Centre frequency, $f_0$</th>
<th>16.5</th>
<th>19.6</th>
<th>GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Phase-noise, $L(100 \text{ kHz})^*$</td>
<td>$-95$</td>
<td>$-102$</td>
<td>dBc/Hz</td>
</tr>
<tr>
<td>Frequency tunability, $T_f$</td>
<td>8.9</td>
<td>3.3</td>
<td>%</td>
</tr>
</tbody>
</table>

$^*$ at centre frequency

VCOs demonstrated phase-noise of $-95$ and $-102 \text{ dBc/Hz}$ at 100 kHz offset, at their respective centre frequencies. Design and further experimental results are presented in section 4.2.1. Considering the combination of high operating frequency and good performance, paper [D] is an important work in showing the potential of ferroelectric varactors for use in microwave oscillators. A later publication [69] presents 25 GHz and 28 GHz low-noise oscillators. The ferroelectric varactors are again integrated on a carrier substrate of HRS on which CMOS modules of cross-coupled cores are flip-chip mounted. The measured phase-noise is reported to $-89 \text{ dBc/Hz}$ and $-80 \text{ dBc/Hz}$ at 100 kHz offset at respective centre frequencies of 24.8 GHz and 28 GHz. In yet another, recent publication [70] combined VCO/radiator modules are designed based on Gunn diodes and ferroelectric varactors. The measured phase-noise is reported to $-95 \text{ dBc/Hz}$ at 100 kHz offset for an operating frequency of 17 GHz.

4.2.1 Experimental results

The VCOs in paper [D] are balanced Colpitt oscillators, Fig. 4.3a and 4.3b show the schematic and layout. All passive elements are formed on the HRS substrate which also carries a pair of flip-chip mounted SiGe heterojunction bipolar transistors (HBTs). The balanced signal is combined into a single-ended microprobe output by a distributed balun. Along the rightmost part of the circuit, microstrip stubs connect to a bias network including contact pads for bias probes. The miniature ferroelectric varactors are located between the transistors. The oscillator module is placed in a brass test fixture providing grounding to the microstrip design. A photograph of an oscillator module placed in a test fixture is shown in Fig. 4.3c, and a close-up photograph of a varactor pair is shown in Fig. 4.3d.

Key results are summarised in Table 4.2. The measured frequency tuning range and phase-noise of the 16 GHz VCO is shown in Fig. 4.4. Although the work rather successfully demonstrates the use of ferroelectric varactors in microwave oscillators, the design suffers from several issues. First of all, the frequency tunability is quite low since the varactor inclusion factor is small. This is mainly due to the choice of topology; the capacitive part of the resonator tank also includes the fixed collector-emitter capacitance of the transistor which reduces the
Fig. 4.3: VCO based on ferroelectric varactors, paper [D], (a) schematic, (b) microstrip layout, (c) module (4.7 × 2.2 mm²) in test fixture, and (d) a pair of varactors.
tunability. For the 19 GHz version the varactor inclusion factor is even lower as seen in the smaller tunability and lower phase-noise. Also, only about half of the available tuning range is be used. Limitations of the experimental process used (only two metal layers, and no via holes nor air bridges available) resulted in a less-than-ideal topology and layout.

Subsequent work of Aspemyr et al. [69] represents a development of the approach taken in paper [D], in using a carrier substrate of HRS with integrated ferroelectric varactors. Better oscillator topology and layout was enabled by using CMOS modules of a cross-coupled transistor pairs, instead of discrete transistor chips. This lead to improved tunability, etc. Furthermore, the inductive elements on the carrier substrate are better optimised (higher quality factor). All in all, it further demonstrates the potential of the integration approach of using carriers including low loss passive elements such as ferroelectric varactors. As
shown in [69], the performance is very competitive in terms of tunability, phase-noise, and FOM when compared to published CMOS VCOs based on traditional varactor technologies.

4.2.2 Conclusions

Ferroelectric varactors are flexible components of good performance, as discussed in section 2.5. In terms of device quality factor and tunability, it was concluded that ferroelectric varactors are comparable to good integrated semiconductor varactors. However, the potential of ferroelectric varactors for use in microwave oscillators is not only a question about device performance, but also a matter of technology integration. Good performance is not enough to make a component useful, there must also be viable means of integration.

Alternatives for implementation of compact microwave oscillators and larger subsystems based on ferroelectric varactors are

1. full integration, where layers of ferroelectric material are included in a semiconductor process,

2. hybrid integration, where low-loss substrates with passive elements only (including ferroelectric varactors) are used as carriers for active ICs, and

3. chip integration, using discrete ferroelectric chip varactors mounted on circuit boards.

The first alternative is a cost-effective solution for large-volume applications. However, in such active processes varactors can be formed in transistor layers (junctions, MOS structures). Although it may be possible to integrate layers of ferroelectric film in say CMOS processing\(^1\), it is doubtful whether the potential advantage in terms somewhat better varactor performance (again, referring to section 2.5) is sufficient to motivate the extra cost of having a more complex process of lower yield.

The second alternative seem much more viable for ferroelectric varactors. To use a low-loss substrate (dielectric or HRS) dedicated for passive elements is beneficial in several ways. The loss of passive elements on dielectric substrates is lower than the loss of analogue elements on a semiconductor substrate, although the difference naturally varies with substrate materials and type of components. Also, there is no need to spend expensive chip area on large-sized passive elements in an advanced sophisticated semiconductor process, when a simple and cheap passive process technology will do. For many applications this is probably a more cost-effective way of making circuits and systems,

\(^1\)Critical points seem to be process temperatures and contamination issues. However, the subject of process integration is beyond the scope of this thesis.
despite extra assembly cost. Furthermore, in this simpler type of processing it is much less expensive to fabricate ferroelectric varactors than semiconductor varactors requiring high-level processing. The performance improvement offered by ferroelectric varactors, although small compared to semiconductor varactors, would come at a low affordable cost. This integration approach is demonstrated in paper [D], and more recently by others [69].

The last alternative of making ferroelectric chip varactors seem interesting in that, again, it requires less sophisticated processing equipment as compared to what is necessary for fabrication of semiconductor chip varactors. However, semiconductor processing is very well developed today, using established techniques to reach a high component yield also for very large wafers, so the cost per varactor chip is already very low. Also, for microwave oscillators, it is probably difficult to get as much performance from chip varactors as compared to the other two alternatives discussed above. Additional assembly is costly and reduces the quality factor and tunability. On a side note, power applications where not only quality factor but also power handling is important, could be an interesting niche for ferroelectric chip varactors tailored for large voltages.

4.3 Oscillators based on TFBARs

TFBARs are compact resonators of very high quality factors, as concluded in section 3.7. As such, these devices are naturally of interest for microwave oscillators. Table 4.3 summarises publications of oscillators based on various types of TFBARs. The majority is fixed-frequency designs, but some of the published oscillators are tuned by some means. Frequency tuning is achieved by using external tuning elements such as semiconductor varactors [75, 93, 96] or switched capacitor banks [95]. In these designs, the TFBARs are operated in their inductive regimes (between the series and parallel resonance points). To benefit from the high quality factor of the TFBARs, the tuning elements (being of lower quality factor) must be lightly coupled to the oscillator in order not to degrade the total quality factor, and the tunability is consequently limited. In [77], the TFBAR is thermally driven by a microheater, reaching a frequency tuning of about 0.3%. In [78] voltage is applied directly across a ZnO-TFBAR, thereby reaching a modest tuning of about 100 ppm through a weak electrostrictive effect. This effect is stronger in BSTO, promising a higher tunability, but oscillators based on BSTO are yet to be demonstrated.

Furthermore, the publications of Table 4.3 represent a range of integration techniques. Most of the works are based on chip resonators

\[\text{As discussed in section 3.6, } Q_m \text{ and } k_{t, eff}^2 \text{ of biased BSTO are still lower corresponding parameters of piezoelectric materials such as ZnO.}\]
<table>
<thead>
<tr>
<th>Year</th>
<th>Ref.</th>
<th>Resonator</th>
<th>Resonator integration</th>
<th>Technology</th>
<th>$f_0$ [GHz]</th>
<th>$P_{out}$ [dBm]</th>
<th>$P_{dc}$ [mW]</th>
<th>$\xi(f_{off})$ [dBc/Hz]</th>
<th>FOM [dBc/Hz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1984</td>
<td>[72]</td>
<td>ZnO, mm</td>
<td>mounted on pcb</td>
<td></td>
<td>0.259</td>
<td>-24</td>
<td>-66 @ 100 kHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td>2001</td>
<td>[73]</td>
<td>ZnO, mm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>0.262</td>
<td>-22</td>
<td>-77 @ 100 kHz</td>
<td></td>
<td></td>
</tr>
<tr>
<td>2002</td>
<td>[74]</td>
<td>AlN, mm</td>
<td>mounted on pcb</td>
<td></td>
<td>2.2</td>
<td>-108 @ 100 kHz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2003</td>
<td>[75]</td>
<td>AlN, mm</td>
<td>mounted on pcb</td>
<td></td>
<td>1.951</td>
<td>10</td>
<td>175 @ 10 kHz</td>
<td>198</td>
<td></td>
</tr>
<tr>
<td>2004</td>
<td>[76]</td>
<td>AlN, mm</td>
<td>mounted on pcb</td>
<td></td>
<td>1.985</td>
<td>10</td>
<td>115 @ 10 kHz</td>
<td>197</td>
<td></td>
</tr>
<tr>
<td>2005</td>
<td>[77]</td>
<td>AlN, mm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>1.9</td>
<td>0.33</td>
<td>-120 @ 100 kHz</td>
<td>211</td>
<td></td>
</tr>
<tr>
<td>2006</td>
<td>[78]</td>
<td>AlN, mm</td>
<td>mounted on pcb</td>
<td></td>
<td>1.9</td>
<td>0.084</td>
<td>-120 @ 100 kHz</td>
<td>216</td>
<td></td>
</tr>
<tr>
<td>2007</td>
<td>[79]</td>
<td>AlN, mm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>4.9</td>
<td>-109 @ 100 kHz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2008</td>
<td>[80]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>5</td>
<td>-109.5 @ 100 kHz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2009</td>
<td>[81]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>130 nm CMOS</td>
<td>5.46</td>
<td>-8.4</td>
<td>13 @ 10 kHz</td>
<td>201</td>
</tr>
<tr>
<td>2008</td>
<td>[82]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>350 nm BiCMOS</td>
<td>5.46</td>
<td></td>
<td>-121 @ 100 kHz</td>
<td></td>
</tr>
<tr>
<td>2007</td>
<td>[83]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>350 nm BiCMOS</td>
<td>5.46</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2006</td>
<td>[84]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>2.1</td>
<td>-120 @ 100 kHz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2008</td>
<td>[85]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>2.1</td>
<td>-144 @ 1 MHz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2009</td>
<td>[86]</td>
<td>ZnO, sm</td>
<td>above-ic</td>
<td></td>
<td>350 nm CMOS</td>
<td>0.9</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>2006</td>
<td>[87]</td>
<td>ZnO, sm</td>
<td>above-ic</td>
<td></td>
<td>2.1</td>
<td>0.6</td>
<td>-144 @ 1 MHz</td>
<td>213</td>
<td></td>
</tr>
<tr>
<td>2008</td>
<td>[88]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>2.1</td>
<td>12</td>
<td>-124 @ 1 kHz</td>
<td>200</td>
<td></td>
</tr>
<tr>
<td>2009</td>
<td>[89]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>90 nm CMOS</td>
<td>6.35</td>
<td>-55</td>
<td>2.95</td>
<td>181</td>
</tr>
<tr>
<td>2008</td>
<td>[90]</td>
<td>AlN, mm</td>
<td>above-ic</td>
<td></td>
<td>180 nm CMOS</td>
<td>2.514</td>
<td>-50</td>
<td>-93</td>
<td>10 kHz</td>
</tr>
<tr>
<td>2008</td>
<td>[91]</td>
<td>AlN, mm</td>
<td>mounted on pcb</td>
<td></td>
<td>0.6, 1.45</td>
<td>22</td>
<td>-151 @ 100 kHz</td>
<td>226</td>
<td></td>
</tr>
<tr>
<td>2009</td>
<td>[92]</td>
<td>AlN, mm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>0.6</td>
<td>22</td>
<td>-151 @ 100 kHz</td>
<td>226</td>
<td></td>
</tr>
<tr>
<td>2008</td>
<td>[93]</td>
<td>AlN, mm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>130 nm CMOS</td>
<td>6</td>
<td>-136 @ 1 MHz</td>
<td>195</td>
<td></td>
</tr>
<tr>
<td>2009</td>
<td>[94]</td>
<td>AlN, mm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>65 nm CMOS</td>
<td>2.2</td>
<td>6</td>
<td>-136 @ 1 MHz</td>
<td>195</td>
</tr>
<tr>
<td>2008</td>
<td>[95]</td>
<td>AlN, sm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>65 nm CMOS</td>
<td>6</td>
<td>0.065</td>
<td>-124 @ 1 kHz</td>
<td>222</td>
</tr>
<tr>
<td>2009</td>
<td>[96]</td>
<td>AlN, sm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>90 nm CMOS</td>
<td>1.7</td>
<td>2</td>
<td>-97</td>
<td>10 kHz</td>
</tr>
<tr>
<td>2009</td>
<td>[97]</td>
<td>AlN, sm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>250 nm BiCMOS</td>
<td>2.516</td>
<td>27</td>
<td>-94</td>
<td>2 kHz</td>
</tr>
<tr>
<td>2008</td>
<td>[98]</td>
<td>AlN, mm</td>
<td>wirebonded to active ic</td>
<td></td>
<td>250 nm BiCMOS</td>
<td>2.516</td>
<td>27</td>
<td>27</td>
<td>2 kHz</td>
</tr>
<tr>
<td>2009</td>
<td>[99]</td>
<td>AlN, mm</td>
<td>integrated with carrier</td>
<td></td>
<td>21.6</td>
<td>21.6</td>
<td>-127 @ 100 kHz</td>
<td>200</td>
<td></td>
</tr>
</tbody>
</table>

1 mm = membrane mounted, sm = solidly mounted
2 with TFBAR, no tuning specified
3 dissipation only for core
4 second harmonic
5 simulated
6 AlN sputtered on Si device layer (30 µm thick)
7 single-ended design
8 balanced design

Table 4.3: Summary of oscillators based on TFBARs.
that are wirebonded or mounted on a circuit carrier or integrated circuit. Other publications [82, 83, 85] have demonstrated the potential for TFBARs post-processed in CMOS, an approach known as above-ic.

Of the appended papers, paper [E] presents yet another oscillator using a wirebonded resonator, while paper [F] presents a development in terms of a resonator integrated with a carrier substrate. Design work and experimental results of both papers are elaborated below.

### 4.3.1 Experimental results

Papers [E] and [F] present a work on fixed-frequency oscillators based AlN-TFBARs, with the main focus of investigating integration of electroacoustic devices with microwave circuitry. Early work presented in paper [E] used a wirebonded resonator chip, the most common approach used by previously published oscillator demonstrators as discussed above. A photograph of a TFBAR wirebonded to a microstrip line is shown in Fig. 4.5a. The experience from the work presented in paper [E] is that the combination of a device having a very high quality factor with hand-made wirebonds is problematic, to say the least. Wirebonds inevitably introduce extra ohmic loss and inductance which may be difficult to predict. From a practical point-of-view it is hard to design good circuits using wirebonded TFBARs. Industrial standard wirebonding is more reliable and predictable, but bonding of individual resonators is still problematic.

After the initial experiments, effort was spent on integrating TFBARs with microwave transmission lines, as shown in Fig. 4.5b. The goal of the work presented in paper [F] was to develop a carrier substrate supporting not only TFBAR devices but also low-loss distributed passive elements and pads for flip-chip mounting of transistor chips. A schematic cross-section of the resonator is shown in Fig. 4.6a. The carrier substrate is oxidised p-type HRS with a specified resistivity of 15–30 kΩ·cm. The first step of processing is fabrication of the Bragg re-
The resonator, consisting of a thin piezoelectric film of AlN combined with bottom and top electrodes of Al, is then defined on top of the reflector. Both the piezoelectric layer and the reflector are unpatterned. On top of this template, a stack of four metals is deposited. All low-loss microwave passive elements, TFBAR excluded, and pads for transistor mounting are defined in this metal stack where a thin layer of Ti provides adhesion for a thick layer of Cu. The thick Cu reduces the metal losses in the passive elements, since the skin depth is comparatively large at 2 GHz ($\delta_s \approx 1500$ nm). Next, layers of Ni (solder stop) and Au are deposited to facilitate flip-chip mounting of transistors. Fig. 4.6b depicts a schematic top-view of the TFBAR, including the coplanar waveguide connecting from the right. The top-electrode employs an apodised geometry to suppress spurious lateral acoustic resonances.

As described in section 3.3, page 25, the TFBARs were characterised by a scalable MBVD model developed from measurements of specially designed and fabricated test structures. The model which also considered the wafer-variation in resonator characteristics was then used when
the oscillators were designed and subsequently positioned on the wafer mask. In order to reduce the fabrication complexity a simple single-transistor topology was chosen, Fig. 4.7a. A oneport TFBAR is located on the emitter terminal of the transistor; the series resonance of the device defines the oscillation frequency. An inductive stub connects to the base terminal, the loop-gain of the oscillator can be changed by adjusting this stub. Finally, the collector is loaded by the 50 Ω output load through a distributed network. Stubs, ac-shorted and close to quarter-wavelength, serve as simple bias lines. The layout is shown in Fig. 4.7b and a photograph of a fabricated and assembled oscillator module is shown in Fig. 4.7c. The lower half of the layout is dominated by pads for dc probes and large-sized ac-grounding capacitors. Because of the distributed elements of the circuit, in combination with a relatively low operating frequency, the dimensions of the circuit are fairly large. Discrete inductor coils were avoided to simplify the fabrication. The dimensions of the TFBAR device are, however, very small, as shown in Fig. 4.7d. The measured oscillator spectrum and phase-noise are shown in Fig. 4.8. The phase-noise was measured to $-125 \text{ dBc/Hz}$ at 100 kHz offset.

4.3.2 Conclusions

As discussed in section 3.7, TFBARs are highly competitive components which already have seen some commercial success. However, the main application so far has been compact duplexer filters which are extremely demanding in terms of insertion loss, power-handling, size and cost. It is an application where the impressive performance of AlN-TFBAR really pays off. This is a general requirement; there must really be a need for quality factors in the order of 1000 or higher, otherwise older simpler and probably cheaper resonator technologies will prevail. TFBARs have not yet been extensively used in oscillators, but applications such as low-power frequency references [99], and innovative radio architectures [100, 101] have been suggested. Compact reference oscillators would certainly benefit from the high quality factors and small sizes TFBARs can offer.

As discussed for ferroelectric varactors in section 4.2.2, there must be viable integration approaches suitable for the considered device technology. Repeating the discussion of ferroelectric varactors, three variations can be considered,

1. full integration, where TFBAR technology is included in a semiconductor process,

2. hybrid integration, where low-loss substrates with passive elements only (including TFBARs) are used as a carrier for active ICs, and
Fig. 4.7: Oscillator based on a monolithically integrated TFBAR device, paper [F], (a) schematic, (b) coplanar layout, (c) a diced module (d) close-up of resonator and flip-chip mounted transistor.
Fig. 4.8: Measured performance of 2 GHz oscillator, paper [F], (a) spectrum, and (b) phase-noise versus offset frequency.

3. chip integration, where discrete TFBARs are mounted on a circuit board.

One example of what can be considered the first approach is the above-ic technology previously mentioned [82, 83, 85]. This approach has, in a way, been proved for AlN which is a material that is easy to process and should be compatible with standard CMOS processing for instance. Although TFBARs are rather small, these devices with reflectors or air cavities still consume a fair amount of chip area (compared to active devices), so it may not represent a cost-effective solution, however. Furthermore, the above-ic process must have a very high yield to keep the cost down.

Paper [F] represents the second approach, while most of the other oscillators presented in Table 4.3 are examples of the last approach. In the case of duplexer filters, TFBAR duplexer chips are successfully sold and integrated on circuit boards in mobile phone applications. Chip
TFBARs could probably be used also for oscillator circuits, because the
device technology is significantly better than what can be accomplished
with integrated resonator tanks, so despite the extra cost it is still an
affordable solution. However, drawing from the experience of the work
presented in paper [E] (wirebonded resonator), and later in paper [F]
(resonator integrated on carrier), it more difficult for a circuit designer
to work with discrete chip components of devices with extremely high
quality factors. From this perspective, the approach presented in paper
[F] seems rather attractive.

All of the above discussion applies to traditional fixed-frequency TFBARs
based primarily on AlN. Tunable TFBARs based on BSTO are
in a different position. The technology is still in its infancy, and state-
of-the-art tunable TFBARs are as of 2009 not good enough to be used
by oscillators, as discussed in section 3.7. The technology will likely
find its first commercial niche in tunable, or switchable filter application
where the device is not necessarily resonant over the full tuning
range. Further development and improvement of tunability and quality
factor could however make the technology interesting also for low-cost
integrated VCOs, where it would be beneficial to replace large-size LC
tanks.
Chapter 5

Summary and future work

The work presented in this thesis investigates and develops thin film devices based on piezoelectric and ferroelectric materials, primarily for the application of microwave oscillators.

Ferroelectric varactors based on paraelectric phase BSTO are investigated and developed. VCOs operating at 16 GHz and 19 GHz are designed and measured to demonstrate the technology in microwave circuits. Furthermore, TFBARs based on AlN are investigated and developed. Fixed-frequency oscillators at 2 GHz using wirebonded resonators, and later carrier-integrated resonators, are designed and measured to demonstrate the technology. Lastly, tunable TFBARs based on bias induced piezoelectricity in paraelectric phase BSTO are investigated and developed. This is a novel technology, and the development presented in this thesis is focused on modelling and model extraction.

Common for all thin film devices discussed in the thesis is the need for viable integration solutions, i.e. practical ways of implementing and using these devices in compact microwave circuits such as oscillators. When it comes to using emerging technologies, hybrid integration solutions are generally preferred since it is too expensive to integrate new materials with established semiconductor processes. This is also the case for the technologies studied in this thesis, and the experimental oscillators presented here are all implemented using hybrid techniques. By developing these device technologies and their associated integration concepts for application in microwave circuits, this work represents an attempt to bring these technologies closer to commercial use.

5.1 Future work

Ferroelectric varactors of practical layouts have been shown to be largely limited by ohmic loss. Future research should include development of ferroelectric materials with lower permittivity (at maintained tunability), allowing wider interconnects and bigger electrodes to reduce the
contribution of ohmic loss. This is necessary to make this technology commercially competitive. Target applications should be integrated microwave systems operating above 10 GHz.

Fixed-frequency TFBAR technology based on AlN has reached some commercial success, and AlN-TFBARs are today used in duplexer filters for mobile telephones. However, although state-of-the-art TFBARs demonstrate $Q \times f$ merits comparable to quartz resonators, the technology has not yet been commercially used for microwave oscillators. TFBARs are potentially interesting for reference oscillators used by PLL synthesisers, or for stand-alone oscillators for low-power applications. Successful commercialisation relies on finding appropriate applications, and academically the AlN-TFBAR technology seems mature and relatively established by now. Instead research effort should be focused toward tunable TFBARs based on the electrostrictive behaviour of paraelectric phase BSTO. It represents a truly emerging technology that will require significant scientific development in the future, including development of materials and device geometries. Better understanding of the electrostrictive behaviour is also necessary in order to improve ferroelectric varactors which currently suffer from increased loss due to this phenomenon. For this purpose, methods to design layer stacks preventing spurious resonances should be further developed.

5.2 Summary of appended papers

Paper A, Layout optimisation of small-size ferroelectric parallel-plate varactors

Parallel-plate varactors of various layouts are measured and modelled in order to optimise the device quality factors. Layouts insensitive to electrode misalignment are considered, and geometry dependencies in the quality factors are investigated. A simple optimisation procedure is also proposed. The varactors were processed and measured at Chalmers.

My contributions: layout design, measurements, theory development with co-authors, writing the paper.

Paper B, Parameter extraction for tunable TFBARs based on $\text{Ba}_x\text{Sr}_{1-x}\text{TiO}_3$

A method for robust extraction of MBVD models for tunable TFBARs is suggested. The method is demonstrated for BSTO-TFBARs, and the resulting bias dependent parameters are presented. The resonators were processed and measured at Chalmers.

My contributions: measurements, theory development with co-authors, writing the paper.
Paper C, Impact of the ferroelectric film surface roughness on the performance of tunable TFBARs

Impact of film roughness on performance of TFBARs is discussed, and analysed through simulations. It is shown that moderate variation in thickness of columnar films has a significant effect on the resonance quality factors.

My contributions: theory development with co-authors, simulations.

Paper D, A low-noise K-band VCO based on room-temperature ferroelectric varactors

Voltage-controlled oscillators based on ferroelectric varactors are designed, fabricated, and measured in order to demonstrate the device technology in microwave circuits. This work was done in collaboration with Ericsson Research, Mölndal, Sweden. The circuits were processed at Chalmers, and measured at both Chalmers and Ericsson.

My contributions: circuit design, measurements, writing the paper.

Paper E, A 2 GHz oscillator based on a solidly mounted thin film bulk acoustic wave resonator

Fixed-frequency oscillators based on wirebonded AlN-TFBARs are designed, fabricated, and measured. This work was done in collaboration with the Uppsala University, Sweden, where the devices and circuits were developed and processed. The circuits were assembled and measured at Chalmers.

My contributions: circuit design, measurements, writing the paper.

Paper F, Oscillators based on monolithically integrated AlN TFBARs

AlN-TFBARs are integrated on a substrate of high-resistivity silicon with other passive elements. Fixed-frequency oscillators are designed, fabricated, and measured to demonstrate the integration technique. This work was done in collaboration with Uppsala University, where the devices and circuits were developed and processed. The circuits were assembled and measured at Chalmers.

My contributions: circuit design, development of integration technique with co-authors, measurements, writing the paper.
Acknowledgements

During my years as a PhD student at Chalmers, I have met a number of persons who in one way or another have helped me in the work toward this thesis.

First of all, I would like to express my gratitude to my supervisor, Prof. Spartak Gevorgian, for giving me the opportunity to work in his group, and for his enthusiastic support during these years. Always inspiring and encouraging, always pushing his students forward.

This thesis is the result of team work. I would like to thank all former and current members of my research group at Chalmers, the colleagues at the Uppsala University and Ericsson Research, Mölndal, for great collaborations.

Last but not least, I would like to thank my co-workers, friends, and family for supporting me during the course of this work.

* 

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Bibliography


Paper A

Layout optimisation of small-size ferroelectric parallel-plate varactors

M. Norling, D. Kuylenstierna, A. Vorobiev, and S. Gevorgian

Paper B

Parameter extraction for tunable TFBARs based on 
$\text{Ba}_x\text{Sr}_{1-x}\text{TiO}_3$

M. Norling, J. Berge, and S. Gevorgian

Paper C

Impact of the ferroelectric film surface roughness on the performance of tunable TFBARs

M. Norling, J. Berge, A. Vorobiev, and S. Gevorgian

Paper D

A low-noise K-band VCO based on room-temperature ferroelectric varactors

M. Norling, A. Vorobiev, H. Jacobsson, and S. Gevorgian

A 2 GHz oscillator based on a solidly mounted thin film bulk acoustic wave resonator

M. Norling, J. Enlund, S. Gevorgian, and I. Katardjiev

in *IEEE MTT-S Int. Microwave Symp. Dig.*, 2006, pp. 1813–1816
Paper F

Oscillators based on monolithically integrated AlN TF-BARs

M. Norling, J. Enlund, I. Katardjiev, and S. Gevorgian